Novel RF MEMS Devices Enabled by Three-Dimensional Micromachining

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Doctoral Thesis
Stockholm, Sweden 2014
Front cover pictures:
The front cover shows two SEM images of devices fabricated in this thesis. The left SEM image is of the prototype 3-D micromachined capacitor. The right SEM image is of the prototype 3-D micromachined coupled line directional coupler.
Abstract

This thesis presents novel radio frequency microelectromechanical (RF MEMS) circuits based on the three-dimensional (3-D) micromachined coplanar transmission lines whose geometry is re-configured by integrated microelectromechanical actuators. Two types of novel RF MEMS devices are proposed. The first is a concept of MEMS capacitors tuneable in multiple discrete and well-defined steps, implemented by in-plane moving of the ground side-walls of a 3-D micromachined coplanar waveguide transmission line. The MEMS actuators are completely embedded in the ground layer of the transmission line, and fabricated using a single-mask silicon-on-insulator (SOI) RF MEMS fabrication process. The resulting device achieves low insertion loss, a very high quality factor, high reliability, high linearity and high self actuation robustness. The second type introduces two novel concepts of area efficient, ultra-wideband, MEMS-reconfigurable coupled line directional couplers, whose coupling is tuned by mechanically changing the geometry of 3-D micromachined coupled transmission lines, utilizing integrated MEMS electrostatic actuators. The coupling is achieved by tuning both the ground and the signal line coupling, obtaining a large tuneable-coupling ratio while maintaining an excellent impedance match, along with high isolation and a very high directivity over a very large bandwidth.

This thesis also presents for the first time on RF nonlinearity analysis of complex multi-device RF MEMS circuits. Closed-form analytical formulas for the IIP3 of MEMS multi-device circuit concepts are derived. A nonlinearity analysis, based on these formulas and on measured device parameters, is performed for different circuit concepts and compared to the simulation results of multi-device nonlinear electromechanical circuit models. The degradation of the overall circuit nonlinearity with increasing number of device stages is investigated. Design rules are presented so that the mechanical parameters and thus the IIP3 of the individual device stages can be optimized to achieve a highest overall IIP3 for the whole circuit.

The thesis further investigates un-patterned ferromagnetic NiFe/AlN multilayer composites used as advanced magnetic core materials for on-chip inductances. The approach used is to increase the thickness of the ferromagnetic material without increasing its conductivity, by using a multilayer NiFe and AlN sandwich structure. This suppresses the induced currents very effectively and at the same time increases the ferromagnetic resonance, which is by a factor of 7.1 higher than for homogeneous NiFe layers of same thickness. The so far highest permeability values above 1 GHz for on-chip integrated un-patterned NiFe layers were achieved.

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To my family
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The presented thesis is based on the following international reviewed journal papers:


2. “High-Directivity MEMS-Tunable Directional Couplers for 10-18-GHz Broadband Applications”,

3. “Analysis of Linearity Deterioration in Multi-Device RF MEMS Circuits”,

4. “MEMS Reconfigurable Millimeter-Wave Surface for V-Band Rectangular-Waveguide Switch”,

5. “Microwave MEMS Devices Designed for Process Robustness and Operational Reliability”,

6. “High-Aspect-Ratio Through Silicon Vias (TSVs) for High-Frequency Application Fabricated by Magnetic Assembly of Gold-Coated Nickel Wires”,
7. “Permeability Enhancement by Multilayer Ferromagnetic Composites for Magnetic-Core On-Chip Inductors”,

The contribution of Umer Shah to the different publications:

1. All design, all fabrication, major part of the characterization, and major part of the writing
2. All design, all fabrication, major part of the characterization, and major part of the writing
3. All design, all fabrication, all characterization, and major part of the writing
4. Part of the fabrication, and part of the characterization
5. Part of the design, part of the fabrication, and part of the characterization
6. All RF characterization, and part of the writing
7. All design, major part of the characterization, and major part of the writing

The work has also been presented at the following international reviewed conferences:

1. (*) “RF MEMS Tuneable Capacitors Based on Moveable Sidewalls in 3D Micromachined Coplanar Transmission Lines”,

2. “Multi-Position Large Tuning-Range Digitally Tuneable Capacitors Embedded in 3D Micromachined Transmission Lines”,

3. “Antenne à ondes de fuite à balayage angulaire à fréquence fixe à 77GHz”,
4. “Basic Concepts of Moving-Sidewall Tuneable Capacitors for RF MEMS Reconfigurable Filters”,

5. “Compact MEMS Reconfigurable Ultra-Wideband 10-18 GHz Directional Couplers”,

6. “TUMESA - MEMS Tuneable Metamaterials for Smart Wireless Applications”,

7. “MEMS 30\mu m-thick W-band Waveguide Switch”,

8. “Millimeter-Wave SPST Waveguide Switch Based on Reconfigurable MEMS Surface”,

9. “Analysis of Linearity Degradation in Multi-Stage RF MEMS Circuits”,

10. “Characterization of High-Q Laterally Moving RF MEMS Tuneable Capacitor”,
The work has also been presented at the following international workshops:

1. “Microwave MEMS Activities at KTH-Royal Institute of Technology”,


4. “RF MEMS Reconfigurable Filters Based on Moveable Sidewalls of a 3D Micromachined Transmission Line”,

5. “Compact MEMS Reconfigurable Ultra-Wideband 10-18 GHz Directional Couplers”,

6. “Basic Concept of Tuneable MEMS Directional Couplers for Ultra-Wideband Applications”,

7. “Nonlinearity Determination and Linearity Degradation in RF MEMS Multi-Device Circuits”,

8. “Monocrystalline-Silicon Microwave MEMS”,

9. “RF to Millimeter-Wave MEMS at KTH”,
The work has also been published in book chapters:

1. “RF MEMS for Automotive and Radar Applications”,

Parts of this work has been awarded the following international prizes:

- (*) Best Student Paper Award (APMC 2010 Student Prize).
  Asia Pacific Microwave Conference 2010, Yokohama, Japan.

- IEEE MTT-S Graduate Fellowship.
  IEEE Microwave Theory and Techniques Society.
Chapter 1

Introduction

1.1 Microelectromechanical systems (MEMS)

Microelectromechanical systems (MEMS) constitute micro devices with the versatility to integrate both electronic and micromechanical components. They consist of sensors and actuators. Sensors can sense force, temperature, motion or any other change in the surroundings which is then converted to a readout signal for interpretation. Actuators stimulate changes in the surroundings by converting the applied signal to force, motion and heat. Examples of MEMS sensors include microphones [1,2], accelerometers [3,4], gyroscopes [5,6], pressure sensors [7,8], and examples of MEMS actuators include switches [9,10], micromirror arrays [11,12], inkjet printheads. These MEMS devices are fabricated using micro fabrication techniques developed and used in the integrated circuit (IC) industry [13]. There has been a substantial increase in research interest in MEMS technology in recent years due to the fact that all MEMS components and control functions can be fabricated together on a micro-scale in a batch process leading to fast, portable and miniaturized solutions with low power consumption and low cost. This has resulted in the commercialization of some MEMS devices such as accelerometers, gyroscopes, pressure sensors and microphones, micromirror arrays for projection and ink jet printheads etc. This in turn has sparked a resurgence of research interest to further the MEMS application areas to other fields such as microfluidics, bio MEMS, optical MEMS and RF MEMS.

1.2 RF/Microwave MEMS

The success of wireless and communication technologies is evident in the success of cellular telephony, cellular data transfer, wireless local area networks, wireless sensor networks, and wireless computer interface etc. The evolution in wireless standards is governed by the growth in consumer demands and the expectations from wireless appliances are getting higher. The steep requirements place stringent
Chapter 1. Introduction

specifications on the conventional RF technology. MEMS is a powerful technology enabling devices to overcome the limitations of conventional RF technology because of their ability of near ideal signal handling behaviour, low power consumption, low loss and large bandwidth [14]. MEMS technology used for high frequency applications is regarded as radio frequency (RF) MEMS. RF MEMS development started in the early 90s with the Hughes Research Labs MEMS microwave switch [15] which showed better performance than GaAs switches. This promising result attracted interest of many companies followed by universities and research institutes which marked the beginning of research in RF MEMS.

RF MEMS provides the opportunity for miniaturization and cost reduction by offering integrated solutions that can be batch fabricated and has the potential to replace the conventional off-chip discrete passive components whose performance is yet to be matched by integrated solutions. Typical RF MEMS devices include RF MEMS switches [16–20], tuneable capacitors [21–25], micromachined inductors [26–30] and resonators [31–35]. These basic devices are used to build integrated complex multi-device RF MEMS circuits such as phase shifters [36–38], tuneable filters [39–41], impedance matching circuits [42–44], and reconfigurable antennas [45–47]. Research in RF MEMS is maturing towards providing batch fabricated packaged integrated solutions with the future of RF MEMS research concentrating more on reliability, packaging and system integration.

1.3 Monocrystalline silicon as structural core material

Operational reliability is a key requirement of any MEMS device. Conventional RF MEMS devices which use thin metallic bridges are susceptible to all kinds of structural reliability issues. In contrast monocrystalline silicon is a very robust MEMS structural material. Monocrystalline silicon is available as a high purity low cost material in different shapes and sizes and crystal orientations. It can be doped at different levels and polished according to requirement. Monocrystalline silicon is the preferred material for MEMS fabrication as it is the base material for IC fabrication technology and has excellent electrical and mechanical characteristics. It has a higher mechanical yield strength than steel and a high elasticity which is comparable to steel with no mechanical hysteresis making it a reliable structural material for MEMS devices.

For RF applications, resistivity and dielectric loss are important substrate parameters that need to be considered. High resistivity silicon or silicon-on-insulator (SOI) substrate has to be used to reduce the overall loss of the RF circuit. Quartz, glass and sapphire substrates could also be used as they provide high resistivity and low dielectric loss but these materials have the disadvantage of having lower thermal conductivities resulting in heating issues and are harder to bulk micromachine [48]. These materials are also difficult to heterogeneously integrate with silicon because of the difference in the coefficient of thermal expansion.
1.4 Three-Dimensional (3-D) micromachined transmission lines

The concept of the 3-D micromachined transmission lines utilizing monocrystalline silicon as core material implemented in this thesis is shown in Fig. 1.1. Fig. 1.1a shows a conventional 2-D coplanar waveguide [49] where the substrate is penetrated by a major part of the electric field lines causing dielectric losses and the current crowding near the gaps at the edges of the conductor leads to increased ohmic losses [50]. In contrast to that, for the 3-D micromachined coplanar waveguide of Type A [51], as shown in Fig. 1.1b, the major part of the electric field lines is above the substrate which decreases dielectric losses and radiation losses into the substrate, and also the metallic losses are decreased since the currents, confined by the skin-depth to just the edges of the conductors, have a much larger volume available due to the extension in the vertical direction. A further variant is the 3-D micromachined coplanar waveguide Type B, shown in Fig. 1.1c, which has metallization only on top of the 3-D topography. Both types are investigated in this thesis. Note that the geometry of the slot widths of the two types is different for achieving the same characteristic impedance, as the geometry of Type A results in higher capacitance.
1.5 Thesis Overview

This thesis presents novel research in the field of RF MEMS. It focuses on the concept and design of three-dimensional micromachined coplanar waveguide transmission line embedded tuneable capacitors and actuators for filter and directional coupler applications, respectively. Fabricated devices are presented and characterized in detail for tuneable capacitors, filters and directional couplers. The thesis further investigates intermodulation distortion in multi-device RF MEMS circuits and the usage of multilayer ferromagnetic composite for permeability enhancement of on-chip integrated inductors.

The thesis is divided into six chapters. The structure is as follows:

Chapter 2 introduces the 3-D transmission line integrated tuneable capacitors. The design strategy, fabrication and implementation in a filter is discussed. Extensive characterization results of the prototype devices is reported consisting of RF, actuator, power handling, linearity and reliability characterization.

Chapter 3 investigates ultra-wideband directional couplers implemented using 3-D transmission line integrated actuators. The novel concept and design used to achieve tunability is discussed. Measurement results including directivity analysis and life-cycle evaluation is presented.

Chapter 4 discusses the linearity analysis of multi-device RF MEMS circuits. Closed form analytical formulas are derived and verified. Nonlinear electromechanical model is applied to real devices and IIP3 is predicted accurately. Finally, design rules are proposed to increase the overall circuit linearity.

Chapter 5 shows the use of ferromagnetic multilayer composites as magnetic core material for on-chip inductances. The proposed composite structure reduces RF induced currents and thus pushes the permeability cut-off beyond 1 GHz.

Chapter 6 summarizes the conclusion of this thesis.
Chapter 2

Tuneable Capacitors integrated in 3-D transmission line

This chapter presents the concept of RF MEMS tuneable capacitors based on the lateral displacement of the sidewalls of a 3-D micromachined coplanar waveguide transmission line. The tuning of a single device is achieved in multiple discrete and well-defined tuning steps by integrated multi-stage MEMS electrostatic actuators that are embedded inside the ground layer of the transmission line. This chapter further analyses and compares three fundamental concepts of moveable-sidewall tuneable capacitors in detail. A tuneable filter created in a 3-D micromachined coplanar transmission line has also been implemented using the moveable-sidewall tuneable capacitor concept [52].

2.1 Introduction

Tuneable capacitors are an important part of frequency-agile microwave systems. Despite recent improvements in the quality factor, the self-resonance frequency [53, 54], linearity and the tuning ratio of solid-state varactor diodes [55], RF MEMS tuneable capacitors have the potential to replace solid-state varactor diodes in applications such as: phase shifters [56], filters [57], voltage controlled oscillators [14] and impedance matching networks [44]. This is because of their ability of near ideal signal handling behaviour, low power consumption, low loss and large bandwidth [14].

Recently, RF MEMS-reconfigurable capacitor banks have achieved some commercial success [58]. Sub-microsecond switching times have also been achieved by utilizing miniature capacitive beams [59, 60]. It has also been shown that using a metal-air-metal (MAM) capacitor configuration, results in low losses [61] and reliability improvement [62, 63].
### Table 2.1: Performance Comparison of Tuneable Capacitors

<table>
<thead>
<tr>
<th>Device Technology</th>
<th>Capacitance ratio</th>
<th>Quality factor (freq.)</th>
<th>Switching time</th>
</tr>
</thead>
<tbody>
<tr>
<td>MEMS [64]</td>
<td>8.4 : 1</td>
<td>35 (2 GHz)</td>
<td>6 ms</td>
</tr>
<tr>
<td>MEMS [65]</td>
<td>1.9 : 1</td>
<td>100 (34 GHz)</td>
<td>NA</td>
</tr>
<tr>
<td>MEMS [59]</td>
<td>2.3 : 1</td>
<td>50 (20 GHz)</td>
<td>400 ns</td>
</tr>
<tr>
<td>MEMS [62]</td>
<td>2 : 1</td>
<td>225 (X-Ku band)</td>
<td>8 µm</td>
</tr>
<tr>
<td>CMOS-MEMS [66]</td>
<td>4.6 : 1</td>
<td>&gt; 300 (1.5 GHz)</td>
<td>NA</td>
</tr>
<tr>
<td>MEMS [61]</td>
<td>3 : 1</td>
<td>10 (10 GHz)</td>
<td>NA</td>
</tr>
<tr>
<td>MEMS [67]</td>
<td>2.8 : 1</td>
<td>8.8 (1 GHz)</td>
<td>NA</td>
</tr>
<tr>
<td>MEMS [60]</td>
<td>3 : 1</td>
<td>90 (20 GHz)</td>
<td>200 ns</td>
</tr>
<tr>
<td>MEMS [63]</td>
<td>9 : 1</td>
<td>100 (C-X band)</td>
<td>50 µm</td>
</tr>
<tr>
<td>MEMS [68]</td>
<td>20 : 1</td>
<td>85 (3.127 GHz)</td>
<td>&lt; 10 µm</td>
</tr>
<tr>
<td>CMOS-MEMS [69]</td>
<td>63 : 1</td>
<td>160 (1 GHz)</td>
<td>600 µm</td>
</tr>
<tr>
<td>CMOS [53]</td>
<td>1.6 : 1</td>
<td>&gt; 100 (24 GHz)</td>
<td>NA</td>
</tr>
<tr>
<td>GaAs [55]</td>
<td>9 : 1</td>
<td>50 (2 GHz)</td>
<td>NA</td>
</tr>
<tr>
<td>SiC [70]</td>
<td>6 : 1</td>
<td>160 (2 GHz)</td>
<td>NA</td>
</tr>
<tr>
<td>CMOS [71]</td>
<td>7.7 : 1</td>
<td>35 (1 GHz)</td>
<td>NA</td>
</tr>
<tr>
<td>MEMS (this work)</td>
<td>1.48 : 1</td>
<td>80 (40 GHz)</td>
<td>140 µm</td>
</tr>
</tbody>
</table>

#### 2.2 RF MEMS tuneable capacitors

There are mainly four different principles to implement RF MEMS tuneable capacitors in the literature:

1. Tuneable parallel-plate capacitors: They consist of a fixed electrode and a moveable electrode using integrated MEMS actuator. This can either be utilized in analog tuning mode or in a switched capacitors mode [59,62].

2. Tuneable interdigital capacitors: They consist of two comb-like electrodes of which one can be moved by a MEMS actuator. These can achieve better tuning linearity and a larger tuning range, but have a lower quality factor at high frequencies, lower self resonance frequencies and occupy a larger area [64] as compared to the parallel plate approach [65].

3. Switched capacitor banks [72,73]: They use MEMS switches to select fixed metal-insulator-metal (MIM) or metal-air-metal (MAM) capacitors from a capacitor bank.
2.3 Tuneable capacitors using laterally moving ground sidewalls

Fig. 2.1 shows the basic concept of the capacitor. The sidewalls in a section of the ground plane of a 3-D micromachined coplanar waveguide can be moved laterally by integrated MEMS actuators and are thus changing the capacitive load of the transmission line. The MEMS actuators are completely embedded in the ground layer of the micromachined transmission line thereby not adding any additional discontinuity in the slots. Tuning of the capacitors is achieved in multiple, well-defined discrete tuning steps by stacking multiple actuators. Asymmetrical operation is possible by actuating both sides alone, creating intermediate overall capacitance steps. The tuneable RF MEMS capacitor is composed of a pure metal-air-metal (MAM) geometry that avoids dielectric charging. Also, the actuators do not require any dielectric layers that are prone to dielectric charging and thus decrease reliability and actuation voltage repeatability [76]. Stoppers are used instead to avoid short circuit. A further reliability-enhancing feature of the design
Chapter 2. Tuneable Capacitors integrated in 3-D transmission line

Figure 2.2: Design I: Illustration of ground sidewall integrated tuneable capacitor with bending sidewalls: (a) 3-D sketch of the design; and (b) actuation sequence.

is the utilization of monocrystalline silicon as the core-structural material for all moving parts, which is then metallized. The signal line of the transmission line is not used for DC-biasing the actuation mechanism. For two of the three presented design concepts, the RF ground shares the same potential as the DC ground and the DC biasing occurs on electrodes outside the RF signal path. For the third design, even the DC ground is decoupled from the RF ground. All transmission line and tuning elements are fabricated using a single photolithographical step.

Table 2.2 lists the general advantages and disadvantages of the three different design implementations for the tuneable capacitor concept. Fig. 2.2a shows a 3-D sketch of the first tuneable capacitor concept Design I. The tuning is achieved by bending of the sidewalls with electrostatic actuators, by applying a DC voltage between the actuation electrode and the ground plane shown in Fig. 2.2b. This design uses a 3-D transmission line (signal line width, gap, height = 80 µm, 90 µm, 30.5 µm) where also the sidewalls are covered by metal, thereby forming a parallel plate capacitor to the ground (plate height: 30.5 µm; gap: 2 µm non-actuated, 4 µm in central part when actuated; length: 600 µm; thickness: 5 µm). For this device concept, only a low capacitance ratio can be achieved as the ground sidewall is not deflected over its whole length.
2.3. Tuneable capacitors using laterally moving ground sidewalls

Table 2.2: Overview of the Advantages and Disadvantages of the Three Design Concepts

<table>
<thead>
<tr>
<th>Advantages common to all designs</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Multi-step digital tuning.</td>
</tr>
<tr>
<td>• Accurately defined and reproducible tuning states.</td>
</tr>
<tr>
<td>• 3-D MEMS transmission line (metal covered silicon core).</td>
</tr>
<tr>
<td>• Actuator embedded in ground layer (actuation elements are invisible to wave propagation).</td>
</tr>
<tr>
<td>• All metal design (no dielectric layer between actuator and electrode).</td>
</tr>
<tr>
<td>• Metallized silicon core (high reliability, temperature compensation).</td>
</tr>
<tr>
<td>• No need for DC bias on the signal line.</td>
</tr>
<tr>
<td>• Single mask fabrication process.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Additional advantages individual to each design</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Design I</strong></td>
</tr>
<tr>
<td>• Extended tuning range by stacking actuators.</td>
</tr>
<tr>
<td>• Number of states independent of required transmission line length.</td>
</tr>
<tr>
<td>• Mechanical suspension is decoupled in its function from the RF signal path.</td>
</tr>
<tr>
<td>• DC ground not shared with RF ground, i.e. de-coupled RF and DC potentials.</td>
</tr>
<tr>
<td>• High self actuation robustness.</td>
</tr>
<tr>
<td>• High reliability.</td>
</tr>
<tr>
<td>• High linearity.</td>
</tr>
<tr>
<td>• High Q-factor.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Disadvantages individual to each design</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Design I</strong></td>
</tr>
<tr>
<td>• Low capacitance ratio due to bending and not displacement of the capacitive element.</td>
</tr>
</tbody>
</table>
Chapter 2. Tuneable Capacitors integrated in 3-D transmission line

Laterally moveable ground sidewall
Stoppers
Actuation electrodes
Springs
Stacked 2-stage actuator
1st stage 2nd stage
Handle wafer
G G S
80µm
90µm
30.5µm

(a)

Stage 1
spring
Stage 2
spring
gap = 
6 µm
Stage 1
spring
actuated
gap = 
4 µm
Stage 1 + Stage 2
spring actuated
gap = 
2 µm

(b)

Figure 2.3: Design II: Illustration of ground sidewall integrated tuneable capacitor with sidewalls displaced uniformly over their entire length, shown for a two-stage actuator achieving 5 discrete tuning steps: (a) 3-D sketch of the design; and (b) actuation sequence.

For Design II, the ground sidewall section is moved uniformly over its entire length as shown in Fig. 2.3a. This design also utilizes 3-D transmission lines (signal line width, gap, height = 80 µm, 90 µm, 30.5 µm) with metal-coverage on the sidewalls forming the capacitor (plate height: 30.5 µm; gap: 6 µm non-actuated, 2 µm when all stages are actuated; length: 600 µm; moveable sidewall width: 20 µm; spring thickness: 5 µm). The RF ground signal is routed via the mechanical springs, which results in limited RF performance due to the increased overall series resistance of the capacitor. Actuator stages can be stacked laterally resulting in larger sidewall movement, in accurately defined discrete steps for each actuation stage. The increased displacement can be achieved at medium actuation voltages as the total movement is split in smaller parts through sequential operation of the actuators, as shown in Fig. 2.3b for a two-stage design. For actuating a subsequent stage, all the previous stages have to be actuated. The device can also be actuated asymmetrically resulting in additional capacitance states. In total, a number of $2^n - 1$ states can be achieved with $n$ being the number of single-side actuation stages.

The main disadvantage of Design II is that the RF signal is routed via the mechanical springs which increases the overall insertion loss. In Design III, the mechanical springs are completely decoupled from the RF signal as shown in Fig. 2.4a. This results in a lower insertion loss and a higher Q for Design III as compared to the Designs II. The 3-D transmission line of Design III is implemented with top metallization only (signal line width, gap, suspension height = 130 µm, 130 µm, 30 µm) which results in reduced substrate losses [77,78]. The transmission lines of Designs I and II had higher losses due to their relatively thin metallization layer on the sidewalls. With top-metallization only, the overall capacitance is
2.4. Fabrication

Figure 2.4: Design III: Illustration of ground sidewall integrated tuneable capacitor with ground capacitive coupling, completely de-coupling the mechanical from the RF functional elements: (a) 3-D sketch of the design; and (b) actuation sequence.

dominated by fringing field components, which results in a lower overall capacitance and capacitance ratio as compared to Design I and II. Fig. 2.4b shows the actuation principle of a Design III tuneable capacitor with a single-stage actuator. More stages can also be integrated for this design as is done for Design II resulting in additional states of the capacitor and extended tuning range.

2.4 Fabrication

All structures are fabricated in a single-mask SOI RF MEMS process originally developed for switches [79] and outlined in Fig. 5.1. A high resistivity > 3000 Ω-cm SOI wafer is used with a device layer thickness of 30 μm, a buried oxide layer of 3 μm and a handle wafer thickness of 500 μm (Fig. 2.5a). The SOI device layer is structured using a single photolithographic step by anisotropic deep reactive ion etching (DRIE) (Fig. 2.5b). This is followed by free etching of the moving structures by wet etching of the buried oxide layer using hydrofluoric acid (Fig. 2.5c). The wafer is coated by a 0.5 μm thick layer of gold in a sputtering process using titanium tungsten as adhesion layer for the prototypes of Design I and Design II. For the prototypes of Design III, a 1 μm thick layer of gold on a titanium adhesion layer is evaporated using high-directivity e-beam evaporation, to achieve top-metallization only (Fig. 2.5d). Finally, the metal coating on the substrate and in the unwanted areas, for instance in the gap between the signal and the ground layer of the coplanar transmission line, is removed by electrochemically assisted selective etching of gold [51] in an electrically biased potassium iodide and sodium sulfite solution [80] (Fig. 2.5e). All the above wet steps are followed by a critical point drying step.
Figure 2.5: SOI RF MEMS fabrication process of the tuneable capacitors [79]: (a) SOI wafer; (b) Deep Reactive Ion Etching (DRIE) of the device layer; (c) free etching of the moving structures by HF wet-etching the buried oxide layer; (d) metallization by high directivity e-beam evaporation; and (e) etching of unwanted metal areas by electrochemically assisted selective etching of gold.

Figure 2.6: SEM pictures of fabricated prototype tuneable capacitors: (a) Design I (3 tuning states); (b) Design II implementation with three-stage actuator (7 discrete tuning states); and (c) Design III, with single-stage MEMS actuator (3 tuning states).

The SEM pictures of the three design implementations are shown in Fig. 2.6.

2.5 Characterization of fabricated tuneable capacitors

The RF measurements of the fabricated capacitors were performed using an Agilent E8361A PNA Vector Network Analyzer calibrated using GGB Industries CS-5 calibration standard and 150 µm GSG coplanar probes and SOLT calibration.
2.5. Characterization of fabricated tuneable capacitors

Table 2.3: Performance Summary of Representative Prototype Implementations of the Three Designs

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Design I</th>
<th>Design II</th>
<th>Design III</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Q$ (40 GHz)</td>
<td>3</td>
<td>$3.8^a/3.6^b$</td>
<td>88</td>
</tr>
<tr>
<td>$C_{max}$ (fF)</td>
<td>127</td>
<td>144.5$^a/134^b$</td>
<td>60</td>
</tr>
<tr>
<td>$C_{min}$ (fF)</td>
<td>104</td>
<td>58.6$^a/57^b$</td>
<td>40.53</td>
</tr>
<tr>
<td>$C_{max}/C_{min}$</td>
<td>1.22</td>
<td>2.46$^a/2.35^b$</td>
<td>1.48</td>
</tr>
<tr>
<td>Number of states</td>
<td>3</td>
<td>5$^a/7^b$</td>
<td>3</td>
</tr>
<tr>
<td>Mechanical resonance frequency (kHz)</td>
<td>-</td>
<td>-</td>
<td>5.3$^a/17.2^d$</td>
</tr>
<tr>
<td>Pull-in time (ps)</td>
<td>-</td>
<td>-</td>
<td>140</td>
</tr>
<tr>
<td>Self actuation pull-in (dBm)</td>
<td>-</td>
<td>-</td>
<td>41.5$^c/47.8^d$</td>
</tr>
<tr>
<td>IP3 (dBm)</td>
<td>-</td>
<td>-</td>
<td>$&gt;68.5$</td>
</tr>
<tr>
<td>Reliability (cycles)</td>
<td>-</td>
<td>-</td>
<td>$&gt;10^b$</td>
</tr>
<tr>
<td>Device length (µm)</td>
<td>600</td>
<td>600</td>
<td>600</td>
</tr>
<tr>
<td>Self resonance frequency (GHz)</td>
<td>-$^e$</td>
<td>-$^e$</td>
<td>46</td>
</tr>
</tbody>
</table>

*a* Device with two stage actuators  
*b* Device with three stage actuators  
$c_1 = 5.8$ N/m  
$c_2 = 27.7$ N/m  
$^d$ Not evaluated due to low $Q$ factor

The displacement measurements for actuation voltage, resonance frequency, self-actuation and response time were performed using a Veeco Wyko NT9300 white-light interferometer. Table 2.3 shows the performance summary of the three designs, discussed in detail in the following subsections. Design III has the overall best performance of all implemented design concepts and is therefore investigated in more detail.

2.5.1 RF characterization

The total insertion and return loss of the actuated and non actuated states for Design I is shown in Fig. 2.7 and is compared to the insertion loss of a transmission line of the same length without a tuneable capacitor. The loss of the transmission line alone is 1.17 dB at 20 GHz which is due to the poor sidewall metal coverage. The capacitance is extracted to be 127 fF in the non actuated state and 104 fF in the actuated state resulting in a capacitance change ratio of 1.22. The capacitance is extracted from the measured $S$ parameters by first de-embedding the probe pads and the transmission line [81]. The measured capacitance ratio is low because the capacitive element is only bending when it is actuated and not displaced over its
entire length.

The insertion and return loss of devices of Design II with two and three actuation stages is shown in Fig. 2.8. The extracted capacitances of the two-stage and the three-stage tuneable capacitor is given in Table 2.4 for all actuation states. The two-stage device can be tuned from 58.6 fF to 144.5 fF ($C_{\text{max}}/C_{\text{min}}$ of 2.46) and the three-stage device can be tuned from 57 to 134 fF ($C_{\text{max}}/C_{\text{min}}$ of 2.35). These capacitance ratios are much higher than for Design I, as the sidewalls in Design II are laterally moved over their entire length in contrast to bending for Design I and due to the larger total travelling distance of 4 µm. The gap between the laterally moved ground plane sidewall and the rigid signal line is 2 and 6 µm in the two extreme positions. For actuating of a higher order stage, all previous stages already have to be pulled in. Different actuators with varying spring constants have been designed, fabricated and evaluated with the actuation voltage of two-stage actuators ranging from 24 V (10.3 N/m spring constant) to 74 V (95 N/m) for stage 1 and 15 V (3.48 N/m) to 53 V (45.3 N/m) for stage 2. The three stage actuation voltage ranges from 23 V (22.5 N/m) to 72 V (211 N/m) for stage 1, 19 V (12.7 N/m) to 53 V (94 N/m) for stage 2 and 18 V (9.5 N/m) to 50 V (73.2 N/m) for
2.5. Characterization of fabricated tuneable capacitors

Table 2.4: Design II: Actuation States with Capacitances Extracted from S-parameter Measurements, for Two-Stage and Three-Stage Actuators.

<table>
<thead>
<tr>
<th>Actuator Actuated Stages (left) 12 21 (right)</th>
<th>Measured Capacitance (fF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Two stage 00 00</td>
<td>58.6</td>
</tr>
<tr>
<td></td>
<td>67.3</td>
</tr>
<tr>
<td></td>
<td>87.1</td>
</tr>
<tr>
<td></td>
<td>114.2</td>
</tr>
<tr>
<td></td>
<td>144.5</td>
</tr>
<tr>
<td>Three stage 000 000</td>
<td>57</td>
</tr>
<tr>
<td></td>
<td>57</td>
</tr>
<tr>
<td></td>
<td>58.6</td>
</tr>
<tr>
<td></td>
<td>61</td>
</tr>
<tr>
<td></td>
<td>69.3</td>
</tr>
<tr>
<td></td>
<td>100.4</td>
</tr>
<tr>
<td></td>
<td>134</td>
</tr>
</tbody>
</table>

Figure 2.9: Measurements of the fabricated prototype of Design III: (a) S-parameters for the unactuated and symmetrically actuated states; and (b) Q-factor: Transmission measurements of a weakly coupled transmission line resonator loaded with the tuneable capacitor.

stage 3. Higher losses are expected for Design II as compared to Design I because of the longer RF ground path length, as the RF ground signal is routed over the mechanical springs of the multi-stage actuators.

The insertion and return loss for Design III for the unactuated and double-side actuated state is shown in Fig. 2.9a. Design III prototype device have been implemented only with single-stage actuators. The capacitance can be tuned from
Chapter 2. Tuneable Capacitors integrated in 3-D transmission line

40.53 fF in the unactuated state to 60 fF in the double-side actuated state resulting in a capacitance ratio \( C_{\text{max}}/C_{\text{min}} \) of 1.48. This tuning ratio can be increased by integrating more stages to achieve more tuning steps and increase the travelling distance of the moving sidewall. The losses of the prototype of Design III are much lower than for Designs I and II since the mechanical springs are no longer used for signal routing and are completely isolated from the RF signal path.

The quality factor of the capacitor was measured by placing the capacitor in a weakly coupled transmission line resonator with a nominal resonant frequency of 40 GHz with a known (measured) unloaded \( Q_r \) of 14.88. The loaded \( Q \) \((Q_l)\) of the resonator/capacitor combination was extracted from the measured resonance behaviour of the transmission line with the capacitor, using \( Q_l = f_o/\Delta f \). The unloaded \( Q \) \((Q_u)\) of the combined structure is then determined by \( Q_u = Q_l/(1 - |S_{21}|) \) which is equal to the loaded \( Q \) \((Q_u = Q_l)\) if \( S_{21} < -20 \text{ dB} \). The unloaded capacitor \( Q \) is then calculated by [14]

\[
\frac{1}{Q_c} = \frac{1}{Q_u} - \frac{1}{Q_r}
\]

This is the recommended method for obtaining the \( Q \) of a low loss capacitor [14]. Fig. 2.9b shows the transmission measurement results of the weakly coupled resonator \((S_{21} < -20 \text{ dB})\) loaded with a capacitor in the two extreme states. For the first measurement, the capacitor gap to the transmission line is 4 \( \mu m \) and the extracted \( Q_c \) is 88 at 40 GHz, and for the second measurement, the capacitor gap to the transmission line is 2 \( \mu m \) leading to an extracted \( Q_c \) of 75.

2.5.2 MEMS actuator characterization

Fig. 2.10a shows the measured actuation and release curve and COMSOL Multiphysics simulated actuation curve for a tuneable capacitor of Design III. The measured average pull-in voltage, for 20 cycle measurements, is 30.70 V with a standard deviation of 1.08 V. For the releases voltage, the average is 21.15 V with a standard deviation of 1.71 V. Fig. 2.10b shows the measured mechanical resonance responses of two fabricated devices of Design III with different spring constants. For a spring constant of 5.8 N/m the mechanical resonance frequency was measured to be 5.3 kHz, and a spring constant of 27.7 N/m results in a mechanical resonance frequency of 17.2 kHz.

The actuator response curve is measured and plotted in Fig. 2.11 for the design with spring constant of 5.8 N/m. The characteristic bouncing behaviour of the actuator during pull-in as shown in Fig. 2.11a. Fig. 2.11b shows the release time measurements of the same actuator. The release is followed by multiple oscillation events implying that there is very low damping of the unpackaged device.
2.5. Characterization of fabricated tuneable capacitors

![Graph](image1)

**Figure 2.10:** Measurements of the fabricated prototype of Design III: (a) Measured and simulated DC-actuation hysteresis of a prototype device with a spring constant of 5.8 N/m.; and (b) mechanical resonance frequencies for two prototype devices with spring constants of 5.8 and 27.7 N/m.

![Graph](image2)

**Figure 2.11:** Pull-in and pull-out transient measurements of multiple actuation events of a tuneable capacitor of Design III (k = 5.8 N/m): (a) Pull-in; and (b) pull-out.

### 2.5.3 Linearity, power handling, and reliability characterization

The linearity characteristic was determined by measuring the two-tone third-order intermodulation intercept point (IP3) with two signal sources at a center frequency of 2.5 GHz, separated by a 12 MHz offset. Fig. 2.12 shows that the IIP3 measurements reach a common limit at 68.5 dBm for all the three states of the capacitor. This value was also found for an IIP3 measurement of a transmission line alone. Thus, it can be assumed that the measured nonlinearity is limited by the transmission line measurement setup, and not the MEMS device [82].

The self actuation robustness was determined by applying a 60 kHz low frequency power signal and the subsequent deflection was measured. Two devices of Design III were measured for increasing power levels up to 50 dBm. Fig. 2.13a shows that for the device with \( k = 5.8 \) N/m, self actuation pull-in occurs at 41.5 dBm,
Figure 2.12: Measured dual-tone fundamental and intermodulation levels versus input power for the three states of a device of Design III \( (f = 2.5 \text{ GHz}, \Delta f = 12 \text{ MHz}) \).

Figure 2.13: Measurements of the fabricated prototype of Design III: (a) Self-actuation behaviour using a high-power signal on the RF transmission line on two prototype devices; and (b) pull-in and pull-out voltages monitored over 1 billion cycles with 34 V unipolar square waveform with a 35% duty cycle and a cycle frequency of 1.6 kHz.

and for the device with \( k = 27.7 \text{ N/m} \) self actuation pull-in occurs at 47.8 dBm. Lifetime measurements were performed on the tuneable capacitor of Design III with a spring constant of 5.8 N/m. The device was cycled with an actuation voltage of 34 V using a unipolar square waveform with a 35% duty cycle and a cycle frequency of 1.6 kHz in an uncontrolled atmospheric environment. The actuation and release voltage values averaged from three actuation cycles at each measurement point are shown in Fig. 2.13b. The measurements were stopped after 8 days with more than 22 hours of accumulated pull-in time, when one billion cycles were reached without observing any failure, fatigue, or altered pull-in hysteresis. Neither the stoppers nor other actuator elements showed any signs of wear.
2.6 Reconfigurable filters based on moving ground sidewalls

The tuneable capacitor of Design III is used to design the filter shown in Fig. 2.14a. The filter is based on three inductive inverters with four tuneable capacitive loads arranged in pairs between the inverters. The tuneable capacitive loads are arranged symmetrically in each slot of the transmission line. Fig. 2.14b shows a SEM picture of the fabricated filter. Fig. 2.14b also shows bond-wires over the ground slots of inductive inverters of the filter, which drastically improve the performance, as shown in Fig. 2.15a. The filter measured before and after wire-bonding has an improved insertion loss from 8.1 to 5.0 dB. Fig. 2.15b shows the measured tuning performance of this filter, for the 5 basic tuning configurations of the capacitive loads. The center frequency can be moved by 1 GHz (5% of the center frequency of 20 GHz) in 5 discrete steps.

Figure 2.14: MEMS tuneable filter concept based on moving the sidewalls of the ground-layer of a 3-D micromachined coplanar-waveguide transmission line: (a) 3-D sketch; and (b) SEM picture.

2.7 Discussion and outlook

This chapter showed the concept of a novel RF MEMS tuneable capacitors based on moving ground sidewalls of a 3-D micromachined coplanar transmission lines with integrated MEMS actuators. The device concept enables multiple tuning steps in discrete and well-defined positions. Embodiments of different device concepts were successfully demonstrated, achieving high-Q, high reliability, high linearity, and high self actuation robustness at medium actuation voltages. The chapter also shows a concept of MEMS tuneable filter based on moving the sidewalls of 3-D micromachined transmission lines for changing the capacitive load of a line.

A weakness in the presented concept is the difficulty in getting a good metallization on the sidewalls, in particular for smaller openings including the
capacitor gap. The sidewall thickness could in principle be improved by using electroplating or electroless plating. However, plating would not work for the narrow gaps and high aspect ratios employed in these designs, which results in large plating nonuniformities.

Design III has the overall best performance of all implemented design concepts which has top metallization only. Here, the overall capacitance is dominated by fringing field components, which results in a lower overall capacitance and capacitance ratio as compared to an implementation with sidewall metallization. The design has the potential for improved capacitance ratio which is largely depending on the travelling distance of the moving sidewall. Thus tuning ratio can be increased by adding multiple stages as for Design II. For all three design concepts, the capacitance ratio could further be improved by moving the sidewall closer to signal line than the implemented 2 \( \mu \text{m} \).

Also by packaging the device in an overpressure or high viscosity gas environment, critical damping \( Q = 0.5 \) can be achieved eliminating the ringing on release. This reduces the response time of the device.

In principle, the stoppers of the actuator should limit the movement to the pull-in position, but it was noted that with high signal power levels of beyond 40 dBm, the moving sidewall can tilt which frequently created a non-reversible short-circuit to the signal line during the power level tests.

Figure 2.15: RF measurements of the filter: (a) With and without the air bridges over the ground slots; and (b) tuning performance.
Chapter 3

Reconfigurable ultra-wideband directional couplers

This chapter presents two new concepts of area-efficient, ultra-wideband, MEMS-reconfigurable coupled line directional couplers. The coupling is tuned by mechanically changing the geometry of the coupled transmission lines utilizing integrated MEMS electrostatic actuators. Concept 1 is based on changing the coupling of each signal line to the ground. Concept 2, on the other hand, is based on simultaneously varying both the ground coupling and the coupling between the two signal lines.

3.1 Introduction

A directional coupler is a passive device used for splitting, combining, sampling, or isolating signals and is one of the most often used components in microwave circuits. Directional couplers and power dividers/combiners are key RF components in modern communication systems which includes beam forming networks [83,84], power control amplifiers [85], MIMO systems, and adaptive antenna feedback mechanisms. With the recent advent of multi-standard frontends in modern telecommunication systems [86], directional couplers capable of operating with different frequency bands [87,88] and with tuneable coupling ratio [89,90] are of particular interest for future reconfigurable architectures. The ratio between the output power on the coupled port and the isolated port is called directivity and is an important measure of the quality of the directional coupler along with the insertion loss. It is difficult to realize integrated directional couplers with high directivities [91]. Typically, directivities between 15 dB and 20 dB are obtained for tuneable couplers operating between 1-10 GHz [85,92].

For tuneable directional couplers, the tuning range of the coupling ratio is usually limited to only a few decibels [93]. When tuned beyond this coupling ratio range, the input matching and directivity degrade drastically. The directional
coupler in [94] tries to overcome these problems but falls short in terms of bandwidth. Tuneable couplers consist of MMIC based active couplers [92, 95, 96] and varactor diode based passive couplers [90, 93, 94, 97, 98], and are predominantly designed for operating below 8 GHz.

Despite MEMS being known for creating very low-loss tuning mechanisms, there have been very few attempts of implementing MEMS-based tuneable couplers. A MEMS power divider concept based on two cascaded hybrid couplers was shown in [99] which occupies a very large area. Coupled line directional couplers on the other hand can be designed more area efficiently and one attempt has been reported in literature on a MEMS-switched coupled line directional coupler concept [100].

### 3.2 Concept and design of the novel tuneable directional couplers

The basic principle of a coupled line directional coupler is shown in Fig. 3.1. RF power is coupled between two unshielded transmission lines due to the interaction of the electromagnetic field when the lines are in close proximity [101]. These lines are called coupled lines and consist usually of three conductors i.e., two signal lines and a common ground layer. A substantial part of the coupling occurs via the common ground layer. These lines are assumed to operate in TEM mode and their electrical coupling characteristics can be completely determined by the effective capacitances between the three conductors, as shown in Fig. 3.1. Even and odd mode excitation of the coupled lines reveal that both the even and odd mode capacitances are strongly influenced by the capacitance to ground of each signal line. The even and odd mode capacitances can be written as

\[
C_{\text{even}} = C_{11} = C_{22} \quad (3.1)
\]

\[
C_{\text{odd}} = C_{11} + 2C_{12} = C_{22} + 2C_{12} \quad (3.2)
\]

where \(C_{11}\) and \(C_{22}\) are the capacitances of the signal lines to the ground and \(C_{12}\) is the capacitance between the two signal lines. If the geometry is symmetrical, i.e., the two signal lines have the same size and location relative to the ground layer, then \(C_{11} = C_{22}\). This results in a strong dependence of the characteristic impedances of the even and odd modes on capacitance of each line to ground [101].

\[
Z_{0,\text{even}} = \sqrt{\frac{L}{C_{\text{even}}}} = \frac{1}{v_p C_{\text{even}}} \quad (3.3)
\]

\[
Z_{0,\text{odd}} = \sqrt{\frac{L}{C_{\text{odd}}}} = \frac{1}{v_p C_{\text{odd}}} \quad (3.4)
\]

where \(v_p\) is the phase velocity of propagation on the line.

This analysis assumes that the lines are symmetric and that fringing capacitances are identical for even and odd modes. Since the overall coupling can be
3.2. Concept and design of the novel tuneable directional couplers

![Diagram of Directional Couplers]

**Figure 3.1:** Basic working principle of: (a) conventional tuneable coupled line directional couplers; (b) novel Concept 1; and (c) novel Concept 2 of the MEMS tuneable coupled line directional couplers.

represented in terms of even and odd mode impedance [101], the total coupling between the lines can be changed by either changing the coupling between the signal lines or the coupling of each signal line to the ground or both:

\[
C = \frac{Z_{0,\text{even}} - Z_{0,\text{odd}}}{Z_{0,\text{even}} + Z_{0,\text{odd}}}
\]  

(3.5)

In the above analysis, it was assumed that the even and odd modes have the same propagation velocities so that the line has the same electrical length for both modes. For a non-TEM line, for instance a quasi-TEM coplanar waveguide transmission line, this condition will generally not be satisfied, which leads to a coupler design with poor directivity [101].

Fig. 3.1a shows the principle of the conventional way of changing the coupling between two coupled lines by varying the capacitance between the two signal lines [100]. Fig. 3.1b and Fig. 3.1c illustrate the two tuning concepts of coupled
Figure 3.2: Illustration of Concept 1 based on geometrically tuning of the signal-to-ground coupling of the coupled lines.

Figure 3.3: Actuation states of Concept 1 (only one side of the coupled signal line illustrated): (a) State 1 (3 dB coupling); and (b) State 2 (6 dB coupling).

line directional couplers presented in this chapter.

### 3.2.1 Concept 1

Fig. 3.2 shows the 3-D illustration of Concept 1. In this concept, the coupling is varied by changing the capacitances of the signal lines to the ground while the capacitance between the signal lines is not changed. This results in a change in both the even and the odd-mode capacitances in contrast to the conventional
3.2. Concept and design of the novel tuneable directional couplers

Figure 3.4: Illustration of Concept 2 based on geometrically tuning of the signal-to-ground coupling simultaneously with the signal-to-signal line coupling.

Figure 3.5: Actuation states of Concept 2 (only one side of the coupled signal line illustrated): (a) State 1 (20 dB coupling); and (b) State 2 (10 dB coupling).

tuning concept where only the odd-mode capacitance is changed. The tuning is achieved by laterally moving the ground sidewalls of the coupled coplanar waveguides transmission line using integrated MEMS electrostatic actuators, as shown in Fig. 3.3. For the prototype device of Concept 1, the nominal coupling is changed from 3 dB in State 1 to 6 dB in State 2 by moving floating ground conductors and connecting them to the RF ground in their end position.

3.2.2 Concept 2

Fig. 3.4 shows the 3-D illustration of Concept 2. In this concept, the coupling is varied by changing the capacitances of the signal lines to the ground and
simultaneously varying the capacitance between the signal lines. This allows to design for a much larger coupling ratio variation by maintaining uniform performance over a large bandwidth. The even and the odd-mode capacitances are controlled independently and thus allows to eliminate any impedance mismatch. The tuning is achieved when two narrow intermediate floating lines are switched to the signal lines of the two coupled lines, and simultaneously the ground sidewalls are moved apart for compensating the ground coupling, as shown in Fig. 3.5. For the prototype device of Concept 2, the nominal coupling is changed from 20 dB in State 1 to 10 dB in State 2. Concept 2 needs actuation voltages to be applied for both states, as the two independent actuators are operated alternatively for maintaining the two states.

3.3 Simulation and characterization

On-wafer 3-D micromachined transmission lines, shown in Fig. 1.1c, which offer reduced substrate and radiation losses as compared to conventional coplanar or microstrip lines [102,103] were used in the design. All structures were fabricated in a single-mask SOI RF MEMS process outlined in Fig. 5.1. Fig. 3.6 shows a SEM picture of the fabricated prototype device based on Concept 1 and Concept 2. To measure the four-port directional coupler on a two-port Agilent E8361A PNA network analyzer, three devices were fabricated of each type, where two ports out of the isolated, coupled, and through port were terminated on-chip with 50 Ω microwave thin film resistors. The termination resistors used are commercially available 0201 SMD resistors [104]. The resistors were mounted on the 3-D-micromachined transmission lines by using conductive silver-paste epoxy. Such kind of SMD resistors have successfully been used previously as termination resistors for millimeter wave application with the resistors being mounted using conductive epoxy [105].
3.3. Simulation and characterization

Figure 3.7: Measured and simulated S-parameters in the two coupling states of 3-to-6 dB directional coupler (Concept 1): (a) through (S_{21}) port; and (b) coupled (S_{31}) port.

Figure 3.8: Measured and simulated S-parameters in the two coupling states of 3-to-6 dB directional coupler (Concept 1): (a) isolated (S_{41}) port; and (b) return loss (S_{11}).

3.3.1 RF characterization

Fig. 3.7 shows the measured and simulated results from the through and coupled ports for the prototype of Concept 1. It is seen that the measurements agree well with the design values except that the measured through and coupled power levels fall short by 1-2 dB from the simulated values. Fig. 3.8a shows that the measured isolation is better than 20 dB in State 2 compared to being better than 16.5 dB in State 1. Fig. 3.8b shows good agreement between simulated and measured return loss trends. It can be seen from the simulated and measured results that both the high isolation and impedance match cannot be maintained over the entire frequency band of 10 to 18 GHz, similar to the conventional tuneable directional couplers [100]. The measured and simulated phase plots are shown in Fig. 3.9. The maximum simulated phase deviation from the ideal response (90°) is 8° in either actuation state. The measured maximum deviation of the phase difference from the ideal
response is 17° in either actuation state.

The measured and simulated results for the through and coupled ports of prototype of Concept 2 are shown in Fig. 3.10. It is seen that the measurements agree well with the design values except that the measured device has higher losses, which is visible by the through-port power falling short by approximately 1-2 dB. The measured isolation, shown in Fig. 3.11a, is better than 40 dB for both states, and the measured return loss, shown in Fig. 3.11b, is better than 15 dB through the entire frequency band of interest. Fig. 3.12 illustrates the measured and simulated phase plots for the through and coupled ports. The simulated phase plot shows the maximum phase deviation from the ideal response (90°) being 15° in State 2. The measured maximum deviation of the phase difference from the ideal response is 27° in State 2. In State 1 the measured phase response difference is not uniform over the whole band, in particular above 15 GHz.
3.3. Simulation and characterization

Figure 3.11: Measured and simulated S-parameters in two coupling states of 10-to-20 dB directional coupler (Concept 2): (a) isolated ($S_{41}$) port; and (b) return loss ($S_{11}$).

Figure 3.12: Measured and simulated phase plots of coupled ($S_{31}$) and through ($S_{21}$) ports in the two coupling states of 10-to-20 dB directional coupler (Concept 2).

3.3.2 Directivity analysis

Fig. 3.13a shows the measured directivity for the prototype of Concept 1 being between 10 and 20 dB and going above 20 dB only in actuation State 1 for the relatively narrow band between 10 and 12.5 GHz. The prototype of Concept 2 achieves a directivity well above 20 dB in the entire frequency bandwidth for both actuation states as shown in Fig. 3.13b. For State 2 the measured directivity is above 40 dB throughout the whole frequency band from 10 to 18 GHz.

3.3.3 MEMS actuator and reliability characterization

Table 3.1 shows the measured and simulated actuation and release voltages of the MEMS actuators. For the prototype devices of Concept 1, State 1 is passive meaning that the actuation voltage is only required to maintain the coupler in State 2. For the prototype devices of Concept 2, both states are active meaning that actuation voltage is required to maintain both State 1 and State 2. Despite utilizing
Chapter 3. Reconfigurable ultra-wideband directional couplers

Figure 3.13: Measured directivity for the fabricated directional couplers in the two coupling states: (a) 3-to-6 dB directional coupler (Concept 1); and (b) 10-to-20 dB directional coupler (Concept 2).

Table 3.1: Measured and simulated actuation and release voltages.

<table>
<thead>
<tr>
<th>Measurement Data</th>
<th>Concept 1</th>
<th>Concept 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>State 1</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Actuation Voltage (V)</td>
<td>0</td>
<td>28</td>
</tr>
<tr>
<td>Release Voltage (V)</td>
<td>0</td>
<td>23</td>
</tr>
<tr>
<td>State 2</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Actuation Voltage (V)</td>
<td>35</td>
<td>0</td>
</tr>
<tr>
<td>Release Voltage (V)</td>
<td>25</td>
<td>0</td>
</tr>
</tbody>
</table>

Simulation Data |

<table>
<thead>
<tr>
<th>State 1</th>
<th>Concept 1</th>
<th>Concept 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Actuation Voltage (V)</td>
<td>0</td>
<td>34</td>
</tr>
<tr>
<td>State 2</td>
<td>Concept 1</td>
<td>Concept 2</td>
</tr>
<tr>
<td>Actuation Voltage (V)</td>
<td>34</td>
<td>0</td>
</tr>
</tbody>
</table>

Spring Constant (N/m) | 49 | 37 | 57 |

stiff mechanical spring designs, reasonable actuation voltages with a maximum around 35 V are sufficient to operate the devices.

This type of all-metal RF MEMS device with a monocrystalline-silicon core for all mechanically moving parts has exhibited high reliability [75, 79, 106] when evaluated for their lifetime. Lifetime measurements were performed with no RF signal applied using an unpackaged chip on a probe station in a non-hermetic, uncontrolled atmospheric laboratory environment. The device was actuated using a unipolar 35 V square waveform with a 50% duty cycle and at a frequency of
3.4 Discussion and outlook

RF MEMS ultra-wideband 10 to 18 GHz coupled line directional couplers presented in this chapter demonstrate promising concepts to achieve tunability in directional couplers. The tuneable directional couplers are implemented using 3-D micromachined coupled transmission lines using SOI high-resistivity wafers.

The biggest concern is that the measured through and coupled port power levels fall short by 1 – 2 dB from the simulated values. For these type of circuits, the right angle bends can lead to performance degradation due to slot-line mode excitation, which increases the radiation loss in the circuit. The reactance of the bend itself can increases radiation loss which can be compensated by using chamfered corners. Furthermore, any step change in the width of the signal lines and ground gap can perturb the normal coplanar waveguide electric and magnetic fields, resulting in radiation loss besides impedance mismatch [107]. However, losses through mode conversion and radiation should also be seen in the simulations, since a multi-mode analysis was carried out. Furthermore, it was found that air bridges placed near the coplanar waveguide bends to suppress the parasitic slot line mode do not significantly improve the loss performance. Current crowding around the etch holes can also be ruled out as a source of the losses as the moving elements are almost current-free in some states. Therefore, it is assumed that the additional losses are mainly attributed to the mounting of the termination resistors, since the simulation model uses ideal port termination for all four ports.
### Table 3.2: Performance comparison of various published directional couplers

<table>
<thead>
<tr>
<th>Device Technology</th>
<th>Through port (dB)</th>
<th>Coupled port (dB)</th>
<th>Return loss (dB)</th>
<th>Isol. (dB)</th>
<th>Dir. (dB)</th>
<th>Freq. (GHz)</th>
<th>BW (MHz)</th>
<th>Size ((\lambda))</th>
</tr>
</thead>
<tbody>
<tr>
<td>CMOS MMIC [92]</td>
<td>1.2</td>
<td>3 to -6</td>
<td>&gt;19</td>
<td>&gt;9</td>
<td>&gt;10</td>
<td>6.8</td>
<td>2300</td>
<td>0.03 \times 0.01</td>
</tr>
<tr>
<td>GaAs MMIC [96]</td>
<td>-</td>
<td>-60 to -6.6</td>
<td>&gt;9</td>
<td>&gt;9</td>
<td>-</td>
<td>2</td>
<td>500</td>
<td>0.02 \times 0.01</td>
</tr>
<tr>
<td>Varactor Diode [90]</td>
<td>-</td>
<td>10 to 6</td>
<td>&gt;15</td>
<td>&gt;15</td>
<td>&gt;10</td>
<td>1.5</td>
<td>1000</td>
<td>0.21 \times 0.12</td>
</tr>
<tr>
<td>Varactor Diode [93]</td>
<td>2 to 6</td>
<td>11 to 3</td>
<td>&gt;20</td>
<td>&gt;20</td>
<td>-</td>
<td>2</td>
<td>400</td>
<td>-</td>
</tr>
<tr>
<td>Varactor Diode [94]</td>
<td>0.6 to 5</td>
<td>15 to 2</td>
<td>&gt;25</td>
<td>&gt;23</td>
<td>-</td>
<td>2</td>
<td>800</td>
<td>-</td>
</tr>
<tr>
<td>MEMS [99]</td>
<td>2 to 25</td>
<td>25 to 2</td>
<td>&gt;16</td>
<td>&gt;17</td>
<td>-</td>
<td>12</td>
<td>400</td>
<td>0.48 \times 0.48</td>
</tr>
<tr>
<td>MEMS [100]</td>
<td>-</td>
<td>17 to 10</td>
<td>&gt;10</td>
<td>&gt;13</td>
<td>-</td>
<td>18</td>
<td>4000</td>
<td>0.08 \times 0.07</td>
</tr>
<tr>
<td>MEMS (Concept 1)</td>
<td>4.15 to 5.37</td>
<td>8.07 to 6.28</td>
<td>&gt;10</td>
<td>&gt;15</td>
<td>&gt;10</td>
<td>14</td>
<td>8000</td>
<td>0.09 \times 0.06</td>
</tr>
<tr>
<td>MEMS (Concept 2)</td>
<td>2.03 to 3.27</td>
<td>18.57 to 11.31</td>
<td>&gt;15</td>
<td>&gt;40</td>
<td>&gt;22</td>
<td>14</td>
<td>8000</td>
<td>0.08 \times 0.06</td>
</tr>
</tbody>
</table>
Chapter 4

Linearity analysis in multi-device RF MEMS circuits

This chapter presents the RF nonlinearity analysis of complex multi-device RF MEMS circuits. Closed-form expressions for the IIP3 of multi-device RF MEMS circuits are derived and presented. This is followed by the IIP3 analysis of different RF MEMS tuneable-circuit concepts, i.e., digital MEMS varactor banks, MEMS switched capacitor banks, MEMS impedance tuners and MEMS tuneable filters. The degradation of the overall circuit linearity with increasing number of device stages is also investigated. Finally, design rules are presented so that the mechanical parameters and thus the IIP3 of the individual device stages can be optimized to achieve a highest overall IIP3 for the whole circuit.

4.1 Introduction

Due to the RF MEMS research effort over the years, tuneable capacitors and switches have achieved impressive performance in terms of insertion loss, bandwidth [14], reliability [63] and tuning range [69]. Although packaging and response time are still major challenges to be addressed, RF MEMS tuneable capacitors find applications as tuning elements in complex multi-device RF MEMS circuits including phase shifters [56], tuneable filters [57], voltage controlled oscillators, and impedance matching networks [44]. These circuits are often operated at moderate to high power levels making linearity an important parameter to avoid signal distortion. The nonlinearity effect for a cascade arrangement of circuits is cumulative [108, 109] and the overall IP3 is limited by the lowest value and the nonlinearity of later stages become more pronounced because of being scaled down by the gain of the previous stages.

A theoretical and experimental study of the nonlinear effects in RF MEMS varactors and capacitive switches for a single device have been carried out in [110], including analytical derivations and an electromechanical model for verification.
Implementations on nonlinearity of the membrane movement of an electrostatically actuated capacitive MEMS switch was also analysed in [111], using a dynamic and parametric model. A mathematical analysis of the nonlinear behaviour of a single MEMS tuneable capacitor based on Volterra series has been shown in [112]. An analytical frequency dependant model for computing the intermodulation distortion in a single RF MEMS capacitor made up of two parallel plates has been proposed in [113]. All these previous RF MEMS theoretical nonlinearity analysis have been performed for single devices only. However, state-of-the-art RF MEMS circuits are composed of an increasing number of MEMS-tuneable/switched stages. WiSpry’s current generation mobile phone antenna tuner, for instance, has an 80-element MEMS-switched capacitor bank [114]. The overall radio receiver has to comply to an overall IIP3 limit set by standardization [115, 116] which cannot be exceeded. This makes it important to analyse the overall linearity of the entire RF MEMS circuit whose IIP3 can differ significantly when compared to a single-stage device. The trend of increase in nonlinear effects in multi-device systems can also be seen in other engineering disciplines [117].

4.2 Nonlinearity in MEMS tuneable capacitors and switches

RF circuits often operate at high RF power levels resulting in high electrostatic forces on the MEMS-tuning mechanisms and thus introduce intermodulation distortion. Although these distortions may be quite small for a single device, they grow significantly for complex, multi-device RF MEMS circuits. IIP3 is a quantitative measure of these distortions. The expression for the IIP3 of a single RF MEMS shunt capacitor loading on a transmission line is provided in [110]

\[
\text{IIP}_3 = \frac{2kg^2}{\phi C Z_0} = \frac{4kg^2}{\omega C^2 Z_0^2}
\]

(4.1)

In the following sections, the equation and the analysis provided in [110] is extended to a two-stage capacitor circuit model with equal stages having the same parameters (capacitance, spring constant and gap). This model is then further extended to a more general case of two capacitors with unequal parameters, i.e. either unequal stages or with the stages in a different configuration. This leads to the general closed form analytical formula for IIP3 calculation for a circuit utilizing N-stage capacitors with stages of unequal parameters. This general model is then used as the basis for the derivation of the overall IIP3 for a very common case where N-stage switched capacitors with fixed series capacitors are loaded on the transmission line to have a better control of the capacitance achieved in various states of the RF MEMS circuit.

4.2.1 Two equal capacitor stages

Fig. 4.1 shows a circuit model with two capacitor stages with the same parameters, i.e. \( C_1 = C_2 = C \) with \( C_{tot} = 2C \), also \( k_1 = k_2 = k \) with \( k_{tot} = 2k \). As both
4.2. Nonlinearity in MEMS tuneable capacitors and switches

4.2.1 Two-stage capacitor circuit with equal stages

When the capacitor values and springs are equal, the moving electrodes experience the same displacement \(g_1 = g_2 = g\). The electrostatic force on each capacitance stage of the circuit can be written as

\[
F_1 = \frac{C_1V^2}{2g_1}, \quad F_2 = \frac{C_2V^2}{2g_2}
\]

and the total electrostatic force is thus

\[
F_{\text{tot}} = \frac{C_1V^2}{2g_1} + \frac{C_2V^2}{2g_2} = \frac{2CV^2}{2g}
\]

The output phase extracted from the transmission coefficient can then be written as [110]

\[
\phi \simeq -\omega C Z_0
\]

Following the derivation in [110], the equation for IIP3 for the two-stage device of equal stages and thus equal behaviour is

\[
\text{IIP3} = \frac{2kg^2}{\phi C Z_0} = \frac{2kg^2}{\omega C^2 Z_0^2}
\]

It can be seen in (4.5) that the IIP3 is reduced by half when compared to (4.1), implying that the overall linearity of circuits is degrading with additional nonlinear device stages. It can be deduced from the above results that by doubling the capacitor with the same gap and spring constant values as in (4.1) reduces the IIP3 by half.

4.2.2 Two-stage capacitor circuit with unequal stages

For the case of unequal stages given by Fig. 4.2 which models either a circuit with two different capacitors or a circuit with two equal capacitors in different operating states [106], the electrostatic force on each capacitor can be written as

\[
F_1 = \frac{C_1V^2}{2g_1}, \quad F_2 = \frac{C_2V^2}{2g_2}
\]
From [110], the output phase can be written as

$$\phi_1 \approx -\frac{\omega C_1 Z_0}{2}, \quad \phi_2 \approx -\frac{\omega C_2 Z_0}{2} \quad (4.7)$$

The total capacitance is

$$C = C_1 + C_2 \quad (4.8)$$

and the total output phase is

$$\phi = \phi_1 + \phi_2 \quad (4.9)$$

Thus, for a small displacement $\Delta x$, the capacitance and the phase changes can be expressed as

$$C(t) = C + \Delta C(t) = C_1 + C_2 + \Delta C_1(t) + \Delta C_2(t) \quad (4.10)$$

$$C(t) \approx C_1 \left(1 - \frac{\Delta x_1(t)}{g_1}\right) + C_2 \left(1 - \frac{\Delta x_2(t)}{g_2}\right) \quad (4.11)$$

$$\phi + \Delta \phi = \phi_1 + \Delta \phi_1 + \phi_2 + \Delta \phi_2 \quad (4.12)$$

$$= \phi_1 \left(1 - \frac{\Delta x_1(t)}{g_1}\right) + \phi_2 \left(1 - \frac{\Delta x_2(t)}{g_2}\right) \quad (4.13)$$

Considering the case when two RF signals are incident on the capacitor, the output can be written as

$$V_0 = V_1 \sin (\omega_1 t + \phi + \Delta \phi(t)) + V_2 \sin (\omega_2 t + \phi + \Delta \phi(t)) \quad (4.14)$$

The displacement given in [110] can be written for two capacitive elements

$$\Delta x_1(t) = \frac{F_1}{k_1} = \frac{C_1}{2 k_1 g_1} [V_1 V_2 \cos (\omega_1 - \omega_2) t] \quad (4.15)$$

and

$$\Delta x_2(t) = \frac{F_2}{k_2} = \frac{C_2}{2 k_2 g_2} [V_1 V_2 \cos (\omega_1 - \omega_2) t] \quad (4.16)$$
4.2. Nonlinearity in MEMS tuneable capacitors and switches

\[ C_1 \neq C_2, \quad k_1 \neq k_2, \quad g_1 \neq g_2 \]

\[ \begin{align*}
C_{S} g_1 + \Delta x_1 & \\
C_{S} g_2 + \Delta x_2 & \\
C_{N} g_N + \Delta x_N &
\end{align*} \]

**Figure 4.3:** N-stage capacitor circuit, modelling either N unequal stages or N stages in different operation states.

where the terms \(2\omega_1, 2\omega_2, \omega_1 + \omega_2\) are neglected. Inserting these equations in the above output equation, the intermodulation level is then derived as

\[
P_{\text{intermod}} = P_{\text{sideband}} = \left[ \frac{V_1 V_2}{2} \left( \frac{\phi_1 C_1}{2k_1 g_1^2} + \frac{\phi_2 C_2}{2k_2 g_2^2} \right) \right]^2 \quad (4.17)
\]

Performing the calculations on the above equation results in

\[
\begin{align*}
\text{IIP}_3 &= \frac{2k_1 g_1^2 k_2 g_2^2}{[\phi_1 C_1 (k_2 g_2^2) + \phi_2 C_2 (k_1 g_1^2)] Z_0} \quad (4.18) \\
\text{IIP}_3 &= \frac{4k_1 g_1^2 k_2 g_2^2}{[\omega C_1^2 (k_2 g_2^2) + \omega C_2^2 (k_1 g_1^2)] Z_0} \quad (4.19)
\end{align*}
\]

**4.2.3 N stages of unequal capacitors**

Practical RF MEMS circuits contain a large number of capacitors, therefore, it is necessary to derive the closed form analytical formula for the IIP3 of N stages of capacitors with unequal capacitances, spring constants and gaps i.e. all stages can be operated independently or are designed differently. Fig. 4.3 shows a circuit model with N parallel capacitor stages which models either a circuit with N different capacitors or a circuit with N equal capacitors in different operating states [106].

Similar to the derivation of a single capacitor in [110], the electrostatic force \(F_i\) on the moving electrode of each capacitor \(C_i\) can be written as

\[
F_i = \frac{C_i V^2}{2g_i} \quad (4.20)
\]

The total capacitance and the phase can be written as

\[
C = \sum_{i=1}^{N} C_i, \quad \phi = \sum_{i=1}^{N} \phi_i \quad (4.21)
\]
Chapter 4. Linearity analysis in multi-device RF MEMS circuits

$N$ being number of capacitors. The capacitance for a small displacement is

$$C(t) \simeq \sum_{i=1}^{N} C_i \left(1 - \frac{\Delta x_i(t)}{g_i}\right)$$

(4.22)

and the phase is

$$\phi + \Delta \phi = \sum_{i=1}^{N} \phi_i \left(1 - \frac{\Delta x_i(t)}{g_i}\right)$$

(4.23)

When two signals are incident on the capacitor, the output is

$$V_0 = V_1 \sin(\omega_1 t + \phi + \Delta \phi(t)) + V_2 \sin(\omega_2 t + \phi + \Delta \phi(t))$$

(4.24)

The displacement for each stage can be written as

$$\Delta x_i(t) = \frac{F_i}{k_i} = \frac{C_i}{2k_i g_i} \left[\frac{V_1^2}{2} + \frac{V_2^2}{2} + V_1 V_2 \cos(\omega_1 - \omega_2) t\right]$$

(4.25)

The terms $2\omega_1$, $2\omega_2$ and $\omega_1 + \omega_2$ are neglected. Inserting this equation in the above output equation and assuming $V = V_1 = V_2$, the third order intermodulation products can be written as

$$P_{\text{intermod}} = \frac{P_{\text{sideband}}}{P_{\text{signal}}} = \left[\frac{V_1 V_2}{2} \sum_{i=1}^{N} \left(\frac{\phi_i C_i}{2k_i g_i^2}\right)\right]^2$$

(4.26)

The two-tone third order intermodulation intercept point is the value of $P_{\text{signal}}$ for which $P_{\text{signal}} = P_{\text{sideband}}$

$$\text{IIP}_3 = 2 \frac{Z_0}{\omega Z_0^2} \left(\frac{1}{\sum_{i=1}^{N} \frac{\phi_i C_i}{k_i g_i^2}}\right) = 4 \frac{Z_0}{\omega Z_0^2} \left(\frac{1}{\sum_{i=1}^{N} \frac{C_i}{k_i g_i^2}}\right)$$

(4.27)

From (4.27), it can be generalized that having $N$ number of identical parallel capacitors reduces the IIP3 of the circuit compared to the single capacitor by a factor of $N$. This is shown in Fig. 4.4 which analyses the overall circuit linearity deterioration with increasing number of tuneable-capacitor devices. For this analysis, for each capacitor, the gap is changed from 1.5 µm to 0.5 µm, spring constant from 12 N/m to 55 N/m and the capacitance from 45 fF to 145 fF in the up and down states, respectively calculated at the center frequency of 2.5 GHz.

4.2.4 N-stage switched capacitors with fixed series capacitors

Capacitor banks are extensively used in tuneable filters [72, 73] where MEMS switches are connected in series with fixed metal-air-metal (MAM) or metal-insulator-metal (MIM) capacitors to achieve a high tuning resolution. By choosing
appropriate capacitor values of the fixed capacitors, the total capacitor variation can be controlled accurately and a very high capacitance ratio can be achieved.

The MEMS capacitor bank can be analysed using the model in Fig. 4.5 shown for two stages. Same derivation applies for this case as for the previous case. The displacement is related to the square of the voltage on the moving MEMS device obtained by using voltage division between the switched capacitor and the fixed capacitor of each stage.

\[
V_{Si} = \left( \frac{C_{Fi}}{C_{Si} + C_{Fi}} \right) V
\]  

(4.28)
Now the displacement can be written as

$$\Delta x_i(t) = \frac{F_i}{k_i} = \frac{C_{Si}}{2k_i g_i} \left( \frac{C_{F_i}}{C_{Si} + C_{F_i}} \right)^2 \left[ \frac{V_1^2}{2} + \frac{V_2^2}{2} + V_1 V_2 \cos(\omega_1 - \omega_2)t \right]$$

(4.29)

where $i = 1, 2, 3, \ldots, N$ and the terms $2\omega_1$, $2\omega_2$ and $\omega_1 + \omega_2$ are neglected. Inserting (4.29) in (4.24), the intermodulation level is then given by

$$P_{\text{intermod}} = \frac{P_{\text{sideband}}}{P_{\text{signal}}} = \left[ \frac{V_1 V_2}{2} \sum_{i=1}^{N} \left( \phi_i C_{Si} \frac{C_{F_i}}{2k_i g_i^2} \left( \frac{C_{F_i}}{C_{Si} + C_{F_i}} \right)^2 \right) \right]^2$$

(4.30)

and the calculation of the IIP3 based on the above equation results in

$$IIP3 = \frac{2}{Z_0} \left[ \frac{1}{\sum_{i=1}^{N} \phi_i C_{Si} \frac{C_{F_i}}{2k_i g_i^2} \left( \frac{C_{F_i}}{C_{Si} + C_{F_i}} \right)^2} \right]$$

or

$$IIP3 = \frac{4}{\omega Z_0} \left[ \frac{1}{\sum_{i=1}^{N} C_{Si}^2 \frac{C_{F_i}}{2k_i g_i^2} \left( \frac{C_{F_i}}{C_{Si} + C_{F_i}} \right)^3} \right]$$

(4.31)

(4.32)

From the above equations it can be concluded that if the value of the fixed capacitor $C_F$ is large as compared to the capacitive switch $C_S$ then most of the voltage drop is across the capacitive switch, and thus the overall linearity is dominated by the linearity of the MEMS capacitive switch. Similarly if the fixed capacitor has a small value as compared to the capacitive switch there is less voltage drop across the capacitive switch and thus the influence of the nonlinearity of the MEMS switch on the overall nonlinearity is reduced.

### 4.3 Linearity analysis of RF MEMS circuit examples

The IIP3 of different multi-device RF MEMS circuit concepts is analysed in this section. These include the digital MEMS varactor bank which is used in distributed MEMS transmission line (DMTL) circuits [14], MEMS switched capacitor bank which is extensively used in tuneable filters [72,73], MEMS impedance tuners [118], MEMS tuneable filters [119] and multi-step MEMS capacitor concept [106]. For the digital MEMS varactor bank and the MEMS switched capacitor bank, the IIP3 is calculated using the derived equations and the values are then compared
4.3. Linearity analysis of RF MEMS circuit examples

Discrete position
varactors
≈ 1×C
≈ 2×C
≈ 4×C

Capacitor I 
Capacitor II 
Capacitor III 
Bit 0Bit 1Bit 2

Figure 4.6: 3-D illustration and equivalent capacitor model of three-stage MEMS varactor bank.

4.3.1 Linearity analysis of digital MEMS varactor bank

This example analyses a 3-bit digital MEMS varactor bank shown in Fig. 4.6. Each varactor has two discrete positions with a capacitance ratio of $\sim 3$. The total capacitance of the circuit can be tuned from 146 fF to 430 fF in eight discrete states. Table 4.1 shows the device parameters used for the calculation of the IIP3 for each state. The capacitance values are based on data provided in [61] with the spring constants in the table being derived from the geometries of the membranes by using a COMSOL Multiphysics simulation model. Equation (4.27) is used to calculate the IIP3 of the varactor bank in each of its 8 operation states. Fig. 4.7 shows the calculated IIP3 for the various states of the bank compared to the simulated IIP3 using a nonlinear electromechanical model [110] implemented in Agilent ADS for each varactor. The simulation data of the IIP3 agrees within $\pm 1.7$ dBm with the calculated values.

4.3.2 Linearity analysis of MEMS switched capacitor bank

Fig. 4.8 shows the analysis example of a 3-bit MEMS switched capacitor bank with a metal-air-metal fixed capacitor in series with each MEMS switch with the values
Table 4.1: Summary of parameters for three stage digital MEMS varactor bank used for linearity analysis

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Capacitor I (fF)</th>
<th>Capacitor II (fF)</th>
<th>Capacitor III (fF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_{up}$</td>
<td>63</td>
<td>45</td>
<td>38</td>
</tr>
<tr>
<td>$C_{down}$</td>
<td>205</td>
<td>145</td>
<td>80</td>
</tr>
<tr>
<td>$g_{up}$</td>
<td>1.5</td>
<td>1.5</td>
<td>1.5</td>
</tr>
<tr>
<td>$g_{down}$</td>
<td>0.5</td>
<td>0.5</td>
<td>0.5</td>
</tr>
<tr>
<td>$k_{up}$</td>
<td>12</td>
<td>12</td>
<td>12</td>
</tr>
<tr>
<td>$k_{down}$</td>
<td>55</td>
<td>55</td>
<td>55</td>
</tr>
</tbody>
</table>

Figure 4.7: Calculated and simulated IIP3 of the eight states of the three-stage digital MEMS varactor bank.

of the fixed capacitors being binary coded. The device parameters used for the analysis are shown in Table 4.2. The up-state capacitance of the MEMS switch is calculated using a $200 \times 100 \, \mu m$ membrane with a gap of $2 \, \mu m$ while the down state capacitance is calculated with a gap of $0.2 \, \mu m$. The up-state spring constant is calculated by performing simulations in Comsol Multiphysics for a membrane with dimensions mentioned above and a thickness of $0.8 \, \mu m$. The down state spring constant is modelled by clamping the membrane at several positions, thus simulating the effect of spacer bumps. Equation (4.31) is used to calculate the IIP3 of the switched capacitor bank in each of the 8 operation states. Fig. 4.9 shows the calculated IIP3 for various states of the bank compared to the simulated IIP3 using the nonlinear electromechanical model of the circuit in Agilent ADS. The simulation of IIP3 agrees within $\pm 2 \, dBm$ with the calculated data.
4.3. **Linearity analysis of RF MEMS circuit examples**

Figure 4.8: 3-D illustration and equivalent capacitor model of three-stage MEMS switched capacitor bank.

Table 4.2: Summary of parameters for three stage MEMS switched capacitor bank used for linearity analysis

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Capacitor I</th>
<th>Capacitor II</th>
<th>Capacitor III</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_{S,\text{up}}$ (fF)</td>
<td>885</td>
<td>885</td>
<td>885</td>
</tr>
<tr>
<td>$C_{S,\text{down}}$ (fF)</td>
<td>885</td>
<td>885</td>
<td>885</td>
</tr>
<tr>
<td>$C_F$ (fF)</td>
<td>160</td>
<td>320</td>
<td>640</td>
</tr>
<tr>
<td>$g_{S,\text{up}}$ (µm)</td>
<td>2</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>$g_{S,\text{down}}$ (µm)</td>
<td>0.2</td>
<td>0.2</td>
<td>0.2</td>
</tr>
<tr>
<td>$k_{S,\text{up}}$ (N/m)</td>
<td>32</td>
<td>32</td>
<td>32</td>
</tr>
<tr>
<td>$k_{S,\text{down}}$ (N/m)</td>
<td>586.5</td>
<td>586.5</td>
<td>586.5</td>
</tr>
</tbody>
</table>

4.3.3 **Linearity analysis of impedance tuner**

A V-band single stub tuner in [118] with 10 MEMS switches in series with MAM capacitors loaded on a transmission line is considered for this analysis. The circuit has a total of 1024 impedance states covering most of the smith chart. The device parameters used for the IIP3 analysis are taken from [118]. The down state spring constant is taken from Table 4.2. The IIP3 of the impedance tuner is simulated at the center frequency of 62.5 GHz with a tone separation of 60 kHz using a harmonic balance simulation of nonlinear electromechanical model for the MEMS switch in Agilent ADS [110]. The results are shown in Fig. 4.10 plotting 15 selected simulated states, the IIP3 of a single switch in the up and down state, and the IIP3 of 10 parallel switches all in either up or down state. It can be deduced from the results that the IIP3 of complex RF MEMS circuits is very different when compared to a
Chapter 4. Linearity analysis in multi-device RF MEMS circuits

Figure 4.9: Calculated and simulated IIP3 of the eight states of the three-stage MEMS switched capacitor bank.

Figure 4.10: Simulated IIP3 for the impedance tuner [118] \((f = 62.5\ \text{GHz, } \Delta f = 60\ \text{kHz})\).

difficult single switch stage and even very different when compared to the same number of switch stages in simple parallel combination. It can also be deduced from the results that for complex RF MEMS circuits, the worst case IIP3 does not necessarily occur for the switches all in either up or down state. This means that a careful analysis of the IIP3 has to be performed in each state of the circuit to extract the worst case IIP3 value.
4.3. Linearity analysis of RF MEMS circuit examples

4.3.4 Linearity analysis of tuneable filter

The example of a three-pole filter from [119] can be tuned from 12.2 to 17.8 GHz with each resonator periodically loaded by four switched MEMS capacitor pairs. Every switch is a series combination of a MEMS switch and a fixed MAM capacitor. The loaded MEMS resonators are coupled through inductive inverters to form a three pole band pass filter. The measured IIP3 is compared to simulated IIP3 at the frequency of 17.8 GHz with the tone separation of 40 and 200 kHz using a harmonic balance simulation of nonlinear electromechanical model for the MEMS switch in Agilent ADS [110] with all the switches in the up state. Fig. 4.11 shows that the simulations agree very well with the measurements. These results confirm the accuracy of the nonlinear electromechanical model implemented in Agilent ADS to precisely predict the IIP3 of the entire RF MEMS circuit.

4.3.5 Linearity analysis of multi-step MEMS capacitor

Fig. 2.3a shows a two-stage multi-step MEMS capacitor whose three-stage version is used for the IIP3 analysis. A total number of 7 states can be achieved with this device. Fig. 4.12 shows the calculated IIP3 for the device in the seven different states at 2.5 GHz, derived from the measured capacitances, measured sidewall displacements and simulated spring constants of the seven states of the device, using (4.1) when the actuators on both sides are in the same state and using (4.27) when the two sides are actuated asymmetrically. For this device type, it was also attempted to measure the linearity characteristic of a fabricated prototype by measuring the two-tone third-order intermodulation intercept point with two signal sources at a center frequency of 2.5 GHz, separated by a 12 MHz offset. The IIP3 measurements were carried out for all the seven states of a multi-step MEMS capacitor. Fig. 4.13 shows that the IIP3 measurements of all states reach a common limit at 63.2 dBm, i.e., the measured nonlinearity is limited by the transmission line measurement setup, and not by the MEMS device, as the IIP3 measurement.

![Figure 4.11: Simulated and Measured IIP3 for the three-pole filter [119] (f = 17.8 GHz, ∆f = 40, 200 kHz).](image)
Chapter 4. Linearity analysis in multi-device RF MEMS circuits

![Graph showing IIP3 and total capacitance for different states of the multi-step MEMS capacitor.](image)

**Figure 4.12:** Calculated IIP3 of the seven states of the multi-step MEMS capacitor ($f = 2.5$ GHz).

![Graph showing measured dual-tone fundamental and intermodulation levels.](image)

**Figure 4.13:** Measured dual-tone fundamental and intermodulation levels versus input power for 7 states of the multi-step MEMS capacitor ($f = 2.5$ GHz, $\Delta f = 12$ MHz).

Of a straight transmission line without any MEMS device was also characterized to be 63 dBm, which is far below the estimate by the analytical model (worst-case nonlinearity of 75 dBm). The literature also confirms the practical measurement limit that arises from the passive intermodulation between the probe tips, the CPW and the substrate [82].
Table 4.3: Summary of parameters for a five stage digital MEMS varactor bank used for linearity analysis

<table>
<thead>
<tr>
<th>No.</th>
<th>Case I</th>
<th>Case II</th>
<th>Case III</th>
<th>Case IV</th>
<th>Case V</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>C (fF)</td>
<td>k (N/m)</td>
<td>C (fF)</td>
<td>k (N/m)</td>
<td>C (fF)</td>
</tr>
<tr>
<td>1</td>
<td>198.4</td>
<td>40</td>
<td>512</td>
<td>266.39</td>
<td>512</td>
</tr>
<tr>
<td>2</td>
<td>198.4</td>
<td>40</td>
<td>256</td>
<td>66.60</td>
<td>256</td>
</tr>
<tr>
<td>3</td>
<td>198.4</td>
<td>40</td>
<td>128</td>
<td>16.65</td>
<td>128</td>
</tr>
<tr>
<td>4</td>
<td>198.4</td>
<td>40</td>
<td>64</td>
<td>4.16</td>
<td>64</td>
</tr>
<tr>
<td>5</td>
<td>198.4</td>
<td>40</td>
<td>32</td>
<td>1.04</td>
<td>32</td>
</tr>
</tbody>
</table>

4.4. Proposed design rules for IIP3 enhancement

As RF MEMS circuits are composed of an increasingly large number of MEMS-tunable/switched stages, the analysis of the nonlinearity of the entire circuit gets increasingly important. The overall circuit linearity is depending on the number of stages of a MEMS-tuning circuit and is degrading with increasing circuit complexity if proper design precautions are not taken. To show the scaling relationship between nonlinearity and the number of device stages, and to provide design rules for optimizing the linearity in multi-device RF MEMS circuits, the examples of a five stage digital MEMS varactor bank and a MEMS switched capacitor bank are analysed. The design variations are implemented in a way that the total capacitance is the same for all designs, and the capacitances and springs are varied as explained below. For simplicity, only the up-state of all capacitances are analysed.

The capacitance and spring parameters of different implementations of a digital MEMS varactor bank are shown in Table 4.3. An initial gap of 1 µm is assumed for the moving membranes. The IIP3 of the digital MEMS varactor banks are calculated for the up-state using (4.27) and are shown in Fig. 4.14 along with the overall capacitance. For Case I, the total capacitance is divided equally among all the stages. The spring constant is also equal for all the stages which results in an IIP3 of 50.15 dBm for each stage as shown in Fig. 4.14. The overall IIP3 for this case is 43.16 dBm. For Case II, the capacitor stages are binary coded and the spring constant is the same for each stage and equal to Case I. It can be seen here that the IIP3 of all the stages are different and the total IIP3 is 40.67 dBm which is lower as compared to Case I. For Case III, the capacitances are again binary coded and the spring constants are adjusted such that the IIP3 of each stage is the same and equal to the value of 50.15 dBm, to compare the circuit of Case I. This results in the total IIP3 for Case III being equal to Case I. For Case IV, the spring constant is divided in the same ratio as the total capacitance is divided among the stages which results in different IIP3 for each stage but the same overall-circuit IIP3 as for Case I and Case III. For Case V, the capacitances are again binary coded and the spring constants are taken from Case III but used in the reverse direction. This
results in an actuation voltage of 49.66 V in the worst case, i.e., a feasible design. It can be seen here that the total IIP3 is 25.78 dBm which is significantly worse as compared to previous cases, which emphasizes the importance of proper circuit design.

For the MEMS-switched capacitor bank, the up-state capacitance is assumed to be 88.5 fF for all switches and the spring constant of the switches is utilized for optimizing the overall nonlinearity. The IIP3 of the MEMS switched capacitor bank was calculated using five cases of different distributions of the total circuit capacitance over five fixed, i.e. switched, capacitances, with the parameters shown in Table 4.4 using (4.31) for the five stage banks. The resulting IIP3 values along with the total capacitances are shown in Fig. 4.15. For Case I, the total capacitance is divided equally among all the stages by having the same value for the fixed capacitor stages. The spring constant of the MEMS switch is also equal for all the stages of this case which results in same IIP3 of 64.39 dBm for all stages as shown in Fig. 4.14. The overall IIP3 for this case is 57.4 dBm. For Case II the fixed

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**Figure 4.14:** Calculated IIP3 for the circuits of the five-stage digital MEMS varactor banks given for all actuators in the up state.

**Table 4.4:** Summary of parameters for a five stage MEMS switched capacitor bank used for linearity analysis

<table>
<thead>
<tr>
<th>No.</th>
<th>Case I</th>
<th>Case II</th>
<th>Case III</th>
<th>Case IV</th>
<th>Case V</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>CF (fF)</td>
<td>k (N/m)</td>
<td>CF (fF)</td>
<td>k (N/m)</td>
<td>CF (fF)</td>
</tr>
<tr>
<td>1</td>
<td>119.42</td>
<td>40</td>
<td>512</td>
<td>40</td>
<td>512</td>
</tr>
<tr>
<td>2</td>
<td>119.42</td>
<td>40</td>
<td>256</td>
<td>40</td>
<td>256</td>
</tr>
<tr>
<td>3</td>
<td>119.42</td>
<td>40</td>
<td>128</td>
<td>40</td>
<td>128</td>
</tr>
<tr>
<td>4</td>
<td>119.42</td>
<td>40</td>
<td>64</td>
<td>40</td>
<td>64</td>
</tr>
<tr>
<td>5</td>
<td>119.42</td>
<td>40</td>
<td>32</td>
<td>40</td>
<td>32</td>
</tr>
</tbody>
</table>
4.5 Discussion and outlook

This chapter has shown that the IIP3 of a multi-device RF MEMS circuit can be significantly lower than the IIP3 of a single device, which requires an analysis of the overall circuit IIP3 rather than just for a single stage. It has also been shown
Chapter 4. Linearity analysis in multi-device RF MEMS circuits

that the IIP3 of complex RF MEMS circuits is very different when compared to a single switch stage and even very different when compared to the same number of switch stages in simple parallel combination. In order to achieve the maximum IIP3 for a multi-device RF MEMS circuit, the circuit should be designed such that each stage has the same IIP3.

Explicit IIP3 equations are derived for specific cases of digital MEMS varactor bank and MEMS switched capacitor bank. For more complex circuits, a nonlinear electromechanical model in Agilent ADS is used. The equations derived in this chapter applies to both digital and analog capacitors in any tuning state. All derived equations for the IIP3 are valid if there is the same voltage across the components in the circuit. If not, the IIP3 of each sub-circuit should be evaluated separately. It should also be noted that the small capacitance approximation $\omega C_0 \ll 1$ used to derive the IIP3 equations is no longer valid for large capacitance values.

There exists a compromise between high tuning range and high IIP3 for the MEMS switched capacitor banks as high tuning range can be achieved with $C_F \gg C_{S,up}$ but at the same time this will result in high voltage drop across the capacitive switch resulting in lower IIP3 as the overall linearity is dominated by the linearity of the MEMS capacitive switch. It should be noted that if instead of using a digital MEMS varactor bank, a single capacitance is tuned in steps as to achieve the same capacitance of various states, the IIP3 is similar for each state only if the cumulative spring constant and the gap are the same for each state.
Chapter 5

Multilayer ferromagnetic composites for permeability enhancement

This chapter discusses un-patterned ferromagnetic NiFe/AlN multilayer composites as magnetic core materials for on-chip inductances. This composite multilayer structure reduces RF induced currents and pushes the permeability cut-off to 3.7 GHz which is significantly higher than for homogeneous NiFe layers. This is accomplished by using a multilayer NiFe and AlN sandwich structure which increases the thickness of the ferromagnetic material without increasing its conductivity.

5.1 Introduction

The integration of passive components on a silicon chip is necessary for the design of high frequency state of the art integrated resonators, filters, impedance matching networks etc. There has been a trend in research efforts to increase the inductance and the quality factor of on-chip RF inductors by integrating magnetic materials with RF integrated circuits. The inductance of on-chip inductors could significantly be enhanced by magnetic materials which reduce the required area and increase magnetic coupling. Permeability of these materials is complex with the imaginary part representing the magnetic loss and the real part representing the ability to confine magnetic flux resulting in enhanced inductance values. It is generally difficult to integrate passive components with magnetic materials on silicon since IC fabrication technology is primarily optimized for digital applications with no need for magnetic materials. Major semiconductor companies have been spending large efforts on developing magnetic films which perform well at frequencies beyond 1 GHz [120]. The relative permeability of sputtered NiFe layers at low frequencies (DC to 10 MHz) can be as high as 3000, and with an optimized electrodeposition process even a permeability of 8500 was recently achieved [121]. However, such homogeneous NiFe films suffer from a number of limitations at frequencies above
100 MHz. Firstly, the magnetic NiFe layers are conductive which increases losses and reduces the effective inductance. Secondly, they have a low ferromagnetic resonance frequency. To suppress the flow of eddy currents in these films only very thin NiFe layers are used. On the other hand thicker layers are desired for inductor cores. Ferrite materials are a good dielectric material that can provide the possibility of extending the ferromagnetic resonance. The magnetic properties of these ferrite materials depend on their crystalline structure whose growth requires high temperature processing [122]. This makes ferrite materials difficult to integrate within the silicon technology as the high temperature step would destroy parts of the integrated circuit. It has been shown that enhanced ferromagnetic resonance is possible by placing the NiFe film between isolation layers [123,124]. Patterning the magnetic film can also be used to enhance the shape anisotropy which increases the ferromagnetic resonance [125]. Another approach to increase the resistivity and ferromagnetic resonance is to use granular films which add to the complexity of processing. A recent study has shown that the eddy currents can be reduced by using a magnetic nanoparticle composite [126].

5.2 Concept and design

Significant inductance enhancement for a magnetic core inductor can be achieved if the field created by the current-carrying conductors is parallel to the hard axis of the magnetic film [122]. A solenoid structure should be used for this purpose since such a field configuration is difficult to realize when using a planar coil. A micromachined microstrip transmission line was chosen as the test structure for the material characterization, consisting of a top copper microstrip line, the composite ferromagnetic material core, and a bottom copper ground layer. Between the layers, silicon dioxide (SiO\textsubscript{2}) and aluminum nitride (AlN) insulation layers are inserted to suppress leakage and induced currents flowing in the ferromagnetic core layer. The ferromagnetic material chosen for this evaluation is nickel-iron (NiFe) because of its common use in semiconductor fabrication, in particular for magnetic sensors and hard disk read-write heads.

5.3 Fabrication

All the fabrication has been done using high resistivity silicon wafers. The fabrication process flow for the microstrip transmission line test structures is shown in Fig. 5.1. The deposited magnetic film consists of 100 nm thick NiFe layers intersected by AlN isolation layers, embedded between a top and a bottom PECVD SiO\textsubscript{2} layer as shown in Fig. 5.2. Vertical interconnects were created by ion beam etching the magnetic/dielectric layers. The NiFe stack was in-situ magnetized during sputter deposition. Fig. 5.3 shows a photograph of the fabricated test structures.
5.4. Experimental results

The RF measurements of the fabricated test structures were performed using an Agilent E8361A PNA vector network analyzer calibrated using a GGB Industries
Chapter 5. Multilayer ferromagnetic composites for permeability enhancement

Figure 5.3: Photograph of part of the wafer showing the fabricated microstrip test structures.

CS-5 calibration substrate and 150 µm ground-signal-ground (GSG) coplanar probes and short-open-load-thru (SOLT) calibration.

From the measured S-parameters of the microstrip lines, the frequency-dependent characteristic impedance and propagation constant are extracted.

\[ Z_0 = \sqrt{\frac{Z_{11}}{Y_{11}}} \]  
\[ \gamma = \tanh^{-1}\left(\sqrt{\frac{1}{Z_{11}Y_{11}}}\right) \]  

where \( Z_0 \) is the characteristic impedance, \( Z_{11} \) and \( Y_{11} \) are extracted from the measured S-parameters and \( l \) is the length of the top strip. The inductance and capacitance per unit length are subsequently derived by using the following formulas.

\[ L = \frac{Im(\gamma Z_0)}{\omega} \]  
\[ C = \frac{Im\left(\frac{\gamma}{Z_0}\right)}{\omega} \]  

where \( \gamma \) is the propagation constant and \( \omega \) is the angular frequency.

The inductance per unit length of the microstrip line with NiFe/AlN multilayer composite is plotted in Fig. 5.4a together with a SiO\(_2\) reference line of equivalent thickness. The higher inductance is directly related to the high effective permeability of the magnetic stack. The capacitance per unit length is plotted in Fig. 5.4b together with the SiO\(_2\) reference line. The larger capacitance per unit length compared to the reference line is due to the NiFe/AlN multilayer composite whose multiple capacitances result in much wider fringing field volume and since the effective total thickness of the capacitance is reduced by the many conducting layers. The capacitance is nearly constant with frequency for both lines.

Fig. 5.5 shows the measured S-parameters and the inductance enhancement factor for the NiFe/AlN multilayer composite line relative to a SiO\(_2\) reference line.
5.4. Experimental results

The enhancement factor is about 4 at 1 GHz. The frequency dependent effective relative permeability of the NiFe/AlN multilayer composite material is shown in Fig. 5.6a and was extracted from the measurements by parameter matching of an Ansoft HFSS simulation model to the measured inductance per unit length for all individual frequency points. An effective permeability of 28 was achieved at 1 GHz, and the ferromagnetic resonance exceeds 3.7 GHz. The extracted permeability of the reference line is also shown with a nearly-constant value of 1 throughout the frequency range of interest, confirming the permeability extraction method. Fig. 5.6b shows the characteristic impedance of the microstrip line with the NiFe/AlN multilayer composite together with the SiO$_2$ reference line. The characteristic impedance for the NiFe/AlN multilayer composite is higher than that for the reference line and the value is not low in either case, resulting in a sufficient sensitivity of the measurements.

The measured relative permittivity of the SiO$_2$/NiFe/AlN/SiO$_2$ multilayer composite material is shown in Fig. 5.7 together with the measured relative permittivity of the SiO$_2$ reference line. The extracted permittivity of the reference line agrees very well with reported permittivities of SiO$_2$. The capacitive structure used to measure the permittivity is shown in an insert in the figure. Disc radii of 75 $\mu$m and 90 $\mu$m are used for the test structures in this frequency range using the following equation [127–129].

$$
\varepsilon_r = -\frac{t \left( \frac{1}{\varepsilon_{a_1}} - \frac{1}{\varepsilon_{a_2}} \right)}{\omega \pi \varepsilon_o \text{Im} \left( Z_1 - Z_2 \right)}
$$

(5.5)

where $a_1$ and $a_2$ are the disc radii, $\omega$ is the angular frequency, $\varepsilon_o$ is the permittivity of free space and $Z_1$ and $Z_2$ are the measured impedances of the discs.

Figure 5.4: Measurement of microstrip lines with and without the ferromagnetic NiFe/AlN multilayer composite: (a) Inductance per unit length vs frequency; and (b) capacitance per unit length vs frequency.
5.5 Discussion and outlook

This chapter proposed using un-patterned multilayer NiFe/AlN composite as magnetic core material to enhance the permeability and ferromagnetic resonance for use with on-chip inductors. Preliminary results show an enhancement in the frequency utilization by a factor of 7.1 as compared to previous work of homogeneous NiFe films of similar total thicknesses.

One of the issues with the fabrication process is the wafer bow due to stress of the deposited films. This results in wafer breakage even during handling also making lithography extremely difficult. This can be compensated by depositing stress compensating films on the wafer. Another issue is that the last SiO$_2$ film
5.5. Discussion and outlook

was deposited at high temperature which may have resulted in the loss of in-plane anisotropy.

The development work was done on a commercial 200 mm MEMS foundry, demonstrating that this process is not only feasible but able to be commercialized rapidly for real world applications. Future work include designing, fabricating and measuring 3-D integrated inductors with un-patterned multilayer NiFe/AlN composite as magnetic core material. This can be accomplished by utilizing copper through-silicon-vias (TSVs) to make the on-chip magnetic solenoids.

Figure 5.7: Relative permittivity of the ferromagnetic NiFe/AlN multilayer composite relative to SiO$_2$ line extracted from the measured S-parameters.
Chapter 6

Conclusions

This thesis has presented RF MEMS devices using monocrystalline silicon as the structural core material. MEMS tuneable capacitors based on the novel concept of moving sidewalls in 3-D micromachined coplanar waveguide transmission lines, enabled by completely integrating the MEMS actuators inside the ground layer are presented. Two different transmission line metallization schemes are presented, and different device designs are fabricated. The characterized devices offer the following advantages: multi-position digitally tuning; large tuning range; number of states independent on required transmission line length; low-loss 3-D micromachined transmission line; single-mask fabrication. Further, the characterized devices achieved high Q, high reliability, high linearity and high self-actuation robustness. Finally, a MEMS tuneable filter realization is shown which uses the presented novel tuneable capacitors as tuning elements.

Furthermore, this thesis presents novel concepts of RF MEMS ultra-wideband 10 to 18 GHz tuneable directional couplers whose coupling is tuned by mechanically changing the geometry of the 3-D micromachined coupled transmission lines. Two different tuning concepts involving tuning both the ground and the signal line coupling are investigated. Designs are fabricated and evaluated using the measured RF and actuator performance. The fabricated devices achieve a large tuneable-coupling ratio while maintaining an excellent impedance match and directivity over a wide bandwidth. In addition, the devices achieved high reliability showing no failure or fatigue even after being cycled one billion times.

Moreover, the thesis presents closed form analytical formals that have been derived for calculating the IIP3 of various multi-device RF MEMS circuits. The RF nonlinearity analysis for three different multi-device RF MEMS circuits i.e. digital MEMS varactor bank, MEMS switched capacitor bank and multi-step MEMS capacitor, was performed along with IIP3 measurements on the prototype of a novel multi-step MEMS capacitor. The scaling of the linearity with the circuit complexity was analysed. It was found that the IIP3 of a multi-device RF MEMS circuit can be significantly lower than the IIP3 of a single device, which requires
a careful analysis of the overall circuit IIP3 in each state rather than just for a single stage. Finally, design rules were proposed to improve the overall linearity of the circuit under investigation with the conclusion that for achieving the maximum IIP3 for a multi-device RF MEMS circuit, the circuit should be designed such that each stage has the same IIP3.

Finally, the thesis shows the utility of un-patterned multilayer NiFe/AlN composites as magnetic core materials for on-chip inductors. The presented approach increases the thickness of the ferromagnetic material without increasing its conductivity, by using a multilayer NiFe and AlN sandwich structure. This suppresses the induced currents very effectively and at the same time increases the ferromagnetic resonance. The NiFe/AlN composite enables an increase of the ferromagnetic resonance and thus an enhancement in the frequency utilization by a factor of 7.1 as compared to homogeneous NiFe films of similar total thicknesses.
Summary of Appended Papers


This paper presents a novel concept of RF MEMS tuneable capacitors based on the lateral displacement of the sidewalls of a 3-D micromachined coplanar transmission line. The tuning of a single device is achieved in multiple discrete and well-defined tuning steps by integrated multi-stage MEMS electrostatic actuators that are embedded inside the ground layer of the transmission line. The highest \( Q \) factor was determined as 88 at 40 GHz. The self-actuation pull-in was measured to be 41.5 and 47.8 dBM for mechanical spring constants of 5.8 and 27.7 N/m, respectively. IIP3 for all discrete device states is above the measurement-setup limit of 68.5 dBM for our 2.5-GHz IIP3 setup. For a mechanical spring design of 5.8 N/m, the actuation and release voltages were characterized as 30.7 and 21.15 V, respectively. Reliability characterization exceeded 1 billion cycles with no degradation in the pull-in/pull-out hysteresis behaviour being observed over these cycling tests.

Paper 2: *High-Directivity MEMS-Tunable Directional Couplers for 10-18-GHz Broadband Applications*

This paper reports on two novel concepts of area-efficient ultra-wideband MEMS reconfigurable coupled line directional couplers, whose coupling is tuned by mechanically changing the geometry of 3-D micromachined coupled transmission lines, utilizing integrated MEMS electrostatic actuators. Concept 1 is based on symmetrically changing the geometry of the ground coupling of each signal line, while Concept 2 is simultaneously varying both the ground coupling and the coupling between the two signal lines. For an implemented 3-6-dB prototype coupler based on Concept 1, the measured isolation is better than 16 dB, and the return loss and directivity are better than 10 dB. For an implemented 10-20-dB prototype coupler based on Concept 2, the measured isolation is better than 40 dB and the return loss is better than 15 dB for both states. The directivities for both states are better than 22 and 40 dB, respectively, over the whole frequency range. Reliability tests were conducted up to 1 billion cycles without device degradation.
Paper 3: Analysis of Linearity Deterioration in Multi-Device RF MEMS Circuits

The paper presents for the first time an RF nonlinearity analysis of complex multi-device RF MEMS circuits. The IIP3 of different RF MEMS multi-device tuneable-circuit concepts including digital MEMS varactor banks, MEMS switched capacitor banks, distributed MEMS phase shifters, MEMS impedance tuners and MEMS tuneable filters, is investigated. Closed-form analytical formulas for the IIP3 of MEMS multi-device circuit concepts are derived. A nonlinearity analysis, based on measured device parameters, is presented for exemplary circuits of the different concepts using a multi-device nonlinear electromechanical circuit model implemented in Agilent ADS. The results of the nonlinear electromechanical model are also compared to the calculated IIP3 using derived equations for the digital MEMS varactor bank and MEMS switched capacitor bank. Design rules are presented so that the mechanical parameters and thus the IIP3 of the individual device stages can be optimized to achieve a higher overall IIP3 for the whole circuit.

Paper 4: MEMS Reconfigurable Millimeter-Wave Surface for V-Band Rectangular-Waveguide Switch

This paper presents for the first time a novel concept of a microelectromechanical systems (MEMS) waveguide switch based on a reconfigurable surface, whose working principle is to block the wave propagation by short-circuiting the electrical field lines of the TE10 mode of a WR-12 rectangular waveguide. The reconfigurable surface is only 30 $\mu$m thick and consists of up to 1260 micromachined cantilevers and 660 contact points in the waveguide cross-section, which are moved simultaneously by integrated MEMS comb-drive actuators. Measurements of fabricated prototypes show that the devices are blocking wave propagation in the OFF-state with over 30 dB isolation for all designs, and allow for transmission of less than 0.65 dB insertion loss for the best design in the ON-state for 60-70 GHz.

Paper 5: Microwave MEMS Devices Designed for Process Robustness and Operational Reliability

This paper presents an overview on novel microwave microelectromechanical systems (MEMS) device concepts which are specifically designed for addressing some fundamental problems for reliable device operation and robustness to process parameter variation. The presented device concepts are targeted at eliminating their respective failure modes rather than reducing or controlling them. Novel concepts of MEMS phase shifters, tuneable microwave surfaces, reconfigurable leaky-wave antennas, multi-stable switches, and tuneable capacitors are presented. This paper summarizes the design, fabrication, and measurement of devices featuring the novel concepts, enhanced by new characterization data, and discusses them in the context of the conventional MEMS device design.
**Paper 6:** High-Aspect-Ratio Through Silicon Vias (TSVs) for High-Frequency Application Fabricated by Magnetic Assembly of Gold-Coated Nickel Wires

In this paper we demonstrate a novel manufacturing technology for high-aspect-ratio vertical interconnects for high frequency applications. This novel approach is based on magnetic self-assembly of pre-fabricated nickel wires that are subsequently insulated with a thermosetting polymer. The high frequency performance of the through silicon vias (TSVs) is enhanced by depositing a gold layer on the outer surface of the nickel wires and by reducing capacitive parasitics through a low-k polymer liner. For evaluation purposes, coplanar waveguides (CPW) with incorporated TSV interconnections were fabricated and characterized. The experimental results reveal a high bandwidth from DC to 86 GHz and an insertion loss of less than 0.53 dB per single TSV interconnection for frequencies up to 75 GHz.

**Paper 7:** Permeability Enhancement by Multilayer Ferromagnetic Composites for Magnetic-Core On-Chip Inductors

This paper reports for the first time about un-patterned ferromagnetic NiFe/AlN multilayer composites used as advanced magnetic core materials for on-chip inductances. The proposed composite structure reduces RF induced currents and thus pushes the permeability cut-off to 3.7 GHz, which is by a factor of 7.1 higher than for homogeneous NiFe layers of same thickness. To the best knowledge of the authors, we achieve the highest effective permeability of 28 at 1 GHz, highest ferromagnetic resonance frequency and highest inductance enhancement factor above 1 GHz ever reported for devices based on un-patterned on-chip NiFe magnetic cores.
Acknowledgments

Some of the most eventful and important years of my life have been spent at the MST Lab as a PhD student. There have been numerous ups and downs during that time. I would like to thank MST seniors, colleagues, friends and family who celebrated with me during good times and supported me during not so good times.

Firstly, I would like to thank my supervisor Joachim Oberhammer for giving me the opportunity to work with him and be his PhD student. I want to thank him for being always available for discussion no matter how busy he was and for extracting the most out of my abilities. None of this would have been possible without his help and vision. I would also like to thank the head of department Göran Stemme for providing an excellent research environment and being an excellent boss. A word of thanks also to Wouter, Hans, Frank, Niclas and Tommy for being supportive or critical as the situation demanded. Not to forget my co-author on most of my publication Mikael Sterner for his help in the cleanroom and his brilliant research input without which majority of the publications would not be possible.

I owe special thanks to Jan Åberg for helping with important measurements and to Thorbjörn Ebefors and Jessica Fredlund for the ferromagnetic materials project. I especially want to thank Alan Cheshire and Gabriel Rouillard for spending late hours with me in the cleanroom to optimize the Centura etch. All the staff at the cleanroom also deserves my gratitude especially Magnus, Reza, Roger and Sven.

Most important gratitude goes to my colleagues. Fredrik Forsberg, for teaching me much in physics, electronics, politics, finance, history, lamb bar-b-que just about everything. Gaspard, for being my oldest friend at work, for being my office mate and for having the same frequency of thought and opinion as me. Fritzi, for the parties, for the cakes, for always having a plan and most of all for being a good friend. Mikael Sterner, for having the calming effect that everything will work out whether at sea, driving on the roads of Pakistan or doing research at MST. Zargham, for having a smile no matter what the situation might be and for his fruitful discussions with me. Simon, for being the combination of Swiss and Japanese flavour, for his support and interesting analysis about stuff. Kristinn, for being a good friend, for being one of the boys despite being a senior and for his engineering approach to problem solving that helped me in my research. Andreas, for his advice which I value a lot whether it be on research or other issues and for his apple pie recipe. Valentin, for his curly hair, for our discussions and
Acknowledgments

exchange of views on all sorts of subjects and for beating me (only once) at arm wrestling which prompted me to get into shape. Martin, for his happy-go-lucky attitude, for the Cuban cigars and for his utter desire to help with finding you what you require on the web. Gabriel, for singing Backstreet’s Back with me :), for introducing me to ‘wraps’ and for providing great company both at work and outside. Nutapong, for his constant complaining and for our hostile conversations which both of us enjoyed. Fredrik Carlborg, for being nice and for discussions about what is happening back home. Hithesh, for providing the opportunity to speak in code words at the department and for our research discussions. Mikael Antelius, for his soft tone and for not discussing work on social occasions. Mikael Karlsson, for providing the means to go on the roof during summer, for feeling for me when being bulldozed by work and for the guitar. Floria, for smiling and being full of expression when telling a story. Nikolai, for his bike which enabled me to fit into my wedding suit, for his Russian plus Finnish style of partying (headache the next day) and for his trademark laugh. Niklas, for always stopping and saying ‘Hi’ and for our discussions on movies and TV series. Henrik, for his cool startup ideas, for my exposure to board games and for the parties at Karolinska. Staffan, for providing up to date information on gadgets and for his humour following our trip to Taiwan. Stephan Schröder, for being the occasional smoking partner and for discussing half marathon run like it was nothing. Stefan Braun, for introducing me to the German way. Björn Samel, for the company during football games. Chianty, for saying whatever is on her mind and inquiring about the weirdest of stuff. Laila, for being the only person in the Lab with correct English. Alexander, for singing Master of Puppets with me. Jonas, for his quick witted responses and for shaving his head which makes me feel good about my ever receding hairline. Carlos, for protecting the penguin on the dance floor and for adding the Spanish touch to the Lab. Farizah, for the car rides and for the not so research oriented discussions. Reza, for always asking how I was doing. Mina, for the pictures and the map. Erika, for putting up with us PhD students and for always providing answers to our naive questions. Kjell, for making the impossible possible with his machining and electronics skill. Mikael Bergqvist, for taking over the job of making impossible possible and for his collection of hunting items which I really admire. I would also like to thank Adit, Sergey, Xiaoxiang, Xuge and Ulrika.

I would also like to thank my friends Fahad, Sheraz, Suhaiib, Shahkar, Omer Khayam, Amir, Jawad and Marco for checking up on me regularly on how I was doing. I would also like to thank my friends in Sweden outside work Mazhar and Björn for being there when I needed them and for their company at the parties.

Last but not least I would like to thank my family who supported me through thick and thin, who has always been there for me, overjoyed at my happiness, shared my burden and loves me with all their hearts no matter what the situation. Thank you Ami Jan, Lala Gi, Lala Gul, Nasir Lala, Quaiser Lala, Ayla, Shahryar, Aminah, Sarah, Muhammad, Zaín, Hassan, Ali, Sophia, Maryam, Sabah, Marwah, Musa, Haider, Suleiman, Mairah, Safia Bhabi, Minhas Bhabi, Roshan Bhabi, Amna Bhabi and my wife Sara. Abu Jan, you would have been proud. Thank You all.
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