PhD Dissertation

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ELECTROSTATIC LOW ACTUATION VOLTAGE RF MEMS SWITCHES FOR TELECOMMUNICATIONS

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Abstract

The present study is inspired by the superior performance of electrostatic RF MEMS switches over the contemporary state-of-the-art solid-state devices and the potential applications in communications field. The prevalent high voltage actuation mode, limits the reliability and applications especially, in wireless communication, therefore the study focuses on the realization of electrostatic low actuation switches with main emphasis on the pull-in voltage and RF response. The actuation voltage optimization is achieved by analyzing the flexure design, beam topology, actuation electrodes and gap height, using analytical models validated by numerical simulations. The RF performance enhancement is done by incorporating the floating metal design, active beam-area reduction and minimization of the associated parasitics. The fabrication is based on surface-micromachining, metal-electroplating and standard IC processing steps. The main features of the fabricated capacitive shunt and ohmic contact switches are: the actuation voltage 3-15V, isolation $\geq 25$dB, and insertion loss $\leq 0.2$dB, at 1-25GHz, for three basic configurations. As a conclusion of the study, an innovative switch design called ‘Symmetric Toggle Switch’ is presented. The microtorsion-actuator based design has actuation voltages of 7-10V, with good RF performance from 8-20GHz. In addition to improved reliability in switching applications, the device can be configured as a MEMS varactor with higher capacitance range or as a tunable filter over a narrow bandwidth. The agreement of 20-60%, between the measurements and simulations is expected to improve further with fabrication process optimization.

Keywords: RF MEMS Switch, Ohmic contact switch, Symmetric Toggle Switch, micro-machining, torsion actuator, Residual stress.
dedicated to

H H S S B
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Chapter 1

Introduction

1.1 MEMS Technology

MEMS - Micro Electro Mechanical Systems also known as micro-system or micro-fabrication technology is ‘miniaturization engineering’, a multi-disciplinary approach to enable batch fabrication of three-dimensional mechanical structures, devices and systems - at least with one of the dimensions in microns or less. The emerging concept is based on the available micro-manufacturing options, material properties and the scaling laws referred to the application under investigation. As the name implies the micro establishes the dimensional scale, electro suggests either electricity or electronics or both and mechanical implies ‘moving’ device components with degrees of freedom in translation, rotation, tilt or a combination of the either. Over the last decade, however the MEMS concept has grown to encompass other micro and sub-micron devices - with or with out moving parts, which respond and measure micro or nano-level changes in physical quantities including thermal, magnetic, piezoelectric, optical and pressure variations.

The origin of MEMS is generally traced to R P Feyman’s hypothesis [1] on miniaturization of devices and systems to the extent till physical laws and material properties impose no limit. According to the size reduc-
tion criterion, two types of systems have been identified - the information storage, computing and atomic level manipulations which require no particular size and ‘machinery’ or mechanical systems that does. The recent advancement in information storage, retrieval and computing ushered by very large scale integration (VLSI) technology is an example of the first type of systems. The scaling down of the basic transistors or logic gates has resulted in unprecedented improvement in general device or system performance, reduced power consumption, increased functionality, packaging density and the economic realization. The more recent culmination of this miniaturization emphasis in micro-electronics is the development of manufacturing and measuring technologies to exploit the quantum mechanical nature of the electrons to realize still smaller computing systems - ushering the field of nano-electronics.

The logical extension of the miniaturization derive in the field of ‘micro-machines’ or micro-systems with moving parts has come to be known as MEMS. The technology is based on the state-of-the-art integrated circuit (IC) fabrication techniques and methodologies and hence exhibits many advantages indigenous to IC technology. A few of those include cost reduction through batch fabrication, device to device consistency from advanced lithography and etching techniques, and general performance enhancement from dimensional down scaling, leading to size and weight reduction. In addition, by using materials such as silicon and fabrication techniques compatible with IC technology, MEMS mechanical components can be made monolithically integrated with electronics, producing a complete smart system-on-a-chip that interacts with the physical world, performs electronic computations and communicates with other systems if necessary. These advantageous characteristics have positioned MEMS to be a winning technology in many application arenas, a few of which are given in Table 1.1.


<table>
<thead>
<tr>
<th>Application</th>
<th>Device</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inertial Measurements</td>
<td>Accelerometers, gyroscopes [3]</td>
</tr>
<tr>
<td>Pressure Measurements</td>
<td>Pressure sensors for Automotive, Medical and Industrial applications [4]</td>
</tr>
<tr>
<td>Optics and communication</td>
<td>Digital micro mirrors, optical switches and displays [8]</td>
</tr>
<tr>
<td>RF Communication</td>
<td>Switches, Inductors, Capacitors, Resonators and systems based on basic RF MEMS [9]</td>
</tr>
<tr>
<td>Others:</td>
<td>Micro relays, disk heads and sensors [10]</td>
</tr>
</tbody>
</table>

Table 1.1: MEMS Applications and Devices.

The applications of the MEMS technology in the radio frequency regime, mostly referred to as RF MEMS, are more recent as compared to the application areas mentioned above. The first MEM switch designed specifically for microwave applications was reported in 1990 [2]. At present, RF MEMS devices such as switches or relays, tunable capacitors or varactors, high Q inductors, resonators and filters, which constitute the ‘fundamental building blocks’ of radio frequency applications, are the most vigorously pursued MEMS devices by academic and industrial research communities. The obvious advantages are the reduction in power consumption, size, superior RF characteristics and compatibility with existing integrated circuit technology and systems. In general the outstanding performance of the RF MEMS is attributed to the physical gap, mostly air gap, between the ‘active’ elements and ‘lossy’ substrate, use of high conductivity metals and high integration compatibility with existing IC fabrication technologies. For example on-chip inductors with \( Q > 100 \) (at 15GHz) have been micro-machined with spirals suspended over a gap of 10 microns from the bottom substrate.
1.1.1 RF MEMS Switch

The paradigm device of the RF MEMS, the switch is one of the first and most studied in this field, since Peterson [11] demonstrated the use of a bulk micromachined cantilever as a switching element. The functioning principle of a RF MEMS switch resembles the “mechanical relay” but on size scale commensurate with semiconductor switches. Similar to RF semiconductor switches, a MEMS switch has two stable states. Switching between the states is achieved through the displacement of a freely movable structural member – essentially a beam or plate anchored at a single or multiple points. The movable member thickness (1 - 5µm), length (500 - 1500 µm) and material (metals e.g. Au, Al, Ni or semiconductor) depend on the switch configuration and available fabrication technology. The displacement is induced via micro-actuators for which various actuation mechanisms are exploited including electrostatic, electromagnetic, electro-thermal and piezoelectric [12]. The electrostatic actuation mode is preferred because of the extremely low power consumption and the easy integration with existing IC fabrication technologies. Also, to date, it is the electrostatic-type switches only which have been demonstrated in the range from 0.1 to 100 GHz [9] with high reliability and wafer scale manufacturing techniques. The electrostatically actuated metal beam based switches out-perform the other solid state switches such as FETs, HEMTs and PIN diodes [13], currently in use. Although, RF MEMS switches have a speed limitation as compared to the semiconductor counterparts, they exhibit very low insertion loss (-0.1dB at 40GHz) due the low resistivity of the metal and high isolation (- 40dB at 10GHz) due to 1 - 3 microns of physical separation between switching elements. The inevitable I-V non-linearities associated with semiconductor junctions in PIN and FET devices are also non-existent, except for the minor hysteresis in C-V in
shunt switches. This improves the distortion characteristics and power handling capabilities in the devices which are important consideration in systems having large number of switches and limited power resources. The RF MEMS switches exhibit no measurable harmonics or intermodulation and require negligible quiescent current consumption; typical power dissipation per cycle is less than 50nJ [14]. The monolithic batch fabrication of switches when used as components in an integrated RF system, improves the overall performance and makes it more cost effective as compared to the design approach with solid state switches.

1.2 The Context and Motivation

In the communication arena whether commercial, defense or space oriented, there is a continuous movement towards smaller, reliable and secure systems that have increased functionality and reduced power consumption. The additional requirements of ‘co-site’ receivers - transmitters and high degree of frequency agility place sever constraints on circuit power dissipation and electromagnetic compatibility. The current state-of-the-art circuit designs use a combination of gallium arsenide FETs, PIN diodes and/or varactors to realize switching, filtering, and tuning functions. The designed systems are characterized by high power consumption and manufacturing costs. In addition, 1 - 2dB loss per switching state and poor RF performance conflicts with the need to co-locate frequency agile communication functions and desired electromagnetic noise immunity. Use of mechanical switches such as coaxial and wave-guides make the overall system bulky and slow. RF MEMS technology addresses these shortcomings and offers the performance advantage of electro-mechanical components on size scales comparable with solid-state devices. In many cases, a single MEMS component can replace and out-perform an entire solid state circuit e.g. low
loss high isolation RF MEMS switches can eliminate the need for an amplifier stage in T/R modules of mobile handset resulting in considerable power saving and size reduction. These are the general motivations driving research and development in RF MEMS and RF MEMS switches in particular.

1.3 The Problem, Approach and Innovation

RF MEMS switches also have their share of problems. Table 2.4 summarizes the major shortcomings of RF MEMS switches, many of which are interrelated or a trade-off between other parameters. Actuation voltage, one of the main characteristics, is intrinsically correlated with the switching speed and power handling capability. In addition, low actuation voltage devices are susceptible to external vibrations, shocks and self-biasing by the RF signal on the transmission line. However, most of the recently developed RF MEMS switching devices are designed for low loss, high isolation applications that do not require very fast switching rates such as air borne or space communication. The actuation mechanism is electrostatic in nature and actuation voltage lie between 20 - 50 Volts. The high voltage actuation mode makes the devices impractical for wireless applications as the additional requirement for up-convertors increases the device/system size and offsets the monolithic integration advantage. In capacitive switches, high electric field (3-5 MV/cm) across the dielectric layer (80-100 nm) leads to charge injection (into the dielectric) which dominates the stiction mechanism and reduces the device reliability. The charge injection is exponential with applied voltage; a reduction of 6V results in a 10-fold increase in the life time of a MEMS switch.

Thus, the aim of the present work is to investigate the issues related to the design and fabrication of “Electrostatically actuated low voltage RF
MEMS switching devices”, suitable for communication applications in the range of 1 - 25GHz. The actuation voltage range of 3-15 V and optimum insertion loss and isolation (-0.1dB and -30dB respectively) is envisaged. In the whole study, the analytically calculated parameters are validated using available commercial simulators and finally compared with experimental results. An effort has been made to investigate the discrepancies and provide appropriate explanations.

The studies on low actuation RF MEMS switches is inspired by the on-going MEMS research in Microsystems Division, ITC-isrt, Povo, Italy. Originally, it was planned to explore low actuation MEMS switches for broad band frequency (1-30GHz) range. The RF response (isolation > -20dB) of capacitive shunt switches is better only above 8-10 GHz, accordingly, to cover the lower frequency range, both capacitive shunt and series ohmic contact type switches (DC-10GHz) have been designed and simulated. To demonstrate a simple application, two single pole double throw (SPDT) switches with resonance frequency at 5 and 8 GHz have also fabricated. However, the ohmic contact switches and SPDT could not be characterized. The inevitable deviation from the original course is due the fact that a new type of design (STS) has been introduced. Therefore, we focus on the capacitive devices only. To some extent the time frame of the study and unforseen process related problems also influenced the course of events in this work.

The main aspects of the study and the adopted methodologies are:

- **Actuation Voltage**: The actuation voltage is a function of the beam spring constant, air gap and actuation electrode area. The spring constant and actuation area combination optimization is used to design and fabricate switches with actuation voltage ranging from 3V to 15V.

- **Isolation and insertion loss**: The switch down-to-up state capaci-
itance ratio determines the RF response of a device. The surface roughness and extent of overlap limit the isolation in down-state. The issue has been addressed by introducing a floating metal layer and reduction in beam active area to optimize the isolation and insertion loss. In ohmic contact switches the insertion loss depends on the contact resistance. A contact flexure design to minimize the loss is also discussed.

• **Deformation and residual stress:** One of the major challenges in MEMS switch fabrication is to have beams or air bridges with no out-of-the-plane deformation. The deformation may arise from the residual stress in the beams or due to the electrostatic force induced bending. Instead of conventional uniform thickness beams, reinforced beams with optimized meander suspension springs have been utilized to limit the extent of deformation and ensuing change in switch parameters.

• **Reliability:** The switch reliability against the drift in actuation voltage and stiction or failure due to dielectric charging is increased by electrostatic actuation below 15 volts. To avoid device failure by external vibrations and shocks, the stiffness is optimized well above the critical acceleration.

• **Innovation:** In the meander based switch design approach power handling capability and reliability against the self-biasing and external shocks can be maximized to a limited extent by increasing the spring stiffness and hence the actuation voltage. Another approach is to incorporate an additional electrode to clamp the air bridge in up state when switch is on. This two air-bridge system increases the processing complexity and adds parasitic capacitances. Instead, we propose a new switch design based on micro-torsion actuators, placed sym-
metrically across the CPW transmission line. Additional electrodes (on the same plane), to clamp the beam in up-state make the device impervious to self biasing and vibrations. The use of micro-torsion springs improves the travel range. The devices can also be configured as MEMS varactor with a wide capacitance range for a given gap and voltage, not achievable with conventional MEMS varactor design. Another, outstanding feature of the device as a RF MEMS switch is its tunability over a wide frequency range.

1.4 Structure of the Thesis

The thesis is divided into seven chapters. The problem is defined in the first chapter. The second chapter presents the state of art and general motivation for research in RF MEMS area. The third chapter focuses on the mechanical design aspects while the fourth chapter deals with the electrical model and related parameters. The fifth chapter presents fabrication, measurements and design optimization of the meander switches. The sixth chapter describes the proposed new switch. It discusses the analytical model, simulations, fabrication issues and measurements. The last chapter provides the summary of the work. Further chapter-wise elaboration of the contents is given below.

The first chapter introduces the M-E-M Systems, starting from the origin, the advantages of MEMS devices and a mention of few concurrent applications. It is followed by the introduction to applications of MEMS technology in the radio frequency field, with emphasis on RF MEMS switching devices. The description outlines the working principle, the shortcomings in state-of-art devices, and the potential applications in the field of communication. This puts the chosen research topic in proper perspective which is described next, outlining the main aspects and the methodology adopted.
1.4. STRUCTURE OF THE THESIS

Chapter 2 further elaborates the RF MEMS switches and presents the state-of-art in this field. The switches are classified based on the contact mechanism or the implementation perspective. The two types of switches namely capacitive and ohmic in shunt or series circuit configuration are the most widely reported devices. The respective merits and shortcomings are briefly described with references. In this work we focus on the electrostatic actuation mode, the most preferred because of the low power consumption and compatibility with existing IC precessing techniques. Two basic RF components e.g. varactor or tuneable capacitor and inductor are also described briefly because of the fabrication process similarities and the importance in RF system design. The RF MEMS in general and switches in particular consist of an air bridge or cantilever realized by low temperature deposition of metals on a sacrificial resist layer. The fabrication process is a combination of surface micro-machining and processing steps selected from IC processing technology. The essentials of the processing steps are also mentioned in this chapter. Some of the reported application and problems associated with RF MEMS switching devices constitutes the last part of the chapter.

MEMS is an inter-disciplinary field which encompasses concepts from mechanical, electrical, electronics and micro-fabrication technology and engineering. Chapter 3 presents the electro-mechanical aspects of the MEMS switch design. The essential electro-mechanical design of a MEM switch consists of the study of an air-gap suspended beam (typically the motion is considered in one direction only), under actuation stimulus. In this chapter we present the general lumped-parameter models of surface micromachined components, including the micro-mechanical equations of motion, spring constants and electrostatic actuation. Efforts have been made to present analytical models whose parameters can be extracted from the layout of the micro-machined components. Analytical expressions are
CHAPTER 1. INTRODUCTION

1.4. STRUCTURE OF THE THESIS

given for spring constants specific to MEMS switches, the effect of residual stress and nonlinear effects. The support flexures for low spring constant structures are of particular interest in this work. A detailed treatment of serpentine flexures is presented, while other flexures are also mentioned. A comparison is presented between the analytically calculated and FEM simulated spring constants for serpentine springs, used to design the devices. The process and material parameters specific to the $ITC$ fabrication process are used to arrive at an optimized design. The actuation voltage is one of the most important switch parameters and is function of the spring constant and other geometrical parameters of the switch. Efforts have been made to describe the models used to derive the actuation voltage, its impact on power handling capabilities and the correlation to external excitation. The chapter also mentions some of the associated aspects such as voltage break down and dynamic response of the beam.

Chapter 4 presents the electro-magnetic aspects of the MEMS switch design. From the electro-magnetic design perspective, the device structure with a movable beam or plate, is treated as a lumped R L C model. The chapter discusses the dependence of electrical parameters on switch geometry, material properties and the fabrication process specifics, with examples from the design of switching devices considered in this thesis, which lead to the final design considerations. Since the dimensions of all the devices considered in the present work are much smaller than the wave length of the routed RF signals (30cm at 1GHz to 0.6cm at 50GHz), electromagnetic interaction of the switch structure with RF signal is negligible over the selected range of $1 - 30$GHz. However, the devices are implemented in standard coplanar wave guide (CPW, 50Ω) configuration, the most preferred connection medium for RF MEMS devices. Though CPW configuration was simulated using a commercial software, a brief introduction on CPW basics is provided in order to present a complete perspective
of the RF MEMS device.

In Chapter 5 we discuss the fabrication, measurement results and design optimization of the serpentine meander based switches. Capacitive shunt switch is taken as a case study. Examples of a SPDT based on shunt switches in CPW configuration and series ohmic contact switch are also discussed. The first design was based on a perforated thin beam with suspension meanders. The measurement results on the major parameters such as actuation voltage and RF response are compared with the simulated and analytical models. The deviation up to 30% in the pull-in voltage is discussed in terms of the process related parametric-variations. The RF response to a large extent is a function of the switch capacitance in on and off state. Beam deformation in the presence of residual stress gradient and electrostatic force induced bending are the main reasons for deviation between the measured and simulated results. The beam deformation before and after actuation is explained with the help of a model. The design optimization consists of the deformation alleviation using reinforced beam design and actuation electrode combination. The capacitance is a function of the surface properties of the underpass, bottom surface of the bridge and active overlap area of the underpass. The capacitance optimization is carried out by introducing an electrically floating metal layer on top of the underpass. Further optimization of the off-state to on-state capacitance ratio can be achieved by optimizing the bridge overlap to floating metal-area ratio. Though, for most of the devices the analytically calculated and simulated values are in excellent agreement, the measured results are off by nearly 30-40%. The switch design was optimized in view of the experimental results from the first trial fabrication run. However, in the second fabrication process, the metal deposition techniques to realize the air bridge has been changed, and are still in optimization phase at the time of writing the thesis. A discussion on this is presented in the chapter on
symmetric toggle switch, where similar kind of problems have occurred.

Chapter 6 describes the proposed new device - symmetric toggle switch (STS). The design overcomes the shortcomings of meander based two-bridge switches and metal-semiconductor composite beam design of cantilever based ohmic contact switches. The proposed analytical model which takes into consideration the device layout parameters, is validated against the simulation results and an excellent agreement is observed. The simulated RF response shows that the switch behaves as single LC tank with well defined resonance peak and the flexibility to change resonance frequency without affecting the actuation voltage. The salient features of the design are: low actuation voltage ($\leq 10V$), configuration adoptability to series ohmic contact switch without affecting other parameters, virtually no abrupt snap-down under electrostatic actuation, wide tuneable range by changing the lever dimension, and impervious to self biasing and external shocks. The potential applications including as a switching device are - as a varactor with very large tunable range, and in tunable filters with high isolation over a determined bandwidth. However, similar to the meander based switches, the fabricated devices show high deformation, accompanied by increased actuation voltage and lower isolation. An attempt has been made to explore the origin of the stress in processing steps used for fabricating the beams. The deformation in cantilever test structures is used to quantify the residual stress gradient. The simulated results agree well with experimental observations. The main fabrication steps which cause the stress are discussed in details with experimental results and suggestions.

Finally, the conclusions are drawn in Chapter 7. During the above mentioned research activities it was possible to obtain some original results of scientific interest, those are the object matter of the following publications and conference presentations.
Related Publications


Communicated Papers


1.4. STRUCTURE OF THE THESIS

CHAPTER 1. INTRODUCTION
Chapter 2

Radio Frequency MEMS

2.1 Introduction

This chapter provides a brief introduction to the basic RF MEMS components, in context with the existing RF systems. Major part of the chapter is devoted to RF MEMS switches, highlighting the working principle, different configurations, salient features of state of the art in this field, a few applications and associated problems. Because of the inherited technological similarities and importance in RF systems, MEMS varactor and inductors are also introduced in brief.

2.2 Basic RF MEMS

In comparison to other mature MEMS technology fields, the radio frequency MEMS are relatively new. The first MEMS switch designed specifically for microwave applications was reported in 1990 [2]. RF MEMS devices that serve as fundamental building blocks are: RF MEMS switches or relays, tunable capacitors or varactors, high Q inductors, resonators and filters, which can substitute the macro off-chip counter parts in existing microwave systems. The replacement criteria may differ from device to device, nevertheless RF functionality per unit volume, integration with
RFIC technology, lower power consumption and low cost, are the evident advantages. In general, from a technological perspective and magnitude of the displacement of movable beams or plates, the RF MEMS devices can be divided into four distinct areas.

- Switches, varactors and inductors that have been demonstrated from DC to 120 GHz. Except for micro-machined inductors, switches and varactors essentially consist of anchored beams or plates which move several microns when actuated.

- Micro-machined transmission lines, high Q resonators, filters and antennas that are suitable for 12–20 GHz. They are generally integrated on thin dielectric membranes or use bulk micro-machining of silicon, but are static and do not move [15].

- Thin Film Bulk Acoustic Resonators (FBAR) and filters that use acoustic vibrations in thin films and that have demonstrated excellent performance up to 3 GHz with very high $Q(>2000)$ [16].

- RF Micro-mechanical resonators and filters that use the mechanical vibrations of extremely small beams to achieve high $Q$ resonance at 0.01 – 200 MHz in vacuum. In this case the mechanical movements are of the order of tens of Angstroms [17].

### 2.2.1 RF MEMS Switch

High frequency switches are essential components in a variety of systems operating in the microwave regime e.g., for mobile phones, wireless local networks, and into millimeter wave regime e.g. for radar and satellite systems. In wireless communications, the switches are mostly used for RF signal routing selecting the right antenna, for switching between the transmit and receive paths or routing the signals to different blocks in
multi-band/multistandard phones. In radar systems, arrays or matrices of switches are used in phase shifters or time delay units, which form the key building blocks for phased array antennas. As previously mentioned the main performance characteristics of a switch are the low insertion loss in the on-state, high isolation in the off state, return loss in both the states, power consumption, power handling capability and linearity.

The state of the art radio frequency systems realize switching, by semiconductor switches such PIN diodes or FETs. Semiconductor switches pro-

<table>
<thead>
<tr>
<th>Actuation Mechanism</th>
<th>Voltage (V)</th>
<th>Current (mA)</th>
<th>Power (mW)</th>
<th>Size</th>
<th>Switching Time (µS)</th>
<th>Contact Force (µN)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electrostatic</td>
<td>20 – 80</td>
<td>0</td>
<td>0</td>
<td>small</td>
<td>1 – 120</td>
<td>50 – 1K</td>
</tr>
<tr>
<td>Thermal</td>
<td>3 – 5</td>
<td>5 – 200</td>
<td>0 – 200</td>
<td>Large</td>
<td>300 – 10K</td>
<td>500 – 4K</td>
</tr>
<tr>
<td>Magnetostatic</td>
<td>3 – 5</td>
<td>20 – 150</td>
<td>0 – 100</td>
<td>Med.</td>
<td>300 – 1K</td>
<td>50 – 200</td>
</tr>
<tr>
<td>Piezoelectric</td>
<td>3 – 5</td>
<td>0</td>
<td>0</td>
<td>Med.</td>
<td>50 – 500</td>
<td>50 – 200</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Movement</th>
<th>Contact Type</th>
<th>Frq. Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vertical</td>
<td>Metal to Metal</td>
<td>DC-60GHz</td>
</tr>
<tr>
<td>Lateral</td>
<td>Capacitive</td>
<td>10-120GHz</td>
</tr>
</tbody>
</table>

**Circuit Configuration**

- **series:** DC - 50 GHz with metal contact and low upstate capacitance
- 10 - 50 GHz with capacitive Contact and Low upstate capacitance
- **Shunt:** DC - 60 GHz with metal contact and low inductance to ground
- 10 - 200 GHz with capacitive contact and low inductance to ground

Table 2.1: RF MEMS Switch characteristics and configuration.
vide the desired performance in terms of switching speed (FETs an order of magnitude faster than PIN diodes) but present power consumption constraints (in particular for PIN diodes) and introduce significant loss (FETs in particular) becoming worse at higher frequencies. In many aspects, the RF MEMS switches resemble conventional mechanical RF switches/relays. The later employ electromagnetic actuation and are built using conventional mechanical manufacturing technologies. Due to the large power consumption and absence of IC-compatibility, the switches are not considered as a true alternative for the semiconductor switches, despite their excellent performance characteristics in terms of the good insertion loss, isolation, large power handling capability and linearity. There is clearly a need for an alternative technology and MEMS may be the way as elucidated in this chapter.

A MEM switch has two stable states just like semiconductor switches. Switching between the states is achieved through the mechanical displacement of a freely movable structural member - essentially a beam or plate anchored at a single or multiple points (Fig. 2.2). The displacement is induced via micro-actuators for which various actuation mechanisms exist including electrostatic, electromagnetic, electro-thermal and piezoelectric [12], as summarized in Table 2.1. The majority of the reported switches use electrostatic actuation. The advantages of electrostatic actuation mode are: extremely low power consumption; power is consumed only during switching, simple fabrication technology e.g. as compared to electromagnetic actuation, compact size, high degree of compatibility with standard IC fabrication processes and easy integration with coplanar waveguide (CPW) and micro-strip transmission lines.
2.2.2 MEMS and Semiconductor Switches

- Basic Difference

Insight into the basic difference in working principles of the semiconductor and MEMS switches is provided by Fig. 2.1, showing the devices in ‘on’ and ‘off’ states. The figure of merit (FOM), equivalent to the reciprocal of cutoff frequency [18], for a semiconductor can be expressed as:

\[
FOM = C \cdot R = \varepsilon_0 \varepsilon_r \rho
\]  

(2.1)

where \( C = \varepsilon_0 \varepsilon_r dx \, dy/dz \) and \( R = \rho \, dz/dx \, dy \), represent the capacitance and resistance of a semiconductor cube with dimensions dx, dy and dz respectively. As shown by above equation the fundamental limit of microwave switching behavior is largely determined by the dielectric constant of the material in the off state and by the conductivity in the on-state, independent of the dimensions which determine the power handling capability of the devices. Using the same criteria for a series MEMS switch with gap height \( g_0 \), contact film thickness \( t \), resistivity \( \rho \), area \( A \) and effective area factor \( a_e \), is given by

\[
FOM( \text{MEMS Series Switch} ) = CR = \varepsilon_0 \rho \frac{t}{g_0} a_e
\]  

(2.2)
### 2.2. BASIC RF MEMS

#### CHAPTER 2. RF MEMS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>RF MEMS</th>
<th>PIN</th>
<th>FET</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage (V)</td>
<td>10 – 80</td>
<td>±3 – 5</td>
<td>3 – 5</td>
</tr>
<tr>
<td>Current (mA)</td>
<td>0</td>
<td>3 – 20</td>
<td>0</td>
</tr>
<tr>
<td>Power consumption (mW)</td>
<td>0.05 – 0.1</td>
<td>5 – 100</td>
<td>0.05 – 0.1</td>
</tr>
<tr>
<td>Switching Time</td>
<td>1 – 300µS</td>
<td>1 – 100nS</td>
<td>1 – 100nS</td>
</tr>
<tr>
<td>$C_{up}$ (Series)(fF)</td>
<td>1 – 6</td>
<td>40 – 80</td>
<td>70 – 140</td>
</tr>
<tr>
<td>$Rs$ (Series)( Ω)</td>
<td>0.5 – 2</td>
<td>2 – 4</td>
<td>4 – 6</td>
</tr>
<tr>
<td>Capacitance Ratio</td>
<td>50 – 500</td>
<td>10</td>
<td>NA</td>
</tr>
<tr>
<td>Cutoff Frequency (THz)</td>
<td>20 – 80</td>
<td>1 – 4</td>
<td>0.5 – 2</td>
</tr>
<tr>
<td>Isolation (1-10 GHz)</td>
<td>Very High</td>
<td>High</td>
<td>Medium</td>
</tr>
<tr>
<td>Isolation (10-40 GHz)</td>
<td>Very High</td>
<td>Medium</td>
<td>Low</td>
</tr>
<tr>
<td>Isolation (1-10 GHz)</td>
<td>High</td>
<td>Medium</td>
<td>None</td>
</tr>
<tr>
<td>Loss (1-100 GHz) dB</td>
<td>0.05 - 0.2</td>
<td>0.3 - 1.2</td>
<td>0.4 - 2.5</td>
</tr>
<tr>
<td>Power Handling (W)</td>
<td>&lt; 1</td>
<td>&lt; 10</td>
<td>&lt; 10</td>
</tr>
<tr>
<td>Third order intercept point (dBm)</td>
<td>+66-80</td>
<td>+27 to -45</td>
<td>+27 to -45</td>
</tr>
</tbody>
</table>

Table 2.2: Performance comparison of GaAs FETs, PIN diodes and RF MEMS.

Eqn.2.2 further demonstrates that if the switch physically moves when changing state from the ‘off-state’ to ‘on-state’, an “engineer-able” degree of freedom can be introduced into the device. This additional degree of freedom and absence of the semi-conducting junctions make MEMS the devices with superior performance as summarized in Table 2.2 [19].

#### 2.2.3 Contact and Circuit Configuration

From a switch contact perspective, there are two types of switches based on MEMS technology: metal contacting or ohmic [20] and capacitive coupling [21]. The metal contacting switches use (Fig.2.3) metal-to-metal direct contact to achieve an ohmic contact between two electrodes. This ohmic contact characteristic allows the device to be suitable for low frequency applications including dc, as well as moderate to high frequency
CHAPTER 2. RF MEMS

2.2. BASIC RF MEMS

Figure 2.2: RF MEMS switch working principle. (a) Top view of the electrostatically actuated capacitive shunt device broad side implementation in CPW. (b) Device cross section along line $A' - A''$. Up - state allows the signal to pass through, while down state couples the input to ground.

applications. The trade-off lies in its contact lifetime, which is a common shortcoming of all metal contacting mechanical switches. Nevertheless, a typical lifetime of $10^8$ cycles is already adequate for some applications [22]. The capacitive coupling switch as illustrated in Fig. 2.4, has a thin dielectric film and air gap between the metallic contact surfaces and the part of the transmission line under the beam called underpass or the active overlap area. The air gap is electro-mechanically adjusted to achieve a capacitance change between the ‘up’ and ‘down’ state. The capacitance ratio (down to up state) is the key parameter for such a device; a high capacitance ratio is always desirable. Because of the coupling nature, these switches are not suitable for low frequency applications. However ‘contact life time’ is typically a smaller issue compared with that in a metal contacting switch.

2.2.3.1 Series and Shunt Implementation

From the application perspective, the MEMS switches are further classified into another two categories i.e. serial and shunt switches. Although
both ohmic and capacitive coupling switches can be used either as a serial or a shunt switch, generally ohmic switches are used in serial mode, while capacitive coupling switches are preferred as shunt switches. The advantage and disadvantage comparison between the switches is primarily within the RF circuit design rather than MEMS component itself. To further indicate the positioning of the armature relative to the RF signal line, the switching device can either be configured as broadside or inline [9]. In an inline switching device, the armature is an integral part of the transmission line (Fig.2.5). For a broadside switching device the long side of the armature is positioned perpendicular to the signal line (Fig.2.4). In addition to the above mentioned basic configurations which lead to eight RF MEMS switching configurations, a multitude of actuation mechanisms can be implemented leading to tens of different configurations. Two types which stand out because of their compatibility with IC fabrication process and easy implementation in CPW or micro-strip line configurations are: electrostatically actuated capacitive shunt and ohmic series relay or switch. The electrostatic and mechanical modeling, design and fabrication issues related to electrostatically actuated capacitive shunt switches are discussed in details in the following chapters. Except for the contact part, the basic structure of the ohmic series switch is similar to capacitive shunt switch.
The meander based switches discussed in Chapter 5 can be easily modified to series configuration.

2.2.3.2 Capacitive Shunt and Series Switch

A basic RF MEMS capacitive shunt switch consists of a movable metal bridge, mechanically anchored and electrically connected to ground of the coplanar waveguide (CPW). In ‘switch’ configuration the dc control voltage and the RF - signal are superimposed (Fig. 2.2) and applied to the signal line. In addition to the classical bridge approach shown in Figs.2.2 and 2.5, other variations of the support beams used to lower the actuation signal are: fixed-fixed flexure, crab-leg flexure, folded flexure and serpentine flexure [19], and are discussed in Chapter 3. To first order the switch can be modeled as a capacitor between the metal bridge (RF ground) and the dielectric coated portion of the signal line under the bridge. In the switch
Figure 2.5: Electrostatically actuated capacitive series switch (inline) implemented in CPW configuration. (a) top view (b) side view showing the two states of the switch.

‘on’ state the bridge is up, resulting in a small capacitance (10-100fF) and negligible effect on the line impedance. Signal attenuation in the up state is termed as the ‘switch on-state insertion loss’. By applying a dc bias, the bridge is pulled down on to the dielectric layer placed locally on top of the signal line. The switch capacitance becomes high (1-10pF); causing an RF short to the ground and the switch is in ‘off’ state. A high down-state capacitance and a low up-state capacitance implies high isolation (in ‘off’ state) and a low loss (in ‘on’ state). For a capacitive series switch (Fig.2.5), the operation is complementary to the shunt switch. When the bridge is up, the switch is in off-state. Pulling the switch down, shorts (capacitive coupling) the interrupted signal line and the switch becomes ‘on’. In contrast, broadside ohmic relays (Fig. 2.3) and capacitive switches (Fig.2.4) are quite compact, which make them more suited for applications at high frequencies, into millimeter wave regime.
2.2.3.3 Ohmic Series Switch

In general, the ohmic contact switches have the actuation bias separated from the RF signal, resulting in a three terminal device called ‘semi-relay’ or ‘semi-switch’ or simply ohmic switch. In the series switch shown in Fig.2.3, the signal line is interrupted. The dielectric armature has localized metallic portions for the switching contacts and the actuation electrodes. The armature consists of a cantilever, but can be a doubly supported structure (like a bridge) with various types of support beams [19] to lower the dc actuation signal magnitude. In absence of actuation bias, the armature is up and the RF input is only weakly coupled to the output via a small capacitance. A sufficiently high dc bias pulls the armature down, closing the gap in the signal line. The relay is in ‘on’ state. To keep the insertion loss small the contact resistance of the electrical contacts should be as small as possible.

2.2.3.4 Three Terminal Implementation

The schematic of a three terminal shunt ‘semi-switch’ or ‘semi-relay’ is shown in Fig.2.4. Two pull down electrodes placed symmetrically on either side of the transmission line contact area, introduce an additional degree of freedom as the actuation and contact electrodes are not the same. The device does not require biasing or decoupling elements; simplifying the design and fabrication process. Also, the actuation voltage can be optimized, independent of the bridge-down state capacitance. The structure can either be implemented with capacitive [23] or ohmic contacts [24]. Both capacitive shunt and series contact type RF MEMS devices discussed in this thesis are based on three terminal configuration except the symmetric toggle switch (STS), a four terminal device. To maintain the nomenclature consistency with literature, henceforth the devices are referred to as
2.2.4 Devices based on Alternative Actuation Modes

Only a few RF MEMS switching devices have been reported which employ modes of actuation other than electrostatic. The electro-thermally actuated ohmic relays [27] are constructed using stress controlled dielectric membranes with patterned metallic contacts. The structure allows the construction of resistive switches useful up to millimeter wave applications with low actuation voltages. The Microlab [28] developed DC-6GHz latching MEMS micro-relay (SPDT), is an ohmic series relay, actuated electromagnetically. The advantage of both electrothermal and electromagnetic actuation is, the lower actuation voltage that can be achieved e.g. 5V. However, a major drawback is the large power consumption except when latching mechanisms are used [28]. Electro-thermal actuation, in addition displays a slow response typically hundreds of seconds as shown in Table 2.1.
2.2.5 RF MEMS Tunable Capacitors

A tunable capacitor also referred to as ‘varactor’ is a capacitor of which the capacitance can be tuned or varied by electrical means, e.g. by a dc (tuning) voltage which makes the capacitance, voltage dependent, \( C = C(V) \). The RF tunable capacitors find applications in tunable matching networks, tunable filters, phase shifters, and as frequency controlling elements for instance in LC tank of a low noise VCO. Semiconductor on-chip varactor diodes or MOS capacitors, suffer from excessive series resistance and non-linearity [29]. RF MEMS varactors on the other hand use highly conducting thick metal layers, with air as dielectric, thus offering substantial improvement over conventional on-chip varactor diodes in terms of power loss. In addition, the RF MEMS capacitors have excellent linearity, wide tuning range and ability to separate the control circuitry from the signal circuit, which greatly simplifies the overall design.

Tuning of the capacitance can be achieved by three fundamental ways: (1) by tuning the dielectric constant, called \( \varepsilon_T \) or “K-tuning”, (2) by tuning the gap spacing called ”gap tuning” and (3) by tuning the area called ”area tuning”. The later two methods define true MEMS varactors. The dielectric based tuning capacitors employ ferroelectric thin films like Barium Strontium Titanate - BaSrTiO3 (BST) or Strontium Titanate - SrTiO3 (STO), which have an electric field tunable dielectric constant. The tuning capacitors based on this principle have the advantage of being rugged as there are no moving parts. The tuning ratios and quality factors, though limited, are acceptable for many applications [30].

The principle of gap-tuning MEMS capacitor controlled by electrostatic means is similar to RF MEMS switches and is shown in Fig. 2.6(a). The top plate is suspended with a support beam of spring constant \( k \), while the bottom plate is fixed. An applied dc bias reduces the air gap, thus
increasing the capacitance. The down capacitance is determined by the thin dielectric film covering the bottom plate and the overlap area. The suspension as well as the plates are fabricated from metal to obtain a low parasitic series resistance and high Q. In a polysilicon implementation for the fixed plate and polysilicon/gold for the suspensions and moveable plate with a nominal capacitance of 2.05pF, a Q factor of 20 at 1 GHz and tuning range near the theoretical 50% at 4V has been reported in [25].

Recently, MEMS capacitor with tuning ratio (ratio between the maximum and minimum achievable capacitance) of 17 and Q as high as 500 have been reported [31]. The area tuning capacitor approach is based on a comb like structure. The movement of one of the combs with respect to a fixed one relies on the electrostatic force between the two. The overlap area is thus a function of the applied dc bias. Unlike the tuning gap capacitor approach the tuning ratio of the comb structure is (to first order) not limited by the pull-in instability. A tuning ratio of 100% at 5V, Q-factor close to 100 (at 400MHz) and a self resonance frequency (SRF) as high as 5GHz for a device with base capacitance of around 3.3pF has been reported by Yao et al [32].

2.2.6 MEMS Inductors

MEMS inductor, another out-standing example of RF MEMS, needs a mention, because of the superior performance and ubiquitous presence in RF communication design. The stringent phase noise requirements in applications like voltage controlled oscillators (VCO) need inductors with quality factor \( Q > 30 \). On-chip inductors implemented in CMOS or bipolar technologies having \( Q < 10 \), make off-chip, discrete inductors the only viable choice. The key parameters that characterize the performance of inductors are the quality factor \( Q \), the inductance \( L \) and self-resonant frequency (SRF) at which the device transforms from the inductive to capacitive char-
MEMS technology improves the on-chip inductor performance by etching away the lossy substrate from underneath the inductor spiral, resulting in a membrane-supported inductor as shown in Fig.2.6(b). The reduced substrate losses and reduced capacitive coupling to the substrate, lead to higher Q and increased SRF. Another approach to build floating or levitated ‘on-chip’ inductors using an IC compatible process is to pattern the spiral in thick electroplated metals like Cu, Al or Au over a sacrificial dielectric or photoresist layer. Inductors ranging from 1.5nH to 18nH with Q from 30-80 at 2GHz have been fabricated and reported [26]. High Q inductors have also been fabricated using low loss, high resistivity substrates such as Si or Alumina without the need to resort to levitated or out of the plane configurations with measured Q values up to 107 at 11 GHz [33].
2.2.7 General Fabrication Process and Materials

The fabrication process design criteria and the choice of materials include the compatibility with standard IC fabrication processes and all the critical switch parameter specifications. In general, RF MEMS switches are fabricated using a combination of surface micro-machining and a set of processing steps selected from standard integrated circuit manufacturing technologies. Excluding the packaging, most of the reported RF MEMS switches are realized using five mask levels with a process flow sequence, which may vary for different implementations. While selection of the wafer, overall thermal budget and material etching and deposition techniques are the general process compatibility issues, contact and structural material considerations determine the switch parameters including contact resistance, metal sticking behavior, life time and environmental and packaging compatibility. In spite of the diverse process designs and implementations, the basic steps are similar, as summarized below and shown in Fig.2.7.

1. For RF MEMS very often low-loss, high resistivity substrate is the starting material. Thermally oxidized high resistivity silicon ( > $3k \, \Omega \, cm$) [23], [14] - [36] is generally preferred, but GaAs [24] or glass e.g. quartz or AF45 [37] have also been used.

2. The transmission or interconnecting lines are realized by depositing and patterning a thick (3 – 5µm) highly conducting layer e.g. Au, Cu or Al, [14] - [37]. Low temperature deposition techniques such as sputtering or electroplating are generally preferred for depositing the metal layers. In three terminal devices the actuation electrodes and biasing resistors are patterned in polysilicon. The capacitive contact area under the bridge may have one or more metal layers (e.g. Ti-Al-1%Si-TiN) in order to provide low resistivity path with smooth surface [38].
3. A thin $(0.1 - 0.3 \mu m)$ dielectric layer is deposited and patterned for the capacitive switch contacts and electric isolation of the actuation electrodes. Low temperature (e.g. $< 350^\circ C$) processing is needed to avoid the adverse effects of high temperature on the bottom metal layers. Mostly PECVD silicon nitride [14]-[40] and PECVD oxide are preferred, however use of high dielectric constant materials like anodized tantalum oxide [41] or sputtered strontium titanate oxide [37] have also been demonstrated.

4. The next step consists of sacrificial layer deposition and patterning. Polymers like positive photo resist [14]-[36] or polyamide [23], [37], $2 - 4 \mu m$ in thickness are spin-coated and patterned. The sacrificial layer thickness determines the ‘gap’ between the bridge and the capacitive contact.

5. The freely moving structural member is defined by depositing and patterning a $1 - 2 \mu m$ thick metal layer on the sacrificial polymer deposited in step 4. High conductivity metals, thermally stable and with low fatigue are preferred. In current designs the metals are aluminum alloys, gold or nickel [14]-[36].

6. The final step consists of the removal of sacrificial layer using a proper release process to avoid stiction [10] and is carried out at low temperature e.g. an isotropic dry etch in oxygen plasma in case photoresist or polyimide is the sacrificial layer.

The above-mentioned steps are the minimum number of steps needed to fabricate an electrostatically actuated capacitive switch. For ohmic switching devices, dielectric over the contact is replaced by a contact metal e.g. gold [20]. In capacitive devices, a thin $(0.1 - 0.3 \mu m)$ and smooth bottom electrode (underpass) is often added locally at the switch contacts to ensure good down capacitance requiring intimate contact with minimal air
2.2.8 Application Areas of RF MEMS

RF MEMS devices, with virtually no mass, insensitive to acceleration, consuming no DC power, having cutoff frequency 30-50 times higher than GaAs devices, outstanding isolation and insertion loss at microwave frequencies and low manufacturing cost are the prime candidates for replacing existing semiconductor or mechanical counter parts, in defense or high-value commercial applications (satellite systems, base stations etc.). However their use in low cost commercial systems is still under investigation due to relatively higher manufacturing cost. The application areas are summarized in Table 2.3. A few of the important applications [19], [42] are briefly described as follows.

<table>
<thead>
<tr>
<th>Area</th>
<th>System</th>
<th>RF Device</th>
</tr>
</thead>
<tbody>
<tr>
<td>Phased Array</td>
<td>Communication and Radar Systems</td>
<td>Switch</td>
</tr>
<tr>
<td></td>
<td>(ground, space, airborne, missile)</td>
<td></td>
</tr>
<tr>
<td>Switching and</td>
<td>Wireless Communication (portable,</td>
<td>switches</td>
</tr>
<tr>
<td>Reconfigurable</td>
<td>base station)</td>
<td></td>
</tr>
<tr>
<td>Networks</td>
<td>Satellite (Communication and Radar)</td>
<td></td>
</tr>
<tr>
<td>Low power</td>
<td>Wireless</td>
<td>Varactors &amp;</td>
</tr>
<tr>
<td>Oscillators</td>
<td>Satellite</td>
<td>Inductors</td>
</tr>
<tr>
<td>and Amplifiers</td>
<td>Airborne</td>
<td></td>
</tr>
</tbody>
</table>

Table 2.3: Application areas of MEMS switches, varactors and high-Q inductors.
2.2.8.1 Switching Networks

Switching networks are used in virtually every communication system and include SPNT switches for filter or amplifier selection, NxN switching matrices and general SPDT and DPDT routing switches. Switch matrices in satellite systems are built using coaxial switches while base station systems are implemented using PIN diodes except for the power amplifiers. PIN diodes require additional amplifier stages to compensate for losses introduced by the switching network. Coaxial switches have excellent overall performance, but are heavy and costly. A typical system needs 100-300 switches. RF MEMS switches with Cu = 2 - 4 fF can easily meet the isolation requirements of NxN switch matrices, but show higher insertion loss (0.5dB for 4x4 to 1.0 dB for 8x8) due to their 1 ohm contact resistance. However they result in much smaller and lightweight systems.

2.2.8.2 Portable Wireless Systems

In order to integrate the whole front end design of a wireless telephone system on a chip, the main bottle-neck are not the switches, but high Q filters and diplexers and also crystal references [15]. Filters are currently being addressed using FBAR technology. Using RF MEMS devices (switches, varactors and inductors) it may be possible to eliminate the off chip inductors in the oscillator circuit or integrate a tunable filter on the silicon chip. However, at present the cost (10-50 times higher) and the reliability as compared to the semiconductor devices prohibit low cost commercial use of RF MEMS components.

2.2.8.3 Phased Arrays

RF MEMS switches are best suited to communication systems e.g. phased arrays that use large number of switching devices. The average loss of the
2.3. **PROBLEM AREAS**

| (1) Low speed | Switching time $2 - 50\mu S$ | A tradeoff with actuation voltage |
| (2) Power Handling | Low RF Power handling capacity (20 - 50 mW) | Dependence on actuation voltage |
| (3) High Voltage drives | Actuation voltage 20-80 V | A tradeoff with switching speed and long term reliability |
| (4) Reliability | $0.1 - 10^8$ cycles | Function of RF power, contact material & packaging |
| (5) Packaging | Hermetic sealing for better reliability | Expensive, still a research & development issue |
| (6) Cost | 10 - 20 times higher than semiconductor devices | Up-convertors, hermetic sealing are the main issues |

Table 2.4: Problems associated with RF MEMS switching Devices.

state-of-the-art 3-bit MEMS phase shifter shows an improvement of more than 3dB over comparable GaAs FET devices [42]. This translates to 6 to 8 dB improvement in radar or a two-way telecommunication system. The improvement is quite significant at Ka, V or W-bands systems. Therefore, one can eliminate a few amplifier stages in T/R chain resulting in 20-100mW DC power reduction per element at X to V-band frequencies.

**2.3 Problem Areas**

Table 2.4 summarizes the major shortcomings of RF MEMS switches, many of which are interrelated or a trade off with other parameters. As shown in the Table 2.4, actuation voltage for most electrostatic devices is greater than 15 volts. The switches need up-converters in order to be integrated with other semiconductor-based systems with standard voltage sources. Also, switches with higher actuation voltage tend to have problems like stiction and low reliability. The low actuation voltage design, the main theme of the thesis is discussed in details in the following chapters. We also
present a novel design approach for higher power applications. Packaging, power handling and reliability are the other important areas mentioned briefly.

### 2.3.1 Device Packaging

RF MEMS switch packaging is crucial not only for proper device functioning, but also for the fast dissemination of the technology. The operational characteristics of the devices are severely affected by the presence of water vapors, oxygen, hydrocarbons and other contaminants. The contact and pulling up forces ($10^{-500} \mu N$) are too small to puncture through contamination layer that may deposit on the contacts in the case of ohmic contact switches or overcome the adhesive forces of water molecules in case of capacitive switches. The hermitic sealing which can solve the above problems is expensive and finally will decide the cost and applications of the RF MEMS.

### 2.3.2 Power Handling and Reliability

The number of on-off cycles of a MEMS device is considered to be one of the major reliability criteria. As shown in Table 2.4 devices have been reported with life time up to a few billions cycles, under low power conditions (0.5-5mW). The failure mechanisms depend on the RF Power used and can be due to thermal stress, dielectric break down, self-actuation or critical current density issues. Dielectric charging, pitting, hardening, dielectric formation typically undermine the reliability of devices.

Capacitive switches with their large contact area can handle more RF power than metal-to-metal contact switches and therefore are preferred for applications requiring 30-300 mW of RF power. However, most capacitive switches operate above 8 GHz due to their relatively small capacitance (2-
5 pF) and DC-contact switches are the only choice at 0.1 to 8 GHz. The lifetime of metal-to-metal contact switches is highly dependent on the RF power. Most of the electrostatic switches result in only 10-1000 million cycle at 10-100 mW of RF power. It is primarily due the small contact and restoring forces (50 – 100µN) that can be achieved using electrostatic actuation. In comparison, thermally actuated switches can handle more than 6W of RF power, but need large amount of DC power (50-330mW) and are not suitable for most applications.
Chapter 3

Mechanical Design Aspects of MEMS Switches

3.1 Introduction

Most electrostatically actuated RF MEMS switches are based on metallic, semiconductor or a combination of the metal-semiconductor beams or plates, either fixed at both ends (bridge configuration) or with one fixed end in cantilever configuration. The electrostatic deflection in a direction perpendicular to the beam plane is used to realize the functional behavior of the devices. In this chapter we discuss the mechanical design aspects of the MEMS switches. We review the equations of motion for a plate fixed at four corners, spring constants for selected mechanical flexures, effects of residual stress and squeeze film damping. The major part of this chapter is devoted to the analysis of the static spring constant for serpentine suspension springs. The calculated spring constant is compared with the FEM-simulations. Serpentine flexures form the bases of low voltage switch design, discussed in the Chapter 5.
3.2 Mechanical Equations of Motion

The mechanical structures can be divided into discrete elements that are modeled using rigid-body dynamics. Finite-element analysis is used to determine the modes that are within the bandwidth of the feedback and external forces. Some structural elements can be modeled simply as rigid body mass, while other models may include the effects of bending, torsion, axial and shear forces. The approach is discussed in following sections where we consider the central capacitive or ohmic contact portion of the bridge as a rigid body and suspension springs, torsion beams and leverage elements, in the case of symmetric toggle switch, are elements subjected to bending and torsion, with appropriate approximations.

A mechanical system with \( n \) degrees of freedom can be described in terms of \( n \) generalized coordinates, \( q_1, q_2, \ldots, q_n \) and time \( t \). A general method of determining the equation of motion involves use of Lagrange’s equation [43].

\[
\frac{d}{dt} \left( \frac{\partial L}{\partial \dot{q}_i} \right) - \left( \frac{\partial L}{\partial q_i} \right) = Q_{nc,i} \ ; \quad i = 1, \ldots, n \tag{3.1}
\]

where \( L = T - V \) is the Lagrangian operator, \( T \) is the total kinetic energy of the system and \( V \) is the total potential energy of the system arising because of the conservative forces. Non-conservative forces such as dissipative forces, are lumped in the terms \( Q_{nc,i} \). If only viscous damping terms (damping proportional to velocity) are present then the Lagrange’s equation can be written as

\[
\frac{d}{dt} \left( \frac{\partial L}{\partial \dot{q}_i} \right) - \left( \frac{\partial L}{\partial q_i} \right) + \left( \frac{\partial F}{\partial q_i} \right) = Q_{ext,i} \ ; \quad i = 1, \ldots, n \tag{3.2}
\]

where \( F \) is the Raleigh dissipation function and \( Q_{ext,i} \) is an external generalized force associated with the coordinate \( q_i \). In general the kinetic energy,
potential energy and the dissipation function have the forms:

\[ T = \frac{1}{2} \sum_{i=1}^{n} \sum_{i=1}^{n} m_{ij} \dot{q}_i \dot{q}_j \]  
\[ V = \frac{1}{2} \sum_{i=1}^{n} \sum_{i=1}^{n} k_{ij} q_i q_j \]  
\[ F = \frac{1}{2} \sum_{i=1}^{n} \sum_{i=1}^{n} B_{ij} \dot{q}_i \dot{q}_j \]  

where \( m_{ij}, k_{ij} \) and \( B_{ij} \) are the inertia, stiffness and damping coefficients respectively.

In this general discussion we apply Lagrange’s equation to a rigid rectangular plate suspended by four springs located at the plate’s corners shown by Fig.3.1. The Cartesian coordinates \( x, y, z \) and three angles of rotation \( \theta, \varphi \) and \( \psi \) are chosen to be the generalized coordinates with the plate center as the origin. However, in specific cases such as bridges and cantilevers, the \( z \)-axis origin is chosen to be the substrate below anchor points. The variable \( \Delta z \) in such a case represents the vertical displacement of the plate.
from its rest (zero mechanical potential) position. Potential energy stored in the springs is determined by the contributions of each spring. Making small angle approximations, we find

\[
V = 2(k_xx^2 + k_yy^2 + k_zz^2 + k_zL_{ky}^2\theta^2 + k_zL_{kx}\phi^2 + k_yL_{kx}\psi^2)
\]

(3.6)

where \(k_x, k_y\) and \(k_z\) are the spring constants in the \(x\), \(y\) and \(z\) directions respectively. The dimensions \(L_{kx}\) and \(L_{ky}\) are the distances along the \(x\) and \(y\) axis from the centroid of the plate to the springs, the springs in Fig. 3.1 are located at \(L_{kx} = L_x/2\) and \(L_{ky} = L_y/2\). The spring force is assumed to vary linearly with the displacement, however nonlinear spring forces can be modeled by substituting stiffness coefficients that are functions of the position into Eqn. 3.6. Assuming massless springs the kinetic energy \(T\) is

\[
T = \frac{1}{2}(m\dot{x}^2 + m\dot{y}^2 + m\dot{z}^2 + I_\theta\dot{\theta}^2 + I_\phi\dot{\phi}^2 + I_\psi\dot{\psi}^2)
\]

(3.7)

where \(m\) is the plate mass and the mass moments of inertia of the plate are:

\[
I_\theta = \frac{m}{12}L_y^2
\]

(3.8)

\[
I_\phi = \frac{m}{12}L_x^2
\]

(3.9)

\[
I_\psi = \frac{m}{12}(L_x^2 + L_y^2)
\]

(3.10)

The viscous damping of the plate can be expressed by the damping function,

\[
F = \frac{1}{2}(B_x\dot{x}^2 + B_y\dot{y}^2 + B_z\dot{z}^2 + B_\theta\dot{\theta}^2 + B_\phi\dot{\phi}^2 + B_\psi\dot{\psi}^2)
\]

(3.11)

where \(B_x, B_y, B_z, B_\theta, B_\phi, B_\psi\), are the damping coefficients of the six modes. The expression for kinetic energy, potential energy and the dissipation function of the mass-spring-damper system are substituted into Equation (3.1) and then solved for each of the six coordinates resulting in
the following equations of motion:

\[ F_x = m \dddot{x} + B_x \dot{x} + k_x x \]  
\[ F_y = m \dddot{y} + B_y \dot{y} + k_y y \]  
\[ F_z = m \dddot{z} + B_z \dot{z} + k_z z \]  
\[ \tau_\theta = I_\theta \dddot{\theta} + B_\theta \dot{\theta} + k_z L_y^2 \theta \]  
\[ \tau_\phi = I_\phi \dddot{\phi} + B_\phi \dot{\phi} + k_z L_x^2 \phi \]  
\[ \tau_\psi = I_\psi \dddot{\psi} + B_\psi \dot{\psi} + k_y L_x^2 \psi \]

where \( F_x, F_y, F_z, \tau_\theta, \tau_\phi, \tau_\psi \), are the external forces and torques that act on the plate. Values of the stiffness and damping coefficients can be determined numerically using finite element analysis or approximated by analytical formulas, as discussed in the following sections. In general most of the simulations and modeling described in the thesis involves only the vertical motion of the suspended plate. Therefore, reference is made to the vertical equations of motion, Eqn.(3.14) - (3.16), which can be expressed in the alternative form as:

\[ F_z = m(\dddot{z} + 2\zeta_z \omega_z \dot{z} + \omega_z^2 z) \]  
\[ \tau_\theta = I_\theta (\dddot{\theta} + 2\zeta_\theta \omega_\theta \dot{\theta} + \omega_\theta^2 \theta) \]  
\[ \tau_\phi = I_\phi (\dddot{\phi} + 2\zeta_\phi \omega_\phi \dot{\phi} + \omega_\phi^2 \phi) \]

where \( \omega_z, \omega_\theta, \) and \( \omega_\phi \), are the resonant frequencies and \( \zeta_z, \zeta_\theta, \) and \( \zeta_\phi \), are the dimensionless damping factors of the \( z, \theta, \) and \( \phi \) modes, respectively. In general, the resonant frequency, \( \omega_i \), and damping factor, \( \zeta_i \), of the mode, \( i \), are given by

\[ \omega_i = \sqrt{\frac{k_i}{m_i}} \]  
\[ \zeta_i = \frac{B_i}{2\sqrt{k_i m_i}} \]
3.3 Squeeze-Film Damping

The viscous flow is the dominant dissipation mechanism for microstructures that operate at atmospheric pressure. Squeeze-film damping illustrated in Fig.3.2, arises from the vertical motion which creates a pressure in the thin film of air between the plate and substrate. A review on squeeze-film damping in accelerometers is given by Starr [44]. We provide a brief review and apply the results to vertical motion of a $300\mu m \times 300\mu m$, plate with a $3\mu m$ air gap.
3.3.1 Viscous flow

Continuum fluid mechanics can be applied to analyze squeeze-film damping if the air gap is much larger than the mean free path, $\lambda$, of the air molecules. Mean free path of a gas is expressed as

$$\lambda = \frac{1}{\sqrt{2\pi d_0^2 n}} \quad (3.23)$$

where $d_0^2 n$ is the collision cross section of the gas molecules, and $n$ is the molecular density, which for an ideal gas is given by $n = P/(k_B T)$, where $P$ is the pressure of the squeeze film, $k_B$ is the Boltzmann’s constant, and $T$ is the absolute temperature. For air at atmospheric pressure and $T = 300^\circ K$, the mean free path is 65nm. The 3µm air gap is about 20× larger than the mean-free-path, so the air can be modeled approximately as a viscous fluid. At pressures below 25T, the mean-free-path is greater than 2µm and the gas film dynamics must be treated as an ensemble of molecules, not as a viscous fluid. In this molecular flow regime the air damping decreases drastically with decreasing pressure, and structural damping will eventually dominate the losses.

The viscous-flow regime is described by the Navier-Stokes equation which, under several assumption can be reduced to [44]

$$\frac{\partial^2 P}{\partial x^2} + \frac{\partial^2 P}{\partial y^2} + \frac{12\mu}{z_0^3} \frac{\partial (\Delta z)}{\partial t} \quad (3.24)$$

where $P$ is the pressure of the squeeze film, $\mu$ is the viscosity of air $^1$, $z_0$ is the air gap height and $\Delta z$ is the plate displacement. The above equation is valid if the squeeze film is isothermal, has small pressure variations and undergoes small displacements with small velocity. Air velocity in the gap can be considered small if the Reynolds number, $R_e$, is much less than 1, where $R_e = (\rho vz_0)/\mu$ and $\rho$ is the density of air $^2$ and $v$ is the air velocity.

---

$^1$The viscosity of air is $1.708 \times 10^{-5}$ Pa·s at atmospheric pressure and $T = 273^\circ K$[45].

$^2$The density of air is $1.3kg/m^3$ Pa·s at atmospheric pressure and $T = 273^\circ K$[45].
With a 2μm air gap, and plate oscillation frequency of 1KHz and oscillation amplitude of 1μm, the Reynolds number is very small \((R_e = 0.0009)\). To first order, both the Reynolds number and squeeze number are independent of pressure, since the air viscosity and density vary linearly with pressure.

### 3.3.2 One dimensional Analysis

We consider the squeeze-film analysis of the plate shown in Fig.3.2(bottom) where the length \(L_y \gg L_x\). The squeeze-film is modeled with a one dimensional version of the Eqn.3.24:

\[
\frac{\partial^2 P}{\partial x^2} = \frac{12\mu v}{z_0^3} \tag{3.25}
\]

Integration and applications of the boundary conditions at the plate edges give:

\[
\Delta P = \frac{6\mu v}{z_0^3} (x^2 - L_x^2/4) \tag{3.26}
\]

where \(\Delta P\) is the pressure difference from the ambient pressure. The average pressure difference across the plate is \(\mu L_x^2 v / z_0^3\), and the total force from the damping exerted on the plate is

\[
F_B = -\left(\frac{\mu L_y L_x^3}{z_0^3}\right) v \tag{3.27}
\]

The squeeze-film damping coefficient for a rectangular plate is

\[
B_z = K_{Bz} \frac{L_x}{L_y} \frac{\mu L_y L_x^3}{z_0^3} \tag{3.28}
\]

where \(K_{Bz}(L_x/L_y)\) is a form factor that is introduced to account for the finite length, \(L_y\). For very long plates \((L_y \gg L_x)\), e.g. the devices considered in present work), we can deduce that \(K_{Bz} = 1\). The two dimensional flow problem can be solved for other geometries; \(K_{Bz}\) is approximately 0.42 for a square plate \((L_y = L_x)\). For the 300μm \(\times 300\mu m\) square plate,
which corresponds to the largest capacitive overlap area in meander toggle switches the damping coefficient is $B_z = 0.0021 \text{ Pa-s}$. However, the switch-beam geometry considered in the present work, the surface micro-machined plates are perforated with holes to facilitate the sacrificial layer etching during release etch. Another effect of the holes is to reduce the squeeze-film damping significantly. Since in present case ($L_y \gg L_x$), we incorporate the effect of perforations in the form factor $K_{Bz} = 1$.

### 3.4 Micro-Machined Flexure Design

In this section we present an overview of the mechanical flexures - the moveable part of many MEMS devices. Metallic (electroplated or sputter deposited $Au$, $Al$ or $Cu$), polysilicon and $Si_3N_4$ based flexures are used in switches, varactors, accelerometers, gripping devices, tuning forks, resonant sensors and many other devices. In most of the cases it is desirable to have a compliant flexure in one direction while being very stiff in orthogonal directions. For example the proof mass of most accelerometers is designed to move easily in the direction normal to the substrate. Motion
in other directions increases sensitivity to cross-acceleration. In MEMS switches, any deflection in direction other than the desired one (invariably the z-axis is the preferred direction of deflection in beam based designs) changes the contact area, lowering the capacitance in capacitive switches, and increases the contact resistance in case of series switches. The flex-ures are in general created as straight beams and with motion constrained in rectilinear direction. The following discussion will emphasize design is-sues for rectilinear-displacement flexures. First, we present the results of small displacement beam theory for simple beams, and discuss the practi-cal limitations on beam dimensions. Rest of the section is devoted to the discussion on static spring constant for a serpentine spring which is used in the designed devices. Finally we compare the results with FEM (ANSYS) simulations.

<table>
<thead>
<tr>
<th>Cantilever</th>
<th>Guided End</th>
<th>Fixed-Fixed Beam</th>
</tr>
</thead>
<tbody>
<tr>
<td>( x = \frac{F_f L}{E I_w} )</td>
<td>( x = \frac{F_f L}{E I_w} )</td>
<td>( x = \frac{F_f L}{4 E I_w} )</td>
</tr>
<tr>
<td>( y = 4 \frac{F_f L^3}{E I_w^3} )</td>
<td>( y = \frac{F_f L^3}{E I_w^3} )</td>
<td>( y = \frac{1}{16} \frac{F_f L^3}{E I_w^3} )</td>
</tr>
<tr>
<td>( z = 4 \frac{F_f L^2}{E I_w} )</td>
<td>( z = \frac{F_f L^2}{E I_w} )</td>
<td>( z = \frac{1}{16} \frac{F_f L^2}{E I_w} )</td>
</tr>
</tbody>
</table>

(a) Concentrated Load

<table>
<thead>
<tr>
<th>Cantilever</th>
<th>Guided End</th>
<th>Fixed-Fixed Beam</th>
</tr>
</thead>
<tbody>
<tr>
<td>( x = \frac{f_x L}{E} )</td>
<td>( x = \frac{f_x L}{E} )</td>
<td>( x = \frac{f_x L}{4 E} )</td>
</tr>
<tr>
<td>( y = \frac{3}{2} \frac{f_x L^3}{E I_w^3} )</td>
<td>( y = \frac{1}{2} \frac{f_x L^3}{E I_w^3} )</td>
<td>( y = \frac{1}{32} \frac{f_x L^3}{E I_w^3} )</td>
</tr>
<tr>
<td>( z = \frac{3}{2} \frac{f_x L^2}{2 E I_w} )</td>
<td>( z = \frac{1}{2} \frac{f_x L^2}{2 E I_w} )</td>
<td>( z = \frac{1}{32} \frac{f_x L^2}{2 E I_w} )</td>
</tr>
</tbody>
</table>

(b) Distributed Load

Table 3.1: General displacement equations derived from small displacement theory.
Figure 3.4: Simulated (solid lines, ANSYS) and calculated (dotted lines linear theory) displacement versus force for two fixed-fixed type beams.

### 3.4.1 Spring Constants for Simple Beams

Cantilever, guided-end and fixed-fixed beams are shown in Fig.3.3. A concentrated force $F(N)$ is applied to the free end of the cantilever beam (a), to the free end of the guided-end beam (b) and to the center of the fixed-fixed beam in (c). A uniform distributed load, $f(N/m)$ (force per unit length) is applied to the surface of each beam in Fig.3.3(d-f). The axial displacement is found from Hooke’s Law: $stress = E \cdot strain$, where $E$ is the Young’s Modulus of elasticity. The displacement equations are summarized in Table 3.1, assuming a beam with small angles of rotation, no axial loading and no shear deformation [46]. The beams with rectangular cross section have width $w$, thickness $t$ and length $L$. For cases where concentrated loads are applied to the beam, linear spring constants are defined as a measure of the beam’s stiffness given by

$$k_x = F_x/x ; \quad k_y = F_y/y ; \quad k_z = F_z/z ;$$

The cantilever beam is the most compliant and the fixed-fixed beam is the stiffest, if the beam dimensions are equal for both cases. The stiffness ratio
of the axial to lateral in-plane motion $k_x/k_y$, is proportional to $(L/w)^2$. For a cantilever with $L/w = 100$ the stiffness ratio is 10000. The stiffness ratio of the vertical to lateral in-plane motion, $k_z/k_y$, is equal to $(t/w)^3$. Thus if restricted vertical motion is desired, the beam width must be larger than the thickness.

3.4.2 Nonlinear Effects

The deflection equations listed in Table 3.1, are derived from differential equations assuming small deflections and small angles of rotation. The exact deflection for a cantilever beam is compared with values using small deflection theory in [47]. Normally the small deflection theory is 10 % in error for deflections greater than 30% of the beam length. Shear deformation, which is neglected in Table 3.1 is small if

$$w << \sqrt{\frac{4}{3(1+\nu)}} L \approx L$$  \hspace{1cm} (3.29)

where $\nu$ is the Poisson’s ratio and is assumed to be 0.42 for Au and 0.3 for polysilicon [47]. Most mechanical flexures are long and narrow, thereby satisfying Eqn.3.29. If axial tensile stress is present in laterally deflected fixed-fixed beams, a nonlinear force-displacement results from the axial stress, where the effective spring constant increases with increasing load. Fig.3.4 compares the values for deflection at the center vs load calculated numerically for two 1.5 $\mu$m wide and 1.5$\mu$m thick fixed-fixed beam with analytical results that assume small deflection. For deflections greater than approximately 1$\mu$m, the effects of tension in the beam becomes significant. The small deflection theory can only be used to accurately predict deflections smaller than 0.2% of the beam length, for a 600$\mu$m long beam [48].
3.4.3 Spring Constant for the cases specific to MEMS Switching Devices

In this section we discuss in brief about the mechanical behavior in terms of the spring constant and stresses originating form the fabrication process, for one of the common switch configurations in RF-MEMS switches: the fixed-fixed beam and its variants in terms of load distribution across the beam. The fixed-fixed beams are widely used because of the ease of fabrication and high spring constant. In MEMS switching devices the operation of the structure is limited to small deflections e.g. 0.2 - 0.5 % of the beam length, therefore the behavior is modeled by using linear spring constant \( k (N/m) \) as mentioned in the section above. In order to account for the biaxial residual stress arising from the fabrication process, the spring constant can be modeled in two parts: (1) stiffness of the structure which accounts for the material properties (Young’s modulus \( E \), and Poisson’s ratio \( \nu \)) and (2) stress \( \sigma \) (Pa) which is due to the fabrication process.

For a beam with rectangular cross section \((w \times t)\), length \( L\), when subjected to concentrated load \( F \), as shown in Fig.3.5, the deflection is given by

\[
\begin{align*}
  z &= \frac{M_A x^2}{2EI} + \frac{R_A x^2}{6EI}, \quad \text{for } x \leq a \\
  M_A &= -\frac{F \cdot a}{L^2} (L - a)^2 \\
  R_A &= \frac{F}{L} (L - a)^2(L + 2a), \quad I = \frac{wt^3}{12}
\end{align*}
\]

where \( M_A (N \cdot m) \) is the reaction moment at the left end, \( R_A \) is the vertical
3.4. MICRO-MACHINED....  CHAPTER 3. MECHANICAL DESIGN ...

Figure 3.6: (a) Variation of the normalized spring constant \((k/w)\) with load distribution for three beam configurations (Au, \(E=108\) GPa), the spring constant for load near the anchor locations is higher, (b) shows the variation with ratio of load reference point \(x\), to beam length \(L\), for the force distributed over the center and at the beam ends.

reaction to the applied load \(F\). Deflection for a beam with concentrated load applied at the center can be expressed in terms of its dimensions by substituting \(x = L/2\) in Eqn.3.31 and is given by

\[
z = \frac{1}{EI} \left[ \frac{F}{48} \left( L^3 - 6L^2a + 9La^2 - 4a^3 \right) \right] \quad (3.33)
\]

In MEMS applications, the load corresponds to the actuation force and is typically distributed across the beam. The deflection at the center is used to determine the spring constant. In electrostatically actuated MEMS switches, the load distribution depends on the location of the actuation electrodes. We present a comparison of three cases, two of which are extensively used in electrostatically actuated RF MEMS switches and one of which has been implemented in the present work.

3.4.3.1 Uniformly Loaded Beams

The deflection for a beam where the load is distributed across the whole beam is found by integrating the deflection given by Eqn.3.33 over the
entire beam length (from L/2 to L and multiplying by 2). The spring constant for load \( f \), the load per unit length \( (F = fL) \) is given by

\[
-k = -\frac{F}{z} = 32Ew\left(\frac{t}{L}\right)^3
\]  

(3.34)

### 3.4.3.2 Load Distributed Over a Central Area

Two terminal MEMS switches with actuation signal superimposed on the CPW central conductor (signal line), have the corresponding actuation force distributed over a small area at the center of the moveable beam. The configuration corresponds to a beam with mechanical load evenly distributed over a centrally located area as shown in Fig.3.6(a), bottom right. Area of the beam subjected to load depends on the CPW geometrical configuration. The spring constant for an area equal to two third of the length is shown in Fig.3.6 and is given by expression:

\[
k = -\frac{F_z}{z} = 32Ew\left(\frac{t}{L}\right)^3 \frac{1}{8(x/l)^3 - 20(x/l)^2 + 14(x/l) - 1}
\]  

(3.35)

### 3.4.3.3 Load Distributed Over Areas Near the Anchor Location

The fixed-fixed beam with load distributed over two equal areas, located towards the anchor points is as shown by the inset in Fig.3.6(a) top, corresponds to the RF MEMS relay or three terminal switch configuration. The area of the electrically isolated actuation electrodes (load area) can be varied to achieve the required spring constant and actuation voltage. The low actuation voltage switching devices discussed in the present work are based on this configuration and are described in Chapter 5. The spring constant is given by:

\[
k = 4Ew\left(\frac{t}{L}\right)^3 \frac{1}{(x/L)(1 - (x/l))^2}
\]  

(3.36)

Fig.3.6(a) shows the variation of spring constant with beam thickness to length ratio for Au beams for three different cases of load distribution as
described above. The spring constant is calculated for the typical dimensions and material used in switch design in the present work. The measured Young’s modulus for electroplated gold with Cr seed layer is found to be 108 GPa [38] and the switch dimensions are: width 150 - 200 \( \mu m \), length 500 - 1000 \( \mu m \), thickness 1.5 \( \mu m \) and the gap or anchor height of 3 \( \mu m \). As illustrated by Fig.3.6, beams with central load are more compliant and the spring constant increases as the load moves towards the anchor location. The switches with two actuation electrodes placed near anchor points have highest spring constant for given dimensions and material. Variation of the load position (electrode location) w.r.t. to the beam center can be used to lower \( k \), subjected to the gap between RF signal line and the central conductor of CPW. Another approach to have a lower spring constant is to anchor the beam with flexures whose stiffness can be controlled independently, resulting in a lower equivalent spring constant for the structure. We have followed the later approach. The beams are anchored with serpentine springs at four locations resulting in devices with spring constant ranging from 0.05 - 15.0 N/m. The following sections present the expression for serpentine spring used in the device design.

### 3.4.4 Contribution of Residual Stress in Spring Constant

MEMS transduction components such as beams and cantilevers are realized in metals (Au, Al, Ni and Cu) either by sputter deposition or electroplating techniques. The different process variants result in structures with intrinsic stress that may vary from compressive to tensile depending upon the material and process parameters e.g. in our process described in Chapter 5 the measured intrinsic stress for electroplated Cr - Au structural layer is 180 MPa. For a beam with cross section \( w \cdot t \) the biaxial residual stress generates a force, \( S = \sigma(1 - \nu)wt \) pulling on the both ends, when beam is modeled as a stretched wire. The contribution to the total spring constant
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Figure 3.7: (a) Variation of effective spring constant ($k/w$), with residual stress contribution, for Al and Au beams in CPW actuation configuration (central load), (b) Critical Stress for Ni, Au and Al fixed - fixed beams.

for the three cases considered above is given by

\[ k_{rs} = 8\sigma(1-\nu)w\left(\frac{t}{L}\right), \quad \text{uniform load} \]  
(3.37)

\[ k_{rs} = 8\sigma(1-\nu)w\left(\frac{t}{L}\right)\frac{1}{3-2(x/L)}, \quad \text{load at center.} \]  
(3.38)

\[ k_{rs} = 4\sigma(1-\nu)w\left(\frac{t}{L}\right)\frac{1}{1-(x/L)}, \quad \text{load near the edge.} \]  
(3.39)

The total spring constant is the sum of the contributions from beam stiffness and the biaxial residual stress. For a beam using CPW central conductor as actuation electrode with length one third of the total beam length, the total spring constant is

\[ k_t = (17.64)Ew\left(\frac{t}{L}\right)^3 + (0.48)\sigma(1-\nu)w\left(\frac{t}{L}\right) \]  
(3.40)

For the switch configurations with two actuation electrodes and length one-third of the beam ($x = (2/3)L$), total spring constant is:

\[ k_t = 54Ew\left(\frac{t}{L}\right)^3 + 12\sigma(1-\nu)w\left(\frac{t}{L}\right) \]  
(3.41)

The results calculated using above equations, for electroplated gold beams with measured tensile stress of 180 MPa [49] are compared with no stress,
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Figure 3.8: Layout of the bridge portion of a fabricated capacitive switch showing the etch release hole-array and serpentine flexure details.

20MPa, and 40MPa in Fig. 3.7(a). The spring constant variations for Al are similar to gold and the residual stress component is observed to dominate for stress greater than 20MPa.

3.4.4.1 Critical Stress

The amount of compressive stress that a beam can withstand before buckling is called critical stress and is given by [47]

\[ \sigma_{cr} = \frac{\pi^2 El^2}{3L^2(1-\nu)} \]  

(3.42)

The variation in the critical stress with the thickness to length ratio for Ni, Au and Al are shown in Fig.3.7(b). For gold beams with critical stress in the range of 3 to 20 MPa and thickness fixed at 1.5\text{\mu}m, the beam length can vary approximately from 250-500 \text{\mu}m. In our process, for some samples the ‘as deposited stress’ is found to vary from mild compressive (25 MPa) to tensile (110 MPa), the final tensile stress after the thermal treatment of Cr-Au structural layer is measured at 180 MPa. However, the fact that holes in the beams and meander suspensions contribute to lower the overall residual stress of the structures, beams with larger length have also been used.
3.4.4.2 Perforated Beams

The beams used for capacitive, ohmic and symmetric toggle switch are patterned with 10 \( \mu m \) x 10 \( \mu m \) holes, also called release etch holes [50]. Fig. 3.8 shows the layout of the bridge portion of a fabricated capacitive device. The advantages of having holes in the beams are: reduction in beam mass, efficient dry etching of photo-resist under the beam, reduced squeeze film effect, higher switching speeds and residual stress reduction. The perforation pattern is characterized by the ligament efficiency, \( \mu = \frac{l}{pitch} \), defined as the ratio of the remaining link width \( l \) to the pitch of the pattern. The reduction in residual stress [50], is approximately equal to \( \sigma = (1 - \mu)\sigma_0 \). For \( l = 10, \mu m \) and pitch of 20 \( \mu m \), the stress in our case is reduced from 180MPa to 90MPa. The effect of holes on up-state capacitance is negligible if the diameter is less than 3 to 4 times the gap height. However, the capacitance in down state is lower as compared to beams without etch release holes.

3.4.4.3 Cantilever Beams

The cantilever or fixed-free beams are particularly useful for “in-line” ohmic contact switches and result in low actuation devices as compared to fixed-fixed bridge configuration. The residual stress is released because of elastic deformation of the free end. However, the stress gradient present over the beam cross section may deflects the beam upon release. To minimize the deflection, an appropriate combination of compressive and tensile materials are typically used [51], albeit this increases the fabrication complexity.
3.4.5 Support Flexure Design for Low Spring Constant
Bridge Structures

The support beam - structures used to anchor the main moveable transducer component of MEMS devices such as switches, varactors, resonators etc. are called support flexures. Fig.3.8 shows details of the serpentine flexures or springs used for the switch design, in the present work. In this section we present the static linear spring-constant analysis of serpentine flexures and compare the results with numerical simulations (ANSYS®). An overview of similar types of flexures used in MEMS device design is also provided.

3.4.5.1 Folded and Crab Leg Flexures

In MEMS devices, in general, the moveable transducer component thickness is small compared to other dimensions. However, the width may be comparable to length in certain cases. The structure when anchored with connecting flexures behaves as a suspended plate and acts as a concentrated load. The spring constant particulary for the bridge configuration, where the beam is fixed at both ends, is generally high and the residual stress contribution is also significant as mentioned earlier. Flexures lower the spring constant and also reduce the residual stress effects. Fig.3.9
shows another three types of flexures [52],[46] used in MEMS devices, (a) shows clamped-clamped flexures which can be modified to crab-leg configuration by adding another small section (L) to reduce the stress and the spring constant. The folded flexure (b) also reduces the axial stress. Each end is free to expand or contract in all directions. For example the original residual stress in a small section \( L_c \) is averaged over the entire beam length, giving a reduced effective residual stress \( \sigma_{reff} = (L_c/L_b)\sigma_r \), where \( L_b \) is the flexure length. Fig.3.8 shows the serpentine flexure together with main switch body. Compliant serpentine flexures can be designed with compact springs by adding more sections. The width of the meanders is adjusted to get the desired stiffness ratio. In these structures the residual and extensional stresses are relieved through bending of the beams.

### 3.4.6 Linear Spring Constant of Serpentine Flexures

This section presents the analytical formula for the linear spring constant of serpentine flexures used in the switch design, in present work. The goal is to find the displacement \( \delta \), resulting from a force \( F \) applied in the appropriate direction. The displacement arising only from bending and torsion are considered. Deformation from shear, beam elongation and beam shortening are neglected. In the procedure for analysis first we consider the symmetry of the geometry. As shown in Fig.3.8, the structure has two fold symmetry. Therefore, we need to analyze only one spring as the resulting spring constant is one fourth of the flexure spring constant. Next is to identify the boundary conditions, at the spring ends. Displacement and rotation of the spring ends are constrained to be zero except in the direction of applied force. Then the beam segments of the spring are considered individually and free body diagram is drawn. The boundary conditions at the end of each segment are determined by equating the sum of forces, moments and kinetic energy to zero. These boundary conditions are ex-
pressed in terms of the reaction forces, moment and torsion at the end of the spring. Then we calculate the moment and the torsion of each beam segment as a function of the position $x$ along the beam. Next, the set of simultaneous equations is solved that describe the boundary conditions to obtain the reaction forces, moments, torsion and displacement at each end of the spring using energy methods. The displacement is expressed in terms of the applied load $P_i$ in terms of the partial derivative of the strain energy of the linear structure, $U$ w.r.t $P_i$ at the point where load is applied, such that the displacement

$$\delta_i = \frac{\partial U}{\partial P_i} \quad (3.43)$$

for an applied moment $M_i$ and the corresponding angular displacement $\theta_i$ is

$$\theta_i = \frac{\partial U}{\partial M_i} \quad (3.44)$$

The last step is to calculate the spring constant, equal to the applied force divided by the displacement.

The serpentine flexure shown in Fig.3.10, is made of $N$ serpentine springs. Each meander has a connecting length $L_c$ or $a$ and span length $L_s$.
or b. The end meanders can be considered half of the span length [19]. The number of meanders can vary for a particular switch configuration. In our switch design we have used one and two meander configurations. Because of the flexure symmetry, the free end of the spring which is connected to the main switch body, has guided boundary conditions, where only motion in the desired direction (z) is allowed. At the anchor point A all six degrees of freedom are assumed to be fixed. A free body diagram of the serpentine spring with N meanders is shown in Fig.3.10. We limit our discussion to the z-directed spring constant. Spring constants for x and y direction can also be calculated in a similar way. At the free end which connects the meander with the main switch body a moment $M_0$ and a torsion $T_0$ are applied to constrain the rotation angles around the x and y axis. The torsion and the moment of each beam are then given by:

$$
M_{a,i} = M_0 - F_z[\xi + (i - 1)a]
$$
$$
T_{a,i} = T_0 + \left[\frac{1+(-1)^i}{2}\right]F_zb
$$
$$
M_{b,j} = (-1)^jT_0 - F_zx + \left[\frac{1+(-1)^i}{2}\right]F_zb
$$
$$
T_{b,i} = (-1)^j(iF_za - M_0)
$$

(3.45)

where $M_{a,i}$ and $T_{a,i}$ are the moment and torsion of the $i_{th}$ connecting beam $a$ and $M_{b,j}, T_{b,i}$ correspond to the span beam $b$, with $i$ and $j = 1$ to $2N$. In these equations $x$ is the longitudinal dimension along each of the beams. Following the virtual work method [53], the total elastic strain energy of the of the meander is given by:

$$
U = \sum_{i=1}^{2N} \int_{0}^{a} \left(\frac{M_{a,i}^2}{2EI_x} + \frac{T_{a,i}^2}{2GJ}\right)dx + \sum_{j=1}^{2N} \int_{0}^{b} \left(\frac{M_{b,j}^2}{2EI_x} + \frac{T_{b,j}^2}{2GJ}\right)dx
$$

(3.46)

where $t =$beam thickness, $w =$ beam width, $E =$ Young’s modulus, $\nu =$ Poisson’s ratio, $G = E/(2(1+\nu))$ is the shear modulus, $I_x = wt^3/12$ is the
Span Beam Length $L_s$  
50 – 260 $\mu$m

Connector Beam Length $L_c$  
20 – 25 $\mu$m

Width (for all Beams) $w$  
10 $\mu$m

Thickness (for all Beams) $t$  
1.5 $\mu$m

Thickness (reinforcing beams) $t_{sw}$  
5.0 $\mu$m

Beam Material (Electroplated Cr – Au)  
$E = 108$ MPa

Poisson’s Ratio Cr – Au  
$\nu = 0.42$

Spring Constant Range $k_z$  
0.05 – 15 N/m

Table 3.2: Dimensions of the springs and switch structure.

$x$-axis moment of inertia, $I_z = tw^3/12$ is the $z$-axis moment of inertia and $I_p = I_x + I_z$ is polar moment of inertia. The torsion constant $J$ for a beam of rectangular cross section is given by

$$J = \frac{1}{3}tw^3(1 - \frac{192t}{\pi^3w}\sum_{i=1,odd}^{\infty} \frac{1}{i^3} tanh\left(\frac{i\pi w}{2t}\right))$$  \hspace{1cm} (3.47)

where $t < w$. If $t > w$, then the roles of $t$ and $w$ are interchanged in Eqn.3.47. For beams with square cross section $J/I_p = 0.843$, the value drops down for higher aspect ratio max($w/t$ and $t/w$). In the present case where $t < w$, $J = 0.413I_p[54]$. Finally, we find the spring constant in $z$-direction, by making use of the equations for boundary conditions and the displacement in $z$-direction given by

$$\phi_0 = \frac{\partial U}{\partial M_0} = 0 \text{,} \quad \psi_0 = \frac{\partial U}{\partial T_0} = 0 \text{,} \quad k_z = F_z/\delta_z = F_z/\frac{\partial U}{\partial F_z}$$  \hspace{1cm} (3.48)

The above equations lead to the following expressions for the reactions moments $M_0$, $T_0$ and the spring constant $k_z$:

$$M_0 = \frac{2Nd}{EL_x} + \frac{(2N+1)b}{GJ} aF_z, \quad T_0 = -\frac{F_zb}{2}$$  \hspace{1cm} (3.49)

$$k_z = \left[\frac{8N^3a^3 + 2Nb^3}{3EI_x} + \frac{abN[3b + (2N + 1)(4N + 1)a]}{3GJ}\right]$$
\[- \frac{Na^2[2Na + (2N+1)b]^2}{2\left(\frac{a}{EI_x} + \frac{b}{GJ}\right)} - \frac{Nb^2}{2}\left(\frac{a}{GJ} + \frac{b}{EI_x}\right)^{-1}\] (3.50)

The expressions for \(k_x\) and \(k_y\) which can be drawn in a similar way, are:

\[
k_x = \frac{48EI_{z,b}}{a^2(\bar{a} + b)N^3}
\] (3.51)

\[
k_y = \frac{48EI_{z,b}}{b^2(3\bar{a} + b)N}
\] (3.52)

where \(\bar{a} = I_{z,b}a/I_{z,a}\) and \(N \gg 3b/(\bar{a} + b)\).

In the expression given by Eqn.3.50 the first two terms represent the contribution to the spring constant that arises from the beam bending and twisting. These terms thus solely depend on the beam geometry and the material of the beam. The other terms are due to the boundary conditions on the moving end and correspond to its inability to rotate around the \(x\) and \(y\) axis.

3.4.7 Spring Constant: Comparison of Analytical and FEM Simulated \(k_z\)

This section presents a comparison between the spring constant values calculated using the analytical expression given by Eqn.3.50 and a FEM tool (ANSYS 7.0). A sufficiently large set of physical dimensions for meanders were considered to finalize the parameters such as spring constant \((k_z)\), actuation voltage, mechanical resonance frequency and the overall switch dimensions. The meander dimensions as summarized in Table 3.2 are used as the input for the FEM simulations and the analytical calculations and form the basis for further switch design. To the free end of the spring (in switch configuration, this end is connected to the main switch body), a concentrated force \((0.01-1 \, \mu N)\) is applied, in \(z\)-direction, with necessary guided end boundary conditions. We use non-linear 8 node shell element
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Figure 3.11: Comparison of analytically calculated spring constant (dotted lines) with finite element analysis approach (solid lines), for serpentine springs with one to six sections. In (a) $L_s = 180\mu m$, in (b) $L_s = 260\mu m$.

Figure 3.12: Comparison of the $k_z$ vs meander span length $L_s$ variation using analytical and FEM approach. Dimensions which are same for all simulations and calculations are indicated in the figure.
Figure 3.13: Layout for two capacitive switch configurations. (a) two meanders (b) single meander, spring constant 0.48 N/m and 5.22 N/m respectively.

(shell 181) with mesh refining till the deflection is constant over subsequent meshing iterations. Fig.3.13 shows some of the meander structures used. Comparison between the simulated and calculated $k_z$ is given by Fig.3.11 with number of meanders ranging from 1 to 6. The agreement between the simulated and analytically calculated $k_z$ is better than 5% for most of the cases except for springs with number of meanders ($\geq 5$) where it deviates by 11% to 16% for the considered two cases $L_s = 180$ and $260\mu m$. Also, as indicated by Fig.3.11 and Fig.3.12 the analytically obtained $k_z$ is 2 to 6% higher over FEM values. The difference for $L_s = 50\mu m$ is about 12%. The discrepancy arises because of the guided end boundary conditions and simplified assumptions of no axial stress. The reduction in the spring force is insignificant in springs with more than two meanders. Larger number of sections also increase the fabrication process complexity [38]. In view of above, we choose the $L_s$ range from $100 - 250\mu m$ with one or two meanders sections. However, capacitive devices with $L_s = 50 - 80\mu m$ have also been fabricated (Fig. 3.13). The comparison between calculated $k_z$ and measured values is presented in terms of actuation voltage in Chapter 5.
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Figure 3.14: Spring massless model.

3.5 Electrostatic Actuation and Pull-in Voltage

Actuation or pull-in voltage is one of the most important parameters of the electrostatically actuated MEMS switches. This section briefly describes the electromechanical model and provides closed form analytical expression for electrostatic parallel plate design. The treatment is mainly based on the analysis provided in [19] and [53].

When a voltage is applied between the beam and pull-down electrode, a capacitive switch can be modeled by a lumped spring mass system as shown in Fig.3.14. The system consists of a moving plate of area $A$ and mass $m$ suspended by a spring with constant $k$, at gap height $g$, above a fixed plate. The electrostatic actuator can be represented by a two port capacitor with voltage $V$ and current $I$ as effort and flow variables in the electrical domain and force $F$ and displacement $z$ in the mechanical domain. Displacement of the plate attached to spring and the gap height variations are in opposite direction and are the flow variables in series; they share the same displacement corresponding to applied voltage. The stored potential energy for a capacitor is given by $W(q_1) = \int_0^{q_1} edq$, where ‘e’ represents the effort (voltage). For a parallel plate capacitor with capacitance $C = \epsilon A/g$, the stored energy is $W(Q) = Q^2/2C$ and the co-energy is $W^*(Q) = CV^2/2$. For a Hook’s spring attached to a fixed support, the effort (force) $F = k \cdot x$, where $x$ (displacement) is the flow variable. The
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stored energy for displacement \( x_1 \) is

\[
W(x_1) = \int_0^{x_1} F(x)\,dx = \frac{1}{2}kx_1^2
\]  

(3.53)

Thus, assuming generalized displacement \( Q \), for the capacitor and \( x \) for the
spring, leads to the same stored energy function, and we can represent the
spring with a capacitance \( C_{spring} = 1/k \).

In a similar way the mass of the moving plate can be represented by an
inductor and the damping by a resistor to arrive at equivalent circuit for
the mechanical domain as shown in Fig. 3.18. In the static analysis we
ignore the mass and damping. In the spring model in Fig. 3.14, we have to
consider both the electrical and the mechanical stored energy. Considering
the gap to be an independent variable, the co-energy is given by

\[
W^*(V, g) = QV - W(Q, g), \quad Q = \frac{\partial W^*(V, g)}{\partial V}
\]

\[
F = \frac{\partial W^*(V, g)}{\partial g} \bigg|_V \quad W^*(V, g) = \int_0^V \frac{\epsilon AV^2}{2g}
\]

(3.54)

(3.55)

From which we find \( Q = \epsilon AV/g \) and in a similar way for force we obtain
\( F = \epsilon AV^2/2g^2 \). With reference to the ‘electro-mechanical’ equivalent cir-
cuit, the applied voltage determines the force, which stretches the spring,
thus determining the charge in the gap such that:

\[
F = \frac{\epsilon AV_{in}^2}{2g^2}, \quad g = g_0 - z, \quad and \quad z = \frac{F}{k}
\]

(3.56)

Using above equations, the gap can be expressed as :

\[
g = g_0 - \frac{\epsilon AV_{in}^2}{2kg^2}
\]

(3.57)

3.5.1 Pull-in Voltage

The voltage controlled parallel plate actuator exhibits important behavior
called pull-in (Fig.3.15). Considering the position of the upper plate sub-
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Figure 3.15: Gap height versus actuation voltage for a capacitive switch with actuation electrode dimensions 150 x 240 microns. The beam snaps down at $V_{pi} = 15V$. The lower dotted line shows the unstable beam position.

jected to spring force in upward direction and the electrostatic pull in the downward direction, the net force is:

$$F_{net} = -\frac{\epsilon AV^2}{2g^2} + k(g_0 - g)$$  \hspace{1cm} (3.58)

For a small perturbation in the position we have

$$\delta F_{net} = (\frac{\epsilon AV^2}{g^3} - k)\delta g$$  \hspace{1cm} (3.59)

where $g_0$ is the gap at zero voltage and no spring extension. For stable equilibrium $\delta F_{net}$ should be negative in Eqn.3.59, $\Rightarrow k > \frac{\epsilon AV^2}{g^3}$. Since equilibrium gap decreases with increasing voltage, there is a specific voltage at which stability of equilibrium is lost. The voltage is called pull-in voltage and denoted by $V_{pi}$. At pull-in, $F_{net} = 0$ and

$$k = \frac{\epsilon AV^2_{pi}}{g^3_{pi}}$$  \hspace{1cm} (3.60)
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Figure 3.16: Actuation voltage $V_{pi}$ versus (a) beam thickness to length ratio (length = 500 µm, thickness = 1.5 µm) and (b) $V_{pi}$ vs gap height(g). Process induced dimensional variations are critical in presence of residual stress higher than 20 MPa.

From Eqn.3.58 and 3.60 we find that pull-in occurs at $g_{pi} = (2/3)g_0$ and the voltage is

$$V_{pi} = \sqrt{\frac{8kg_{0}^{3}}{27\epsilon A}}$$

(3.61)

The variation of the pull-in voltage with beam length and gap height for given residual stress is shown in Fig.3.16 (a) and (b). The residual stress above 20-30 MPa makes the beams more stiffer and may change the $V_{pi}$ from 50 - 300% [55], for gap heights more than 3 µm. Stress induced deviation in $V_{pi}$ can be minimized by use of strict dimensional control, suspension springs and reinforced beam design as discussed Chapter 5.

3.5.2 Pull-out Voltage

The minimum voltage required to keep the switch in ‘on-state’ is called pull-out ($V_{po}$) or the threshold voltage. If the applied actuation is decreased below $V_{po}$ the beam restores back to its original zero bias position. In capacitive MEMS switches, with a 0.01 - 0.02 µm thick isolation layer of $SiO_2$ or $Si_3N_4$ between the beam and actuation electrode, the electrostatic
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Figure 3.17: Pull-out voltage versus gap height for one of the fabricated devices at various spring constant values.

The electrostatic force is

\[ F_e = \frac{V^2}{2} \frac{\varepsilon \varepsilon_0 A}{(g + (t_d/\varepsilon))^2} \quad (3.62) \]

where \( \varepsilon \) (\( = 1 \) for \( g \neq 0 \) and 0.4 - 0.8 for \( g = 0 \)), and accounts for parallel plate capacitance reduction due to roughness of the metal-dielectric interface [19]. The mechanical restoring force acting on the beam is

\[ F_r = k_{es}(g - g_0) \quad (3.63) \]

where the spring constant also includes the effect of beam stretching and is applicable to bridge structures. For the switch to stay in down position the electrostatic force (Eqn.3.62) must be larger than the restoring force given by Eqn.3.63, and this is achieved when applied voltage is

\[ V_h = \sqrt{\frac{2F}{\varepsilon\varepsilon_0 A}} \frac{(g + (t_d/\varepsilon))^2}{(g_0 - g)(g + (t_d/\varepsilon))^2} = \sqrt{\frac{2k_{es}}{\varepsilon\varepsilon_0 A}} \]

(3.64)

In general devices require a pull-out voltage of 4-10 volts less than the pull-in voltage (Fig. 3.17). It is because of the complex nature of the contact between beam and actuation electrode involving adhesion and repulsion...
forces between metal and dielectric layers. However, the reduction in actuation voltage once the beam has been pulled-down, is important in order to reduce charge injection into the thin dielectric layer and improve the switch reliability.

3.5.3 Power Handling

The RF power handling capability of MEMS switch depends on the device geometry, circuit configuration and switch contact type. Capacitive switches are more sensitive to the RF power on the transmission line, because of the large overlap area (150 × 90 μm² – 300 × 300 μm² in present work). Series switches are insensitive to RF power as the overlap area is small (90 × 20μm²). The electrostatic force corresponding to an incident wave \( V^+ = V_{pk} \sin(\omega t) = \sqrt{2PZ_0} \sin(\omega t) \), (where \( Z_0 \) is the characteristic impedance of the transmission line), the on the switch, assuming reflection coefficient \( S_{11} << -10dB \), is

\[
F_e = -\frac{1}{2} \varepsilon_0 A \left( \frac{1}{2} V_{pk}^2 (1 + \sin(2\omega t)) \right) = -\frac{1}{2} \varepsilon_0 A \frac{V^2_{dc-equ}}{g^2} \quad (3.65)
\]

Because of the lower mechanical resonance frequency, switches respond only to DC component given by:

\[
V_{dc-equ} = \frac{V_{pk}}{\sqrt{2}} = \sqrt{PZ_0} \quad (shunt \ switch) \\
V_{dc-equ} = \sqrt{2}V_{pk} = 2\sqrt{PZ_0} \quad (series \ switch) 
\]

Eqn.3.66 determines the extent to which \( V_{pi} \) can be reduced for given RF power. In order to avoid self actuation \( V_{pi} \) should be \( \geq V_{dc-equ} \). The maximum power that a capacitive switch can handle in shunt and series configuration is

\[
P_{shunt} = \frac{V_{pi}^2}{Z_0}, \quad P_{series} = \frac{V_{pi}^2}{4Z_0} \quad (3.67)
\]
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The actuation voltage range for the capacitive switch considered is 3 - 15V, which corresponds to 0.18 - 4.5W. However, by using additional up-state clamping electrode, RF power handling capability can be increased. In the symmetric toggle switch Chapter 6, this depends on the voltage on pull-out electrodes.

3.5.4 Effects of Non-periodic Excitation, Gravitational Acceleration and Acoustic Waves

Non-periodic excitation such as impulse excitation and shocks are caused by forces that are generally large in magnitude and of very short duration. The acceleration caused by such an impulse should be less than the critical acceleration for a switch and is given by

\[ a \ll a_c = \frac{k g_0}{D} = \frac{w_0^2 g_0}{D} \]

where D denotes the dynamic factor \((1 - 2)\) of shock and depends on exact shape of the shock pulse [56]. For a switch with resonant frequency of 200kHz and gap 2.5µm, the threshold acceleration \(\approx 2 \times 10^5 \text{g}\) which is well above the acceleration specifications \((5000g - 50,000g)\) for cellular handsets [57]. In switch down state, the induced acceleration threshold is given by

\[ a_0 = \frac{F_c}{m} \approx \frac{\varepsilon_r \varepsilon_0 A V^2}{2m t_c} \]

The threshold for the fabricated devices range between \(2.5 \times 10^6 - 10^7\), which is above the threshold given 3.69.

In comparison MEMS device with a larger spring constant \((\geq 5N/m)\) and lower mass are less sensitive to the acceleration due to gravity and acoustic waves. Typically for Au-beam with mass ranging from \(5 - 12 \times 10^{-11} \text{kg}\), the acceleration due to gravity results in a force \(4 - 11 \times 10^{-9} \text{N}\) and corresponding deflection of few nanometers for spring constant \(\geq 8N/m\).
CHAPTER 3. MECHANICAL DESIGN ... 3.5. ELECTROSTATIC ACTUATION...

For a gap of $3\mu m$ acceleration of 5000g (Au) - 37000g (Al) is needed to close the switch. The deflection due to acoustic waves is given by $\Delta x = \frac{F}{k} = \frac{PA}{k}$ where $P$ is the pressure difference between the top and bottom plates and $A$ is the area. The resulting force for pressure levels up 0.1Pa, $A = 300 \times 100 \mu m^2$ is $3 \times 10^{-9} N$. With device packaging further lowering the pressures, the beams experience pressures negligible to cause reliability problems.

3.5.5 Voltage Break Down in MEMS Switches

The electric field in MEMS devices may vary from $1 - 3 \times 10^5$ V/cm for a gap height of $3\mu m$ and applied voltages ranging from 20 - 60 volts. The ionization break down does not occur because the effective distances are much smaller. The electron mean free path, $\lambda_e$ is given by

$$\lambda_e = \frac{T}{273} \frac{p}{P_c(V)}$$

(3.70)

where $T$ is the absolute temperature in Kelvin, $p$ is the pressure and $P_c(V)$ is the probability of collision. For $T = 273K$, and $p = 760$ torr, we have

$$\lambda_e = \frac{14}{P_c(V)}$$

(3.71)

For $g \leq \lambda_e$ (effective gap under actuation) there is no interference of electrons in device functioning. The measured values of $P_c(V)$ for oxygen and nitrogen are 20 - 40 for an electron energy corresponding to 10-100V, which gives $\lambda_e = 0.7 - 0.35 \mu m$, which is around the effective gap distance. Therefore ionization breakdown does not occur. However, at contact the threshold voltage of 5-10 volts gives rise to electric field $> 1 - 2$ MV/cm across the dielectric layer, which results in electron injection and dielectric charging. This leads to shifts in the actuation voltage and degrades the reliability of capacitive devices as discussed in Chapter 5.
3.5. ELECTROSTATIC ACTUATION...

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3.5.6 Dynamic Response of the Beams

The expressions derived for pull-in voltage and pull-out voltage of the beam describe the static response of the switch. Inclusion of finite mass and mechanical damping forces that arise from the viscosity of the air which must be squeezed out when the top plate moves down and drawn in, when the plate moves out, are necessary to describe the dynamic behavior of the switch in terms of resonance frequency and switching and release time. The enhanced model is shown in Fig.3.18 is based on the treatment given in [53], [58], and [59]. Due to the inevitable delays in device fabrication no dynamic measurements have been performed. However, this section is added because of its importance in switch design and fabrication.

Figure 3.18: Electrostatic actuator (a) with elements representing the inertia of the movable beam, mechanical damping and source resistance of the electrical network. (b) shows the equivalent circuit model in electrical and mechanical domain.
CHAPTER 3. MECHANICAL DESIGN ...  3.5. ELECTROSTATIC ACTUATION...

<table>
<thead>
<tr>
<th>Length ($\mu m$)</th>
<th>Width ($\mu m$)</th>
<th>$k$ (N/m)</th>
<th>$V_{pi}$ (V)</th>
<th>$Q_{no-holes}$</th>
<th>$Q_{holes}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1008</td>
<td>150</td>
<td>22.41</td>
<td>15.0</td>
<td>0.12</td>
<td>4.1</td>
</tr>
<tr>
<td>928</td>
<td>150</td>
<td>15.5</td>
<td>13.14</td>
<td>0.098</td>
<td>3.5</td>
</tr>
<tr>
<td>768</td>
<td>150</td>
<td>7.8</td>
<td>10.63</td>
<td>0.092</td>
<td>3.1</td>
</tr>
</tbody>
</table>

Table 3.3: Calculated quality factor with and without holes ($10\mu m\times10\mu m$) for three fabricated devices with electroplated Au-beams.

3.5.6.1 Damping and Quality Factor

The damping for a rectangular plate geometry is given by [58],

$$b = \frac{3}{2\pi} \frac{\mu A^2}{g_0^3}$$

(3.72)

where $A$ is the area of the device. Damping can be decreased by the use of holes in the top membrane. The effect of holes on damping coefficient as described in [60] is given as

$$b = \frac{12}{N\pi} \frac{\mu A^2}{g_0^3} \left( \frac{p}{2} - \frac{p^2}{8} - \frac{\ln(p)}{4} - \frac{3}{8} \right)$$

(3.73)

where $N$ is the total number of holes and $p$ is the fraction of open area on the plate. The formula for the quality factor ($Q = k/\omega_0 b$) is given by

$$Q = \frac{\sqrt{E\rho t^2}}{\mu (w/2)^2} g_0^2$$

(3.74)

where $\mu$ is the effective viscosity of air [59]. Table 3.3 gives the quality factor for three membranes of meander-based switching devices. The high Q-structures do not effect the switching speed of the devices if $Q \geq 3$. However it effects the release time of the switching. The large displacement correction for the damping coefficient has been given in [61]. The
corresponding $Q$ using above equations is

$$Q_e = Q \left(1.1 - \left(\frac{x}{g_0}\right)^2\right)^{3/2} \left(1 + 9.638\left(\frac{\lambda}{g}\right)^{1.159}\right)$$ \hspace{1cm} (3.75)$$

where $Q$ is the nominal small displacement quality factor at $g = g_0$ and the second term is a multiplication factor introduced by [61] to account for large displacement. This results in $Q_e \neq 0$ for $x = g_0$ and a permissible solution when the beam touches the pull down electrode.

### 3.5.6.2 Switching Time

The dynamic behavior of the beams is modeled by considering the beam of mass $m$ suspended by a spring with constant $k$ and squeeze-film effects represented by the damping coefficient $b$. The resulting one dimensional model is shown by Fig.3.18 and the governing equation is

$$m\frac{d^2x}{dt^2} + b\frac{dx}{dt} + kx + k_s x^3 = F_e + F_c$$ \hspace{1cm} (3.76)$$

where $k$ includes the bending and stress spring constant and $k_s$ the stretching effects of the beam. The damping factor is $b = k/w_0Q$ where $Q$ is given by Eqn.3.75. The voltage on the switch and external forces (for $x = g_0 - g$) are given as:

$$V = V_s - i(t)R_s = V_s - \left(C\frac{dV}{dt} + V \frac{dC}{dt}\right)R_s, \quad C = \frac{\varepsilon_0 A}{g_0 + \frac{t_d}{\varepsilon_r} - x}$$ \hspace{1cm} (3.77)$$

$$F_e = \frac{1}{2} \frac{\varepsilon_0 AV^2}{\left(g_0 + \frac{t_d}{\varepsilon_r} - x\right)^2}, \quad F_c = \frac{C_1 A}{(g_0 - x)^3} - \frac{C_2 A}{(g_0 - x)^{10}}$$ \hspace{1cm} (3.78)$$

where $F_e$ is the electrostatic pull-down force and $F_c$ represents the attractive *van der Waals* forces (first term) and repulsive nuclear forces (second term) between the metal and dielectric layer [62]. The constants $C_1 = 10^{-75} \text{Nm}$ and $C_2 = 58A^0$ determine the surface energy due to the
van der Waals attraction and equilibrium distance from the surface respectively. The switching time is obtained using Eqn.3.76 at \( x = g_0 \). The voltage on the switch is constant \( (R_s = 0) \) and the damping factor is given by Eqn.3.75. A closed-form solution for the switching time can be obtained for inertia limited system (acceleration limited) i.e. for the beams with a small damping coefficient and \( Q \geq 2 \). Under this approximation, the equation of motion becomes:

\[
\frac{m}{g_0^2 x} + k x = -\frac{1}{2} \frac{\epsilon_0 A V^2}{g_0^2}
\]

where force is considered to be constant and equal to the initial applied force. The solution gives the switching time \( t_s \) as

\[
t_s = 3.67 \frac{V_p}{V_s w_0}
\]

It has been shown that this closed form agrees well with numerical modeling and the measurements [63]. In a similar way the equation for damping limited system \( (Q \leq 0) \), can be derived with assumptions of constant damping with height and neglecting the acceleration and spring constant [64]. For electrostatic pull-in force we have \( F_e = b(dx/dt) \). The solution using integral methods for \( F_e = (\epsilon_0 A V^2)/2g^2 \) gives

\[
t_s = \frac{2b g_0^3}{3 \epsilon_0 A V^2} \approx \frac{9 V_{pi}^2}{4 w_0 Q V_s^2} \text{ for } V_s \gg V_{pi}
\]

Another estimate can be made by assuming a constant \( F_e \) and constant velocity approximation \( (dx/dt = g_0/t_s) \). This switching time estimate is given by

\[
t_s = \frac{2b g_0^3}{\epsilon_0 V_s^2} \approx \frac{27 V_{pi}^2}{4 w_0 Q V_s^2}, \text{ for } V_s \gg V_{pi}
\]

For the damping limited case the Eqn.3.82 tends to over-estimate the switching time and Eqn.3.81 tends to underestimate the switching time. Clearly, the switching response is a trade-off with the actuation voltage.
For meander based low actuation switches with $V_{pi}$ 3 - 15 V, the switching down time ranges from 35-75 $\mu$secs, assuming the actuation voltage of 25 volts. For the overall performance a reasonable balance between the actuation voltage and switching time should be envisaged.
Chapter 4

Electromagnetic Design Aspects of MEMS Switches

4.1 Introduction

In this chapter we present the electromagnetic design aspects of the RF MEMS capacitive switching devices under consideration. From the electrical perspective, both the meander based capacitive switches and symmetric toggle devices have essentially the same electromagnetic behavior, except for the structural implementation of the beams, actuation electrode configuration and dimensional parameters. In capacitive shunt configuration, together with parts of CPW transmission line, the microswitch can be represented by a lumped R L C or ‘T’ model [36]. The electrical model is used to characterize the switch RF performance by determining its resistance, capacitance and inductance. We discuss the methods of extracting switch resistance, inductance and capacitance values using s-parameters from simulated and/or measurement results. The dependence of the switch R L C, on its dimensions, material properties of the micro-bridge and transmission line characteristics are used to achieve optimum isolation and minimal return loss over the desired frequency bandwidth. As mentioned earlier,
the dimensions of RF MEMS switches are much smaller than the wave
length of the routed RF signals (30cm at 1GHz to 0.6cm at 50GHz). The
electro-magnetic interaction of the RF signal with switch structure being
negligible over the selected range of 8-30GHz, can be neglected. The device
layout and design is further discussed in Chapters 5 and 6. The following
section describes the design aspects of coplanar wave guide (CPW), ‘the
collector’ commonly used for RF MEMS switches.

4.2 Coplanar Wave Guide

Microstrip and CPW are the most commonly used connection media in RF
MEMS devices. The choice is guided by the switch type and configuration.
For example, microstrip implementation of a inline series switch results
in a more compact device. The basic structure of CPW is illustrated in
Fig.4.1(a) where the arrangement is assumed to be symmetric with strip
width $w$ and equal longitudinal gap $s$. A variant of the conventional CPW
is the finite ground coplanar strip line shown in Fig.4.1(b) which results in a
lower coupling of the adjacent lines as the ground (signal return current) is
not shared by two or more lines. All of the devices reported in this work are
implemented in conventional CPW. In general CPW is preferred because
of (1) the easier access to ground for MEMS shunt devices and surface
mounted devices in general (2) photolithographically defined structures
with relatively low dependance on substrate thickness (3) lower fabrication
cost and (4) reduced dispersion and radiation loss.

4.2.1 Characteristic Impedance - Synthesis Approach

The CPW line model and analysis has been reported in [65] and [66].
The discussed approach is based on [65] and can be easily implemented
in common mathematical tools for quicker synthesis of CPW. The basic
4.2. COPLANAR WAVE GUIDE

Figure 4.1: Structure of the coplanar waveguide (CPW): (a) conventional and (b) finite ground CPW (FGCPW). For conventional CPW, the specifications are: conductor material - Au, skin depth $\delta$ at 2GHz = 1.76$\mu$m, $w = 90\mu$m, $s = 75\mu$m, $h = 525\mu$m, CPW Ground = 4W.

Expression for characteristic impedance is

$$Z_0 = \frac{30\pi}{\sqrt{\varepsilon_{\text{eff}}}} \frac{K'(k)}{K(k)} \tag{4.1}$$

where $K(k)$ is the complete integral of first kind with modulus $k$ and is expressed as

$$K(k) = \int_0^{\pi/2} \frac{d\Phi}{\sqrt{1 - k^2 \sin^2 \Phi}} \tag{4.2}$$

$$k = \frac{a}{b} = \frac{w}{w + 2s}, \quad k' = \sqrt{1 - k^2}, \quad K'(k) = K(k') \tag{4.3}$$

The effective permittivity for a CPW is defined as $\varepsilon_{\text{eff}} = (c/v_p)^2$, where $c$ is the free space velocity and $v_p$ is the phase velocity or propagation velocity of the dominant mode on the CPW [67] and is given by

$$v_p(\text{vacuum}) = \frac{1}{\sqrt{\mu_0\varepsilon_0}}, \quad v_p(\text{medium}) = \frac{1}{\sqrt{\mu_0\varepsilon_0\mu_r\varepsilon_r}}, \tag{4.4}$$

where $\varepsilon_0 = 8.854 \times 10^{-12}$ F/m, $\mu_0 = 4\pi \times 10^{-7}$ H/m are the permittivity and permeability of free space and $\varepsilon_r, \mu_r$ are that of the medium respectively. An expression in terms of the dimensional parameters of CPW [68],
accurate approximately within 1.5% and valid for $h/s \geq 1$ is given as

$$
\varepsilon_{eff} = 0.5(\varepsilon_r + 1)\left(\text{tanh}[1.785\log(h/s) + 1.75]\right) + \\
(k_s/h)[0.04 - 0.7k + 0.01(1 - 0.1\varepsilon_r)(0.25 + k)]
$$

(4.5)

$K'(k)$ and $K(k)$ can be easily determined by using tables of elliptical integrals. However, expressions given in [68] which depend on the range of $k$ can be used and are

$$
\frac{K(k)}{K'(k)} = \pi \ln\left(\frac{2(1 + \sqrt{k})}{1 - \sqrt{k}}\right), \quad 0 \leq k \leq 0.707
$$

(4.6)

$$
\frac{K'(k)}{K(k)} = \frac{1}{\pi} \ln\left(\frac{2(1 + \sqrt{k})}{1 - \sqrt{k}}\right), \quad 0.707 \geq k \geq 1
$$

(4.7)

$k$ is obtained by using Equation 4.3 which express the $k$ in terms of gap $s$ and central conductor width $w$. Using Equations 4.1 and 4.6 we have

$$
x = 2\left(\frac{1 + \sqrt{k}}{1 - \sqrt{k}}\right) = \exp\left(\frac{Z_0\sqrt{\varepsilon_{eff}}}{30\pi^2}\right)
$$

(4.8)

In a CPW the field extends to the substrate ($\varepsilon = \varepsilon_0\varepsilon_r$) and partially to air ($\varepsilon = \varepsilon_0$), therefore $\varepsilon_{eff} = (\varepsilon_r + 1)/2$ can be used, to find $k$ from Eqn.4.8. To synthesize the CPW parameters, an initial value of the central conductor $w$ is selected based upon the switch geometry and fabrication process, to get the gap $s$ using Eqn.4.3. A more elaborate iteration is completed by using the effective permittivity given by Eqn. 4.5. The above analysis provides a first approximation of the impedance, lower by $2 - 5\%$ compared to the simulated values for a simple CPW configuration shown in Fig. 4.1.

4.2.2 Thickness of CPW and Mobile Beams

The intended actuation voltage is the primary criterion for selecting the mobile beam thickness. Other considerations are the mechanical resonance
frequency and out-of-the plane deflection resulting from the process induced stresses. In the flexure and torsion spring based designs considered in this thesis, actuation voltage is independent of the beam thickness as the spring constant depends on the flexure spring thickness. On the other hand the minimum, thickness of the CPW central conductor and ground area is a function of skin depth - a well known manifestation of RF in microwave frequency regime [69] and defined as the distance it takes the field to decay exponentially to $e^{-1} = 0.368$ or 36.8% of its value at the air conductor interface. The skin depth ‘$\delta$’ is given by

$$
\delta = \frac{1}{\sqrt{f \pi \mu \sigma}}
$$

(4.9)

where $f$ is the signal frequency, $\mu$ is the permeability of the medium surrounding the conductor, and $\sigma$ is the conductivity of the metal conductor. The internal impedance of the conductor for a unit length is $Z_s = R_s + X_s$, with $R_s = \sqrt{\pi f \mu / \sigma}$, and $X_s = \omega L_i$, as surface resistivity and internal reactance of the conductor. Skin depth represents the frequency depen-

Figure 4.2: (a) Skin depth $\delta$ vs frequency for commonly used metal layers and (b) decay of electric field vs conductor thickness in terms of $\delta$. 

![Skin depth graph](image1)

![Electric field decay graph](image2)
dent energy loss due to propagation in the resistive region within a skin depth and can be minimized by choosing materials with higher conductivity. Another important implication of skin depth is the exponential decay of energy within the conductor. In order to avoid energy dissipation into wafer the CPW conductor thickness is kept at four skin depths, where the amplitude decays to 1.8% of its incident value, at the lowest working frequency as shown in Fig.4.2(b). The skin depth for Au (resistivity = 2.4×10^{-6}\,\Omega\text{cm}) is 1.24\mu m at 4GHz. Therefore, thickness of electroplated gold CPW in all the devices is 5\mu m. The width of the ground lines is four to five times the central conductor width. Other parameters are shown in Fig.4.1.

4.3 MEMS Switches - Electrical Model

This section presents the circuit model of a capacitive shunt switch. The model is used to extract the switch parameters: capacitance (C) in bridge up and down state, resistance (R) of the metallic beam and inductance (L) from the simulated and measured s-parameters. The dependence of R, L and C on the bridge geometry gives insight into the device RF response and is helpful for the optimization in desired frequency range. The well known ‘T’ model (Fig. 4.3) [36] describes the RF behavior of a capacitive shunt switch by considering the metallic bridge membrane and CPW line impedance. The capacitive switch is placed in shunt configuration between the transmission line and ground. In up-state (on - state) it leaves the line almost undisturbed, resulting in zero insertion loss and ideally infinite isolation in the down-state (off-state), when the actuation bias is applied. In practice, (reported) switches have low insertion loss (−0.04 to −0.3dB) at 5 to 50GHz in the up state and acceptable isolation (more than −20dB at 10−50GHz). The model does not take into account the second order effects
4.3. Capacitive Switch

Figure 4.3(a) shows the top view of a MEMS capacitive shunt switch in CPW configuration. In the electrical equivalent model shown in Fig.4.3(b), the metallic bridge is represented by lumped R, L, C, elements with two
small sections of the transmission line representing the line impedance. In beam down-state the capacitance $C$ is mainly constituted by the bridge - dielectric - transmission line active overlap area, while in up-state the air gap between the beam and active overlap area results in a very small capacitance. The finite resistance of the bridge material is represented by $R$. The inductance $L$ is mainly dominated by the bridge portion above the gap between CPW ground and active overlap area. As discussed in Chapters 5 and 6, in the down state inductance increases the isolation and can be used to optimize the RF response over a narrow frequency range. The length of the transmission line sections depend on the definition of the reference plane and is equal to $W/2 + l$ where $l$ is the distance form the reference plane to the bridge edge and $W$ is the bridge width. The shunt switch total impedance is given by,

$$Z_s = R_s + j(\omega L - \frac{1}{\omega C})$$ (4.10)

with $C = C_{up}$ or $C_d$ depending on the switch position. The LC series resonance frequency of the switch is given by

$$f_0 = \frac{1}{2\pi} \frac{1}{\sqrt{LC}}$$ (4.11)

Depending on the frequency range of the application the switch impedance can be approximated as:

$$Z_s = \begin{cases} \frac{1}{j\omega C}, & \text{for } f \ll f_0 \\ R_s, & \text{for } f = f_0 \\ j\omega L, & \text{for } f \gg f_0 \end{cases}$$ (4.12)

From Eqn.4.12, it can be seen that the RLC model behaves as a capacitor below the LC series resonant frequency and as an inductor above this frequency. At resonance the model reduces to the bridge resistance $R$.  

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CHAPTER 4. ELECTROMAGNETIC DESIGN 4.3. ELECTRICAL MODEL...

<table>
<thead>
<tr>
<th>Overlap Area (µm²)</th>
<th>Air Gap (µm)</th>
<th>C_{up-cal} (fF)</th>
<th>C_{up-sim} (fF)</th>
<th>C_{fr} (fF)</th>
<th>Difference %</th>
</tr>
</thead>
<tbody>
<tr>
<td>150 x 90</td>
<td>3</td>
<td>33.65</td>
<td>43.7</td>
<td>10.05</td>
<td>30</td>
</tr>
<tr>
<td>250 x 150</td>
<td>3</td>
<td>110.70</td>
<td>135.05</td>
<td>24.35</td>
<td>22</td>
</tr>
<tr>
<td>300 x 300</td>
<td>3</td>
<td>265.65</td>
<td>312.4</td>
<td>46.75</td>
<td>17.5</td>
</tr>
<tr>
<td>150 x 90</td>
<td>1</td>
<td>100.9</td>
<td>125.4</td>
<td>24.15</td>
<td>23.9</td>
</tr>
<tr>
<td>250 x 150</td>
<td>1</td>
<td>332.0</td>
<td>400.4</td>
<td>68.4</td>
<td>20.6</td>
</tr>
<tr>
<td>300 x 300</td>
<td>1</td>
<td>796.9</td>
<td>918.4</td>
<td>121.6</td>
<td>15.25</td>
</tr>
</tbody>
</table>

Table 4.1: Up-state capacitance for three capacitive switches. The capacitances with a lower gap value are shown for comparison only.

For low spring constant meander switches (capacitive area = 150x90 µm) with $C_{up}=33.65fF$ and $C_d=3.9$ pF and $L=5pH$ the resonance occurs at 338GHz and 36GHz respectively. The inductance thus is unimportant in up-state. In off-state the inductance increases the isolation. The cut-off frequency $f_c = 1/2\pi C_u R_s$, e.g. for switch resistance $R_s = 0.15\Omega$ and $C_{up} = 33fF$ is 31 THz. However, since the inductance limits the down-state performance to much lower frequencies than $f_c$, the better value for upper frequency of operation is $2 f_0$, since MEMS switches result in reasonable isolation up twice the resonance frequency in down-state as shown in Fig.4.7. At resonance frequency, the minimum isolation is provided by the bridge resistance $R$. The electrical quality factor $Q_e$ of a capacitive microswitch is given by, $Q_e = 1/(\omega_0 R C) = 1/R\sqrt{L/C}$.

4.3.1.1 Scattering Parameters

The electrical performance of capacitive switches modeled as a two port network with a shunt connection, is characterized by the simulated and measured scattering parameters. In unactuated state the scattering parameters, $S11$ and $S21$ represent the return and insertion loss. In actuated state $S21$ gives the isolation of the switch. The scattering parameters (or...
reflection and transmission coefficients, in dB) defined in terms of switch and CPW characteristic impedance $Z_s$ and $Z_0$ respectively are [71]

$$S_{11} = -20 \log \left| \frac{-Z_0}{2Z_s + Z_0} \right|$$  \hspace{1cm} (4.13)

$$S_{21} = -20 \log \left| \frac{2Z_b}{2Z_s + Z_0} \right|$$  \hspace{1cm} (4.14)

Insertion loss (in unactuated state) and isolation (actuated state) are the important characteristics of a switch and are described in the following sections in terms of the design parameters of the switches.

4.3.2 Capacitance

The up-state capacitance of a RF MEMS capacitive switch determines the insertion loss, while the capacitance in down-state determines the isolation over the frequency range of application. In general a high $C_d/C_{up}$ ratio is desired; the ratio optimization is presented in Chapter 5. In the following paragraphs we discuss the factors affecting the capacitance ratio and present the simulation results for devices.

4.3.2.1 Up-state Capacitance

The up-state capacitance determines the reflected (return loss) and transmitted (insertion loss) power. In unactuated state, the bridge capacitance is ideally constituted by a air capacitance $C_{air}$ and $C_{diele}$, the capacitance due to dielectric layer on the signal line, in series. The total capacitance is:

$$C_{up} = \frac{A_0 \varepsilon_0}{g_0 + t_d/\varepsilon_r}$$  \hspace{1cm} (4.15)

where $A_0 = WL_0$, is the overlap area between the bridge and dielectric layer (Fig.4.3(a)), and $t_d = 1000^0$ is the dielectric ($SiO_2$, $\varepsilon_r$=3.9) layer thickness. However, the presence of fringing fields at the beam boundaries,
Figure 4.4: Insertion loss vs frequency for a RF capacitive switch in unactuated state, for a range of capacitances, with constant bridge inductance and resistance. The dashed curves represents the devices under consideration.

give rise to a fringing capacitance $C_{fr}$. This makes the actual microswitch capacitance substantially higher than given by Eqn.4.15. Table 4.1 summaries the calculated and simulated bridge capacitance for some of the fabricated device structures ordered by the area of overlap. The simulated capacitance $C_{up-sim}$ is higher than the $C_{up-cal}$, because in the simulations the effect of fringing capacitance $C_{fr}$ is also included. Also, as shown by tabulated values, the fringing field capacitance increases as the overlap area is increased. The larger gap height, $g_0$ reduces the fringing field effect, resulting in a smaller $C_{fr}$.

The effect of $C_{up}$ on return and insertion loss, in the unactuated switch is shown by the simulations in Figs. 4.4 and 4.5 for a range of $C_{up}$ values including those for the fabricated devices. The bridge resistance, inductance and line impedance are held constant. As shown by Fig.4.4 below 8-10GHz, the insertion loss is similar for all $C_{up}$ values. Also, the smaller up-capacitance has less insertion loss at frequencies above 10GHz, while larger capacitance values are useful at for lower frequency range only (insertion loss better than $-0.15$ dB at 8GHz). The return loss dependance
on the up-state capacitance is shown by Fig. 4.5, calculated using Eqn. 4.13. Devices with larger capacitance have a higher reflection coefficient ($C_{up} = 33\text{fF}$ results in $-40\text{dB}$ where as the device with $C_{up} = 265\text{fF}$ has $-22\text{dB}$ at 10 GHz). The reflected power can thus be reduced either by decreasing the overlap area or by increasing the bridge gap. Reduction in reflected power using gap height, increases the actuation voltage further. In Chapter 5 on design optimization we present the floating capacitor design where the capacitance in unactuated-state can be minimized further to have better return loss.

Although the up-state capacitance $C_u$ can be computed by using Eqn.4.15, it does not contain the fringing field capacitance $C_{fr}$. As such it is advantageous to extract the total up-state capacitance from the simulated or measured return loss plots. In the unactuated state, the bridge and dielectric layer on the signal line are separated by the air gap. Since the current flow though the bridge is negligible, the bridge impedance arising form the inductance and resistance can be neglected. Also considering the fact that below LC resonance frequency (Eqn. 4.12) the switch behaves as
a capacitor, the return loss given by Eqn. 4.13 can be rewritten as

\[ S_{11}|_{f \ll f_0} = -20 \log \left| \frac{-j\omega C_{up}Z_0}{2 + j\omega C_{up}Z_0} \right| \quad (4.16) \]

The simulated up-state return loss shown in Fig. 4.6(a) and described by Eqn. 4.16 can be used to extract the up-state capacitance, at a frequency below resonance. Considering the return loss (−30dB, for device with calculated \( C_{up} = 265.6 \text{fF} \)) at 2GHz in Fig.4.6(a), the corresponding \( C_{up} \) extracted using Equation 4.16 is 308fF, which agrees well with simulated value (312 fF) shown in the Table 4.1.

4.3.2.2 Effect of the Holes on Up-State Capacitance

All of the devices in the present studies, have been fabricated with array of holes in the bridge membrane. The holes facilitate dry-etching of thick photo-resist (\( \geq 3\mu m \)) under the beam, reduce the air damping and also decrease the mass of the resulting beams. The fringing fields cover the holes do not affect the capacitance if the hole dimensions are less than \( 3g_0 \), i.e. the simulated capacitance for a perforated and continuous sheet is same
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4.3.2.3 Capacitance in Actuated State

In the actuated state, the metal-insulator-metal capacitor formed by bridge-dielectric-transmission line central conductor connects the input port to ground. The capacitance is given as:

\[
C_d = \frac{A_0 \varepsilon_0 \varepsilon_r}{t_d} \tag{4.17}
\]

The dielectric layer thickness \(t_d = 80 - 100\text{nm}\) in general is small enough to ignore the fringing field capacitance. The simulated down state capacitance agrees within 0.2-2.0\% with the calculated values using Eqn.4.17 [72]. The effect of \(C_d\) on the isolation characteristics and resonance frequency is illustrated by Fig.4.7, at various values of capacitance with bridge inductance (L), resistance (R) and line impedance (\(Z_0\)) held constant. The dashed curves correspond to the fabricated devices. The isolation improves as the frequency increases and switches with larger capacitance show bet-

Figure 4.7: Isolation vs frequency (actuated state), for various values of down state capacitance with constant bridge inductance and resistance. The dashed curves represent the fabricated devices.
Overlap Area | Oxide thickness | $C_{d-cal}$ | $C_{d-sim}$ | Difference
---|---|---|---|---
150 x 90 | 100 | 3.9 | 3.74 | 4.1
250 x 150 | 100 | 12.9 | 12.35 | 4.2
300 x 300 | 100 | 30.10 | 28.6 | 4.9

Table 4.2: Calculated and simulated down-state capacitance for three capacitive switches.

As described earlier, the down state capacitance can also be extracted from the simulated or measured s-parameters. At operational frequencies below the resonance frequency, the bridge impedance $Z_s$ is mainly characterized by the down state capacitance $C_d$ as shown by Eqn.4.12. For $f \ll f_0$, the switch isolation given by Eqn.4.14 can be rewritten as

$$S_{21}|_{f \ll f_0} = -20 \log \left| \frac{2}{2 + j \omega C_d Z_0} \right|$$

Figure 4.6(b) shows the simulated isolation characteristics of a meander-based low-actuation voltage device with bridge length = 620 $\mu$m, overlap area = 300 $\mu$m x 300 $\mu$m, thickness = 1.5 $\mu$m and oxide layer thickness = 1000 nm. The values of $C_d$ extracted from the scattering parameters in Fig. 4.6(b) are shown in Table 4.2. As shown by the tabulated values the difference is within 5%. The accuracy of the extracted values can be further improved by selecting ‘S21’ at much lower operating frequency. In the table, $f \approx f_0/4$.

Based on Figs.4.4, 4.5 and 4.7, it can be deduced that to achieve a low insertion loss and better return loss when the capacitive switch is unactuated, the up-state capacitance, $C_u$ should be small as possible. To achieve a small $C_u$, the bridge width over the active area should be small too (Table 4.1). On the other hand, in actuated state, $C_d$ should be high as possible to achieve high isolation at lower frequencies. Higher $C_d$ implies larger
overlap area. These conflicting requirements lead to a compromise. In fabricated devices $C_d$ is further lowered by the surface roughness and extent of the overlap. For this reason, the conventional capacitive shunt design is useful only above 10 GHz. The following section further elaborates the issue. A possible solution to capacitance design optimization is suggested in Chapter 5.

4.3.3 Operating Frequency Regime and Capacitance Ratio

The capacitance ratio calculated using Eqn.4.15. and 4.17. is given as

$$\frac{C_d}{C_{up}} = \frac{\varepsilon_r g_0}{t_d}$$

(4.19)

where $\varepsilon_r$, $t_d$ and $g_0$ are the dielectric constant, dielectric thickness and gap height respectively. Below resonance frequency, a shunt configured switch (Eqn. 4.12) represents a low pass filter with corner frequency (-3dB point) $f_{-3dB} = 1/\pi C Z_0$. A series switch forms a high pass filter with corner frequency $f_{-3dB} = 1/4\pi C Z_0$. In switching from on (bridge - up) to off (bridge - down) state the corner frequency changes from $f_{IL}$ to $f_I$ (IL - Insertion Loss, on state $S21$, I - Isolation, off-state $S21$). The operating
regime characterized by a low insertion loss and high isolation is bound by the frequencies $f_{IL}$ and $f_I$, as shown in Fig.4.8(a). A wide operating regime, thus requires a high ratio of the corner frequencies in ‘on’ and ‘off’ state, expressed as

$$\left. \frac{f_{IL}}{f_I} \right|_{\text{shunt}} = \left. \frac{f_I}{f_{IL}} \right|_{\text{series}} = \frac{C_d}{C_{up}} \quad (4.20)$$

For a specified insertion loss ($IL_{sp}$) and isolation ($I_{sp}$) over a given frequency band bounded by $f_u$ and $f_l$, the condition is

$$\frac{C_d}{C_{up}} > \frac{f_u}{f_l} \sqrt{\frac{10^{0.1I_{sp}} - 1}{10^{0.1IL_{sp}} - 1}} \quad (4.21)$$

For instance, $IL_{sp} < 0.2\text{dB}$ and $I_{sp} > 30\text{dB}$ requires a capacitance ratio $> 146$ for a narrow band application, but increases to $> 450$ for $3:1$. The required capacitance ratio for a given bandwidth expressed as the ratio of upper and lower frequency bounds, for commonly used isolation specifications is illustrated by Fig.4.8(b).

A wide operating regime requirement, with a low insertion loss and better isolation, (Figs.4.5 and 4.7) thus leads to a high capacitance ratio $C_d/C_{up}$, as already concluded in previous section. The material and process
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Figure 4.10: Simulated (a) isolation and (b) return loss for actuated capacitive switch (250 × 150 µm) with different bridge resistances.

parameters, $\varepsilon_r$ and $t_d$ (Eqn.4.19) can be optimized to achieve higher ratio. However, high electric field across the dielectric layer limits the reduction in thickness to 100 – 150nm. High dielectric materials such as tantalum oxide ($\varepsilon_r = 25$) and strontium tantalum oxide (STO, $\varepsilon_r = 30 – 120$) [37] can be used with added process complexity. A ratio of 60 – 120 : 1 is common for devices with standard CMOS process compatible dielectric materials which is further lowered by the quality of the involved surfaces. As mentioned earlier, we the alternative approach is presented in the next chapter.

4.3.4 Switch Resistance

In an RF capacitive switch the current passes to ground through the bridge from the transmission line. In most of the capacitive shunt implementations, the bridge is anchored at CPW ground. A qualitative view of the average current distribution on the bridge and CPW is shown in Fig.4.15. The equivalent circuit model for the resistance $R_s$ of a capacitive microswitch therefore, is the resistance of the bridge in series with CPW line resistance
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<table>
<thead>
<tr>
<th>Bridge length (µm)</th>
<th>Overlap Area (µm²)</th>
<th>S21 (dB)</th>
<th>f₀ (GHz)</th>
<th>Rs (Ohms)</th>
</tr>
</thead>
<tbody>
<tr>
<td>770(ribbs)</td>
<td>150 x 90</td>
<td>-32</td>
<td>37.5</td>
<td>0.62</td>
</tr>
<tr>
<td>610(ribbs)</td>
<td>150 x 90</td>
<td>-45</td>
<td>22</td>
<td>0.28</td>
</tr>
<tr>
<td>670</td>
<td>250 x 150</td>
<td>-42.5</td>
<td>12.8</td>
<td>0.34</td>
</tr>
<tr>
<td>820</td>
<td>300 x 300</td>
<td>-39.5</td>
<td>8</td>
<td>0.42</td>
</tr>
</tbody>
</table>

Table 4.3: Bridge Resistance extracted from simulated (HFSS) s-parameters, for 5V devices.

\[ R_s = R_b + R_{cpw} \quad \text{where} \quad R_b = \frac{1}{2} \frac{\rho_b (L/2)}{A_{cr}} \quad (4.22) \]

and \( \rho_b, A_{cr}, L \) are respectively, the resistivity, cross-sectional area and length of the bridge. If current is assumed to be uniform at characteristic depth of penetration \( \delta \), (Eqn. 4.9), then the cross-sectional area of the bridge can be written as:

\[ A_{cr} = 2\delta(w_b + t_b) \quad (4.23) \]

where \( w_b \) and \( t_b \) are the bridge width and thickness respectively. The effect of total bridge resistance on the insertion and return loss, in unactuated and actuated states is demonstrated by the simulations shown in Fig.4.9 and 4.10. The values of capacitance (\( C_{up} \) and \( C_d \)), bridge inductance \( L \), CPW impedance \( Z_0 \), corresponding to one of the fabricated devices, are held constant. The bridge resistance is varied from 0.1 to 10 ohms. As seen in Fig.4.9(a), in the unactuated state, the bridge resistance has negligible effect on the return loss, below 10GHz. But it deteriorates from -50dB at 1GHz to about -10dB at 40 GHz, both for \( R= 0.1 \) and 1 ohm. Similarly insertion loss characteristics are also not altered by the bridge resistance below 10GHz. However, the effect of resistance on insertion loss is
more pronounced for $R$ greater than 5Ω, at frequencies higher than 10GHz (Fig.4.9(b)). In the actuated state, as the resistance becomes smaller, the isolation at resonance becomes sharper and deeper (Fig.4.10(a)). This is due to the fact that at resonance frequency, the bridge resistance is the only effective resistive component loading the CPW line. For $R_s$ less than a few ohms, the isolation curves are indistinguishable at frequencies lower than $3f_0/4$ GHz. Similarly, the return loss of the switch is better for lower switch resistance.

4.3.4.1 Determination of Bridge Resistance from measured S - parameters

In order to extract the bridge resistance either from the simulated or measured s-parameters of the devices the resonance frequency can be calculated using Eqn.4.11, or from the isolation plot itself, if the switch inductance and capacitance are unknown. As shown by Eqn.4.12, at resonance the isolation given by Eqn.4.14 can be expressed as:

$$S_{21}|_{f_0} = -20 \log \left| \frac{2R_s}{2R_s + Z_0} \right|$$  \hspace{1cm} (4.24)
Figure 4.12: (a) Isolation vs frequency for 150 μm x 90 μm actuated capacitive switch for various bridge inductance values. (b) Return loss vs frequency, the indistinguishable plots show the insensitivity of return loss to different inductance values.

The bridge resistance extracted using the simulated s-parameters for the fabricated devices are shown in Table 4.3. For the first two devices, though the overlap capacitive area is same, the area over the CPW gap is different. The resistance of the second bridge in Table 4.3 is calculated using Eqn.4.9, for gold conductivity \( \sigma_b = 4.0 \times 10^7 \Omega m^{-1} \) and resistivity \( \rho_b = 2.5 \times 10^{-8} \Omega m \) is 0.05Ω. The extracted resistances shown in the table are higher, because of the contribution from CPW and bridge.

4.3.4.2 Length of CPW

The CPW length affects the switch characteristics by changing the effective switch resistance. In Fig. 4.11(a) a comparison is made between two capacitive switches with same dimensions except the length of the CPW. Fig.4.11(b) shows one of the switches with the bridge suspended over a 325 μm long standard 50Ω CPW (75/90/75 μm), while the length of CPW is approximately three times more in the second case. The simulated isolation vs frequency response clearly demonstrates that the longer CPW lines
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Figure 4.13: (a) Return loss vs frequency plots for actuated switch, showing that at high frequencies the bridge impedance is dominated by bridge inductance. (b) Effect of the number of connecting ribs on isolation characteristics for 250x150 $\mu m$ actuated capacitive switch (simulated).

lead to inferior isolation and resonance occurring at lower frequencies as compared to the shorter CPW connections. In order to accurately extract the bridge resistance from the measured isolation plots it is necessary to have CPW lines as short as possible. The bridge resistance extracted from the isolation curves in Fig. 4.11(a) and using Equation 4.24 for the device with CPW lengths of 325$\mu m$ and 975$\mu m$ are 1.2$\Omega$ and 4.1$\Omega$ respectively.

4.3.5 Bridge Inductance and Average Current Distribution

The bridge of a capacitive RF MEMS switch presents a small inductance in series with the resistance $R_s$ and the bridge capacitance as shown in Fig.4.3. The bridge inductance is of significant importance when the switch is actuated, but has little effect in the unactuated state.

The effect of the bridge inductance on the switch characteristics in the actuated state is shown in Fig.4.12, for a device with capacitive area of $150 \times 90\mu m^2$. The bridge is assumed to be in contact with $SiO_2$ dielectric layer, to resemble actuated device. For instance, bridge inductances $L_b =$
4, 8, 12 and 16 pH are used while the down state capacitance $C_d$ and total bridge resistance are kept constant at 3.9 pF and 0.28Ω, respectively.

As shown in Fig.4.12(a), when the inductance is increased from 8 to 16 pH, the resonance frequency $f_0$ shifts from approximately 32 GHz to 22 GHz. The isolation (in dB) also changes with inductance, though this change is not very significant. However, for better isolation characteristics at lower frequencies higher inductance is preferred. The effect of inductance variation on the return loss is negligible as shown by four indistinguishable curves in Fig.4.12(b).

A simple and accurate method to model the bridge inductance, in the actuated state is to assume that the down state capacitance is large enough to represent a bridge short circuit with the CPW conductor line. At frequency $f \gg f_0$, where $f_0$ is the resonance frequency, $Z_s = j\omega L_s$ (Eqn.4.12). This is demonstrated graphically by Fig.4.13(a). The isolation vs frequency curves with $R_s = 0$ and 1Ω uses large capacitance, $C_d = 2 \times 10^9$ pF and inductance $L_s = 15$ pH. At frequencies $\geq 40$GHz, the curves for bridge
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Table 4.4: Bridge inductance extracted from simulated s - parameters, for 5V capacitive device with $C_d = 12.9pF$.

<table>
<thead>
<tr>
<th>Number of Ribs</th>
<th>Overlap Area (area = $50 \times 10\mu m^2$)</th>
<th>$S_{21}$ (dB)</th>
<th>$f_0$ (GHz)</th>
<th>$L_s$ pH</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>250 x 150</td>
<td>-35.934</td>
<td>13.8</td>
<td>2.49</td>
</tr>
<tr>
<td>3</td>
<td>250 x 150</td>
<td>-36.146</td>
<td>13.9</td>
<td>2.43</td>
</tr>
<tr>
<td>4</td>
<td>250 x 150</td>
<td>-36.257</td>
<td>14</td>
<td>2.40</td>
</tr>
</tbody>
</table>

resistance $R_s = 0$ and $1\Omega$ coincide, proving that at higher frequencies, $Z_s = j\omega L_s$. Therefore, at $f \gg f_0$, the isolation given by Eqn.4.13 can be rewritten as

$$S_{21} = -20 \log \left| \frac{2j\omega L_s}{2j\omega L_s + Z_0} \right|$$

(4.25)

Similar to capacitance and resistance, the bridge inductance can be extracted using simulated or measured isolation curves and the above equation. Table 4.4 shows the extracted bridge inductance for one of the studied devices with active overlap area of 250x150$\mu m^2$ and the number of connecting ribs varied from 2 to 4.

Figs. 4.13(b) and 4.14 show the simulated isolation characteristics of 250x150 $\mu m^2$ capacitive device, when the number of connecting ribs is varied from 1 to 4. The simulated average current distribution for the device in on and off state is shown in Fig. 4.15. The ribs, which are above the CPW gap connect the capacitive area to rest of the bridge on both sides and have length and width of 50$\mu m$ and 10 $\mu m$, respectively. The effect of inductance change with the number of ribs shown in Fig. 4.13(b) is further demonstrated on a finer scale in Fig. 4.14. There is a noticeable change in isolation and resonance frequency with change in the number of ribs (2–4), where as in the case of a single connecting rib it is quite large (Fig.4.13(b)). This is due to the fact the bridge inductance is mainly determined by the bridge sections above the CPW slots and not by the portion of the bridge over the central conductor [35]. The inductance in turn depends on the
current distribution on the bridge. As shown in Fig. 4.15 by the simulated average current distribution, the current concentration is higher on the outer ribs. This leads to a smaller difference in the inductance as the number of ribs is increased beyond three (Table 4.4). Also the shift in resonance frequency and change in isolation are less. In all meander-based switches the number of connecting ribs, thus has been chosen mainly to obtain a stiffer bridge without any warping and minimum overall mass. In the case of a single rib Fig.(4.13(b)) the isolation at 13GHz is −0.42 dB, making it a poor choice both for electrical and mechanical behavior of the switch. However, the area over the CPW, i.e. the number of ribs or width of strip can be effectively utilized for tunable filter applications over a narrow frequency band as further discussed in Chapter 6.
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Chapter 5

Meander based RF MEMS Switches

5.1 Introduction

In this chapter we present the design, fabrication and the characterization of serpentine meander based, electrostatically actuated low voltage RF - MEMS switches. The essential electro-mechanical design has already been presented in the previous two chapters. The device actuation voltage - a function of the flexure spring constant, fabrication process specifics and overall device geometry was discussed with the essential theoretical background - relevant to low actuation devices in Chapter 3. The electrical behavior of switches is characterized by RF response described by the S-parameters e.g. insertion loss, reflection coefficients, and transmission loss of the devices in, on and off - state. The dependence of s - parameters on the basic device electrical parameters e.g. bridge capacitance, resistance and inductance has been presented in Chapter 4. This chapter supplements the mentioned two chapters by presenting the overall switch design and design-optimization, fabrication, measurements and the problems encountered in the implementation during the study.
5.2 Actuation Voltage

The mechanical design of the electrostatic MEMS switches starts by considering the required DC actuation voltage. Most of the recently reported RF MEMS switching devices are designed for low loss applications that do not require very fast switching rates such as airborne or space communication [73] - [76]. The actuation voltages are in the range of 20-50 volts. The high voltage actuation mode makes the devices impractical for most applications especially in wireless communication as the additional requirement for high voltage increases the device/system size and offsets the effective monolithic integration advantages. In addition, the resulting high electric field, also leads to charge injection which strongly affects the long term reliability of capacitive switches. Therefore, reduction in pull voltage is always desired.

The actuation voltage for a fixed-fixed beam or air bridge is given by the well known equation described in Chapter 3, (Eqn.3.61) i.e.

\[ V_p = \sqrt{\frac{8K_z g_0^3}{27\epsilon_0 A}} \]  

(5.1)

where \( K_z \) is the equivalent spring constant of the suspended structure in the direction of preferred motion (z-direction), \( g_0 \) is the air gap between the beam and the actuation electrode, \( \epsilon_0 \) is the free space permittivity and \( A \) is the switch area where the electrostatic force is applied.

5.2.1 Actuation Voltage Optimization

As indicated by Eqn. (5.1), the possible approaches to optimize the actuation voltage are: to increase the area of actuation electrodes, diminishing the gap, \( (g_0) \) between the bridge and the bottom actuation electrodes, and designing the structure with low spring constant flexures. The actuation
area can only be increased to the extent before the compactness of the switch becomes a prevailing issue. The restriction on the reduction of gap $g_0$ is imposed by the return loss associated with the RF signal on the CPW transmission line (Chapter 4). Though gap reduction can be applied to low frequency applications ($< 10$ GHz), it adversely affects the high frequency on-state switch performance by compromising the insertion loss for a capacitive switch and off-state isolation for a series switch. The dependence of actuation voltage on device geometry is illustrated graphically by Figs.5.1 - 5.3. For this reason the most viable approach consists of the reduction in spring constant, which can be achieved almost without affecting the other switch parameters.

5.2.1.1 Meander Spring Design

The change in spring constant of the structure offers more flexibility since it does not considerably impact the size, weight and RF performance of a switch. As mentioned in Chapter 3 and shown by Fig. 5.6, the considered switches are anchored to the substrate by four serpentine springs, used
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Figure 5.2: (a) Actuation voltage as a function of $L_s$ and (b) resonance frequency vs pull-in voltage, at various actuation electrode areas.

to substantially lower the switch stiffness. If $k_z$ is the z-directed spring constant of each spring, the equivalent switch spring constant is then given by

$$K_z = 4k_z$$  \hspace{1cm} (5.2)

Compared to simple cantilever beams of equal total length, the springs have the additional advantage of occupying less space, but show higher spring constant [48]. However, as shown by Figs. 3.11, 3.13 and by Eqn. 3.50, the total spring constant of a switch structure can be significantly lowered by adding more meanders, without excessively increasing the required space. Each meander of the spring is defined as a set of four beams, two primary beams or connector beams denoted by $L_c$ (or simply by $a$ in Chapter 3) and two secondary beams, $L_s$ or $b$. The meander beam dimensions and material properties are given in Table 3.2. Suspension springs with constant ranging from 0.05 N/m to 15 N/m, for more than twelve combinations of switches with actuation voltages from 3 to 25 V, have been simulated.

The stiffness of a switch can be easily optimized by varying the meander dimensions. However, certain dimensions of meander beams have to be chosen in compliance with the adopted fabrication process. For example,
the selection criteria for the minimum width and thickness of the meanders (10 and 1.5 µm respectively) is based on the design rules and compatibility with fabrication process in this case ITC-irst ‘multi-user’ process [38] - [49].

Fig.5.1(a) shows the variation of actuation voltage with actuation electrode length (at constant width = 150 µm) for the secondary spring length ($L_s$) varying from 30 - 110 µm. Typically, for $L_s > 70µm$, the change in actuation voltage with change in electrode dimensions is small; < 25% for electrode length varying from 200-400µm. Fig.5.2(a) further illustrates the fact that for meander length $L_s$ in the range of 70-120 µm, actuation electrode dimensions can be optimized below 150 µm x 150 µm, to achieve comparatively compact devices. However, as illustrated by Fig.5.1(b) for $L_s > 70µm$, the mechanical resonance frequency ($f_0$) is less than 20kHz. The reduction in actuation voltage by increasing the actuation area also results in longer and slow devices, as compared to optimization using spring dimensions (Fig.5.2(b)). The above comparison suggests a explicit compromise between the speed and low actuation. Mathematically, the tradeoff can be expressed as

$$t_d = \sqrt{27/2} \left( V_p/2\pi f_0 V \right)$$

$$t_{up} \approx 1/f_0$$

where $f_0$ is the mechanical resonance frequency of the structure, $t_d$ is the switch down time and $t_{up}$ is the time for the beam to restore to its zero bias position. Normally, $t_{up}$ is quoted as the switching speed of the device.

The more compliant devices are also susceptible to external shocks. Assuming that the shock dynamics can be modeled by a spring-mass system (Fig.3.14), the condition for the shock resistance is that the induced acceleration is much smaller than the critical acceleration for the device i.e.

$$A_{cr} \ll \frac{4\pi^2 f_0^2 g_0}{D}$$

$$109$$
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Figure 5.3: Different types of meanders with simulated spring constant shown at the bottom of figure (a). (b) shows the simulated spring constant for type (4), with meander length (Ls) varying from 200 to 260 µm and Lp = 15, 25 µm, width w = 5, 10 µm.

where $D$ is the dynamic shock factor and depends on the shape of the shock pulse ($1 < D < 2$) [56] and $g_0$ is the gap height. The specifications of the critical acceleration vary for different applications (e.g. cell phone 5000g to 50,000g). In present design approach, for a switch with mechanical resonance frequency of 20KHz and gap height of 3 µm, the threshold acceleration, is greater than $2 \times 10^4 g$, which is sufficient for most applications. In general, it is desired to have pull-in voltage as low as possible and resonance frequency as high as possible. As discussed above and shown by Fig.5.2(b), a tradeoff is inevitable. By incorporating a third electrode, on top of the existing bridge, the switch can be made insensitive to external shocks and vibrations [55]. The fabrication process for a two bridge structure is comparatively complex. In order to overcome the problem, we propose a symmetric toggle device, with a single bridge structure and additional ‘hold’ electrodes, as described in the next chapter.
5.2.1.2 Other Meander Structures

The other, investigated meander structures, along with the structures used in switch design, are shown in Fig.5.3. All the meanders have been simulated as free-end cantilevers, with a concentrated \( z \)-directed force applied to the tip of the spring. The type-4, which finally has been used and discussed in details in Chapter 3, has the lowest spring constant for comparable dimensions. The selected type is also easier to modify and fabricate. The out-of-plan deflections or the warping levels arising because of the residual stress in the fabricated switches, can be minimized by stiffer structures; particularly, type-2 and type-3. However, as discussed in the section on beam design optimization, the same can be achieved by selectively changing the thickness of the main switch body. Therefore, in all the devices, type - 4 meanders have been used.

5.3 Device Fabrication

In this section, we describe in detail the device fabrication process for the MEMS switches. The basic process is essentially the same for all devices. The modifications or the optional steps are mentioned in the appropriate sections. Some processing steps such as electroplating and physical vapor deposition are particularly important, hence are elaborated more as compared to other steps.

5.3.1 Device Specifications

The present studies of the low actuation RF MEMS devices are motivated by the on-going RF-MEMS activities in *ITC-irst, Povo Italy*. The main aim of the first ‘test’ run devices was to explore the fabrication feasibility, within the existing process constraints, which actually was devoted to
micro-machined high actuation voltage, fixed-fixed beam based switches with comparatively shorter bridge-lengths. The primary specifications of devices were to have actuation voltage below 20 volts, isolation better than -20 dB, and insertion loss less than -0.2dB over the intended frequency range of 10 - 30GHz. As mentioned earlier, with low actuation voltage (which implies low spring constant) the resonance frequency is a trade off. Therefore, in this meander based approach switching speed has been of secondary importance. However, for the first device run mechanical resonance frequencies above 8 KHz are envisaged. The specifications of the devices for the first fabrication-run are summarized in Table 5.1.

### 5.3.2 First Fabrication Process

As mentioned in the chapter on introduction to the state-of-art, the fabrication process for the RF-MEMS switches is based on surface micro-machining and modified CMOS processing steps. The process design/flow and the choice of movable structural beams is based on the process - compatibility with standard IC technology and the critical switching parame-
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5.3. DEVICE FABRICATION

Figure 5.4: RF-MEMS switch fabrication process flow - I

Figure 5.5: RF-MEMS switch fabrication process flow - II
Excluding the packaging, most of the reported RF MEMS switches are realized using five mask levels with a process flow sequence, which may vary from one implementation scheme to other. While, selection of the wafer, overall thermal budget, material etching and deposition techniques are the general process compatibility issues, the contact and structural material considerations determine the switch parameters including the contact resistance, metal sticking behavior, life time and environmental and packaging compatibility.

The first version of the process consists of seven mask levels. The additional mask levels are required to incorporate biasing resistors and DC-blocking capacitors. In the reported devices only biasing resistors are used in conjunction with the actuation electrodes. Since all the switches are ‘three’ terminal devices, the DC blocking capacitors are not needed, but are a part of the ITC-first ‘multiuser’ process run, under which the switches have been fabricated. The processing sequence for the realization of switches with gold as structural layer is shown in Figs. 5.4 and 5.5 and is described as follows.

- In general a low-loss high-resistivity substrate is the starting material for RF-MEMS devices including the switches. Thermally oxidized high resistivity silicon wafers (> 3kΩcm) are preferred [14] [23]-[36]. In present work, for all the devices 5kΩ, p-type, (Si < 100 >) silicon wafers have been used. In the first step (Fig. 5.4(a)) an isolation layer of 1000 nm thermal field oxide (FOX) is grown (at 975°C, for 10 minutes in N₂ ambient and at 975°C, for 385 min, in H₂O ambient). In order to reduce the fixed charges, the field oxide is then annealed at 975°C for one hour in N₂ ambient.

- The next step defines the polysilicon actuation electrodes, connection lines and underpass (Fig. 5.4 (a)). The polysilicon - titanium nitride
(TiN) underpass has only been used in the first fabrication run. For the switches with reinforcing ribs and symmetric toggle switch, it was replaced by multilayer metal consisting of Ti - TiN - Al:1%Si - Ti - TiN. The un-doped polysilicon layer (630 nm) is grown by LPCVD (low pressure chemical vapor deposition process) at 620°C. The polysilicon deposition is followed by boron ion implantation (BF$_2$ at 120 KeV, dose = 5.0E15, resistivity 245 Ω after the drive in cycle). The first lithography at this stage defines the polysilicon actuation electrodes for the switches, resistors and connection lines. The polysilicon dry etching is followed by the resist removal (plasma ashing) and the boron drive-in at 925°C in N$_2$ ambient for one hour.

- The electrical isolation between the switch actuation electrodes and the metallic beam structure (to be electro-deposited) is provided by a layer of 300 nm thick oxide deposited by pyrolytic oxidation of tetraethylorthosilane (TEOS) using LPCVD at 718°C. The electrical properties of TEOS oxide such as fixed charges and impurities have a strong impact on the reliability of the devices. The presence of impurities and fixed charges coupled with high electric fields across the oxide layer, may result in a drift in the pull-in during the subsequent actuation cycles or total device failure as discussed in next section. The contact holes opened in TEOS, by the second lithography connect the polysilicon actuation electrodes and resistors to the TiN - gold contact pads.

- After ashing the photoresist mask, the next step is to realize the multilayer underpass - which joins the two portions of the CPW central conductor or signal line under the bridge. The multilayer is also used to realize the metal capacitors. Wetting of the wafers in isopropyl alcohol(IPA), deionized water (DIW) and pre-metal deposition clean
in 8% HF proceeds the sputter deposition of Ti (30 nm) and TiN (50 nm, reactive sputtering, Ti sputtered in \( N_2 \) ambient) both at 400°C. The layers act as adhesion promotor and diffusion barrier respectively (Fig.5.4(b)). In the next step, layers of \( Al1\%Si \) (450 nm) followed by Ti (30 nm), are deposited by sputtering at room temperature. Finally a smooth capping layer of TiN (80 nm) is sputter deposited at 300°C. The total thickness of the multilayer structure has to match the polysilicon actuation electrode height. Under actuation, the height difference between the multilayer structure and actuation pads lead to complex, non-uniform bending of the beams as described in the section on measurements. The height difference eventually results in lower down state capacitance and a shift in resonance frequency of the capacitive shunt devices. The next lithography step followed by metal dry etch defines the underpass lines and the diffusion barrier (TiN) on the polysilicon contacts (Fig.5.4(b)).

- The insulation layer on the multilayer metal underpass is provided by the LPCVD low temperature oxide (LTO, 100 nm). The down state capacitance of the capacitive shunt switches depend on the surface roughness and the dielectric properties of this oxide layer. Via holes which connect Al in the multilayer to the gold CPW central conductor are subsequently defined by lithography and dry etching. A sufficient ‘over-etch’ time is incorporated into the process in order to ensure the removal of TiN barrier layer and exposure of the Al underneath.

- At this stage, the process for devices with an electrically floating metal layer (an additional layer of metal, on top of the LTO, which corresponds to the capacitance area of the shunt switches), has another lithography and etching step to define the floating metal area. In the seven mask process (for conventional shunt switches) this step is not
required. In the case of ohmic contact switches this extra step is the precursor for noble metal deposition on the exposed electrical contacts which provide a low resistance electrical path.

- The sacrificial layer used for the construction of the suspended metal structure is defined in this step using thick positive photoresist (HPIR 6517HC). The resist is baked (pre-bake thickness 3685 nm, post-bake thickness 3000 nm) at 200°C for 30 minutes in order to get well rounded contours for better step coverage (Fig.5.5(c)). As a seed layer for electrochemical Au-deposition, a 10/150 nm thick Cr/Au layer is deposited by PVD. The Cr/Au deposition step is critical in order to get metal structures with minimal residual stress. The deposition of Cr is followed by Au, in the same evaporation chamber without air exposure, to avoid any chromium oxide formation. However, Cr gets oxidized does during the resist dry-etching process. And the diffusion of chromium into gold and formation of oxide layer change the Young’s modulus of the layers. This may also resulting in significant warping of the free structural members. This processing step is further discussed in Chapter 6.
The movable bridge structure is realized in two electroplating steps. In the first step, the springs and the main switch body is electroplated - up to a thickness of 1500+ nm. The bridge is defined on a 4µm thick photoresist (maP225). At this stage, the 3µm thick ‘spacer layer’ already exists. Though the required bridge thickness is only 1.5 µm, thicker resist is used to achieve better step coverage over the spacer area. After a quick flash in oxygen plasma to remove any organic residue a 1.5µm thick layer of gold is selectively grown in gold sulphite (Ammonium gold sulphite) bath. The quality of the electroplated gold layers is a strong function of the electroplating bath parameters and therefore need precise monitoring. The electroplating current is maintained at 68.5 mA, using general purpose potentiostat (2051 AMEL). The pH of the solution is adjusted between 6.8 - 7.4, by adding $H_3PO_4$ solution. The temperature of the bath is maintained at 55°C.

After the wet removal of the first electroplating mask, the CPW lines and anchor posts for the bridge structure are defined in 5 µm thick resist (maP225) baked at 100°C for 3 minutes. The resist is exposed in three steps of 10 seconds each followed by a cool down time of another 10 seconds, to avoid excessive heating of the resist. In the following electroplating step, using the same solution and deposition parameters for the bath, the bridge thickness is selectively increased by approximately 3.8 µm. The CPW total thickness, including the selective bridge portions, thus becomes 5 µm plus. The last electroplating mask and the seed layers are removed by wet etching. The removal of the Cr/Au whiskers is done by a dip in diluted (1:3) *aqua regia* for 5 seconds. The structure then is sintered at 190°C, for 30 minutes in $N_2$ ambient. The sintering step is important to minimize the residual stress gradient in the beams. As an optional step, the
wafers are coated with a 2.5 \( \mu \text{m} \) thick photoresist and pre-diced. This allows the easy separation of the dies, while the wafer can still be handled as a whole during the final processing steps. Finally, the movable structures are released by ashing of the spacer resist with a modified plasma ashing process (for 20 minutes in oxygen plasma at 200\(^\circ\)C). The dry etch release is preferred in order to avoid the sticking of the structures to the wafer surface. A SEM micrograph of a fabricated device switch with details of the suspension spring and underpass is shown by Fig.5.6. The whole process consists of about 120 steps, and the turn around time of about 10 weeks.

5.4 Measurements and Results

In this section we present the measurements together with discussions on the test structures and meander based low actuation switches. The test structures are the precursors to the low actuation voltage switches under consideration and are typically used to extract the fundamental electro-mechanical properties of the materials.

5.4.1 Measurement Setup

The experimental setup used for studying the electro-mechanical behavior and the experimental verification of the parameters such as actuation voltage and capacitance is built around a Karl Suss PM8 probe station as shown by the block diagram in Fig.5.7. With minor modification the setup can also be used for measuring the dynamic response of the devices and test structures. The set-up consists of an impedance analyzer (HP 4292A), a semiconductor parameter analyzer (HP 4145B), a RF signal generator and general purpose instrumentation to generate the bias pulses for the device actuation and monitoring of the output signal. The
bridge driver is used to generate ‘dual pulse’ actuation wave form with amplitude ranging from 10 - 200 volts and variable time period (10e-5 to 1 sec). For capacitance versus voltage measurements of the switches and test structures, only impedance analyzer and the actuation sweep source (semiconductor parameter analyzer) are needed. The impedance analyzer provides a DC sweep voltage that can be adjusted from 0 to ±35V with 10mV resolution, and a variable frequency synthesizer for test frequencies from 5KHz to 13MHz. The analyzer input is applied across the variable capacitor constituted by the movable structural member of the bridge and CPW transmission line (signal ‘high’ - to the CPW central conductor and ‘low’ to the bridge, which is connected to the CPW ground). For dynamic testing the test structures and the devices are actuated by voltage pulses with adjustable amplitude and duty cycle. In order to reduce the charging of the dielectric layers, actuation is achieved by a high magnitude pulse (1.0
5.4.2 Test Structures

The test structures provide a good insight into the electro-mechanical behavior of beam based devices, with easy access and shorter measurement cycle. Fig. 5.8(a) shows one of the test structures consisting of five micro-bridges with length ranging from 450 $\mu$m to 850 $\mu$m. The width (150 $\mu$m), thickness (1 $\mu$m) and the air gap (3 $\mu$m) are kept constant for all the structures. All the micro-bridges are actuated contemporaneously, by a single actuation electrode and corresponding capacitance vs voltage characteristics are obtained for all the five micro-bridges. Since there is no surrounding CPW ground, the test structure configuration provides capacitance measurements, closer to the designed values.

The measured CV plots for three test structures are shown in Fig. 5.8(b). The measurements related to the smallest micro-bridge, which...
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Figure 5.9: (a) Electric charge induced shift in pull-in voltage, the second measurement is made after 30 minutes. (b) capacitance change ascribed to abnormal bending of the micro-bridges, only left half of the switch is shown.

has actuation voltage greater than 35V has been excluded as it needs the external source with sweep voltage of larger range. The agreement between the measured and calculated pull-in is close (1-3%), except for the longest bridge where difference is more than 16%. The variation in the pull-in voltage and associated capacitance especially for bridges (Fig. 5.8(b)) with larger length is attributed to the variation in electroplated gold thickness over different locations in the wafer, which changes the spring constant of the structures.

5.4.2.1 Charging and Beam Deformation

The effect of accumulated charges in the dielectric layer and the complex deformation of the micro-bridges is illustrated by Fig.5.9. The measurements for a test structure in Fig.5.9(a) are performed over gap of 20 - 30 minutes. After the initial actuation, the shift in pull-in was observed even after a few hours. This kind of charging effects were especially observed in the devices with nitride as dielectric layer, which subsequently abandoned for its intrinsic instabilities.
The charge injection and trapping in the dielectric layers is a major cause of stiction in capacitive devices and undermines the reliability especially in high actuation voltage switches. The three areas where the charges can be trapped in a MEMS capacitive switch are the interface traps between the metal and the dielectric layer, the bulk traps inside the dielectric layer and the surface-state traps on the top of the dielectric layer [77]. The dielectric layer charging is a complex phenomenon and may arise because of the applied mechanical, ionizing, or electrical field stresses. The crystallographic imperfections or defects due to tooling (dislocations, non-stoichiometry) or irradiation under ionizing beams may also give rise to charges and charge trap centers [78]. For an actuation voltage of 30-60 volts and oxide thickness of 1500Å, the electric fields can be as high as 2-4 MV/cm. Under such field strength it is possible for the charges to tunnel into the dielectric layer under a phenomenon similar to Frankel - Poole injection [79], [80]. The recombination time for the trapped charges is very slow, of the order of seconds to days. It is for this reason that low actuation RF - MEMS switches are important.

A deformation giving rise to multiple steps in the CV curve, as shown in Fig.5.9(b) arises because the beam bends in steps instead of a uniform snap down over the transmission line. In this situation, the central area comes in contact first, followed by the rest of the beam as the actuation voltage is increased further. The meander based beams, which are more compliant as compared to fixed-fixed test structures, are designed to overcome this kind of subsegment bending. However, the residual stress gradient can lead to more complex bending profiles as will be discussed in the following sections.

The transfer and distribution of charges in metal - dielectric and dielectric - air interface, as shown in Fig.5.10(a) can also make the devices insensitive to further actuation or cause a complete failure [81]. In this case
the charges are transferred from the metallic beam to the surface states and the force applied on the beam is reduced. After the actuation voltage is removed, the charge remains in the surface states, causing an increase in the pull-in voltage. In high actuation devices the effect of charge injection can be minimized by reducing the hold-down voltage to 10 - 15 volts once the device has been actuated. This is shown in Fig. 5.10(b) for a bridge with pull-in voltage of 45 volts. The bridge is actuated with a pulse, whose amplitude is reduced to half once the device is actuated. In this case the drift in the pull-in for more than 1000 actuation cycles is less than 4 volts.

5.4.3 Capacitance and Voltage Measurements: RF MEMS switches

In this section we present the measurements on the first type of meander based low actuation switches. A comparison between the measured and analytically computed pull-in voltage (using Eqn. 5.1) is shown in Table 5.2. The table also shows the measured on and off-state switch capacitances.
The discrepancies between the designed and measured capacitance and pull-in voltages arise mainly because of the process related parameters such as: (1) thickness of the structural gold layer; the change in $V_{pi}$ is proportional to change in meander thickness, (2) oxide thickness; capacitance in down state is inversely proportional to the low temperature thermal oxide (LTO) thickness on the underpass and is also a function of surface roughness, moreover the measurements also include the parasitic capacitances present in the real structures and are ignored in the ideal case (3) variation in spacer (sacrificial photoresist) resist thickness which changes the effective gap height (4) the height difference between the underpass and actuation electrodes and (5) the residual stress gradient in the structural layer.

As seen in the table (5.2), the measured actuation voltage differs from 5 to 40% from the analytically calculated values and in most of the devices measured values are lower. Fig. 5.11(a) shows the location of the devices on the wafer, while (b) shows the measured gold thickness and air gap in a fabricated wafer. The measured thickness of the first Au-layer, which defines the springs and the movable beam structure, varies from 1550 nm

### Table 5.2

<table>
<thead>
<tr>
<th>Bridge Length (µm)</th>
<th>Actuation Cal. (V)</th>
<th>Actuation Meas. (V)</th>
<th>% Deviation (Act. Meas. Voltage)</th>
<th>$C_d$ (pF)</th>
<th>$C_{up}$ (pF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1008</td>
<td>15</td>
<td>12.8</td>
<td>3.3</td>
<td>3.54</td>
<td>3.0</td>
</tr>
<tr>
<td>688</td>
<td>16</td>
<td>12.6</td>
<td>37.5</td>
<td>2.75</td>
<td>2.1</td>
</tr>
<tr>
<td>928</td>
<td>13.5</td>
<td>11.0</td>
<td>24.07</td>
<td>3.16</td>
<td>2.8</td>
</tr>
<tr>
<td>688</td>
<td>14</td>
<td>10.6</td>
<td>42.8</td>
<td>3.74</td>
<td>2.5</td>
</tr>
<tr>
<td>1048</td>
<td>10.3</td>
<td>8.7</td>
<td>19.9</td>
<td>3.34</td>
<td>2.66</td>
</tr>
<tr>
<td>768</td>
<td>10.8</td>
<td>8.5</td>
<td>21.3</td>
<td>3.27</td>
<td>2.60</td>
</tr>
</tbody>
</table>

The discrepancies between the designed and measured capacitance and pull-in voltages arise mainly because of the process related parameters such as: (1) thickness of the structural gold layer; the change in $V_{pi}$ is proportional to change in meander thickness, (2) oxide thickness; capacitance in down state is inversely proportional to the low temperature thermal oxide (LTO) thickness on the underpass and is also a function of surface roughness, moreover the measurements also include the parasitic capacitances present in the real structures and are ignored in the ideal case (3) variation in spacer (sacrificial photoresist) resist thickness which changes the effective gap height (4) the height difference between the underpass and actuation electrodes and (5) the residual stress gradient in the structural layer.

As seen in the table (5.2), the measured actuation voltage differs from 5 to 40% from the analytically calculated values and in most of the devices measured values are lower. Fig. 5.11(a) shows the location of the devices on the wafer, while (b) shows the measured gold thickness and air gap in a fabricated wafer. The measured thickness of the first Au-layer, which defines the springs and the movable beam structure, varies from 1550 nm
at the center to 1930 nm at the edges of the wafer, against the nominal 1500 nm. As shown in Fig.5.11(b), the spacer thickness, which defines the air gap, was also measured to be different at the center and the wafer edges. The increase in meander thickness and gap height results in stiffer devices with higher pull-in (Chapter 3 Eqn.3.47, Fig.3.17). However, as shown by Table 5.2 and the C-V curves in Figs. 5.12 and 5.13, the pull-in in 90% of the switches is lower than expected for spring thickness of 1.7 µm. The low pull-in is probably caused by a residual stress gradient related deformation of the bridges as they are released from the spacer. The stress arises because of the composite effect of Cr (used as seed layer) diffusion into Au and oxidation of bottom Cr layer, during plasma etching and sintering process at 200°C. The resulting tensile stress lowers the air gap height and hence the pull-in voltage. The residual stress (which is further discussed in Chapter 6) in micro-fabricated beams is a process specific phenomenon and can be minimized by optimization and precise monitoring especially, of the Cr-Au sputter deposition and Au-electroplating.
Figure 5.12: The measured switch capacitance as a function of applied voltage for two devices, from different locations on the same wafer.

Figure 5.13: Measured C-V curves, (a) reduction in the off-state capacitance $C_d$ as voltage is increased beyond the pull-in. (b) reduction and multiple actuation as the beam overlaps contact area.
5.4.3.1 Change in Capacitance

In this section we discuss the behavior of the switch membrane in presence of residual stress and the parasitic impedance arising because of the CPW and finite substrate resistance. We present a model to discuss the beam deformation and ensuing changes in actuation voltage and on-state capacitance.

The measured down-state capacitance for the switches is shown in Table 5.2. Figs.5.12 and 5.13 show measured C-V curves, demonstrating the peculiarities of some of the switches. The down state capacitance, is lower by 10 to 30% as compared to the simulated values. The main factors which contribute to the reduction are: increase in oxide thickness (Fig. 5.11(b)) above 14%, surface roughness of the underpass, deformation of the beam and parasitic capacitances. The large increase in up-state capacitance is mainly due to the complex deformation of the bridge, reducing the gap height and due to the parasitic capacitances, which can even offset the overlap capacitance of the bridge and underpass. Surface profile measurements before the sacrificial layer removal show a uniform profile. However, residual stress gradient results in a pre-deflected beam when released e.g. after the sacrificial layer removal. The membrane profile, before and after actuation is depicted by schematic in Fig. 5.14 with exaggerated y-dimensions. In view of the symmetry only half of the bridge membrane is considered. The reduction in the gap height is more at the central area which contributes to the increased bridge capacitance in up-state, together with parasitic capacitances.

In conventional ‘dielectric on active overlap area’ design, the membrane should be held flat over the electrodes, under the bridge, by the electrostatic force of attraction. But in the present case it probably takes a curved shape as it falls over the underpass area. The difference of height between
the actuation electrodes and underpass (Fig.5.11(b)), significant rigidity of the beam and the absence of any force to pull it down (in the center) makes the beam bend upwards at the center. The residual air gap ‘f’, thus introduced, contributes significantly to the bridge down state capacitance. The gap height can be estimated by considering the moments generated by the forces, $F_f$ and $F_t$ w.r.t the turning point O. At equilibrium (Fig.5.14),

$$F_f a_2 = F_t f \quad (5.6)$$

In the expression, $F_f$ is the vertical force at the midpoint between the actuation electrode and the underpass generated by the flexure of the membrane, while $F_f$ is the horizontal force at the mid point of the underpass, due to the tensile stress in the bridge layer. To first order, $F_f$ can be estimated by considering the membrane at the mid point between the actuation electrode and the underpass, as a cantilever beam, fixed at one end. The force necessary to bend the free end by half the step height can be expressed as:

$$F_f \left( \frac{a}{2} \right)^3 = \frac{H_s}{2} 3EI_y \quad (5.7)$$

where $E$ and $I_y(= h^3w/12)$ are the Young’s modulus and moment of inertia of the beam respectively. The horizontal force $F_t$ is the traction force.
exerted by the tensile stress in the bridge and is given by

\[ F_t = \sigma w h \]  

where \( \sigma \) is internal tensile stress and \( w \) and \( h \) are the width and the thickness of the beam. Using the above two expressions we have, for the residual air gap,

\[ f = \frac{E}{1 - \nu} \frac{1}{\sigma_{e}} \left( \frac{h}{a} \right)^2 \frac{H_s}{2} \]  

As the actuation voltage is increased the air gap height increases because of the turning force at point O, reducing the down state capacitance further. The effect is shown by measured CV curves in Fig.5.13(a) for actuation above 11 Volts. Some of the membranes are observed to make multiple actuations (Fig.5.13(b)), probably a cumulative effect of the internal stress, applied electric field and defects local to the devices, placed near the wafer edge. The capacitance ratio in switches is typically less than 2, implying a very low air gap.

The residual air gap and the associated residual capacitance ‘\( C_{resi} \) can be estimated by comparing the capacitance of a ‘fabricated as actuated’ switch and a normal air-gap device, having same dimensions. Fig.5.15(a) shows the switch cross section with intended and unintended parasitic capacitances. Considering the shunt switch as a purely capacitive circuit (Fig.5.15(b)), in up-state it can be represented by capacitance \( C_g \), the capacitance due to the intended air gap, in series with \( C_{ox} \), the active overlap area oxide capacitance and a parallel capacitance \( C_p \), mainly the parasitic capacitance of the CPW ground plane. In the ideal unactuated case, \( C_p = 0 \); \( C_{up} \approx C_g \), assuming the fringing contribution to be negligible. In presence of parasitics, since \( C_{ox} \gg C_g \), and \( C_p \gg C_g \); the series capacitance is essentially dominated by \( C_g \); the up-state capacitance can be expressed as, \( C_{up} = C_p + C_g \approx C_p \).
The equivalent capacitance of an actuated switch with residual air gap can be obtained by adding a series “residual air gap capacitance” $C_{resi}$, to $C_{ox}$, in the ‘ideal overlap’ circuit shown by Fig.5.15(c). The equivalent capacitance of the switch thus becomes $C_d = C_p + (C_{resi}C_{ox})/(C_{resi} + C_{ox})$. Thus, having a test switch, designed ‘as actuated’ to get $C_d$, the capacitance $C_{resi}$ can be found. An estimate of the residual air gap can be made by assuming parallel plate capacitor such that $f = \varepsilon_0 S/C_{resi}$. The extracted air gap for a set of test structures agrees within ±3% with the calculated using the model (Eqn. 5.9).

The parasitic capacitance and the residual gap introduced by the uneven height of the actuation electrodes and underpass area, dominates the shunt switch capacitance. The primary effect of these ‘extra capacitances’ is the shift in resonance frequency of a switch as shown by measured RF response in Fig.5.16. As indicated by the model, residual air gap capacitance can be eliminated by having the underpass and actuation electrodes of the
same height. In practice, when the underpass and the actuation electrodes are realized using two different material layers, the height difference will depend on the tolerances of both the layers. In addition to $C_{up}$, parasitic capacitance and resistance arising because of the finite oxide - substrate resistance, also limit the quality factor of the devices. For frequencies above $f_{sub} = 1/(2\pi R_{sub}C_p)$, the substrate resistor shunts the MEMS capacitor, becoming a primary limit on achievable $Q$ [13]. As a result the electrical model is completed by adding the dotted branch in Fig.5.15(d).

5.4.3.2 RF Measurements

The RF response is characterized by S-parameters measured using an Vector Network Analyzer. The measurement setup is similar to the one, described in Chapter 6. The measured and simulated RF performance for two representative switches in actuated state is shown in Fig. 5.16. The devices are similar except for the bridge length and the meander spring dimensions. As shown by Fig.5.16, in both cases, there is a shift of 2 - 2.5 GHz in resonance frequency and a lower isolation in measured response as compared to the simulated values. The longer device (Fig.5.16(a), bridge
length = 1048 \, \mu m \) in off-state has measured isolation better than -20 dB, over 20 - 28 GHz and a return loss of -0.2 dB. In Fig. 5.16(b), for a bridge length = 688 \, \mu m, isolation is -20 dB over the range 18 - 24 GHz. As described above, the overlap between the beam and the under-pass, limited by the residual air gap and the surface roughness of the underpass (Fig. 5.20), shifts the measured resonance frequency where as in simulations, ideal contact is assumed. In the RF measurements shown in Fig. 5.16, additional bias of 20 - 30 V has to be superimposed on the transmission line, in order to have better contact/overlap. The frequency shift in this case is primarily caused by the parasitic component \( C_p \). The finite resistance of the bridge (0.819 ohms and 0.487 ohms, for the long and the short bridges, respectively) and the CPW line also result in isolation loss of 8-10 dB as compared to simulated values.

Nonetheless the measured response of the devices is satisfactory, with a well defined resonance peak, between 22 - 24 GHz. However, during the characterization of the devices it was observed that: (1) majority of the structures had a residual stress gradient resulting in deformed beams with a lower gap between the bridge and underpass, (2) the down to up capacitance ratio \( C_d/C_{up} \) is small (2 - 4). To reduce the residual air capacitance, additional bias has to be superimposed on the transmission line. Based upon the results, it was felt that beam and underpass design has to be reconsidered to improve \( C_d/C_{up} \). The adopted approach is discussed in the following sections.

5.5 Design Optimization

To eliminate any undesirable deformation in a fabricated membrane and keep it planar over the underlying circuitry, are the major challenges in low actuation RF MEMS switch development. The high aspect ratio mov-
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Figure 5.17: Simulated \textit{(ANSYS\textsuperscript{®})} z-deflection of the bridge vs the meander length, for uniform z-directed load on the actuation electrodes. Schematic at the bottom, represents side-view for two extreme cases (a) and (d).

Viable structural members of RF MEMS, are generally realized using metals deposited over a resist layer. When a thin metal film is deposited on a sacrificial layer at a temperature lower than its reflow temperature then, intrinsic stress develops in the ‘film-sacrificial layer system’, which causes warping or curling of the structures upon release from sacrificial layer. A detailed discussion on this is presented in the \textit{Chapter 6} - sections on internal stress in the beams, electroplating and structural properties.

The metallic movable structural member, which is invariably referred to as beam, bridge or membrane in this work, is subjected mainly to two kind of forces, which can cause undesirable deformation also referred to as buckling, curling or warping. Warping level is defined as the maximum vertical distance between any two points of a movable structure. In long thin beams, the electrostatic force between the beam and the actuation electrodes, coupled with the height difference between the underlying electrodes, can deform the beam, resulting in a nonuniform contact/overlap.
area. In switches, based on torsion springs, beam may bend, instead of rotating about the springs [82]. The residual stress in beams in the absence of external actuation is known to cause more deformation. It finally results in higher actuation voltage, low down state capacitance, and sometimes complete failure of the devices. In present work, warping levels of 30 - 90 $\mu m$, have been observed in the fabricated switches and cantilever test structures with length ranging from 300 - 1500 $\mu m$.

The deformation caused by the residual stress and electrostatic force induced beam bending can be minimized by optimizing the combination of meander spring constant, actuation electrode area and the stiffness of the main bridge. However, based on the observations of deformation in the test structures and devices fabricated under different batches, it is clear that residual stress is more process specific and its alleviation needs an integrated design and fabrication approach.

### 5.5.1 Deformation alleviation using meanders

The mechanical behavior of the bridge membrane is studied using the switch topology shown in Fig.5.17. The main aim of the bridge design optimization is to have a planar beam structure over the actuation electrodes and the capacitive contact area, with optimum stiffness to mass ratio. The simulated (ANSYS®) membrane is meshed using a shell element (length, width and thickness of the membrane are: 688, 150 and 1.5 $\mu m$ respectively) and is subjected to a uniform load on the areas corresponding to the actuation electrodes. The simple boundary conditions with $z$-directed load are more closer to the real actuation force on the beam. The resulting deflection profile with highly exaggerated $y$ - dimensions is shown by the side-view, in Fig.5.17. The equivalent spring constant of the bridge structure is four times the spring constant of each spring. In order to alleviate the deformation introduced by residual stress, stiffer beam
structures are preferred [83]. However, the bridge structure with highest spring stiffness (Fig.5.17(a), k for a single spring = 0.6N/m) shows the maximum out-of-the plane deformation (excluding spring deflection), under the electrostatic actuation force (in this case the equivalent pressure), as compared to the more compliant bridge in Fig.5.17(d). The deflection profile of the stiffest structure also agrees with the profile presented by the model in Fig.5.14, showing beam curvature over the actuation electrodes. In addition to reduced deformation, more compliant meanders result in low actuation voltage as discussed in section 5.2.1 and shown by graphs in Figs. 5.1 and 5.2. The meanders of three other types (Fig.5.3) were found to be stiffer or less compact. In presence of residual stress, the beam deformation is comparable for same spring dimensions, hence simple U-type spring structure with spring constant 0.15 - 1.5 N/m are preferred.
5.5.2 Deformation alleviation using stiffening ribs

As discussed in the previous section, meander design is one of the factors that determines the unwanted deflections of a switch. Another major factor is the switch thickness. As shown by Eqn. 5.1, actuation voltage is a function of the spring constant \(k\), and theoretically independent of the bridge thickness. Consequently, the thickness of the bridge can be increased to minimize undesired deflections due to a residual stress gradient and bending induced by the electrostatic force over the length of high aspect ratio beams (the stiffness of a beam is given by \(k_b = \frac{Ewt^3}{2L^2}\), where \(L, w\) and \(t\) are the length, width and thickness respectively). However, increasing the thickness of movable structure lowers the mechanical resonance frequency (\(f_0 = \frac{1}{2\pi} \sqrt{\frac{K_b}{m}}\)), \(K_b = \text{equivalent spring constant of the bridge, } m = \text{mass}\) resulting in a slower device. In order to get an optimum stiffness to mass ratio, six types of structures, under known fabrication process constraints, were investigated by using a commercial FEM simulator (\(\text{ANSYS}^\text{®}\)).

Fig. 5.18 shows the simulated and fabricated part of the bridge together with the spring constant to mass ratio for the implemented beam structures. The \(k/m\) ratio is the highest for the beam structure of type (B) - which is a \(5\mu m\) thick Au plate (\(E = 108 \text{ MPa},[38]\)) perforated with \(10 \mu m \times 10 \mu m\) holes at a pitch of \(20 \mu m\). The holes in the beam reduce the squeeze film damping and increase the switching speed. Holes are also necessary to facilitate the complete removal of sacrificial-photo-resist layer (\(3\mu m\)) during the final bridge release by reactive ion etching in oxygen plasma. Therefore, solid plates of type (A) - \(5\mu m\) thick plate and type (C) - with outer and inner frames in 5 and 1.5 \(\mu m\) thick gold, are not considered. The implemented structure (type D) consists of a perforated main switch body in \(1.5\mu m\) thick Au, selectively reinforced by \(5\mu m\) beams. Fig.5.19(a)
Figure 5.19: (a) Stress distribution (simulated, (ANSYS\textsuperscript{®})) on a reinforced switch structure, stress is confined to the meanders only, main switch body is stress free. (b) deflection profile of the switch (simulated), showing no warping over the entire suspended structure except the springs, when subjected to uniform load on the actuation area.

shows the simulated (COVENTOR\textsuperscript{®}) stress distribution corresponding to a ‘load’ of 5V on the actuation electrodes of a switch (bridge length = 608 \(\mu\text{m}\)), of type D structure. As expected, the springs are subjected to maximum stress (peak value of 3.9 MPa), while the reinforced rectangular plates and connecting ribs are nearly stress free. The deflection behavior of the switch with the same dimensions is shown by Fig. 5.19(b). The bridge area corresponding to actuation electrodes is subjected to a z-directed, uniformly distributed load of 30 Pa. The maximum deflection occurs in the springs, where as the z-deflection of the bridge is similar to a “rigid plate” with no warping. The stress behavior of a 5 \(\mu\text{m}\) thick perforated plate (type B, Fig. 5.18(c)) is similar, though more rigid (K = 53 N/m) as compared to type -D. The implemented structure (type D, K = 20 N/m) is approximately 4 - 20 times stiffer, than the suspension meanders. It consists of a switch frame and selected reinforcing beams in 5 \(\mu\text{m}\) while rest of the switch body and the meander springs are in 1.5 \(\mu\text{m}\) thick Au. The surface micro-machined structure is realized in two electroplating steps. In
the first step, the main switch body including the meanders is electroplated over Cr(10 nm)- Au(150nm) seed layer to a thickness of 1.5 \( \mu m \). In the next gold electro-deposition step, the thickness is selectively increased to 5 \( \mu m \).

### 5.5.3 Capacitance optimization

As mentioned previously (Chapter 4), a capacitive shunt switch can be modeled to the first order, as a capacitor between the metal bridge and the signal line. In the up-state, capacitance is too small to affect the line impedance. However, in the actuated state, the capacitance between the bridge and underpass becomes high and the switch is in isolation mode. The down/up capacitance ratio quantifies the RF response of a switch and is one of the important figures of merit. The ratio in terms of the switch geometry and material properties is expressed as:

\[
\frac{C_{\text{down}}}{C_{\text{up}}} = \frac{\varepsilon_0 \varepsilon_r A_{\text{overlap}}/t_{\text{die}}}{\varepsilon_0 A_{\text{overlap}}/d_{\text{air}}} = \frac{\varepsilon_r d_{\text{air}}}{t_{\text{die}}} \tag{5.10}
\]

where \( \varepsilon_r \), \( A_{\text{overlap}} \), \( d_{\text{air}} \) and \( t_{\text{die}} \) are the constant of the dielectric material, overlap area between the bridge and the signal line, air gap and thickness of the dielectric material respectively. According to Eqn.5.10, the freedom to have high a \( C_{\text{down}}/C_{\text{up}} \) ratio is highly constrained. The second problem encountered in capacitive switches is the degradation of the effective down-state capacitance as a result of the surface roughness preventing intimate contact between the beam and the dielectric on signal line [35]. The sketch in Fig. 5.20(a) illustrates the surface conditions on the central part of a conventional capacitive switch, discussed earlier. Part (b) shows the dielectric surface of a fabricated device, with hillocks and particulate of the size approximately 100 nm to 250 nm. Additional problem may arise because of the surface roughness and other artifacts on the bottom surface of the electroplated bridge. Fig.5.20(c), shows whiskers on the bottom
surface of the bridge, due to uncomplete removal of the seed layer which inhibit the proper bridge - dielectric contact. The cumulative effect may reduce the capacitance more than 50% [19].

In addition to residue removal and minimizing the surface roughness below 5 nm, the common approach is to use thin refractory metal layer under the oxide [34], [57]. Instead, in the modified design, we use an electrically floating layer of metal covering the dielectric. Fig.5.21(a) shows a SEM micrograph of such a device. The floating metal provides an optimal contact without resorting to smooth surfaces. On the signal line a few, contact points (5µm × 5µm, Fig. 5.21(b)) are sufficient to have an optimal down capacitance given by equation:

$$C_d = \varepsilon_0 \varepsilon_r A_{float}/t_{die}$$  \hspace{1cm} (5.11)

where $A_{float}$ is the area of the floating metal. However, the bridge to float metal contact impedance (combination of the contact resistance and capacitance due to native oxide) should be sufficiently low. For this reason, Au and Pt have been used together with optimal contact points.

The ratio $C_d/C_{up}$ can be further optimized by using the fact that the down capacitance depends on $A_{float}$, whereas $C_{up}$ still depends on the overlap between the bridge and dielectric on the signal line. $C_{up}$ can be lowered
by having a narrow bridge over the dielectric layer, without affecting, $C_d$. The geometrical factor that increases the ratio is given as:

$$
\frac{C_d}{C_{up}} = \frac{\varepsilon_0 \varepsilon_r A_{\text{float}}/t_{\text{dielectric}}}{\varepsilon_0 A_{\text{overlap}}/d_{\text{air}}} = \varepsilon_r \frac{A_{\text{float}} d_{\text{air}}}{A_{\text{overlap}} t_{\text{dielectric}}} \quad (5.12)
$$

In Fig. 5.21(a), the central capacitive area in a conventional switch is $250\mu m \times 150\mu m$, when converted to a single beam ($150\mu m \times 10\mu m$), shown as dotted strip, the $C_d/C_{up}$ ratio can be improved by 25 times over the full plate design, rest of the parameters are the same for both the devices.

5.6 Reinforced Switches

5.6.1 Device Specifications

Mainly, three kinds of serpentine meander based RF MEMS switches have been designed and fabricated. As described in the section on fabrication, the whole movable structural member including the springs is electroplated to a gold thickness of $1.5\mu m$, followed by the second electroplating step which increases the thickness on selected portions to another $3.5\mu m$. Thus,
 thickness of the reinforcing ribs, in all of the switches, including the symmetric toggle switches discussed next chapter is 5\( \mu m \), where as the suspension springs are in 1.5\( \mu m \) thick gold. The basic three types differ in the size of the central active overlap area i.e. 150\( \mu m \times 90 \mu m \), 250\( \mu m \times 150 \mu m \) and 300\( \mu m \times 300 \mu m \) which corresponds to the intended frequency range of 22 - 26 GHz, 12 - 14 GHz and 8 - 10 GHz, with isolation better the -20dB and insertion loss less than -0.2 dB. The flexure spring design and the actuation electrode area combinations, for the three main types are similar. Each type has further seven variations in terms of the meander length - actuation area combination, thus resulting in actuation voltages ranging from 3 to 15 V, and more than 40 types of shunt and ohmic switches in all. Table 5.3 summarizes the meander based switches, with active area of 150 \( \mu m \times 90 \mu m \). The other types are similar except the bridge length. Both the conventional i.e. ‘dielectric - on - underpass’, and the floating metal type have been designed and fabricated. Fig. 5.22(a) shows the SEM micrograph of a single meander 150\( \mu m \times 90 \mu m \) type device, \( V_{pi} \) =

<table>
<thead>
<tr>
<th>Bridge Length (( \mu m ))</th>
<th>Sec. Mea. Length (( \mu m ))</th>
<th>No. of Meanders</th>
<th>Actuation Electrode area 2( \times )(( \mu m ))^2</th>
<th>Actuation Voltage Cal.(V)</th>
<th>Actuation Voltage SIM.(V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>408</td>
<td>100</td>
<td>1</td>
<td>100 ( \times ) 150</td>
<td>15.30</td>
<td>14.50</td>
</tr>
<tr>
<td>488</td>
<td>120</td>
<td>1</td>
<td>140 ( \times ) 150</td>
<td>10.00</td>
<td>10.25</td>
</tr>
<tr>
<td>528</td>
<td>120</td>
<td>1</td>
<td>120 ( \times ) 150</td>
<td>10.90</td>
<td>10.50</td>
</tr>
<tr>
<td>608</td>
<td>170</td>
<td>1</td>
<td>200 ( \times ) 150</td>
<td>5.16</td>
<td>5.25</td>
</tr>
<tr>
<td>688</td>
<td>220</td>
<td>1</td>
<td>240 ( \times ) 150</td>
<td>3.24</td>
<td>3.75</td>
</tr>
<tr>
<td>528</td>
<td>140</td>
<td>2</td>
<td>160 ( \times ) 150</td>
<td>5.10</td>
<td>x</td>
</tr>
<tr>
<td>688</td>
<td>190</td>
<td>2</td>
<td>200 ( \times ) 150</td>
<td>3.00</td>
<td>x</td>
</tr>
</tbody>
</table>

Table 5.3: Actuation voltage, calculated and simulated (COVENTOR®) for reinforced conventional and floating metal devices. For all the devices primary meander length \( L_p = 25 \mu m \), spring thickness = 1.5 \( \mu m \), overlap area = 150\( \mu m \times 90 \mu m \).
5V, (b) is a similar device with two meanders and lower actuation voltage (3V) and (c) is a 250\(\mu\text{m} \times 150\mu\text{m}\) type switch with \(V_{pi} = 5\)V. The device of the third type (300\(\mu\text{m} \times 300\mu\text{m}\)) is shown in Fig.5.23(a). However, due to unavoidable process problems the measurements could be performed on the floating metal type of switches only. As will be discussed in the chapter on the symmetric toggle switch and briefly mentioned in following sections, some of the devices and test structures were highly deformed to make any meaningful measurements.

5.6.2 Simulated RF response

Figure 5.23(a) shows a SEM micrograph of a reinforced floating metal switch with active overlap area of 300\(\mu\text{m} \times 300\mu\text{m}\). The simulated (HFSS) qualitative R F response of the switch at 9 GHz, in bridge-up and down state is shown in Fig.5.23(b) and (c). The figures show a 2D average current density distribution, on the transmission line and the surrounding CPW ground area, in response to a wave incident form the right side in figure (b) and (c). Insertion loss (-2dB) in up state and isolation (-40dB) in down state are clearly indicated. A brief introduction to the simulation
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Figure 5.23: (a) SEM micrograph of 300 μm x 300 μm overlap area switch. (b) simulated average current density distribution for the switch in bridge-up and (c) in bridge-down state.

procedure is given in the chapter on symmetric toggle switch (Chapter 6). The response for other devices is similar and has been presented in Chapter 4.

All of the switches are implemented in standard 50 Ω CPW configuration (75μm−90μm−75μm) and fed through 300−350μm long line sections. Fig.5.24 shows the simulated response of the switch with top floating metal. A comparison of the insertion loss and isolation with a reduced bridge geometry is also presented in same figure. In the on-state, the insertion loss (−20 log(S21)) lies between -0.1 to -0.25 dB (Fig.5.24(b)) over the frequency band of interest and falls off sharply as the frequency is increased beyond 9.5 GHz. The insertion loss depends on the up-state capacitance and can be improved by reducing the overlap area, as shown by the insertion loss curves, for normal and reduced area devices, in Fig.5.24(d). In the case, where the active overlap area has been replaced by a single strip of 300μm × 10μm the insertion loss remains up to -0.25dB, over the whole frequency range till 11 GHz, where as with full overlap area design it degrades after the resonance. The switch off-state RF behavior is shown by Fig.5.24(a) and is compared with the reduced area device in
Figure 5.24: Simulated RF characteristics of a 300\(\mu\)m x 300\(\mu\)m switch (a) switch isolation in switch off-state for 2 - 5 connecting ribs, (b) insertion loss in the switch on-state, (c) comparison of switch off-state isolation for normal (300\(\mu\)m x 300\(\mu\)m) and reduced (300\(\mu\)m x 10\(\mu\)m) overlap area. (d) comparison of switch on-state insertion loss.

In the simulations, between the bridge and active overlap area, a perfect contact is assumed, which is valid for the floating metal design, except for the contact resistance, which can affect the off-state switch response. The switches show an isolation better than -25dB (Fig.5.24(a)) over a frequency range of 7 to 12 GHz. Also in confirmation with earlier observations, the meander based switches behave as a single LC tank with well defined resonance frequency. The multiple curves in Fig.5.24(a), show the effect of the beams (inductance), connecting the active overlap area to the rest of the bridge on both sides. The reduction in isolation for curve S21-2, Fig.5.24(a), is due the increased bridge resistance as the number of the connecting beams is reduced to two. The effect of the reduction in bridge overlap area with respect to the normal area on the switch isola-
tion characteristics is compared in Fig.5.24(c). Though, the isolation has improved, the resonance has shifted to higher frequency. The shift is due to the change in bridge inductance. In devices fabricated with reduced overlap area, the isolation may further deteriorate because of the limited contact, local oxidation or presence of the organic residue.

5.6.3 S-parameter Measurements

The S-parameter measurement setup and measurement procedure for the meander based RF MEMS switches is essentially the same as that for ‘Symmetric Toggle Switches’ described in the next chapter. The measurement set-up is shown in Fig. 6.18 Chapter 6. The measurement set-up availability also influenced some of the design parameters and measurements, especially on the high frequency devices. The main constituents of the measurement system are the Vector Network Analyzer (HP VNA 8719D) and the RF probe station. The pitch of the probes is an important CPW design consideration. The devices described earlier have been measured with a 250 \( \mu m \) pitch RF probe setup, while all rib-reinforced devices are characterized using a setup with 150\( \mu m \) pitch RF probes. Accordingly, the CPW design have been changed but keeping the characteristic impedance (50 Ω) the same in all the devices discussed in the present work. The essential features of the Vector Network Analyzer are the signal frequency range and the sweep voltage. The characterization of the high frequency devices \((150 \times 90 \ \mu m, \text{resonance frequency} = 22-24 \text{ GHz})\) is particularly limited by the available frequency range (50MHz - 13.5GHz) of VNA.

The switches are considered as two-port devices and the magnitude and phase for all the four S-parameters (S11, S21, S12, S22) are recorded simultaneously at a predefined sweep voltage step. The Labview\textsuperscript{(TM)} software controlled system provides the real time display of isolation and return loss.

\[^1\text{RF characterization was done at ARCES-DEIS, University of Bologna and CNR-IMM, Rome}\]
Table 5.4: Comparison of actuation voltages - calculated, simulated (COVENTOR®) and measured for reinforced floating metal devices. Simulated and measured isolation response is also summarized. For all the devices primary meander length $L_p = 25 \mu m$, spring thickness = 1.5 $\mu m$.

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<tr>
<td>480</td>
<td>$150 \times 90$</td>
<td>15.30</td>
<td>42.2</td>
<td>15.5(12GHz)</td>
<td>12.0(12GHz)</td>
</tr>
<tr>
<td>528</td>
<td>&quot;</td>
<td>10.90</td>
<td>38.5</td>
<td>16.4(12GHz)</td>
<td>11.8(12GHz)</td>
</tr>
<tr>
<td>608</td>
<td>&quot;</td>
<td>5.20</td>
<td>18.5</td>
<td>16.0(12GHz)</td>
<td>20.2(12GHz)</td>
</tr>
<tr>
<td>688</td>
<td>&quot;</td>
<td>3.2</td>
<td>16.2</td>
<td>15.8(12GHz)</td>
<td>15.5(12GHz)</td>
</tr>
<tr>
<td>470</td>
<td>$250 \times 150$</td>
<td>15.30</td>
<td>49.0</td>
<td>32(9GHz)</td>
<td>18.3(9GHz)</td>
</tr>
<tr>
<td>550</td>
<td>&quot;</td>
<td>10.90</td>
<td>26.5</td>
<td>35(9GHz)</td>
<td>17.0(9GHz)</td>
</tr>
<tr>
<td>670</td>
<td>&quot;</td>
<td>5.20</td>
<td>17.0</td>
<td>34(9GHz)</td>
<td>21.5(9GHz)</td>
</tr>
<tr>
<td>750</td>
<td>&quot;</td>
<td>3.24</td>
<td>x</td>
<td>x</td>
<td>x</td>
</tr>
<tr>
<td>620</td>
<td>$300 \times 300$</td>
<td>15.50</td>
<td>52.0</td>
<td>37(8GHz)</td>
<td>20.0(8GHz)</td>
</tr>
<tr>
<td>700</td>
<td>&quot;</td>
<td>10.00</td>
<td>40.7</td>
<td>33(8GHz)</td>
<td>17.5(8GHz)</td>
</tr>
<tr>
<td>820</td>
<td>&quot;</td>
<td>5.20</td>
<td>23.5</td>
<td>35(8GHz)</td>
<td>21.7(8GHz)</td>
</tr>
<tr>
<td>900</td>
<td>&quot;</td>
<td>3.00</td>
<td>x</td>
<td>x</td>
<td>x</td>
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of a switch corresponding to the applied sweep actuation, enabling quick analysis. The accuracy of the pull-in voltage can be improved by refining the step voltage. As explained in the following paragraph, in presence of the high deformation in all the fabricated devices and cantilever based process-parameter extraction test structures, the analysis is limited to the essential details only.

A comparison between the simulated and measured pull-in voltages and switch isolation is presented in Table 5.4, for four type of switches e.g. with actuation voltage of 15, 10, 5 and 3V. The measurements, averaged over a few devices of each type, are for floating-metal-on top (FMT) type of switches only. The measured isolation characteristics for the represen-
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Figure 5.25: Measured isolation and return loss for a 250 $\mu$m x 150 $\mu$m, 300 $\mu$m x 300 $\mu$m active area floating metal switches.

tative device from each main type i.e. with active area of 250x150 $\mu$m, 300x300 $\mu$m and 150x90 $\mu$m, are shown in Fig 5.25 (a), (b), (c), (d) and Fig. 5.26. As seen in Table 5.3 and Table 5.4, a good agreement is observed between the calculated and simulated pull-in voltages. The analytical model ignores damping, nonlinear bending and the beam stretching effects, which accounts for the small difference between the two. A further improvement in the agreement can be expected by fine tuning the simulation in the pull-in range, once the approximate values have been found. The adopted simulation procedure is similar to the ‘Symmetric Toggle Switch’ and a brief description is given in Chapter 6. However, as shown in Table 5.4 and measured response curves in Figs.5.25 and 5.26, the difference between the measured and simulated values is significantly high. The isolation though acceptable, is also lower by 20 to 50% in most of the switches. The higher difference is due to the deformation of the electro-
plated structural elements. The warping or out of the plane deformation is caused by the residual stress in the Au-layers. Most of the ‘two meander’ spring combination switches, which are more compliant and larger in width, could not be actuated or failed during testing. Also, the higher deformation obscured any meaningful measurements on the conventional ‘oxide-on-overlap area’ type switches. The reason for this is the unexpected high deformation gradient, probably due to the seed layer and changes in depositing it.

In Fig.5.27 the optical profiles (2D line scans and a 3D profile) clearly show the deformation gradient for a 300$\mu m \times 300\mu m$, single meander switch. As shown by Fig.5.27(a) and (d), in the upper actuation electrode area of the bridge, along the y-axis, difference between the lowest and the highest point is 27.9 $\mu m$. The lowest point is about 6 $\mu m$ above the actuation electrodes on the wafer. The effective air gap between the actuation electrodes and the bridge, thus, varies from 1 to 23$\mu m$ against the intended value of 3$\mu m$. This explains the higher actuation voltage for all the devices and complete failure in some of the devices.

The central overlap area shows a curvature in both x and y directions, as seen in Fig.5.27(b) and line scan in Fig.5.27(c). The maximum warping
is 6 $\mu m$. In most of the conventional switches, reduction in active overlap area leads to poor isolation or already ‘actuated’ switches, with portion of the bridge touching the bottom dielectric. Such devices show complex deformation under actuation, accompanied by a little or no change in isolation characteristics. The isolation in floating metal (FMT) devices is lower by 20 to 50%. This is because of the poor contact between the bridge and the contact points on the underpass.

The observed out-of-the plane deformation is not uncommon in MEMS switches [39]. However, as compared to the first process, the deformation in the second process run is significantly high. In the first process, beams longer than 1000 microns were deflected downwards under the tensile stress by a small fraction of the intended air gap. The main processing steps which contribute to the residual stress and hence the deformation of
the fabricated structures, are the physical vapor deposition of Cr-Au (10 and 150 nm) seed layer, Au-structural layer electrodeposition (in two steps) and the thick resist lithography defining the spacer - on which bridge is patterned. The seed layer deposition is carried out in vacuum by thermal evaporation (sublimation) of Cr, followed by Au evaporation. The amount Cr deposited, Cr-oxide formation during sintering process, diffusion of Cr into gold layer on top and electro - deposition of gold in two steps - results in a non-homogenous multilayer structure. The exceptionally high deformation is probably the manifestation of the change in composition along the thickness of the structure. In the last fabrication process, a new electron beam evaporation system has been used, this may have altered the thickness and properties of the deposited layers in a way not yet fully understood. The electro-deposition and Cr-Au seed layer deposition process are discussed further in the Chapter 6 on symmetric toggle switch, where similar kind of deformation has occurred.

5.7 Series Ohmic Contact Switch

*Series Ohmic Contact* RF MEMS switch configuration has also been studied because of its importance in communication and inherent structural similarities to the shunt capacitive devices discussed earlier. However, because of the inevitable delay in fabrication of the devices and other time-constraints, measurements and characterization could not be completed. The main advantage of the series ohmic configuration is that it can be used from DC to 8GHz with excellent isolation and low insertion loss. In series configuration, low voltage, metal to metal ‘direct-contact’ devices thus, are more suitable for hand held wireless communication applications in C-band and below. In this section, we present the design, layout and simulated RF response of the series switches, for which the actuation volt-
5.7. SERIES OHMIC

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Figure 5.28: Series ohmic contact switch, (a) An optical micrograph of a fabricated switch, \( V_{pi} = 5V \), (b) detailed view of the contact flexure

...age range is similar to the shunt capacitive switches i.e. 3 to 15V.

5.7.1 Mechanical Design and Layout

The switch geometry of the ohmic switch implemented in series configuration follows the same definitions as for the capacitive shunt switches. Fig.5.28(a) shows an optical micrograph of a fabricated low actuation voltage (5V) series device. The switches are implemented in 50 Ω standard CPW configuration similar to the capacitive shunt devices. The main switch body (electroplated gold) is anchored by meander flexures to achieve actuation voltages in the range of 3-15V and optimized in a way similar to shunt switches. The bridge structure is patterned with etch release holes and also reinforced with beams of thickness 5\( \mu \text{m} \). Both series and shunt switches are fabricated using the process already presented in one of the previous sections.
The main difference between a shunt capacitive and a series ohmic contact also referred to as ‘DC-contact switch’, is the way RF signal is coupled from the input to the output port. In series ohmic configuration instead of the ‘capacitive overlap’ of the bridge and transmission line, an ‘ohmic contact flexure’ as shown in Fig.5.28(b) and by schematic sketch in Fig.5.29(b) is used to transmit the signal over a gap in the microwave transmission line. The gap which is about 150 $\mu$m in all the devices under consideration is also called the ‘inline segment’ of a switch. The contact flexure is suspended above the interrupted transmission line by a vertical gap height of 3 $\mu$m. This combination offers an excellent isolation in the switch off-state. The series switches, in the present work are implemented as ‘relays’ i.e. with two separate actuation electrodes as seen in Fig.5.28(a). Under the actuation bias, the contact flexure falls down on the transmission line and creates a short circuit between the open ends. This corresponds to the switch on-state. Unlike the shunt switch, the bridge is connected to ground through a high value resistor (40-80 k$\Omega$). Except for the contact flexure, the switch inline length and the intended frequency range, rest of the design parameters such as the meander flexures, the actuation voltage considerations, CPW/bridge layout and the fabrication process are similar to the shunt capacitive switches discussed earlier.

5.7.1.1 DC Contact Flexure

As compared to the prevalent cantilever based metal-semiconductor design approaches, the switches are based on metallic bridge, isolated from the ground plane of CPW and suspended above the interrupted transmission line. The performance of such a configuration is mainly characterized by the transmission loss; isolation is comparatively easy to obtain because of the larger air gap in the transmission line and the vertical air gap between the contact flexure and the contact points on the transmission line. The
transmission loss is dominated mainly by the metal to metal contact resistance and to some extent by the leakage through the biasing resistors, via holes joining the transmission line and underpass and the substrate itself. Thus, it is imperative to optimize the contact resistance and the layout of the biasing resistors.

Transmission loss due the parasitic contributions of biasing lines and coupling between the actuation electrodes is to a larger extent function of the layout strategy and the actuation voltage of the devices. Loss can be minimized by avoiding the overlap between the biasing and RF transmission lines and by placing the actuation electrodes at a distance of 40-80 μm from the RF line [19]. In the present design, depending on the total bridge length, the actuation electrodes are 50-70 μm away from the transmission line and there are two unavoidable ‘DC-bias - transmission line’ overlap segments with area of 15 μm × 70μm each.

The contact resistance is a function of the contacting metal properties, contact force and contact area. A detailed contact resistance model is given in [84]. Following the approach, the contact flexure design as shown in Fig.5.29(b) consists of a row of ‘bumps’ on the transmission line and a compliant contact flexure fixed to the ‘rigid’ bridge plate. The mechanism
ensures a point-like contact between the bumps and the bridge with a controlled amount of force. The bumps or dimples (5µm × 5µm) are defined in polysilicon, under the underpass area of the transmission line. The contact force control is achieved by considering the height of the bumps, flexure dimensions and the ‘contact gap’ of the bridge and dielectric on the actuation electrodes. The force generated on the contact bumps is given by

\[ F = \frac{3E I_y h}{1 - \nu l^3} \]  

(5.13)

where \( E, \nu, I_y, l \) and \( h \) are, respectively the Young’s modulus, Poisson’s ratio, moment of the flexure, length of the flexure and height of the bumps [85] as shown in Fig.5.29(b). The total force generated by the flexure wings in Fig.5.29(b) is about 100 µN i.e. a medium force regime. With forces of this kind an elastic deformation of the two contacting bodies can be assumed. The minimum contact force depends on the applied actuation voltage and the number of contact bumps. In practice, the surface conditions of the two contacting areas and the deposition of organic residue also alter the contact resistance behavior and hence the loss characteristics of a switch.

5.7.2 RF Response

The electrical equivalent circuit of a series switch is shown in Fig.5.29(c). The RF response of a device is mainly characterized by the switch capacitance in up-state and the inline segment resistance which also includes the contact resistance, in the down state. Similar to capacitive shunt, the parasitics, arising from the wafer, CPW and DC interconnects also effect the performance. As shown by the equivalent circuit the up-state capacitance is composed of (a) series capacitance \( C_s \), the capacitance between the flexure and transmission line overlap area, (b) parasitic capacitance between
the open ends of the transmission line. So that the total capacitance is:

\[ C_{up} = \frac{C_s}{2} + C_p \]  

(5.14)

The series capacitance, \( C_s \), since arises because of the overlap of metal portions of the transmission line and the flexures, separated by an air gap (3\( \mu m \)), it can be estimated as a parallel plate capacitance given by \( C_{pp} = \varepsilon A/g \). Additionally, the capacitance due the fringing fields should also be considered. The parasitic capacitance \( C_p \) depends on the CPW width and the separation between the two ends of the transmission line. In all the switches considered in this work, the CPW width is 75 \( \mu m \), and the gap between the two ends of signal line is about 150 \( \mu m \). This gives a parasitic capacitance of about 1.5fF and less. The total up-state capacitance for the considered switch configurations lies between 3 - 5 fF, which provides good isolation as shown by Fig. 5.31(a) for one of the
representative devices.

The current distribution on the series switch is also similar to the shunt configuration is shown in Fig. 5.30 for the bridge in up and down states. The simulated distribution shows that current is concentrated on the edges of the inline segment of the bridge and therefore not affected by the ‘etch’ holes on the bridge. In the on-state, the switch is simply a continuation of the CPW transmission line and the insertion or transmission loss is mainly due to the contact resistance ($R_c$) and the resistance of the inline segment ($R_l$). The total resistance of the switch thus, can be written as, $R_s \approx 2R_c + R_l$. Similar to shunt capacitive switches, the CPW resistance should also be considered. Fig.5.31 shows the simulated isolation and insertion loss for a representative device. As seen in figure (a) isolation varies from -50dB to about -30dB over a frequency range of 1-10GHz. The insertion loss is better than -0.3dB over the considered range of 1-10GHz. In practice, the transmission loss and reliability of the series switches are mainly limited by the contact mechanism and needs thorough investigation related to the materials involved, the contact force and the surface properties. Nevertheless, switches show highly promising characteristics in the
As an example of the shunt capacitive switch application, a 5GHz single pole double through switch (SPDT) has also been designed and fabricated. In Fig. 5.32 the simulated current density distribution shows the full SPDT, along with the enlarged view of the switches, S1 in off-state and S2 in on-state. The switch is implemented in standard 50 Ω CPW configuration similar to other devices in this work. Meander based capacitive shunt switches (S1 and S2) have capacitive overlap area of $300 \mu m \times 300 \mu m$ and actuate at 5 V, as described in earlier sections. The switches are at quarter wave length from the reactive junction resulting in a ‘tuned’ configuration. In order to transfer power, for example from port 1 to the port 2, repre-
CHAPTER 5. MEANDER BASED RF MEMS

5.8. SPDT

Figure 5.33: Simulated S-parameters for SPDT shunt MEMS switch, (a) 5GHz SDPT, distance from reactive junction \((\lambda/4) \approx 5634 \, \mu m\), (b) 7GHz SPDT \((\lambda/4) \approx 3521 \, \mu m\).

Presented by switch S2, the switch S1 must be in down state. This results in an open circuit in the arm of S1 at the reactive junction and input power is transferred to port 2 or switch S2 side. In Fig.5.32 this is shown by an isolation of more than -35dB in the case of switch S1, in off-state.

Fig.5.33(a) shows the simulated isolation and insertion loss of the SPDT. The isolation curve corresponds to switch S1 in off-state. The insertion loss curve which corresponds to the switch S2 up-state also includes the CPW line loss. The device configuration presents a fairly low return loss over the entire frequency range. It is mainly because of the quarter-wave transmission length, coupled with the down-state capacitance of switch S1, which results in a high impedance at the reactive T junction due to the quarter wave transformer effect. The isolation can be further improved by incorporating a series/shunt switch configuration in each arm as suggested by Pachecco et. al in [86]. The bandwidth will still be limited by the quarter-wave sections, in each arm. Another, 7GHz SPDT configuration with shunt switches of capacitive overlap area \(250 \, \mu m \times 250 \, \mu m\), has also been fabricated and has similar response as shown in Fig.5.33(b). As mentioned
earlier, due to unavoidable fabrication process related problems, the devices could not be characterized. However, a reasonably good performance is expected when residual stress gradients in the electroplated beams are low enough to ensure beam warping levels below about half a micron.
Chapter 6

Symmetric Toggle Switch

6.1 Introduction

In continuation with our pursuit on low actuation RF MEMS switching devices, in this chapter we propose a new device called ‘Symmetric Toggle Switch’ (STS), which overcomes the problems associated with the meander and cantilever based design approach. As mentioned previously, in meander based switches, reliability against self-biasing, external shocks and power handling capability can be improved by incorporating a third electrode, at the cost of added process complexity. Another approach is the cantilever based push-pull configuration, also with an additional electrode, to keep the device in off state, independent of the restoring force of the beam [87]-[88]. In cantilever fabrication, the residual stress control is more critical as it requires an optimal combination of dielectric (with compressive stress) and conducting (tensile stress) layer thicknesses, resulting in complex and stringent fabrication process requirements. These fixed - free asymmetric structures are more sensitive to the residual stress gradients in the constituent layers as compared to the beam structures fixed on both sides.

In addition to a low actuation and tuneable resonance frequency, the proposed symmetric toggle switch has a large vertical travel range which
makes it a varactor with a wide capacitance range not achievable by the conventional MEMS approach. The design is based on electrostatic torsion microactuators and consists of a single electro-plated gold beam used to realize the on-off and hold functions [89]-[90]. The devices are fabricated using the same process as described in Chapter 5 for meander based switches.

### 6.2 Operating Principle and Device Topology

Figure 6.1(a) shows the schematic of the proposed switch implemented in standard 50Ω CPW configuration. The switch is symmetric about the central CPW conductor. The movable bridge structure consists of a central
contact area, placed directly above the CPW conductor line and connected to torsion microactuators on either side by levers. A torsion microactuator consists of a top electrode anchored at the center by torsion springs and separated from the bottom actuation electrodes by a gap height of $3\mu m$. Fig.6.1(b) shows the side-view of the switch.

The outer and inner pairs of the actuation electrodes of the two microactuators are electrically shorted together by polysilicon lines and are called “pull-out” and “pull-in” electrodes respectively. In shunt-configuration as shown by Fig.6.1(b) the pull-in electrodes when biased to a voltage $\geq$ pull-in threshold, bring the central area of the beam in contact with the oxide layer on the transmission line, capacitively coupling the RF signal to ground. The switch is in off-state. The devices are implemented both in conventional ‘dielectric-on-contact area’ and ‘floating metal-on-contact area’ configurations. The pull-out electrodes are biased to keep the beam clamped in up-state, corresponding to the switch on-state. The switch is impervious to the external vibrations and self-biasing if the pull-out voltage is higher than the RF-signal magnitude. Transmission losses in the on-state are mainly determined by the switch up-state capacitance ($C_{up}$) which is a function of the beam area above the transmission line, and vertical gap height. In present design approach the vertical gap can be varied continuously from the zero to twice the nominal or zero pull-in bias gap height. This, coupled with the reduced beam area in floating-metal configuration, results in negligibly small switch on-state transmission losses. In a series ohmic-contact implementation, the roles of the pull-in and pull-out electrodes are reversed. At pull-in, the gap between the two interrupted sections of the transmission line are electrically shorted by the beam, which corresponds to the switch on-state. The pull-out bias switches off the device by breaking the connection between the two segments of the transmission line. The three configurations mentioned above have been designed
6.3 Electrostatic Torsion Microactuators

In contrast to the parallel plate actuators, the torsion actuators used in STS have one degree of rotational freedom around the spring axis, defined by tilt angle $\alpha$ (Figure 6.4). An important property of the torsion actuators is the pull-in voltage, beyond which the electrostatic torque overcomes the mechanical torque and the movable plate snaps abruptly to the fixed electrode. The aim of a typical design is to determine the spring and electrode parameters for a chosen working point near the pull-in corresponding to a desired maximum controllable angle at a given bias voltage [89] - [93]. Determination of the above parameters for torsion elements is important to have an first order approximation of toggle switch design parameters. In this section we provide an outline of the methodology based on [89] - [90] essential to highlight the torsion based design.
6.3.1 Theory

Figure 6.2 shows the half schematic view of an electrostatic rectangular torsion actuator. In the diagram $L_1$ is the horizontal distance between the axis of rotation and the edge of bottom electrode. $L_3$ and $L_2$ are the electrode lengths from axis of rotation. The length, width and thickness of the springs are $l, w$ and $t$ respectively. Before pull-in, the mechanical torque $M_m$ equals the electrostatic torque $M_e$ [89], i.e.,

$$K_{\alpha} = \frac{\varepsilon_0 V^2 b}{2\alpha^2} \left[ \frac{1}{1 - \frac{\alpha\beta}{\alpha_{\max}}} - \frac{1}{1 - \frac{\alpha\gamma}{\alpha_{\max}}} + \ln \left( \frac{\alpha_{\max} - \alpha\beta}{\alpha_{\max} - \gamma\beta} \right) \right] \quad (6.1)$$

where $\alpha$ is the angle between top plate and the bottom electrode, $\alpha_{\max} = g/L_3$ is the maximum constrained tilt angle, $\beta = L_2/L_3$ is the electrode length ratio, $\gamma = L_1/L_3$, and $V, \varepsilon_0$ and $K_{\alpha}$ are the applied voltage, dielectric constant of vacuum, and the torsional spring constant, respectively. Neglecting $\gamma$ being very small, Eqn. 6.1 reduces to:

$$K_{\alpha} = \frac{\varepsilon_0 V^2 b}{2\alpha^2} \left[ \frac{1}{1 - \frac{\alpha\beta}{\alpha_{\max}}} - 1 + \ln \left( 1 - \frac{\alpha\beta}{\alpha_{\max}} \right) \right] \quad (6.2)$$

At the pull-in point, mechanical spring constant $K_{\alpha}$ (i.e. $dM_m/d\alpha$) is equal to the electrostatic spring constant ($dM_e/d\alpha$). Thus differentiating Eqn. 6.2 w.r.t $\alpha$, multiplying by $\alpha$ and subtracting from Eqn.6.2 yields:

$$\frac{1}{1 - \beta\theta_{\text{pin}}} - 1 + \ln(1 - \beta\theta_{\text{pin}}) - \frac{\beta\theta_{\text{pin}}}{3(1 - \beta\theta_{\text{pin}})^2} + \frac{\beta\theta_{\text{pin}}}{3(1 - \beta\theta_{\text{pin}})} = 0 \quad (6.3)$$

where $\theta = \alpha/\alpha_{\max}$ is the fractional deflection of the top plate, and $\theta_{\text{pin}}$ is the corresponding fractional deflection at pull-in. Solving Eqn.6.3 yields

$$\beta\theta_{\text{pin}} \approx 0.4404 \quad (6.4)$$

When the electrode length ratio $\beta = 1$, $\theta_{\text{pin}} \approx 0.4404$. It implies that for equal length top plate and actuation electrode design, the voltage controlled vertical displacement of the beam is 44.04% of the total gap height,
6.3. ELECTROSTATIC TORSION

which is sufficient for RF switching but may be not enough e.g. for the optical switching applications such as micro-mirrors. At $\beta = 0.4404$, tilt angle $\alpha = \alpha_{\text{max}} = g/L_3$, provides the maximum tilt range for a given length of the top plate and vertical gap. For lower electrode ratios pull-in does not occur. For $\gamma \approx 0$ and $\beta \theta_{\text{pin}} \approx 0.4404$ the pull-in voltage [89], can be easily obtained from Eqn.6.1

$$V_{\text{pin}} \approx \sqrt{\frac{9.68K_\alpha \alpha_{\text{pin}}^3}{\varepsilon_0 b}} \approx 0.909 \sqrt{\frac{K_\alpha d^3}{\varepsilon_0 b L_2^3}}$$ (6.5)

where $\alpha_{\text{pin}} \approx 0.440 \, g/L_2$. An additional generalization can be made by normalizing the applied voltage and tilt angle by their respective pull-in values in Eqn.6.2 and 6.5. Thus taking $X = \alpha/\alpha_{\text{pin}}$ and $Y = V/V_{\text{pin}}$, the equations become,

$$\frac{Y_2}{X^3} \left[ \frac{1}{1 - 0.44X} - 1 + \ln(1 - 0.44X) \right] \approx 1, \quad \text{and}$$

$$V_{\text{pin}} \approx Y \sqrt{\frac{9.68K_\alpha \alpha^3}{\varepsilon_0 b X^3}}$$ (6.7)

The resonant frequency of the torsion mode is determined using

$$f_\alpha = \frac{1}{2\pi} \sqrt{\frac{K_\alpha}{I_\alpha}}$$ (6.8)

where $I_\alpha$ is the mass moment of inertia about the rotation axis, which can be expressed as

$$I_\alpha = \int \int \int \rho(y^2 + z^2) \, dx \, dy \, dz \approx \frac{1}{3} \rho L_3 b w (L_3^2 + w^2)$$ (6.9)

where $\rho$ is the material density. When $t < w$, the spring constant $K_\alpha$ can be written as

$$K_\alpha = \frac{2Gwt^3}{3l} \left[ 1 - \frac{192t}{\pi^5 w} \sum_{n=1,3,5,...} \frac{1}{n^5} \tanh\left( \frac{n\pi w}{2t} \right) \right]$$ (6.10)
Figure 6.3: (a) Flow-chart for calculating the electrostatic torsion microactuator parameters.

where \( G \) is the shear modulus for the microactuator material. As shown by the flow chart in Fig.6.3, Eqns.6.4, 6.6 and 6.7, can be used to arrive at the design of torsion actuator. The obtained design parameters (spring length, thickness, width, plate and electrode dimensions, gap height \( g \)) provide a first order approximation of the design parameters, needed to arrive at a feasible design for the full Symmetric Toggle Switch. The design parameters thus obtained are used in the analytical model developed for STS and finally verified using standard FEM tools like Coventorware® or ANSYS®.

### 6.4 Symmetric Toggle Switch: Analytical Model

In this section we derive a closed form analytical expression for the switch actuation voltage in terms of its geometrical parameters. The switch being
6.4. ANALYTICAL MODEL

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Figure 6.4: (a) Schematic diagram representing the left half of the switch, (b) simplified one dimensional view showing the reaction forces corresponding to the electrostatic actuation force $F_e$ at equilibrium.

Symmetric about the CPW central line, we consider only the left half of the switch, represented by the torsion actuator, the connecting lever of length $l$ and half of the bridge-transmission line overlap area as shown by the schematic in Fig. 6.4. For small tilt angles, $(\tan \alpha \approx \alpha)$, we have

$$\alpha = \frac{h_L}{L}, \quad \alpha_B = \frac{h_l}{l}, \quad \text{and} \quad \alpha_A = \alpha - \alpha_B \quad (6.11)$$

where $h_L$ is the height by which the top-right corner of the actuator plate moves downwards under actuation, $h_l$ is the height difference between the actuator and the central capacitive overlap plate and $L$ is length of bottom electrode. To simplify the analysis, the dimensions of the top and bottom electrodes are assumed to be equal i.e. $L_2 = L_3 = L$. Considering the torsion actuator under applied bias and before pull-in, the mechanical torque of the spring $(k_T \alpha)$ at the anchor point $F$ and the reaction moment of plate, at point $E$, is equal to the electrostatic torque $M_e$, such that at
equilibrium we have,

\[-k_T \alpha + M_e - L F_B = 0 \quad (6.12)\]

where \(k_T\) is the constant of the torsion spring joining the top plate to anchor point \(F\). In terms of the moment \(M' = l F_B\), the equation can also be expressed as:

\[-k_T \alpha + M_e - \frac{L}{l} M' = 0 \quad (6.13)\]

The torsion angles \(\alpha_A\) and \(\alpha_B\) expressed in terms of the moment of inertia \(I_y\) of the beam \(l\) are \([85]\):

\[\alpha_A = \frac{M'l}{3EI_y}, \quad \alpha_B = \frac{M'l}{6EI_y}, \quad \Rightarrow \quad \alpha = \frac{1}{2} \frac{M'l}{EI_y} \quad (6.14)\]

where \(E\) is the Young’s modulus. Substituting for \(M'\), Eqn. 6.13 becomes

\[-k_T \alpha + M_e - 2 \frac{EI_y L}{l} \alpha = 0 \quad (6.15)\]

The equilibrium condition for the structure in Eqn. 6.12 thus can be rewritten as:

\[M_e = (k_T \alpha + 2 \frac{L}{l^2} EI_y) \alpha \quad (6.16)\]

### 6.4.0.1 Electrostatic Moment

From the schematic sketch in Fig.6.4, we have \(G = G_0 - m x\), where \(m = \tan \alpha\). The electrostatic force acting on a length element \(dx\) of the strip is \(dF_e(x) = dQ(x)E(x)\), where \(E(x) = V/G_x = V/(G_0 - m x)\). The force and electric moment in terms of the applied actuation voltage \(V\) and structural dimensions can be expressed as

\[dQ(x) = dC(x)V = \varepsilon \varepsilon_0 \frac{Wdx}{G_0 - mx} V\]

\[dF(x) = -\varepsilon \varepsilon_0 Wdx \left(\frac{V}{G_0 - mx}\right)^2\]

\[dM(e) = x dF(x) = -\varepsilon \varepsilon_0 Wdx \left(\frac{V}{G_0 - mx}\right)^2 \quad (6.17)\]
where $W$ is the actuator width; in this analysis the torsion plate (top) and actuation electrode (bottom) are assumed to have same dimensions ($\beta = 1$). The electrostatic torque exerted on the plate with length $L$ becomes:

$$M_e = \varepsilon \varepsilon_0 W \frac{V^2}{m^2} \int_0^L \frac{xdx}{(x - \frac{G_0}{m})^2}$$

(6.18)

Using the standard integral,

$$\int \frac{X}{(X + a)^2} dx = \frac{a}{X + a} + \ln(X + a), \text{ we have}$$

(6.19)

$$M_e = \varepsilon \varepsilon_0 W \frac{V^2}{m^2} \left[ \frac{1}{1 - \frac{Lm}{G_0}} - 1 + \ln \left(1 - \frac{L \alpha}{G_0}\right) \right]$$

Rearranging the above equation for small torsion angles ($m = \tan \alpha \cong \alpha$), we obtain

$$M_e = \varepsilon \varepsilon_0 W \frac{V^2}{\alpha^2} \left[ \frac{1}{1 - \frac{L \alpha}{G_0}} - 1 + \ln \left(1 - \frac{L \alpha}{G_0}\right) \right]$$

(6.20)

By using Equation 6.20, the condition for equilibrium in Eqn. 6.16 thus can be rewritten as:

$$(k_T \alpha + 2 \frac{L}{l^2} EI_y) \alpha = \varepsilon \varepsilon_0 W \frac{V^2}{\alpha^2} \left[ \frac{1}{1 - \frac{L \alpha}{G_0}} - 1 + \ln \left(1 - \frac{L \alpha}{G_0}\right) \right]$$

(6.21)

For rearranging the above equation we consider:

$$K = k_T + 2 \frac{L}{l^2} EI_y, \quad A = \varepsilon \varepsilon_0 W, \quad a = \frac{L}{G_0} = \frac{1}{\alpha_{max}}, \quad y = \frac{\alpha}{\alpha_{max}}, \quad \gamma = \frac{\alpha_{max}^3 K}{AV^3}$$

Substituting in Eqn.6.21 we get

$$\gamma y^4 - \gamma y^3 + y + (1 - y) \ln(1 - y) = 0$$

(6.22)

Expanding the last term into series form, we have

$$(1 - y) \ln(1 - y) = -y + \frac{1}{2} y^2 + \frac{1}{6} y^3 + \frac{1}{12} y^4 + \cdots \cdots$$
Substituting in Eqn. 6.22, we arrive at the familiar quadratic equation which describes the model for the STS,

$$ (\gamma + \frac{1}{12})y^2 - (\gamma - \frac{1}{6})y + \frac{1}{2} = 0 $$ (6.23)

The roots of the Eqn.6.23 are:

$$ y_{1,2} = \frac{\gamma - \frac{1}{6}}{2(\gamma + \frac{1}{12})} \pm \frac{\sqrt{(\gamma - \frac{1}{6})^2 - 2(\gamma + \frac{1}{12})}}{2(\gamma + \frac{1}{12})} $$ (6.24)

### 6.4.0.2 Pull-in Condition

For an electrostatic actuator the pull-in corresponds to an applied voltage at which the electrostatic torque is no longer balanced by the mechanical torque, and the beam snaps down. The equilibrium condition before the sudden snap-down of the beam implies a unique voltage for given device dimensions - and is called the pull-in voltage. From a mathematical point of view this implies that the discriminant of Eqn.6.24 must be zero, i.e.

$$ (\gamma - \frac{1}{6})^2 - 2(\gamma + \frac{1}{12}) = 0, \Rightarrow \gamma_{1,2} = 2.3914 \text{ or } -0.0581 $$ (6.25)

Discarding the negative solution, which in our case has no physical meaning, we have the relation for the pull-in voltage of the toggle switch

$$ \gamma = \frac{\alpha_{\text{max}}^3 K}{AV^3}, \text{ which gives } V = \sqrt{\frac{\alpha_{\text{max}}^3 K}{2.3914 A}} $$ (6.26)

The ratio of the instantaneous value of the tilt angle $\alpha$ to the maximum tilt angle $\alpha_{\text{max}}$ is given by the gradient of the Eqn.6.23. At pull-in, $\alpha = \alpha_{\text{pull-in}}$ implies

$$ y_{\text{pull-in}} = \frac{\alpha_{\text{pull-in}}}{\alpha_{\text{max}}} = \frac{(\gamma - \frac{1}{6})}{2(\gamma + \frac{1}{12})} = 0.44949 $$ (6.27)

which is in excellent agreement with reported travel range [89] for torsion based actuators with upper and lower electrodes of same dimensions.
6.4. ANALYTICAL MODEL

6.4.0.3 Vertical deflection range and Capacitance

The vertical deflection of the toggle structure in terms of the gap height, length of the actuation electrodes and lever length can be further expressed as (Figure 6.4)

\[ H = h_L + h_t \cong \alpha L + \alpha_B l = (L + \frac{1}{3} l) \alpha \]

Therefore, \[ H_{max} = (L + \frac{1}{3} l) \alpha_{pull-in} = (L + \frac{1}{3} l)0.44949 \frac{G_0}{L} \] (6.28)

The condition for a precise overlap between the signal line and bridge, thus becomes \[ \frac{l}{L} = 3.6742 \]. This ratio also implies a maximum gap of \[ 2G_0 \], in the switch on-state, at \[ \beta = 1 \]. The minimum capacitance, for given area \[ S \], thus reduces by a factor of two, i.e.

\[ C_{min} = \frac{\varepsilon_0 S}{2G_0} \] (6.29)

The large vertical travel range is important as it reduces the insertion loss in shunt capacitive switches by decreasing the up-state capacitance \[ C_{up} \]. To express the actuation voltage explicitly in terms of the structural parameters, we rearrange Eqn.6.26 as

\[ V = \sqrt{\frac{\alpha_{max}^3}{2.3914 \varepsilon_0 W} \left( k_T + 2 \frac{L}{l^2} EI_y \right)} \] (6.30)

In the case of the symmetric double beam configuration we have:

\[ k_t = \frac{2GI_t}{l_t}, \quad G = \frac{E}{2(1 - \nu)}, \quad \text{and} \quad \alpha_t k_t = M_t \] (6.31)

where \( E \) is the Young’s modulus, \( \nu \) is Poisson ratio, \( k_t \) is the constant of the spring and \( M_t \) is the spring torque. Eqn 6.30 thus can be rewritten as

\[ V = \sqrt{\frac{\alpha_{max}^3}{2.3914 \varepsilon_0 W} \left( \frac{E}{(1 - \nu)} \frac{I_t}{l_t} + 2 \frac{L}{l^2} EI_y \right)} \] (6.32)
### Table 6.1: Designed dimensions of the toggle switch structures.

<table>
<thead>
<tr>
<th>Structure →</th>
<th>Torsion Electrode ((L_3)(Au))</th>
<th>Actuation Electrode ((L_2)(poly))</th>
<th>Torsion Spring ((Au))</th>
<th>Connecting Lever (l(Au))</th>
<th>Central Area(^a)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length</td>
<td>200-300</td>
<td>0.44L_3, L_3</td>
<td>60</td>
<td>0.75 − 2L_3</td>
<td>150-300</td>
</tr>
<tr>
<td>Width</td>
<td>200-300</td>
<td>0.44L_3, L_3</td>
<td>20</td>
<td>70</td>
<td>90-300</td>
</tr>
<tr>
<td>Thickness</td>
<td>1.5, 5(^b)</td>
<td>poly-layer</td>
<td>1.5</td>
<td>1.5</td>
<td>1.5, 5(^b)</td>
</tr>
</tbody>
</table>

\(a\) - three types of capacitive devices with area \(150 \times 90\), \(250 \times 150\) and \(300 \times 300\mu m^2\).

Substituting for the moments of inertia for the lever and torsion elements i.e. \(I_t = C_1 h_t b_t^3\); \(C_1 = f(h_t/b_t) = 0.33\) \([85]\), and \(I_y = (b h^3)/12\), where \(h_t, b_t, l_t, h, b,\) and \(l\) are the height, width and length of the spring and lever respectively, the final closed form analytical expression for the actuation voltage in terms of STS physical parameters can be written as:

\[
V = \sqrt{\frac{E}{2.3914 \varepsilon_0 W} \left( \frac{G_0}{L} \right)^3 \left( \frac{C_1}{(1-\nu)} \frac{h_t b_t^3}{l_t} + \frac{L b h^3}{l^2} \right)}
\]

(6.33)

The above analytical model has been used to calculate the design parameters for the fabricated devices. As shown in Table 6.2, a good agreement between the simulations and the model is observed. However, the model does not consider non-linear bending of the springs, squeeze film damping and residual stress in the structure. The small angle approximation changes the pull-in by less than 1% over even \(10^\circ\) \([82]\). Of all the limitations, the residual stress and stress gradients in the structure effect the device performance more seriously. The electrostatic nonlinear bending can be avoided by the use of shorter springs.
6.5 Design and Layout

Similar to meander based switches, for STS also, the switching voltages and RF performance are the primary criteria. In order to achieve low actuation voltages, the geometrical dimensions of the torsion actuator and connecting lever need to be optimized. The RF performance mainly depends on the capacitive area, the parasitic impedances arising because of the silicon wafer, finite resistance of the bridge material, the CPW ground and connection lines. In capacitive shunt configuration, the RF behavior is described by the ‘T’ electrical model mentioned in Chapter 4. For ohmic switches the contact mechanism is one of the major limitation both for performance and device reliability.

Equation 6.33 and Figs. 6.6 - 6.9 show that there can be numerous combinations to arrive at specified actuation range of 5 – 15 volts. The device dimensions optimized for the specified range are summarized in Table 6.1. The layout for two representative devices is shown in Fig.6.5. The selection criterion is based on: (1) the intended operating characteristics, (2) the fabrication process and material constraints and (3) over-all device dimensions. For example, the thick resist photolithography, used for defining the bridge and resist etch rate in the sacrificial layer removal step limit the vertical gap height and the pitch of the holes in the electro-plated gold structures. The two process dependent parameters influence the actuation and isolation characteristics of the switches. The total length of the movable bridge structure is also critical because of the residual stress in the bridge layer. Similar to meander based devices, bridge fabrication consists of electron-beam evaporated Cr and Au seed layers followed by thick electro-deposited gold in two steps. The structure thus obtained develops residual stress gradient along the beam thickness. The ensuing out-of-the plane deflection is proportional to the length of the structure.
Figure 6.5: Layout of a capacitive shunt switch (left) and an ohmic switch (right) in CPW configuration. The dimensions for the fabricated devices are summarized in Table 6.1 [94]. Therefore, optimized minimum dimensions are preferred.

6.5.1 Electrostatic Actuation

6.5.1.1 Torsion Actuator

The electrostatic actuation part of the STS mainly consists of micro-torsion actuators. The design parameters obtained using the methodology outlined in Fig.6.3 provide a good first approximation. The obtained parameters are used in the analytical model given by Eqn.6.33, and fine tuned to arrive at desired device geometry consistent with the operating characteristics.

The graphical representation in Figs.6.6. - 6.9 provides an insight into the inter-dependency of the parameters. For example, Fig.6.6(a) shows the pull-in voltage dependence on connecting lever length \( l \) and torsion actuator length \( L_3 \). For \( l \leq 20\mu m \), \( V_{\text{pull-in}} \) strongly depends on \( L_3 \) (\( \leq 180\mu m \)) and results in an impractical range for low actuation voltages or...
Figure 6.6: (a) Calculated $V_{\text{pull-in}}$ vs torsion electrode length at various connecting lever lengths. The useful range for $L_3$ lies between 200 - 300 $\mu$m with $l > 50\mu$m. (b) $V_{\text{pull-in}}$ varies only by 3% - 6% for lever lengths $\geq 300\mu$m.

larger dimensions. However, for $l > 50\mu$m, $V_{\text{pull-in}}$ is less sensitive to the actuation electrode dimensions, as shown by almost coinciding curves "e" in Fig.6.6(a) and by the expanded view in Fig.6.6(b), for $L_3 = 200\mu m - 600\mu m$. The electrode length $L_3 = 200\mu m - 300\mu m$ with lever length $l > 50\mu m$ and width $= 70\mu m$, provide a low pull-in range (< 15 Volts) with compact dimensions. The lever length also affects the pull-out behavior and RF response of the device as discussed in the following sections. The “parasitic” inductance of the lever improves the isolation and can also be used for tuned filtering applications [75].

Torsion actuator spring dimensions are critical for the overall dimensions of the device. Fig.6.7(a) shows the variation of the actuation voltage w.r.t. the ratio of spring width to length, at a given top electrode length ($L_3$). In particular, for $L_3 \leq 200\mu m$, a spring width to length ratio less than 0.3-0.4 is required. Also, as shown in Fig.6.7(b), actuation voltage is more sensitive to spring thickness as compared to other dimensions, particularly for shorter electrode dimensions. Variations in spring thickness due to drift
in electro-plating bath parameters were observed as one of the main causes for out-of-the-plane deflections and change in actuation voltage. In general, to get a large tilt, the spring needs to be as thin and long as possible for a given plate thickness. However, if the spring is too long and thin, the resulting electrostatic force induced bending - mode movement decreases the effective gap height and pull-in angle, thus limiting the tilt range.

6.5.1.2 Gap Height Optimization

The gap height, denoted by \( g_0 \) or \( G_0 \), is the air gap in vertical direction, by which the whole micro-machined moving structure is separated from the conducting layers on the silicon wafer. The gap height is an important parameter as both the actuation voltage and the RF response depend on it. A larger gap height implies better isolation in series configuration and less transmission losses in shunt capacitive case. However, as seen in Fig.6.8(a), a larger gap affects the actuation voltage adversely. The push-pull configuration of STS implies that \( G_0 \) can be lowered while maintaining the same
RF performance. In the switch ‘on-state’ in the absence of pull-out electrodes, the maximum height difference (for a capacitive-shunt) between the contact area and the CPW central conductor is fixed at zero bias height $G_0$. However, as shown by the simulated response (ANSYS) of the bridge in Fig.6.8(b), with a bias on the pull-out electrodes the gap height can be increased to $2G_0$, improving the RF-performance, as the reflection losses are much smaller. On the other hand, in a standard design for a zero bias gap $G_0 = 3\mu m$, the $V_{\text{pull-in}}$ is $13.25V$ volts. In presence of a pull-out mechanism, $G_0$ can be reduced to $2\mu m$ ($V_{\text{pull-in}} = 7.25V$), while maintaining a ‘on-state’ gap of $G = 4\mu m$ which is larger than zero bias position. With a lower gap, the actuator dimensions can be further reduced to get a compact design. However, in the ITC – irst ‘multiuser’ process [49] the spacer resist thickness which determines the gap height has been fixed at $3\mu m$. 

Figure 6.8: (a) $V_{\text{pull-in}}$ vs gap height for top electrode length varying from 100 - 300 $\mu m$. Gap height of $3\mu m$ with $L_3 = 200 \mu m – 300 \mu m$, provides actuation voltage range of 5 - 15V. (b) Simulated (ANSYS®) cross-sectional view of the bridge corresponding to pull-in and pull-out bias.
6.5.1.3 Travel Range and its Applications

In the case of a STS travel range is defined as the maximum tilt angle of the torsion actuator or the maximum vertical displacement of the bridge overlap area, proportional to the applied actuation voltage (Eqn.6.27). When the electrode dimensions are not equal, the equation is modified to \( \beta \alpha_{\text{pull-in}} = 0.44949 \alpha_{\text{max}}, \) where \( \beta = \frac{L_2}{L_3}, \) is the length ratio of the bottom actuation electrode to the corresponding top torsion electrode, assuming the widths are equal. The normalized travel range for \( G_0 = 3\mu m \) is shown in Fig.6.9. The tilt angle or the vertical movement of the movable bridge structure can be controlled over a large percentage of the gap height with maximum of 44% at \( \beta = 1. \) When \( \beta = 0.44, \) \( \alpha_{\text{pull-in}}/\alpha_{\text{max}} = 1, \) which implies that tilt over the entire zero bias gap height is proportional to the applied actuation voltage. For example, with \( G_0 = 3\mu m, \) (\( \beta = 1 \)) the beam movement is proportional to applied bias, till the beam has moved by 1.32\( \mu m \) from its zero bias position. A further increase in the bias sets-in the nonlinear electrostatic attraction and the beam snaps-down to the...
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Figure 6.10: 3-D model of the symmetric toggle switch with exaggerated z-dimensions, showing the absolute displacement $|U_z|$ of the bridge corresponding to an applied z-directed load on the inner-actuation electrodes.

bottom plate. Though not important for switching applications, STS can be configured to form a MEMS varactor with a higher capacitance range. In electrostatic parallel plate MEMS varactors the range is limited to 33% [19].

6.5.2 Pull-in Voltage Simulations

In order to validate the proposed analytical model, the calculated pull-in voltages are compared with the results obtained using a commercial MEMS simulator ($CoventorWare^TM$). In this section we briefly describe the adopted simulation methodology.

The typical design-simulation cycle proceeds in a linear flow and can be adjusted to refine parameters or design accuracy. As shown in Fig.6.10, a 3 – D solid model of the device is constructed from the device mask layout (gds or CIF format) in conjunction with fabrication process parameters using the ‘process’ and ‘layout’ modules. The material properties and
fabrication process parameters such as material thickness, deposition type (stacked, conformal, or planar), sidewall angles, and mask polarity can be adjusted so as to emulate the real process.

The 3-D model of a switch shown in Fig. 6.10 is subjected to a z-directed uniform load on the inner actuation electrodes of the bridge and is meshed using two kinds of meshing elements. Since electrodes on the silicon wafer are used for electrostatic analysis only, they are surface meshed with quadrilateral element type (element size = 5\(\mu\)m). The device model being a orthogonal geometry, Manhattan mesh option with parabolic elements is preferred. Since the devices are intrinsically high aspect ratio structures, the element size of 15\(\mu\)m in x and y direction, and 1\(\mu\)m in z-direction has proven to be sufficiently accurate. Smaller element size increases the mesh density and results in much longer computational time. The \textit{MemElectro} and \textit{MemMech} solvers provide the first approximation of the electrostatic (capacitance corresponding to applied voltage) and mechanical behavior (displacement and stress) of the device.

The \textit{CosolveEM} couples the electrostatic and mechanical solvers and is used to calculate the pull-in voltage and hysteresis characteristics of the
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<table>
<thead>
<tr>
<th>No.</th>
<th>Lever Length ($\mu$m)</th>
<th>$V_{\text{pull-in}}$ (Model) (Volts)</th>
<th>$V_{\text{pull-in}}$ (Simulated) (Volts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>150</td>
<td>9.23</td>
<td>9.5</td>
</tr>
<tr>
<td>2</td>
<td>200</td>
<td>8.48</td>
<td>9.0</td>
</tr>
<tr>
<td>3</td>
<td>300</td>
<td>7.9</td>
<td>8.5</td>
</tr>
<tr>
<td>4</td>
<td>400</td>
<td>7.7</td>
<td>8.25</td>
</tr>
</tbody>
</table>

Table 6.2: Calculated and simulated ($CoventorWare^{TM}$) $V_{\text{pull-in}}$ for devices with torsion electrode length and width = 200$\mu$m, spring length = 60$\mu$m, width = 20$\mu$m, thickness = 1.5$\mu$m, gap height = 3$\mu$m and lever width = 70$\mu$m.

device structure. In CosolveEM pull-in voltage mode, the growth of the electrostatic force becomes dominant over the linearly increasing mechanical restoring force - and the beam quickly snaps to the ground plane. Once the beam has touched the contact surface - a mathematical definition of the rigid plane, in present case 3$\mu$m below the beam, beyond which beam doesn’t move, a release voltage can be found. At this voltage the electrostatic force exactly balances the spring force of the pulled-in beam. Between the pull-in and the release voltages the beam has two valid solutions and exhibits what is known as hysteresis. Figure 6.11 shows the simulated pull-in and pull-out voltages for two representative devices with lever length of 150$\mu$m and 400$\mu$m. Corresponding to a single ramp 0 – 15 V, the simulation tool automatically generates the increasing and decreasing segments from this specification. Both pull-in and release voltage are found in a single simulation pass. In the results shown in Fig. 6.11, voltage trajectory step of 0.25 volts has been used. However, smaller ramp voltage increment steps can be used to find more accurate pull-in and release values. Table 6.2 shows the comparison between the actuation voltages calculated by using Eqn.6.33, and from the $CoventorWare^{TM}$ simulation procedure. The difference of 3 – 7% in the values is attributed to simplifications and approximations assumed in the model. The accuracy of
the model can be improved by considering the nonlinear bending of the 
beams, and squeeze film damping and incorporating the process induced 
residual stress gradients in the fabricated devices. Among all these effects 
the residual stress is more critical as discussed in the following sections.

6.5.3 Reinforced Beam Design

The structural design of the main beam in a symmetric toggle switch is 
similar to the meander based switches as described in Chapter 5, section 
5.5, except the anchor springs and connecting levers. STS is also a high 
aspect ratio structure with dimensions $\geq 1000\mu m \times 200\mu m \times 5\ \mu m$ and 
compliant torsion springs. Both, STS and meander based devices are re-
alized using the same fabrication process. The main bridge frame and 
springs are fabricated using surface micro-machining and gold electroplat-
ing of thickness $1.5\ \mu m$. The combination of higher aspect ratio and use of 
the metals deposited below their reflow temperature makes STS structure 
more susceptible to residual stress related warping and buckling. Also, 
the long structure with a large area actuation electrodes and thin springs, 
can bend under the electrostatic force instead of the rotational movement 
around the springs. Hence, the STS bridge structure is also reinforced 
with additional beams electroplated to an additional thickness of $5\ \mu m$, in 
the second electroplating process, to ensure a flat structure without any 
warping.

Similar to the meander switches the actuation voltage given by Eqn.6.33, 
is independent of the micro-torsion actuator top electrode thickness. 
Hence, a stiffer structure with higher spring constant (spring constant is 
proportional to cube of the beam thickness) can be realized without af-
fecting any of the other switch dimensions. Out of the various structural 
geometries simulated for lower mass and higher stiffness ($K/m$ ratio) a 
structure with reinforcing beams of thickness $5\mu m$ has been implemented.
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Figure 6.12: (a) Active contact area of a fabricated device, showing the reinforcing ribs and perforated area. (b) shows the simulated stress distribution corresponding to an uniform z-directed load on the inner actuation electrodes of a similar device, which shows that deformation is essentially confined to the springs and levers.

Fig.6.12(a) shows the central part of a fabricated toggle switch with reinforcing beams. The spring and connecting lever thickness is maintained at 1.5μm. The actuation electrodes and the central area are realized in two Au electroplating steps. In the first step, the main bridge structure is electroplated up to a thickness of 1.5μm over the Cr – Au seed layer. In the second step, the thickness of the selected portions is increased to 5μm.

Fig. 6.12(b), shows the simulated stress distribution corresponding to a z-directed uniformly distributed load on the inner electrodes of torsion actuator in this structure. The central area is smaller as compared to the one shown in Fig. 6.12(a). The reinforced electrode plates are nearly stress free. The springs being thin and narrow are subjected to maximum stress which at central portion is $\approx$ 8000 times the stress on top electrodes. The connecting lever has low stress (4 - 8 times less than the spring stress) as it bends in reaction to the rotational movement of the bridge. The
combination of more compliant springs as compared to connecting levers, which in turn are compliant as compared to rest of the bridge results in a structure which is impervious to the electrostatic force induced bending and ensures an optimal contact with the transmission line, as shown by simulations (ANSYS® and CoventorWare™) in Fig. 6.12 and Fig. 6.10.

6.5.4 RF performance - Simulations

The basic symmetric toggle switch can be configured (Fig. 6.1) as a capacitive shunt or series ohmic contact switch. For simplicity we consider the capacitive configuration only. Three designs with capacitive area of 150µm x 90µm, 250µm x 150µm and 300µm x 300 µm have been studied. In this section, the simulated RF-response of a representative device with capacitive area 250µm x 150 µm (corresponding to calculated $C_{up} = 110$ fF
Figure 6.14: Simulated average current density distribution on the bridge in (a) on-state (b) off-state for a 250µm × 150µm capacitive switch. Wave is incident from left side.

and $C_{down} = 13.9\text{pF}$) is given.

As discussed in Chapter 4, a toggle switch in shunt configuration can also be represented as a capacitor between the metal bridge and the signal line with a series bridge resistance $R_b$ and bridge inductance $L_b$. The capacitance depends on the state of the bridge. Similar to meander switches, the STS therefore, can also be modeled by a transmission line with impedance $Z_0$ and a lumped series resistor-inductor-capacitor (RLC) model of the bridge, with capacitance varying from minimum $C_{up}$ to maximum $C_d$. The impedance of the bridge can be approximated by $Z_b = R_b + j(L_b - 1/C_b)$, where $C_b$, the bridge capacitance depends on the switch state. At resonance frequency, $f_0 = 1/(2\sqrt{2\pi L_b C_b})$, the switch impedance is closely approximated by the bridge resistance $R_b$. The on-state resonance frequency of the switch with active capacitance area of $250\mu m \times 150\mu m$, lever length $l = 150\mu m$, and extracted inductance and capacitance = 2.56$\mu \text{H}$ and
Figure 6.15: (a) Simulated insertion loss of a switch (250µm × 150µm) at zero bias gap ($G_0$) and 2$G_0$, corresponding to pull-out bias. (b) return loss at $G_0$ and at 2$G_0$. (c) Isolation (off-state) for four lengths of the connecting levers, effect of inductance change is well resolved as compared to meander switches (d) return loss has negligible inductance change effect.

$140fF$, respectively, is 265GHz, where as off-state resonance frequency is 14.5GHz. The inductance is especially important in STS as it can be effectively utilized to shift the resonance frequency and improve isolation in off-state, as discussed in the following section and shown in Fig. 6.15.

The basic substrate and structure parameters used to simulate the switch are shown in Fig.6.13. The simulator has a two layer system - the dielectric layers and the metallic layers. By default the layer at $z = 0$ is the ground plane which theoretically extends to $-\infty$ and has conductivity of $4.9e^7s/m$. The topmost layer is air and extends to $1.0e^{18}\mu m$. The interface in the dielectric layer system is defined as a metallic layer.
The substrate parameters include the number of dielectric layers, the z-coordinate of the top surface, the substrate permittivity, permeability and conductivity. Complex permittivity and permeability are accepted in the simulations. The meshing depends on the highest frequency and the number of cells per wavelength (Ncell). A higher discretization rate is preferred for higher accuracy. However, 15 – 20 cells per wavelength already provide a stable solution in this method of moment simulator. For structures much shorter than the wavelength finer meshing is required [95].

The simulated (HFSS) RF characteristics of a representative device (250µm x 150 µm) for four lever length - combinations are shown by Figs. 6.14 and 6.15. The devices, ranging in length from 1275 - 1775 µm, are implemented in 50Ω standard CPW (75 – 90 – 75µm) configuration and are fed through 300 µm long line sections. The simulated average current density distribution in the bridge up and down state is shown by Fig.6.14. It is seen that the current is concentrated on one edge (Fig.6.14(b), left edge) of the bridge since this edge presents a short circuit to the incoming wave. Similar to the meander switches, the current density distribution
Table 6.3: Inductance and resistance for a switch with capacitance area of 250µm x 150 µm, extracted from the simulated response in bridge down state.

<table>
<thead>
<tr>
<th>No.</th>
<th>Lever Length( µm)</th>
<th>Resonance Frequency ( GHz)</th>
<th>Extracted Inductance(pH)</th>
<th>Extracted Resistance(ohms)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>150</td>
<td>14</td>
<td>2.56</td>
<td>0.22</td>
</tr>
<tr>
<td>2</td>
<td>200</td>
<td>13</td>
<td>3.49</td>
<td>0.28</td>
</tr>
<tr>
<td>3</td>
<td>300</td>
<td>11.5</td>
<td>4.85</td>
<td>0.33</td>
</tr>
<tr>
<td>4</td>
<td>400</td>
<td>10</td>
<td>6.57</td>
<td>0.41</td>
</tr>
</tbody>
</table>

qualitatively demonstrates the RF behavior of the device, highlighting the transmission loss in on-state and isolation in the off state.

Fig.6.15(a) shows the transmission loss \((-20 \log(S21))\) of the switch in on-state. In normal configuration, i.e. at zero pull-out bias, transmission loss is \(\approx -0.36\, \text{dB}\) at 10GHz. With bias on the pull-out electrodes, the transmission loss improves to \(-0.15\, \text{dB}\). The pull-out increases the effective gap height to twice the zero bias gap. As shown in Fig.6.15(b), the return loss also improves as the capacitance is decreased under pull-out bias. In the simulation curves shown by Fig.6.15(c) and (d), which correspond to the switch off-state, a perfect contact between the bridge and dielectric layer on the transmission line is assumed. In fabricated devices it is limited by the surface roughness and beam deformation. In the floating metal configuration, to some extent the limitation can be the impedance of the metal to metal contact. The isolation given by scattering parameter \(S21 (-20 \log(S21), \text{Fig.6.15(c)})\) and return loss \((-20 \log(S22), \text{Fig.6.15(d)})\) are shown for a device with four different lever lengths (150, 200, 300, and 400µm, respectively). The effect of the lever inductance is clearly shown by the shift in ‘LC’ resonance dip at frequencies from 10GHz to 14.5GHz, as the lever length decreases from 400 µm to 150 µm. A similar effect in another type of STS (active area = 150 µm x 90 µm) is shown by Fig.6.16. Compared to a meander switch with same capacitance in Fig.5.16, Chap-
6.6 Measurements

In this section we discuss the measured pull-in voltage and RF response of the fabricated devices. The fabrication process steps for the symmetric toggle switch are essentially the same as for the serpentine spring based

Figure 6.17: (a) Fabricated symmetric toggle switch (lever length = 300 µm). (b) SEM micrograph showing the details of the spring, anchor and reinforcing ribs.

*However, the response of STS device shows better isolation (-50dB vs -40dB) at lower frequency. In general, in all STS devices, the shift in frequency peak, is larger and well resolved when compared to the meander switches with same calculated capacitances. The reduction in isolation is caused by the increased bridge resistance as lever length is increased. Modifying the width of lever also leads to similar shift but smaller in magnitude. The extracted bridge inductance and resistance values for a representative device are shown in Table 6.3. Since the switches behave as a single LC tank with a well-defined resonance peak, the tuning of the LC resonance frequency by reshaping the connecting levers thus offers the flexibility to greatly improve the isolation in a determined bandwidth. This is of particular interest for low frequency applications, where the shunt switch by nature, has poor performance.*
low voltage actuation switches, described in the previous chapter. The high aspect ratio (1275 -1750 \( \mu m \) x 200 \( \mu m \) x 1.5 - 5 \( \mu m \)) moving structures are realized in electroplated gold as before. Fig.6.17(a) shows one of the fabricated devices with lever length of 150\( \mu m \), while (b) shows the details of the anchor and spring portion of the device. Other dimensions are summarized in Table 6.1. The critical steps which affect the measured response are discussed in following sections, after the measurement results.

### 6.6.1 S-parameter Measurement Setup

The measurement set-up used for characterizing the switches is shown in Fig. 6.18. The PC-based set-up is built around a RF-Probe station (probe pitch = 150\( \mu m \)) and consists of a Vector Network Analyzer (HP VNA 8719D) and a precision semiconductor parameter analyzer (Agilent-4156B)\(^1\). The setup measures the de-embedded s-parameters of the DUT. The response of the connecting standard 50\( \Omega \) coaxial interface to non-coaxial CPW environment and RF-probes is removed from the final analysis. The programmable actuation sweep (0 - 50 - 0V ) voltage is provided by Agilent 4156C (signal/monitor unit, SMU1). The actuation bias can also be superimposed on the transmission line using the bias-Tee and SMU2. Both units are controlled by Labview\textsuperscript{TM} based control-software and the characterizing input parameters such as frequency used for real time display of the device s-parameters and actuation voltage incremental steps, can be interactively controlled. The 50\( \Omega \) airline TRL calibrated analyzer (VNA 8719D) provides 50 MHz to 13.5 GHz frequency range and measures the reflected and transmitted RF signal at the device input and output ports, in response to a predefined stimulus. Similar to the meander switches, STS is also considered as two port (input- port1 and out-put port2) switch. All the four s-parameters (S11, S21, S12 and S22) with magnitude and phase, are

\(^1\)ARCES-DEIS, University of Bologna
recorded at predefined sweep voltage steps over the entire frequency range. This facilitates the s-parameter analysis at any voltage within the selected sweep and frequency range. The real time display of the measured return loss and transmission characteristics versus the actuation voltage at a predetermined frequency enables a quicker analysis of the pull-in, pull-out voltages and RF characteristics, which can be further examined with finer voltage increments and at appropriate frequency.

### 6.6.2 Measurement Results

The measurements on switch actuation voltages, isolation and return loss, are summarized in Table 6.4. Figures 6.19 and 6.20 show the measured isolation and return loss vs actuation voltage curves for two basic types i.e. active overlap area of 250 $\mu m \times 150 \mu m$ and 300 $\mu m \times 300 \mu m$, with different lever lengths.
In the table, the pull-in voltage is quoted as the mean of voltages at the onset of actuation and when the device is fully actuated e.g. average of the voltages at point ‘A’ and ‘B’ in Fig. 6.19(c). As seen in the Table 6.4, the measured actuation voltages, differ significantly from the simulated values. For example, in Fig.6.19(a), for a device with lever length of 150μm, the pull-in occurs at ≈ 25 volts, against the calculated pull-in voltage of 9.5V, which when corrected to increased spring thickness becomes 14V. In addition, in some of the devices, isolation vs voltage curves depart from the normal, snap down profile as the voltage is increased (Figs. 6.19(a) and 6.20(c)), indicating the actuation in ‘steps’. The behavior is similar to
meander based switches mentioned in Chapter 5. As further discussed in the following paragraphs, the deviation in pull-in and pull-out voltage is due to the deformation of bridge structures under residual stress gradient. In general devices with smaller lever length (150 µm) have less deviation in actuation voltage. In Fig. 6.20(a) for example, for a switch with lever length of 300 µm, the deviation is more than 500%, where as isolation is acceptable at -23dB.

In most of the devices the measured isolation is better than -20dB measured at 8 and/or 12 GHz. The simulated isolation, in Table 6.4 is mentioned at the resonance frequency of the device. As seen in the simulated response in Fig. 6.15, the isolation above and below the resonance fre-
quency, falls down sharply. For example in Fig. 6.15(c) curve S21-L3, which corresponds to a lever length of 300 \(\mu m\), isolation is \(\approx -22\) dB at 12GHz, where at resonance it is \(\approx -38\)dB. The measured value as shown in the table is \(\approx -21\) dB at 12GHz, showing a closer agreement. The deviation from a normal ‘snap-down’ behavior of the devices shown in Figs 6.19(a) and 6.20(c) arises probably because of the bridge - floating metal contact resistance variation with increase in applied electric field. For devices with floating metal layer on the dielectric, capacitance is constant. The isolation in this case is limited by contact resistance only. The contact resistance is a function of the area, surface conditions and contact force [84]. With increase in the electric field, the ‘deformed’ bridge makes a contact with larger area on the float metal and lowers the resistance. The increased field also overcomes the effect of insulation arising because of the atomic layers of organic matter between the contact points and the bridge in down state. In Fig.6.20(c), the deviation in actuation voltage and isolation are small as compared to other devices. However, the multi-step actuation is achieved over a large voltage range. In the present experimental setup, RF characterization is limited till 13GHz only. Therefore, devices with higher resonance frequency can not be characterized appropriately. Also, the actuation and RF response are expected to improve with smaller increments in actuation voltage sweep.

On the other hand, in conventional devices with no floating metal layer, the change in capacitance under actuation was insignificant, as the overlap between bridge and dielectric is highly constrained by the deformation. The additional bias applied on the signal line through bias tee arrangement, in most of these devices, resulted in failure of the oxide insulation and/or local micro-welding of the bridge to electrodes underneath.
### 6.6. MEASUREMENTS

#### 6.6.2.1 Deformation Measurements

Similar to the meander type devices, the large drift from the expected parameters and the inconsistency of measurements within the similar type of devices, is due to exceptionally high out-of-the plane deflections of the electroplated bridges. This was revealed by the scanning electron microscopy and optical profilo-metric analysis of the wafers. Fig.6.21 shows a 2-dimensional optical profile and x-y line scans of a severely curled-up device, obtained by using Wyko NT100 Optical Profiler. The 3-dimensional view of the same device is shown by Fig.6.22(a). In Fig.6.21, on the top-inner electrode of torsion actuator (with markers), the height of the points on the left and right corners in x-direction, with respect to the center of the electrode, is $3.4\mu m$ and $4.5\mu m$ respectively. The x-deformation of the outer actuation electrodes is also of the same magnitude. The warping levels along the length (y-scan) range from 16.8 to 24 microns for inner and outer electrodes respectively. The connecting lever warping is more sever in case of longer levers. The CPW metal thickness is approximately

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</tr>
</thead>
<tbody>
<tr>
<td>250x150</td>
<td>150</td>
<td>9.5</td>
<td>7.2</td>
<td>25.0</td>
<td>22.5</td>
<td>42.5(14)</td>
<td>16.0(12)</td>
</tr>
<tr>
<td>‟</td>
<td>200</td>
<td>9.0</td>
<td>6.5</td>
<td>45.2</td>
<td>34.6</td>
<td>39.5(13)</td>
<td>23.8(12)</td>
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<td>‟</td>
<td>300</td>
<td>8.5</td>
<td>5.25</td>
<td>72.0</td>
<td>60.5</td>
<td>37.5(11)</td>
<td>21.3(12)</td>
</tr>
<tr>
<td>‟</td>
<td>400</td>
<td>8.25</td>
<td>4.5</td>
<td>85.2</td>
<td>67.5</td>
<td>35.8(10)</td>
<td>24.0(12)</td>
</tr>
<tr>
<td>300x300</td>
<td>150</td>
<td>9.5</td>
<td>6.25</td>
<td>31.0</td>
<td>24.0</td>
<td>36.5(9.0)</td>
<td>25.3(8)</td>
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<tr>
<td>‟</td>
<td>200</td>
<td>9.0</td>
<td>5.75</td>
<td>48.7</td>
<td>36.4</td>
<td>35.2(8.5)</td>
<td>24.9(8)</td>
</tr>
<tr>
<td>‟</td>
<td>300</td>
<td>8.5</td>
<td>4.0</td>
<td>57.2</td>
<td>47.0</td>
<td>33.8(7.0)</td>
<td>22.8(8)</td>
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Table 6.4: Comparison between the simulated and measured pull-in voltage and isolation characteristics for two basic types of symmetric toggle switches.
6 microns. Thus, on the average the curling up of the actuation electrodes have effectively increased the air gap height from 10 to 18 microns, which coupled with the increase in spring thickness explains the high actuation voltages as predicted by graph in Fig.6.8. The warping of the central contact area (3.5 - 4.5 microns) and connecting levers, limits the overlap area between the transmission line and the bridge. It lowers the capacitance in switch off state whereas increasing it in the switch on-state. Thus, the deflection gradient results in lower isolation and higher transmission loss, as shown by the experimental results in Fig.6.19 and 6.20 and the fact that devices without floating metal layer show no significant change in behavior under applied actuation. The high deformation makes further characterization difficult and at times resulted in a complete failure of the device. The uneven contact between bridge and oxide on the transmission line, also generates electric field gradients between the metal beam and transmission
6.7. INTERNAL STRESS IN BEAMS

In the present work, the analysis and the models developed for the symmetric toggle switch and the serpentine switches allow for an first estimate of the spring constant and the actuation voltages. These do not take into account any internal stress on the metallic structural elements. The internal stress or residual stress, a manifestation of the fabrication...
process, however, develops in most microstructures and it presents one of the major challenges in the development of MEMS devices, which has been amply highlighted in the previous sections. A thorough investigation of the phenomenon perhaps can be better addressed by test structures dedicated to stress measurements and fabrication process parameter extraction/monitoring. It could not be incorporated into the present work because of the time constraints. However, in this section, we discuss the issues related to the internal stress and present the measurements and simulation results for cantilever based test structures fabricated along the devices, on the same wafer.

6.7.1 Origin of the Internal Stress

The internal stress (in absence of any external force) in a thin film deposited over a substrate, can be divided in two kinds: extrinsic and intrinsic stress. The origin of the extrinsic stress mainly lies in the adhesion of the film to the substrate; it can be introduced by the mismatch between the thermal coefficients of the film and substrate (thermal stress), by the lattice misfit with its substrate, or due to chemical reaction with its substrate when the intermetallic compounds formed are coherent to the film but have a slight lattice misfit (coherency stress). The intrinsic stress comes from the intrinsic growth factors such as substitutional or interstitial impurities, dislocations and the dynamics of the film growth (cluster formation and coalescence).

From MEMS device perspective, a general uniaxial residual stress field in a thin film can be represented by the polynomial,

$$\sigma_{total} = \sum_{k=0}^{\infty} \sigma_k \left( \frac{y}{h/2} \right)$$

(6.34)

where $y \epsilon (-h/2, h/2)$ is the coordinate across the thickness h, with origin
at the film mid-plane, as shown in Fig. 6.22(b) and (c). For a first order approximation the total stress can be calculated as

$$\sigma_{\text{total}} = \sigma_0 + \sigma_1 (2y/h)$$

(6.35)

The equation implies that the total stress can be expressed as the superimposition of the mean stress $\sigma_0$, negative or positive depending on whether the film is in tension or compression, and a gradient stress $\sigma_1$, about the mid-plane. In physical terms, the mean stress corresponds to the extrinsic stress described above and is primarily caused by the thermal mismatch between the film and the substrate; for switch movable structural element, excluding the anchor posts, it is the Cr-Au-resist interface. The gradient stress (intrinsic) arises from localized effects including the atomic diffusion through film thickness $h$, which in our case, constitutes evaporated Cr-Au film and gold layers deposited in two electro-plating steps. When the structure is released from the supporting substrate or sacrificial layer, the structure becomes free to deform out-of-plane, following the release of the internal stress. The extent and direction of the deflection depends on the sign of $\sigma_1$ and $\sigma_0$ and both can be calculated from the measured deflection profile of a cantilever [96].

The surface micro-machining of metal beam based MEMS structures particularly RF MEMS switches is a low thermal budget fabrication process. In order to protect the underlying metal layers and circuitry, the movable metallic structures are realized using low temperature metal deposition and etching processes. Use of thick ($\geq 3\mu m$) photoresist as sacrificial layer further limits the process temperature. The structural layer deposition has to be performed below the reflow temperature of the resist (typically $\leq 200^\circ C$). Electroplating techniques provide both cost effective and low processing temperature for structural metal layers. However, when a thin metal film is deposited on a sacrificial layer at temperature lower
than its reflow temperature, the intrinsic stress or residual stress develops in the ‘film-sacrificial layer’ system [97]. Under this type of stress the thin film structures experience undesirable deformations which are significant, particularly for high aspect ratio structures [55], as has been observed in the fabricated devices mentioned in above sections (Fig.6.21, 6.22) and Chapter 5. A number of studies have been reported to theoretically explain the mechanism of the stresses [98]- [99] and to experimentally measure their effects [96], [100]. Nonetheless, in general, thin film stress is a complex phenomenon that depends on the specifics of the fabrication process. There is also very little information available on the metallic microstructures built by the thin-film depositions and effective ways to control the stress and its effects.

Therefore, in this section we outline the fabrication process steps which need careful monitoring to minimize the stress development and present the estimation of the residual stress deducted using the measured optical-profile of the cantilever test structures and compared with simulated results.
6.7.2 Electroplating and Structure Properties

The fabrication steps critical to stress behavior of the metallic structural layers are: the sacrificial layer (thick photoresist) lithography and etching, sputter deposition or evaporation of Cr-Au seed layer and gold electroplating. Among these, the Cr-Au composition and gold electroplating realized using gold sulphite solution \( [(NH_4)_3Au(SO_3)_2] \) Ammonium-sulfite-gold(I) in a home-made electroplating bath, are more crucial, as these determine the mechanical characteristics of the electroplated bridge structure.

The characteristics of the electro-deposited metals are mainly influenced by the environment in the immediate vicinity of the cathode. In addition, in MEMS structures, the post-deposition process, which influence the stress behavior are the beam ‘release’ and sintering. The electrodeposits are crystalline in nature and the form of the deposit depends largely on two factors: (i) the rate of formation of the crystal nuclei by the discharge of the ions at the cathode and (ii) the rate at which these nuclei grow into large crystals. If the conditions favor the rapid formation of a fresh nuclei on the cathode, the deposit will tend to consist of small, fine grained crystals. The deposited metal will be smooth and relatively hard. On the other hand, if the nuclei increase in size rapidly, the deposit will consist of relatively large crystals and the surface is rough in appearance [101]- [102]. The main parameters that influence the aforementioned factors, namely the crystal nuclei and their increase in size, are mentioned briefly.

6.7.2.1 Current Density

At low current densities, the discharge of ions happens at a slow rate, allowing for ample crystal-nuclei growth time. The deposits under these conditions exhibit a coarse crystalline structure. As the current density is
increased, the rate of discharge of the ions also increases and fresh nuclei will tend to form. The resulting deposit consists of smaller crystals. The increase in current, within certain limits yields deposits that are fine grained. However, at very high currents densities, the crystal tend to grow out from the cathode towards the region where the solution is more concentrated and hence create trees or nodules in the film. Fig. 6.23(a) shows the stress behavior of the electroplated gold layer with variation in current density, for the gold sulphite solution used for electrodeposition. Minimum stress conditions and good surface quality is obtained when the current density is maintained at $3.0 - 3.5 mA/cm^2$ [103].

### 6.7.2.2 Concentration of the Electrolyte

Increasing the gold concentration of the electrolyte solution can largely offset the side effects of the electroplating at high current densities. Likewise, the use of agitation in the electrolyte also de-accelerates the nodule formation at higher current densities.
6.7.2.3 Temperature

Increasing the temperature promotes the diffusion of ions towards the cathode, thereby preventing the depletion of ions which leads to roughness of the deposits. On the other hand, it also increases the rate of growth of the nuclei, so that the deposits have a tendency to be more coarse. In gold sulphite bath at moderate temperatures (50 – 60°C) the first of the above mentioned effects predominates thus deposited layers have better surface quality. The stress is found to be compressive and can be varied with current density, as shown in 6.23(b). However, at higher temperatures the surface quality of the deposit deteriorates.

6.7.2.4 Impurities

Electro-deposited films normally contain various types of inclusions or impurities. The sources of these impurities may be from one or more of the following: added chemicals (brightener, levelers etc.), cathode products (complex metal ions), hydroxides (of the depositing metals), and bubbles (hydrogen gas etc.). Though the effect of a particular additive is frequently specific for a metal, the additive agents in general have a high surface activity i.e. they tend to adhere to or be absorbed by the surface. Therefore, if the substance covers the nucleus, the further growth of the nucleus will be prevented. The subsequently discharged ions create fresh nuclei resulting in a fine grained deposit. However, excess of the additives and impurities cause the deposit to become brittle and rough where there is a relatively thick layer of impurities and increasing the internal stress.

6.7.2.5 pH

The pH of the solution influences the discharge of hydrogen ions, thus causing the solution in the cathode layer to become alkaline and precipitate
hydroxides and basic salts. In the case of a cyanide gold solution based deposition, the solution locally turns basic, thus causing damage to resist resulting in definition loss. On the other hand, inclusion of the compounds in the deposited film lead to surface roughness. Additionally, the evolution of hydrogen gas is often accompanied by the formation of spots and streaks in the film. The pH of gold sulfite solution is maintained between 6.52 to 7.40, by adding $H_3PO_4$, with temperature fixed at $55^\circ C$. Fig.6.24(b), shows a section of the bridge with poor adhesion between the deposited Au-layers and surface roughness. The aging solution and drift in temperature is believed to be the cause. The deposition with fresh electroplating solution and commercial electroplating bath are still in a trial phase at the time of reporting.

### 6.7.3 Cr-Au Seed layer

The Cr-Au thin film, also referred as ‘seed layer’ is deposited on ‘spacer’ (HPIR 6517HC positive photoresist) and precedes the first gold electrodeposition. Chromium (10nm) is used as an adhesion layer and a diffusion barrier for Au, while Au (150nm) acts as a seed layer for further gold elec-
troplating in gold-sulfite bath. As mentioned in Chapter 5, section 5.3.2, in the first process Cr was sublimated by resistive heating (heating current = 380 Amp, for 30 minutes) immediately followed by Au evaporation, in the same chamber, to avoid chromium-oxide formation. However, as shown in the schematic of the structure in Fig.6.25, the bottom Cr-surface is exposed to oxygen during the removal of sacrificial photoresist in oxygen plasma (at $190 - 200^\circ C$, for 20 min).

Though the Cr film is quite thin, the heating of Cr-Au interface, during sintering at $200^\circ C$, for 30 minutes, in nitrogen ambient, also leads to diffusion of Cr into Au grain boundaries. The resulting structure has - four different kind of ‘layers’ - bottom chromium oxide layer, Cr, Au with diffused Cr - concentration varying from 100 at the interface to 0 %, and gold, electroplated in two separate steps. The oxide formation and diffusion thus changes the mechanical properties of the structural layer [104] and may contribute significantly to the total stress change of the multilayer, due to the high internal stress of the chromium deposit.

In the second process, in which reinforced meander based and toggle switches are fabricated, the Cr-Au seed layer deposition process (resistive heating) has been replaced by electron beam evaporation. But the amount of Cr (0.05 gm) and Au (2.0 gm) are kept the same. Possibly, the amount of Cr sublimated was less. And the Cr residue left in the crucible subsequently evaporated as Cr-Au mixture, resulting in a seed layer with different composition than the first process. Another change in the second process, is the use of the old sulphite solution, probably with impurities and depleted concentration of the constituents. The cumulative effect is believed to be the main reason for high deformation. The precise control of electroplating bath parameters and optimization of Cr deposition can lead to better stress control in the beams.

The effect of sacrificial photoresist etching on Cr-Au multilayer stress
has been studied in [103] for a blank wafer in which chromium oxidation effect is not present, but Cr diffusion is expected due to high temperature. As shown in Fig. 6.24(a), the change in stress of the multilayer, after the removal of the resist is considerable, with final tensile stress of about 180 MPa. This extrinsic stress can be eventually explained by the assumption of complete relaxation of the gold layer during the annealing cycle: if at high temperatures the internal stress is released, generating a low stress state, during the cool down to ambient temperature internal stress is generated which is proportional to the thermal gradient and is given by [105]

\[
\sigma_{th} = \frac{E_f}{1 - \nu_f} (\alpha_f - \alpha_s) (T_2 - T_1)
\]

(6.36)

where \((T_2 - T_1)\) is the thermal gradient, \(\alpha_f\) and \(\alpha_s\) are the temperature coefficients of expansion for the film and substrate respectively, and \(E_f\) and \(\nu_f\) are the Young’s modulus and Poisson’s ratio of the film. According to the formula, the resulting stress is 216 MPa, which is close to the observed stress of the samples.

**6.7.4 Estimation Residual Stress in Test Structures**

Figure 6.26 shows the optical profiles, line scan and simulated deflection profile for the test structures consisting of a set of cantilevers. As shown in figure (a) the z-deflection is particularly high (30-90 µm) for test structures near the wafer edges as compared to the one at the center in figure (c). In order to estimate the residual stress from the measured deflection profile, three cantilevers (150, 250, 300 µm long) from the test structure in Fig. 6.26(c) are considered. The cantilevers are supposed to have two layers structure: first the electroplated gold and Cr-Au seed layer and second, the thicker electroplated gold layer. Assuming isotropic growth of electroplated gold structures (\(\sigma_x = \sigma_y\), and \(\sigma_z\) is very small) stress, in one of the layers
Figure 6.26: (a) 3D optical profile of a cantilever test structure near the wafer edge. (b) simulated deflection profile of a cantilever (250 $\mu m$ x 30 $\mu m$ x 6 $\mu m$), (c) test structure at the center of wafer, used for stress estimation, (d) measured optical profile of the cantilevers in (c).

is varied till the simulated deflection is equal to the measured maximum deflection. The second layer is assumed to be stress free. The selection of the top or bottom layer for variable stress will only change the sign of the resulting deflection. As shown in Fig. 6.26(b) and (d) and summarized in Table 6.5, the measured and simulated deflection are closer when $\sigma_x = \sigma_y = 285$ MPa. The stress estimated on a blank wafer, for the second process under which the symmetric toggle switches and reinforced devices have been fabricated, is $\approx 320$ MPa. Also, as seen in the table, after assigning relative stress components to the layers, the difference in Young’s modulus of the two layers have little effect. The measured and simulated deflections closely agree for small cantilever lengths. For lengths $\geq 300\mu m$, simulation
### Chapter 6. Symmetric .... 6.7. Internal Stress in Beams

<table>
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<tr>
<th>Lever Length $\mu m$</th>
<th>$\sigma_x = \sigma_y$ (MPa)</th>
<th>E(YM) Layer1 (GPa)</th>
<th>E(YM) Layer2 (GPa)</th>
<th>Defl. Sim. $\mu m$</th>
<th>Defl. Meas. $\mu m$</th>
</tr>
</thead>
<tbody>
<tr>
<td>150</td>
<td>285</td>
<td>108</td>
<td>80</td>
<td>3.4</td>
<td>3.1</td>
</tr>
<tr>
<td></td>
<td>285</td>
<td>108</td>
<td>108</td>
<td>3.3</td>
<td>3.1</td>
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<tr>
<td>250</td>
<td>285</td>
<td>108</td>
<td>80</td>
<td>9.4</td>
<td>9.4</td>
</tr>
<tr>
<td></td>
<td>285</td>
<td>108</td>
<td>108</td>
<td>9.1</td>
<td>9.4</td>
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<tr>
<td>300</td>
<td>285</td>
<td>108</td>
<td>80</td>
<td>13.5</td>
<td>13.7</td>
</tr>
<tr>
<td></td>
<td>285</td>
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Table 6.5: Estimation of residual stress by comparing the simulated and measured deflection profiles of cantilever test structures. The second layer is assumed to be stress free.

Convergence problems were observed. Probably the deflection is too large a fraction of the cantilever dimensions (cantilever width = 30$\mu m$, deflections are 30 - 90 $\mu m$ for larger cantilevers).

The other fabrication process steps which change the stress behavior are - gold electroplating and thick resist photolithography. Over the same device, electroplated gold thickness variation of 50 - 70 $\%$ were observed. Thickness profile of the 'spacer' photo-resist (3.5 microns) after a hard bake at 200$^\circ$C, reveals variations of more than 15$. The effect coupled with the change in concentration of electrolyte constituents may leads to non-uniform thickness on the same structure. The surface roughness and adhesion between the first (1.5$\mu m$ thick on Cr-Au seed layer) and second (3.5$\mu m$ thick) electroplated gold layers depend on the quality of electrolyte and operating parameters of the electro-chemical bath. In Au-layers electroplated using sulfite electroplating solution (Ammoniumsulfite-gold(I), [(NH4)3Au(SO3)2] ) internal stress can vary from compressive to tensile with change in the current density. In the first process run, using the same fabrication steps, for devices and test structures with nearly same aspect ratios, out of the plane deflections were comparatively small. However, in
the present fabrication run drift in electrodeposition bath parameters, aging of the solution and Cr-Au seed layer composition probably have affected the resulting surface quality, adhesion and obtained thickness. Recent optimization experiments with Cr deposition show that with reduction in quantity of Cr metal and change of electroplating solution, the deformation in the structures has reduced significantly. The improved control of the fabrication process - is expected to result in devices with minimal drift from designed parameters.
Chapter 7

Conclusion

The present research work is concerned with the study of electrostatically actuated RF MEMS switches. The inspiration stems from the highly promising RF characteristics of the MEMS switches and the ever-increasing need in communications industry for linear, low actuation and less power-consuming switches, compatible with existing IC fabrication technologies. The current state-of-the-art RF MEMS switches mostly, have been developed for high value defence oriented applications, with excellent RF characteristics. However, the commercial applications are still under investigation mainly because of the operational voltages, long term reliability and the prohibitive cost in comparison to solid-state devices.

The main goal of the present work thus, has been to explore the design and fabrication feasibility of low actuation RF MEMS switches with standard IC fabrication techniques. Most of the reported RF MEMS switching devices have been designed for higher operational voltage because of its interdependence with switching speed. Nevertheless, in many wireless communication applications, high performance - low power consumption switching systems are preferred, where switching time of 50-100 microseconds is adequate. In view of above, present work is focused on the switch design and fabrication with actuation voltages 3 - 15V and fre-
quency range of 1-25 GHz. Of the various configuration - capacitive shunt and series ohmic contact implementations in CPW have been studied. Inherently, series ohmic contact devices show better RF performance from few MHz till 10GHz, where as capacitive shunt devices have reasonably good isolation and transmission characteristics above 8 GHz. For this reason both capacitive and ohmic contact switch have been considered. The basic flexure design has been configured in such a way that it can be modified to any of the above mentioned types by changing the suspended bridge design. Similarly, symmetric toggle switch can also be configured as a capacitive or ohmic contact device.

From the design perspective, the switch encompasses both the electrical and mechanical domains. In the mechanical domain, the switch is considered as micro-mechanical structure with a free member constrained to move in z-direction and is represented by the lumped parameter models of surface micromachined components. After discussing the general equations of motion - analytical models has been presented for the cases specific to RF MEMS switches. In order to achieve low equivalent spring constant for the switch, various flexure-support designs have been investigated. In view of the achievable equivalent spring constant, stress alleviation characteristics and fabrication complexity, serpentine meander flexures have been implemented. A closed form analytical expression is discussed to calculate the spring constant for a meander spring. The analytically calculated and numerically simulated values agree within 5%. A brief treatment is also provided on models, used to calculate the actuation voltage, the correlation to power handling capabilities and effect of external vibrations on the devices.

The electrical behavior of the capacitive shunt and ohmic switches is discussed in terms of the available R L C model. The model correlates the geometry of the device to its scattering parameters, which are used to
CHAPTER 7. CONCLUSION

characterize the device RF response. The devices are implemented in CPW configuration. A brief discussion on CPW, provides a closed form analytical expression to synthesize CPW dimensions. From the RF applications perspective, three basic configurations categorized by the bridge active overlap area e.g. $150\mu m \times 90\mu m$, $250\mu m \times 150\mu m$ and $300\mu m \times 300\mu m$, have been considered and numerically analyzed.

The first exploratory fabrication run consisted of meander based capacitive shunt devices with actuation voltages ranging from 10 - 15V, and similar off-on state capacitance ratio for all the devices. The fabrication process is based on Au-electroplating (for the structural elements), surface micromachining, and modified CMOS processing steps. The capacitance-voltage measurements of the devices show that the actuation voltages are lower than the expected. The capacitance ratio was also found to be lower than the calculated and simulated values. The discrepancies are discussed using a model - which depicts the pre and post deflection behavior of the switches. The low pull-in and capacitance ratios are caused by the residual stress gradient in the Au-beams, which lowers the effective gap height. The height difference between the actuation pads and the underpass is found to give rise to additional parasitic capacitance, which dominates the switch behavior in off-state. The measured RF response though lower than expected is satisfactory; off-state isolation is better than -30dB at resonance, and -20dB from 18 - 25GHz. In order to include the parasitic capacitance - arising from the surrounding CPW ground and finite resistance of the substrate, which shift the measured frequency, improvements in the RLC model are also suggested.

In view of the pre-deflected beams under the residual stress and the parasitic capacitances arising because of height mismatch between the actuation electrodes and underpass, the bridge fabrication process and design were optimized. The electroplated beam structures have been reinforced by
raising the bridge thickness to 5µm in selected areas. The simulated behavior shows that the beams behave like rigid plates with equivalent spring constant much higher than the suspension springs. This ensures planar beam profile over the underlying electrodes to achieve optimum capacitance. The down state capacitance is also limited by the surface roughness of the dielectric layer and beam bottom side. In order to optimize the switch capacitance ratio, a floating metal layer design is introduced. This not only solves the surface roughness problem but also increases the capacitance ratio depending upon the beam area over the transmission line.

The inherent susceptibility of the low spring constant switch design to RF signal self-biasing and external vibrations can be circumvented by adding another bridge electrode on the top. However, the ‘two bridge’ structure leads to a more complex fabrication and stringent process requirements. Instead, we have proposed an innovative switch configuration called Symmetric Toggle Switch. The device is based on micro-torsion actuators and has additional pair of ‘hold’ electrodes to avoid self biasing and actuation by external vibrations. The analytical model which correlates the device geometry to actuation voltage shows excellent agreement with numerically simulations. The simulated RF response for three types of switches show isolation better than -30dB and isolation of ≤ -0.15dB, over a frequency range of 8-20GHz. In addition to an improved insertion loss and low actuation (< 10V) in switching applications the device can also be configured as a MEMS varactor with a capacitance range much higher than the conventional MEMS design approach. The well defined resonance peaks which can be shifted by modifying the beam dimensions without affecting the actuation - makes it suitable for filtering applications over a selected narrow bandwidth.

The measured response of the reinforced meander devices and toggle switches however shows large discrepancies in actuation voltage as com-
pared to the simulated values. The RF response for the devices with floating metal option is satisfactory; isolation better than -25dB and insertion loss $\leq 0.2$dB over a frequency range of 8-25 GHz. The conventional devices with dielectric on the underpass segment of the transmission line show higher deviation from the simulated behavior. The difference arises because of the residual stress gradient in the bridge structures. In conventional capacitive and ohmic contact switches the deformation prohibited a proper overlap/contact of the bridge with transmission line. In view of this it was not possible to extract any meaningful information. In electroplated gold layers which constitute the mobile beam, the stress is generated because of the varying Cr concentration along the beam thickness, process parameter drift and the fact that metallic layers are deposited below their reflow temperature. Initial process optimization experiments - show that the stress can be reduced by optimizing the Cr seed layer deposition and precise control of the electroplating bath parameters. With stress gradient related deformation, limited to half a micron or less, the device performance close to the simulated can be achieved.

In summary, this work, has demonstrated the feasibility of fabricating low actuation RF MEMS switches with actuation voltages ranging from 3V to 15V with acceptable RF response, in the frequency range of 1 - 25GHz. The fabrication process is based on surface-micromachining, Au-electroplating and standard IC fabrication steps. An innovative switch design concept has been presented and validated using commercial simulation tools. Device fabrication feasibility based on this design has also been demonstrated. The devices are very likely to find use as low actuation RF switches, MEMS varactor or in tunable MEMS filters. The low actuation voltage devices (3-15V), in a 50 $\Omega$ system need 0.02 -10 watts of RF power for self biasing. Therefore, low voltage MEMS bridges can also be used as protective circuitry in front of sensitive amplifiers and electronics.
7.1 Future Work

Over a decade or more, a lot has been accomplished in the field of RF MEMS switches. Similar to CMOS technology, the present phase could be the “fine-tuning” era for MEMS devices. The general understanding and accessibility of electrical and mechanical modeling and design tools especially for students community has improved over last few years (ITC-irst and University of Bologna acquired MEMS simulation tools and RF measurement equipment last year). However the fabrication process needs more efforts than is generally perceived. The topics of future work related to low actuation voltage include the yield, reliability and packaging. Few of the points specific to present work are:

- Study of the stress mechanism in electroplated metals.
- Study of the metal to metal contact mechanism and materials, in direct contact switches for higher power applications.
- Dynamic properties of the symmetric toggle switches.
- Low actuation $\leq 10$ V, high speed 100-500ns RF MEMS switches.
- Varactor applications of symmetric toggle switch.
- Low-cost, high yield hermetic packaging techniques
- Reliability studies of RF MEMS switches under different temperatures and radiation effects.
- RF MEMS switch VHDL models and switch library similar to the digital design.
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