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Micro- and nano-scale switches and tuning elements for microwave applications

Thomas P. Ketterl
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Micro- and Nano-Scale Switches and Tuning Elements for Microwave Applications

by

Thomas P. Ketterl

A dissertation submitted in partial fulfillment
of the requirements for the degree of
Doctor of Philosophy
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Dedication

To Marianne for all the support, patience and love she has given me.
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Micro- and Nano-Scale Switches and Tuning Elements for Microwave Applications

Thomas P. Ketterl

ABSTRACT

In this work, various components for low power RF telemetry applications have been investigated. These designed, fabricated and tested devices include radio frequency (RF) micro-electro-mechanical systems (MEMS) switches, single-pole-double-throw (SPDT) RF MEMS switches, nano fabricated capacitors and switching devices, and micromachined microstrip patch antennas.

Coplanar waveguide (CPW) RF capacitive switches in shunt and series configuration were designed for high isolation, low insertion loss, and fast switching speed. Switches with > 35 dB isolation, < 0.3 dB insertion loss and switching speeds in the 10’s of μs were fabricated and measured. These switches were packaged using photo-imagable resists and flip-chip bonding techniques. The MEMS shunt switch topology was also implemented into a single-pole-double-throw (SPDT) design by utilizing two such switches in a series and a shunt configuration, offset by a quarter wavelength section to provide a RF short at the input of the shunt switch in the off state. This type of design has the advantage of requiring a simple on-off (0 V and 35 V) bias supply to select the switch state.
Also, the use of a focused ion beam (FIB) tool to mill sub-micron gaps in CPW transmission line structures was investigated. Nearly ideal capacitors in the micro- and mm- frequency range with capacitance of 8-12 fF were obtained using this milling technique. The FIB’s capability to mill such small gaps at an oblique angle was also utilized to fabricate RF nano switches. These devices were switched with speeds of less than 300 ns with voltages of less than 20 V.

Finally, solid state and packaged MEMS switches were integrated into a novel binary amplitude shift keyed (BASK) modulating RF telemetry system to provide the modulation of a redirected 10 GHz continuous wave (CW) signal. A pair of cross-polarized micromachined microstrip patch antennas was used in the system to receive the CW signal and re-transmit the modulated signal. A transmission range of over 25 m was demonstrated with the solid state switch reflectenna.
1.1 Overview

Micro-electro-mechanical systems (MEMS) technology is growing at a rapid rate and is already finding a multitude of applications in optical engineering [1], microfluidics and biochemical analytic systems [2, 3], mechanical resonators and accelerometers [4, 5], and environmental sensing and biomedical products [6, 7]. The attractiveness of MEMS devices can be attributed to their small size (typically 1 µm to 1 mm), low power consumption, manufacturability using well characterized integrated circuit (IC) fabrication techniques, and ease of integration with other system devices. However, recent research in this technology has not been limited to the micro-scale level. Already many manufacturing techniques are being investigated to produce devices and device components with sub-micron dimension, i.e. in the nanometer scale [8, 9].

Another field where MEMS technology is finding great interest is radio and microwave frequency (RF/Microwave) engineering. Due to their characteristics of small size and excellent performance at high frequencies, RF MEMS devices have the potential to replace solid state IC devices in many micro and millimeter wave applications. This has already been demonstrated in such circuits as switches and switching networks [10], phase shifters [11, 12], tunable circuit elements [13], and configurable devices
including planar transmission lines and antennas [14, 15]. MEMS devices also have the potential to greatly reduce the power requirements in today’s communication systems; this is especially desirable in the ever expanding technologies of mobile communication and wireless sensor networks.

The motivation of this research was to design and develop micro- and nano-fabricated devices as building blocks for a low power RF telemetry system. These components will be integrated with other microwave devices into a novel communication system, called the reflectenna, where a continuous wave microwave signal is received, modulated, and then redirected to the interrogating source. Figure 1.1 shows the mode of operation and layout of the reflectenna system. A pair of micromachined patch antennas will be used to receive and retransmit the carrier signal. As shown, a micro- or nano-switch will modulate the signal with data before retransmission. To differentiate between the send and receive signal, the patch antennas are oriented 90 degrees with respect to each other to provide orthogonal polarization diversity.

![Figure 1.1. Simplified diagram of the reflectenna with modulating switch and transceiver.](image)
In this investigation, RF MEMS switches in shunt and series configuration, as well as a metal-to-metal contact series switch, with coplanar waveguide (CPW) topology were designed, fabricated and tested. The operating frequency was tuned down to 10 GHz using fixed inductive lines in the shunt sections that connect a fixed-fixed beam to the ground planes of the CPW lines. The bridge is suspended 1.5 to 1.8 µm over the center conductor in the un-actuated state. The switches were shown to possess an insertion loss of less than 0.25 dB and isolation of greater than 35 dB at the operating frequency. Switching speeds as fast as 40 µs were measured with corresponding actuation voltages of less than 30 V.

The MEMS switch design was also implemented into a single-throw-double-pole (SPDT) switching device were two MEMS switches were implemented in series and shunt configuration. In this design, a signal can be transmitted through one of the two outputs with the other port in isolation with no bias applied to the switches. When the switches are biased, the signal gets routed through the other port. Port isolations of 15 dB and 35 dB were measured in the actuated an un-actuated states, respectively. Insertion loss in both states was less than 1 dB. The advantage of this design over previously demonstrated multi-port switches is that the switch is operational in one of the states with no applied bias and is switched to the other state when a 35 V bias is applied. This reduces the power required to bias the switch by 50% as well as the complexity of bias network.

The use of a focused ion beam (FIB) tool was also investigated to fabricate microwave devices and switches with submicron features. Nanometer-wide gaps were milled across the center conductor and shunt lines of CPW transmission lines to provide
low value capacitive coupling. FIB milling has the advantage of producing devices where high yield and great precision is needed. Also, FIB milled capacitors have the advantage over their metal-insulator-metal (MIM) counterparts in that they can be processed without the need of a dielectric layer to produce low value capacitors, they can be fabricated with a single layer process without the need masking and corrosive etching steps, and use up much less real state due to their extremely narrow milled area. Using this process, 70 nm wide trenches were milled and capacitances of 8-12 fF with very low parasitic inductance and resistance were extracted from the measured data using equivalent circuit models.

The FIB’s ability to mill at oblique angles was also utilized to develop a nano-fabricated MEMS switch. This switch uses two electrodes to alternate the actuation of a pair of cantilevers which were created by cutting the angled nano gap across a fixed-fixed suspended beam. The gap in the cantilevers could be separated or closed, depending on the biasing, to provide isolation or transmission of the signal. Due to the small distances of travel involved when the cantilevers are actuated, a switching time of less than 300 ns has been obtained with a bias voltage of less than 20 V. The switching speed is about an order of magnitude faster than that of currently reported MEMS switches. An isolation of greater than 25 dB at 10 GHz was obtained with a measured insertion loss of less than 0.3 dB at the corresponding frequency.

Packaging techniques for the MEMS and nano devices were also investigated. A polymer photoresist was used as a sealing ring to encapsulate the switches with a quartz lid. SU-8 was used for the sealant and was patterned on the quartz samples. These lids were then bonded using a flip-chip-bonder at 110° C and a force of 10 N.
A novel wireless telemetry system, called the reflectenna, was demonstrated. A dielectric lid was used to package and seal the reflectenna. A transceiver system was also designed and built to send the 10 GHz CW signal and demodulate the retransmitted signal after reception. A dual polarized reflector antenna was used to transmit and receive. Due to a finite port isolation of the reflector antenna of 55 dB, switches were placed in the transmission and reception paths of the transceiver. In this way, the receiver is turned off when a signal is transmitted and turned on when the CW signal is cut off at the transmitter. This enabled the reflectenna to communicate using binary amplitude modulation (BASK) at distances over 25 m with a data rate of 2 kHz. The benefit with this type of design is that range of communication is mostly limited by the power level and sensitivity of the transceiver thereby greatly reducing the power requirements of the telemetry device at the sensor location.

1.2 Dissertation Organization

Chapter 2 introduces a background to various RF MEMS switching technologies. The current state of technology of high performance switches in this category is discussed.

In Chapter 3, the design and fabrication of a MEMS shunt and series switches in CPW configuration is presented. Simulated and measured results of the RF characteristics are shown as well as the switching speed performance. RF MEMS packaging methods are also discussed and the packaging technique used in this research is presented.
The implementation of the RF MEMS switch design, in a SPDT configuration is presented in Chapter 4. The fabrication and biasing technique is discussed. Measured and simulated frequency response data is compared in this chapter in the actuated and un-actuated switching states. With the use of optimization techniques, an equivalent circuit model with extracted component values for the tested devices in both switching states is presented.

The investigation of nano-fabricated capacitive devices is introduced in Chapter 5. Here, the focus is shifted to the design, fabrication, testing, and modeling of submicron-milled gaps in CPW transmission lines using a FIB tool. Measured results are again compared to those of corresponding simulations and an equivalent circuit model. The milling technique is then applied to nano-fabricated, electro-statically tuned RF switches. The design and testing of these devices is discussed.

The final chapter is focused on the design of a novel RF telemetry system using low power RF MEMS or solid state switches, as well as micromachined patch antennas to modulate and redirect a 10 GHz CW signal. A discussion on BASK communication techniques and link budget simulations is included. Finally, a transceiver design for long range communication is presented and measured results of the link are shown.
Chapter 2

An Introduction to RF MEMS Switches

2.1 Introduction

In recent years, the utilization of MEMS technology in radio and microwave frequency (RF/Microwave) and wireless engineering applications has seen rapid growth and development. MEMS fabricated switches are an example of devices in this innovative field and have already shown superior electrical performances (isolation, insertion loss, DC power consumption, and intermodulation products) to solid state p-i-n and field effect transistor (FET) switches at high frequencies [16]. Due to these qualities, as well as their small size and their manufacturability using well characterized semiconductor processing techniques, RF MEMS switches have the potential to be a viable replacement to their solid state switch counterparts. Examples of components where the tuning capability of RF MEMS switches plays an intrinsic role are tunable elements such as resonators [17], capacitors [18], inductors [19], transmission lines [20] and antennas [21].

In this chapter, RF MEMS switching devices with various topologies and actuation mechanisms will be introduced. An overview of electro-statically actuated MEMS structure operation will also be given. Finally, a review of important MEMS switching parameters followed by summary of the current state of electrostatic RF
MEMS switching technology including examples of fabricated and tested switches will be presented.

2.2 RF MEMS Switch Technology

RF switches are probably among the most widely researched and fabricated MEMS devices to date. As already mentioned, this can be attributed to the fact that RF MEMS switches have the potential to significantly outperform their solid state counterparts at RF and microwave frequencies in such characteristics as insertion loss, isolation, linearity, and power consumption. But there are still areas where solid state switches currently outperform MEMS switches. These include switching speed, power handling, high drive voltage, reliability, and packaging. Table 2.1 displays a general comparison between MEMS, P-I-N diode, and field effect transistor (FET) switches [22].

| Table 2.1. Performance comparison between FET, P-I-N Diodes, and RF MEMS switches [22]. |
|-------------------------------------------------|----------------|----------------|
| RF MEMS                                        | PIN            | FET            |
| Actuation Voltage (V)                          | 20 - 80        | 3 - 5          | 3 – 5       |
| Power Consumption (mW)                         | 0.05 - 0.1     | 5 - 100        | 0.05 - 0.1  |
| Switching Time                                 | 1 - 300 µs     | 5 - 100 ns     | 1 - 100 ns  |
| Isolation (1-10 GHZ)                           | Very High      | High           | Medium      |
| Isolation (10-40 GHZ)                          | Very High      | Medium         | Low         |
| Loss (1-40 GHZ) (dB)                           | 0.05 - 0.2     | 0.3 – 1.2      | 0.4 – 2.5   |
| Power Handling (W)                             | <10            | <10            | <10         |
Many different kinds of RF MEMS switch designs have already been realized and demonstrated and there are probably as many different types of MEMS switches as there are fabrication techniques and materials available. What differentiates RF MEMS switches from solid state switches is the utilization of a movable mechanical structure that can be actuated and is used to either block or transmit an RF signal. The different types of MEMS switches can generally be categorized with respect to the following properties: actuation mechanism, actuation structure movement, contact type, circuit configuration, and transmission line topology. Actuation mechanism refers to how the movement of the mechanical actuation structure is obtained while the structure movement itself refers to the actual direction of displacement. Contact type describes how RF energy is transferred when the switch is actuated. Whether the actuation structure is placed in series or shunt configuration and in what type of transmission line topology it is implemented completes the list of possible switch categories.

A physical force is required to move a mechanical structure, usually either fixed-fixed membranes or cantilevers, in MEMS switching devices. This force can be obtained by either applying an electrostatic or electromagnetic field, by introducing thermal changes to the device, or by using a piezoelectric material in the movable structure. Figure 2.1 describes the actuation mechanism of these types of switches in more detail. Electrostatic and thermal actuated devices tend to consume more power since it is necessary to use a current in the actuated states. Thermally actuated switches generally also have a longer switching time since this is dependent on the time it takes to heat a resistive metal. Electrostatic and piezoelectric actuated switches both share the trait of virtually zero power consumption but electrostatic actuation seems to be the most
prevalent mechanism used today due to smaller size, manufacturability using inexpensive material and well characterized IC processing techniques. Electrostatic actuation will be described in greater detail in the following chapter. Table 2.2 shows a comparison of important features of these different types of actuation mechanisms.

Figure 2.1. Actuation mechanism for RF MEMS switches: (a) Electrostatic – an electrostatic force is created by applying a voltage potential between the upper and lower conductors. This causes the upper conductor to move towards the lower conductor. (b) Electromagnetic – A current in the electric coil around one of arm of a magnetic yoke produces a magnetic flux in the yoke. This results in a magnetic force in the gap of the yoke arms and causes the conductors to be pulled together. (c) Piezoelectric – A voltage applied to the top and bottom of a piezoelectric material causes an expansion of the material due to the induced electric filed. Since the piezoelectric is attached to the upper electrode, a bending moment of the upper conductor towards the lower one is produced. (d) Electrothermal – A current passed through a resistive material on top of the upper conductor causes the resistive layer to expand due to thermal heating. This induces a bending moment in the upper electrode.
Table 2.2. Relative comparisons of MEMS switch actuation mechanism. (Actuation voltage and switching time relative to IC switches.)

<table>
<thead>
<tr>
<th>Actuation Mechanism</th>
<th>Actuation Voltage</th>
<th>Current</th>
<th>Power</th>
<th>Switching Time</th>
<th>Size</th>
<th>Fabrication Complexity</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electrostatic</td>
<td>medium to high</td>
<td>zero</td>
<td>zero</td>
<td>medium to slow</td>
<td>small</td>
<td>low</td>
</tr>
<tr>
<td>Electromagnetic</td>
<td>low</td>
<td>medium to high</td>
<td>high</td>
<td>slow</td>
<td>medim</td>
<td>high</td>
</tr>
<tr>
<td>Piezoelectric</td>
<td>medium</td>
<td>zero</td>
<td>zero</td>
<td>medium</td>
<td>medium</td>
<td>high</td>
</tr>
<tr>
<td>Electrothermal</td>
<td>low</td>
<td>medium to high</td>
<td>medium to high</td>
<td>very slow</td>
<td>large</td>
<td>low</td>
</tr>
</tbody>
</table>

The direction of movement of the mechanical structure when actuated can either be vertical or horizontal as shown in Figure 2.2. Vertical movement has been the most widely used in today’s RF MEMS switches since it tends to result in devices with smaller footprints. This is due to the vertical orientation of the structure and smaller distance of movement that can be obtained when fabricating vertically-actuated switches. The gap between the static and movable portions of the switch can be as small as half a micrometer or even less, while the fabrication of lateral gaps is limited to the lowest resolution of the fabrication technique used and increases the overall lateral size of the switch.

Figure 2.2. Direction of movement of vertical (left) and horizontal (right) fabricated actuation structures.
When the movable structure is actuated, RF energy can either flow by means of metal-to-metal contact or through capacitive coupling as describe in Figure 2.3. Metal-to-metal switches are generally used for frequency applications from DC to 30 GHz since isolation in the un-actuated state is highest toward the DC range. These types of switches use RF-extrinsic actuation, meaning that the actuation area is separated from the RF contact area. Most applications of capacitive coupled switches are found in the 20 GHz and higher range, even though the operating frequency can be tuned lower with the use of inductive lines in series with the capacitive structure in the switch design. The isolation response of both types of switches is illustrated in Figure 2.4. Capacitive coupled switches can use RF-extrinsic or intrinsic actuation. In the RF-intrinsic case, the actuation and the coupling structure are the same. The disadvantage of metal-to-metal contact switches is their susceptibility to failure through pitting and hardening of the contact layer due to the impact forces that occur [23]. Also micro-welding of the contact area that can occur during high current applications will cause these types of switches to fail prematurely. Mechanical failure can also occur in capacitive coupled switches when electric charges get trapped in the dielectric layer during actuation, resulting in the actuation structure remaining in contact with the dielectric layer [24]. This phenomenon, known as stiction, is probably the main limiting factor in RF MEMS switch reliability.
Capacitive and metal-to-metal contact switches can either be implemented in a series or shunt circuit configuration, as shown in Figure 2.5. In the series configuration, the signal path is isolated from the output port when no bias voltage is applied to the actuation structure, i.e. the switch is in the off-state. During actuation, usually accomplished with capacitive coupled electrodes using high resistivity DC bias lines, the mechanical structure is pulled down and makes contact with the lower signal line. This
causes the signal to flow through the switch in the on-state. In shunt configuration, the signal is transferred with low insertion loss when the actuation structure is un-actuated. Capacitive coupling of the signal to ground is achieved when the switch is actuated, i.e. the signal is shorted and the switch is in the off-state. The coupling is provided either through separate high resistivity DC bias lines or through the capacitive structure itself. In the latter case, a bias network needs to be designed to isolate the DC voltage from the RF signal.

Finally, capacitive and metal-to-metal contact switches, whether in series or shunt configuration can be designed using either microstrip or coplanar waveguide (CPW) transmission line topology. (Waveguide switches have also been demonstrated using MEMS devices [25] but are mainly applied to systems using high power signals, such as
At micro- and millimeter wave frequencies, shunt switches in a CPW design are most often used due to their lower dispersion characteristics at high frequencies and easier integration with other lumped elements. Also, via holes through the substrate are not required since the ground planes are planar with the center conductor. Due to the latter, it is much easier to implement a shunt RF MEMS in a CPW line. For shunted microstrip switches, either via holes would need to be fabricated or large butterfly stubs have to be added to the design to provide the RF short in the off-state.

2.3 MEMS Switch Parameters

RF MEMS switch have generally been designed with the optimization of the following parameters in mind: insertion loss, isolation, actuation voltage, switching speed, and reliability. These are described in more detail below.

*Insertion Loss* is simply the loss in signal power of the incident signal through the RF coupling mechanism, either through absorption, reflection or both, when the switch is in the on-state. In series DC contact switches, metal-to-metal contact resistance will be the main contributor to signal power reduction due to heat dissipation at the contact area. In shunt capacitively coupled switches, leakage of the RF signal through the capacitive air gap will reduce the strength of the transmitted signal. And in both cases, signal reflections due to impedance mismatches between the characteristic impedance of the transmission line and switch circuit will also contribute to a lower output power level. Insertion loss is generally stated in literature as a ratio of the transmitted power to incident power in decibels, or the
magnitude of S21 in S-parameter terminology*. It is desirable to design MEMS switches with less than 1 dB of insertion loss over the operating bandwidth. 

*Isolation* describes how well the incident signal is prevented from being transmitted to the output in the off-state. In the series DC contact case, capacitive coupling through the air gap between the electrodes can cause undesired signal leakage. In the shunt case, high contact resistance, for extrinsic DC contact switches, or low down state capacitance, for capacitive switch devices, will impede the incident signal from being shunted completely to ground and therefore increase signal leakage. Just as in the case for insertion loss, isolation is usually given as the magnitude of the power ratio of the input to output signal levels, in decibels. An isolation greater than 20 dB is considered adequate in most switching applications. Figure 2.6 illustrates how isolation and insertion loss results are generally shown in literature.

*S-parameters, or scattering parameters, define a relationship between the amplitudes of incident and reflected waves into an n-port network. These amplitude waves are defined as:

$$a_n = V_n^+/(Z_{0n})^{1/2}$$ and $$b_n = V_n^-/(Z_{0n})^{1/2},$$

where $$a_n$$ and $$b_n$$ are the incident and reflected waves at the $$n^{th}$$ port, respectively. $$V_n^+$$ represents the incident voltage amplitude and $$V_n^-$$ the reflected voltage amplitude at the $$n^{th}$$ port. The characteristic impedance of the $$n^{th}$$ port is represented by $$Z_{0n}$$. The complete set of amplitude waves of a n-port network can then be implemented into S-parameter matrix with the following relationship:

$$[b] = [S][a],$$

where the $$m^{th}$$ and $$n^{th}$$ elements of the matrix are defined as $$S_{mn} = b_m/a_n$$. For a two-port device, $$S_{11}$$ and $$S_{22}$$ represent the ratios of the incident to reflected waves at ports 1 and 2, respectively, with the opposite port terminated with a matched load. $$S_{21}$$ defines the ratio of incident wave at port 1 to the reflected wave at port 2 and $$S_{12}$$ the ratio of incident wave at port 2 to the reflected wave at port 1. $$S_{11}$$ and $$S_{22}$$ are also referred as reflection coefficients while $$S_{21}$$ and $$S_{12}$$ are known as transmission coefficients.
Figure 2.6. Example of a $S_{21}$ plot showing the response of an RF MEMS switch in the on and off states. The isolation of the switch is shown on the left axis while the insertion loss can be seen on the right axis, although both parameters are stated as positive values.

*Actuation Voltage* is the amount of voltage required to fully actuate a MEMS structure. In electrostatic MEMS switches, the actuation voltage tends to be in the 10s of volts range due to the relatively large air gap required in the switch design. Reducing the gap would translate to higher undesired signal coupling, reducing the quality of the switch’s RF performance. New schemes of MEMS biasing design are being investigated to reduce the actuation voltage to less than 5 V to make the technology more compatible with IC circuitry.

*Switching Speed* can be described as the time required for MEMS structure to switch from un-actuated to actuated position, and vice-versa, when a corresponding bias condition is applied. Although there still are inconstancies in literature on the exact definition of this parameter, the majority of time it is defined as the time it takes for the RF signal to rise from 10% to 90% or fall from 90% to 10% of the highest input level when the switch is turned actuated or un-actuated, respectively; this is also referred as the rise and fall time of the switch. The on- and off-time of MEMS
switches will also be given on occasion. This refers to time it takes the signal to rise to 90% of the highest signal value when the control voltage is turned on or back to 10% of the signal value when the control voltage is turned off, respectively. These switching speed parameters are illustrated in Figure 2.6. As already mentioned, the switching speeds of MEMS switches are still confined in the micro second range and are therefore not as suitable as solid state switches in applications where high switching speed are required.

![Figure 2.6. Illustration of the switching speed parameters.](image)

Reliability translates directly into the usable lifetime of a MEMS switch. The limiting factors are usually fatigue of the mechanical actuation structures, dielectric charging of the capacitive area, pitting and micro-welding at metal-to-metal contact points as well as stiction due environmental effects. Many switches in literature have already been presented with lifetimes of up to billions of switching cycles.
Other parameters such as intercept point, power handling, and bandwidth are also important but are more application specific.

2.4 Electrostatic RF MEMS Switches

2.4.1 Electrostatic MEMS Actuation

Electrostatic controlled RF MEMS switches seem to show the most promise for usability and integration in microwave systems. The actuation mechanism of an electrostatically actuated CPW MEMS switch can be observed in Figure 2.8. Whether the switch is in series or shunt configuration, it always consists of a two electrodes; one fixed to the substrate, the other movable but fixed either at one or at both ends and separated by an air gap. A dielectric layer is usually coated on top of the lower electrode to provide DC isolation during actuation.

![Figure 2.8 Parallel plate capacitor model of MEMS switch actuation.](image)

By applying a voltage between the top plate and the lower metal, a deflection of the beam towards the bottom plate can be produced due to the electrostatic force between the two metal plates. The amount of deflection is a function of the voltage applied and
this relationship can be modeled by representing the beam actuation as a parallel capacitor. The capacitance of a parallel plate capacitor is given by:

\[ C = \frac{\varepsilon_r \varepsilon_0 A}{d}, \]  

(2.1)

where \( \varepsilon_r \) is the dielectric constant of the dielectric between the plates, \( \varepsilon_0 \) the electric permittivity of free space, \( A \) the overlapping area of the plates, and \( d \) the separation distance between the plates. If a voltage were to be applied between the top and bottom plates, a force would have to be applied to the top plate to keep it from moving towards the bottom plate. This electrostatic force is given by:

\[ F = \frac{\varepsilon \cdot A \cdot V^2}{2 \cdot d^2}, \]  

(2.2)

where \( V \) is the applied voltage between the two plates. To convert (2.2) to an approximate beam deflection versus voltage relationship, the structural and material properties of the electrode have to be taken into account. These properties can be represented analytically by the spring constant of the movable electrode that is either fixed at one or both ends. Since the restoring force is given by \( F=kx \), this force can be represented as a function of the distance the beam travels toward the lower electrode when the bias is increased. This yields:

\[ F = k(d_0-d), \]  

(2.3)

where \( d_0 \) is the beam height with zero bias and \( d \) the distance traveled. When the electrostatic (2.2) and restoring (2.3) forces are equated, the following relationship for the voltage as a function beam height \( x \) is obtained:

\[ V = \sqrt{\frac{2kd^2(d_0-d)}{\varepsilon_0 A}} \]  

(2.4)
Once the bias voltage is applied to initiate the deflection of the beam towards the lower plate, the bias can be increased to reduce the distance between the plates. It has been shown that a threshold voltage exists, at which point the beam actuation overcomes the natural restoring spring force of the cantilever material and starts to deflect toward the lower plate until the gap is completely closed. This spontaneous deflection occurs due to the continuous increase in the concentration of the electrostatic force at the beam plates as it bends downwards. The threshold voltage, also known as pull-down voltage, can be derived from (2.4) by taking its derivative with respect to x and setting it equal to zero, which results in:

\[ d = \frac{2}{3} d_0 \]  

Substituting (2.5) into (2.4) gives the pull-down voltage as:

\[ V_{\text{close}} = \sqrt[3]{\frac{8k\varepsilon_0}{27\varepsilon A}} \]  

After the bias is decreased, the deflected beam will not immediately reflect back to the original position and therefore will release at a much lower voltage than the pull-down voltage. This is due to a hysteretic effect caused by the very high capacitance that occurs as soon as beam moves slightly from the dielectric layer. Since this force is much greater than the force required to initially pull down the beam, a lower voltage is required to hold the beam in the down-state. The voltage required to release an actuated beam was calculated by Zackry, et. al, [26] and given as:

\[ V_{\text{open}} = (d_0 - d) \sqrt[3]{\frac{2kd}{3\varepsilon A}} \]
2.4.2 MEMS Switch Dynamics

The switching time, i.e. the time to fully collapse the moveable plate after the bias is applied, can be derived from a mechanical beam model of a MEMS switch. This model is shown in Figure 2.9.

![Mechanical MEMS beam model](image)

Figure 2.9. Mechanical MEMS beam model. For cantilever beams, $\frac{1}{2} k$ will be used.

The dynamic response for this model is given by the force equation:

$$m \frac{d^2 x}{dt^2} + b \frac{dx}{dt} + kx + k_s x^3 = F$$

(2.8)

where $m$ is the mass of the movable structure, $b$ the dampening coefficient, and $F$ the external applied force to the system. $k_s$ is a spring constant component due to the stretching effect of fixed-fixed beams that can be neglected for cantilever structures. If damping is assumed to be very small and the stretching effect of the actuated beam is neglected, the switching time can be derived by setting (2.8) equal to (2.2), with $d = d_0$. This results in an approximate expression for the switching time as
where $\omega_0$ is the mechanical resonant frequency of the actuation beam and $V_A$ the applied bias voltage. From the (2.9), it can be seen that the switching time can be improved with applied voltages that are higher than the pull-down voltage of the actuation structure. This approximate expression for switching time should only be used as a rule of thumb for switch design since it does not take into account dampening and spring elasticity of fixed-fixed beams. More precise equations can be found in the literature [27].

### 2.3.3 MEMS Switch Failure Mechanisms

RF MEMS switches have been found to be extremely susceptible to premature failure that is usually caused by the movable membrane not releasing properly after the actuation voltage is removed. This can be due to changes in the switch material as well as environmental effects. This type of switch failure is known as stiction and its causes have been studied extensively [28].

Dielectric charging has been found to be a major cause of stiction for MEMS switches with capacitive actuation structures. This is due to charge injection into the dielectric while the switch is held in the down-state with an applied voltage. The injected charges become trapped within the dielectric and tend to hold the movable electrode in the down state if the electric potential of the trapped charges is larger than the restoring force of the beam. The release time can therefore be significantly delayed and is dependent on the dissipation time of the trapped charges.
Solutions to reduce switch failure due to charge trapping include the use of a dielectric material with lower trap densities, such as PEVCD silicon dioxide [24]. The use of a bipolar waveform for the actuation voltage can also be implemented. The period of the polarity change needs to be much smaller than the charging time of the dielectric and fast enough so the movable electrode will not react with the polarity reversals in the down state. Many designs also now use pull down electrodes that have been separated from the capacitive coupling structure [29]. These are usually placed symmetrically on both sides of the RF contact region.

With metal-to-metal contact switches, stiction of the metal layers is predominately caused by micro-welding between the contact areas of the metals. This can occur when the RF signal power is large enough to heat the metal to the point of fusing the two contact metals together. This heating effect results due to the contact resistance between the two metals in the down-state. Due to significant surface roughness at the contact area, localized heating can occur and can therefore cause switch failure even at medium power levels.

Environmental conditions can also have a detrimental effect on switching performance. If the humidity if high enough, surface tension between the contact materials due to water vapor can be larger than the restring force of the switch membrane. Therefore, it is critical that the MEMS switch will be packaged in an inert environment that includes a hermetic seal at the package interface layer.
2.5 Current State of Technology

What follows are examples of capacitive MEMS switches that have already been developed and characterized at frequencies from 1 GHz up to 60 GHz. These RF MEMS switches have been generally designed with the optimization of the parameters described in the previous section in mind.

At frequencies below 30 GHz, RF MEMS switches are usually of the metal-to-metal contact type. The advantage of this type of configuration is the open circuit state that results when the switch is in the up-state (un-actuated) for the series configuration. This is equivalent to placing a low value capacitor in the signal path, providing high isolation from DC to microwave frequencies. By having separate pull-down electrodes, as in the extrinsic case, metal-to-metal contact can be achieved when the movable structure is actuated. This provides a conductive signal path where the loss is only limited by the contact resistance of the conductors. If placed in shunt configuration, DC-contact switching can also be achieved by again separating the actuation electrodes from the signal coupling area. In this case, the switch is in the on-state with no actuation and off with actuation. Table 2.3 summarizes the important characteristics of examples for these types of switches.
Table 2.3. DC-contact MEMS switches.

<table>
<thead>
<tr>
<th>Author</th>
<th>Type</th>
<th>Actuation Structure</th>
<th>Frequency of Operation&lt;sup&gt;1&lt;/sup&gt;</th>
<th>Insertion Loss</th>
<th>Isolation&lt;sup&gt;2&lt;/sup&gt;</th>
<th>Actuation Voltage</th>
</tr>
</thead>
<tbody>
<tr>
<td>Muldavin [30]</td>
<td>Series</td>
<td>Bridge</td>
<td>DC–30 GHz</td>
<td>&lt;0.1 dB</td>
<td>30 dB</td>
<td>18-22 V</td>
</tr>
<tr>
<td>Tan [31]</td>
<td>Shunt</td>
<td>Bridge</td>
<td>DC–18 GHz</td>
<td>0.1-0.15 dB</td>
<td>25 dB</td>
<td>45-55 V</td>
</tr>
<tr>
<td>Schauwecker [32]</td>
<td>Series</td>
<td>Cantilever</td>
<td>DC–7 GHz</td>
<td>&lt;0.28 dB</td>
<td>17 dB</td>
<td>17 V</td>
</tr>
<tr>
<td>Oberhammer [33]</td>
<td>Series</td>
<td>Cantilever</td>
<td>DC–20 GHz</td>
<td>2 dB</td>
<td>34 dB</td>
<td>20 V</td>
</tr>
<tr>
<td>Sloan [34]</td>
<td>Series</td>
<td>Cantilever</td>
<td>DC–25 GHz</td>
<td>&lt;0.2 dB</td>
<td>27 dB</td>
<td>10-12 V</td>
</tr>
<tr>
<td>Mihailovich [35]</td>
<td>Series</td>
<td>Cantilever</td>
<td>DC–90 GHz</td>
<td>&lt;0.1 dB</td>
<td>40 dB</td>
<td>60 V</td>
</tr>
<tr>
<td>Shen [36]</td>
<td>Shunt</td>
<td>Bridge</td>
<td>DC–40 GHz</td>
<td>&lt;0.1 dB</td>
<td>25 dB</td>
<td>9-16 V</td>
</tr>
<tr>
<td>Hyman [37]</td>
<td>Series</td>
<td>Cantilever</td>
<td>DC–40 GHz</td>
<td>&lt;0.15 dB</td>
<td>33 dB</td>
<td>30-40 V</td>
</tr>
</tbody>
</table>

<sup>1</sup> Upper frequency at 25 dB isolation
<sup>2</sup> Isolation measured at 10 GHz

MEMS switches with capacitive coupling in CPW configuration have been generally used for circuits operating at frequencies greater than 10 GHz. When the actuated structure is placed in shunt configuration, the signal is shorted to ground by capacitive coupling in the down-state. But the capacitive section of the switch can also be placed in series, where the capacitive coupling transmits the signal in the down-state. A disadvantage of the capacitive switches is that the operational frequency at the upper end is limited by a parasitic inductance that occurs due to the short transmission line length of the actuation structure. However, this effect can also be utilized to tune the resulting resonance to a desired frequency of operation by increasing the parasitic inductance. High impedance and meander line structures in the beam geometry have been used this implement this effect. Table 2.3 list up to date RF MEMS switches using capacitive coupling with their stated performance.
Table 2.3. Capacitive coupled MEMS switches.

<table>
<thead>
<tr>
<th>Author</th>
<th>Frequency of Operation</th>
<th>Insertion Loss</th>
<th>Isolation 1</th>
<th>Actuation Voltage</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ramadoss [38]</td>
<td>12-30 GHz</td>
<td>0.3-0.4 dB</td>
<td>50 dB</td>
<td>110 V</td>
</tr>
<tr>
<td>Rizk [39]</td>
<td>75-110 GHz</td>
<td>&lt;0.25 dB</td>
<td>30 dB</td>
<td>55 V</td>
</tr>
<tr>
<td>Ulm [40]</td>
<td>18-28 GHz</td>
<td>0.2 dB</td>
<td>34 dB</td>
<td>N/A</td>
</tr>
<tr>
<td>Goldsmith [41]</td>
<td>24-40 GHz</td>
<td>0.3-0.6 dB</td>
<td>40 dB</td>
<td>30-50 V</td>
</tr>
<tr>
<td>Park [42]</td>
<td>1-20 GHz</td>
<td>&lt;0.15 dB</td>
<td>44 dB</td>
<td>8 V</td>
</tr>
</tbody>
</table>

1) Isolation at resonant frequency

Switching time is also a very important parameter for RF MEMS devices. This is generally defined as the time it takes for the RF signal to rise from 10% to 90% or fall from 90% to 10% when the switch is turned on or off respectively due to the transition of the actuation structure. Switching speed is dependent on the spring constant of the actuation structure, the gap between electrodes, and the damping force seen during actuation, and has been shown to lay in the 10 to 300 µs range for most switches presented today, even though MEMS switches with transition times between 1 and 10 µs have also been demonstrated. Examples are the Raytheon MEMS capacitive shunt switch [43], with an actuation time of 3 µs and 5 µs release time, the Rockwell Scientific DC-contact series switch with switching times of 8-10 µs [44], the Radant MEMS inline DC contact series switch with 2-3 µs switching speeds [45], and the Motorola DC contact MEMS series switch which has a reported transition time of 2-4 µs [46].
2.6 Conclusion

MEMS switching technology for RF and microwave applications has been introduced. A description of various actuation mechanisms, which include electrostatic, electromagnetic, electrothermal, and piezoelectric, as well as the direction of movement, was given. A comparison of the benefits and disadvantages between these methods of actuation was made. Also, a description of how an RF signal is transferred when a MEMS switch is actuated, either by DC direct contact or through RF capacitive coupling, was shown. These different types of switches can then be implemented in various transmission line topologies, with microstrip and CPW being the most common. Finally, examples of the current state of technology of fabricated and tested MEMS switches at frequencies above 1 GHz were presented.
Chapter 3

CPW MEMS Shunt and Series Switches

3.1 Introduction

Part of this research includes the design of a switching device for a telemetry system that is able to modulate a continuous wave signal. This switch needs to possess good RF characteristics and mechanical performance, as well as low power consumption. And in the previous chapter, it was shown that MEMS switches using electrostatic actuation with coplanar waveguide topology have been demonstrated to be best suitable for switching applications requiring high RF and mechanical performance when compared to MEMS switches with different actuation mechanisms. This is due to their relatively simpler design using well known fabrication techniques and low cost materials. Implementing the switch as a CPW component also gives it the advantage of easier integration with other microwave circuitry. For this reason, various CPW electrostatic MEMS switch designs, in shunt and series configurations, were investigated and reported upon in this chapter.

Following a description of RF performance characteristics, the MEMS switch design methodology utilized in this chapter will be described. This discussion includes the design of the CPW transmission lines, the capacitive RF coupling component and the inductive line sections of the switches for X-band operation. The various switch design
will be introduced, and include shunt switches and series switches with capacitive coupling as well as a shunt contact switch using separate actuation structures. Initial design iterations will be shown that include a dual cantilever shunt switch with a large actuation pad at the tip of the suspended beam structure and series switches, and shunt switches with double beams, where a thin metal pad was fabricated directly on top of the dielectric to improve the RF coupling when the beam is in the down state. Also shunt switches with single beams and with various geometries of thin meander line sections will be shown. Finally, a shunt DC contact switch will be introduced.

After the switches were fabricated and measured, isolations between 30 and 45 dB were typically measured at their respective resonant frequencies in the off-state. The series capacitive switches showed isolation of about 30 dB measured at 10 GHz. Insertion losses in the on state were measured to be 0.3 to 0.5 dB for most switches. The results were also compared to simulated data. Equivalent circuit models for the MEMS shunt switches were derived and circuit component values were extracted from simulated and measured data. Circuit models for both the on-state and off-state cases are shown. Switching speed measurements were performed and results as fast as 40 µs were achieved. A metal-to-metal contact shunt switch with an isolation of 20 dB and insertion loss of 0.8 dB at 10 GHz was also measured.

Packaging techniques using photo-imagable resist material for the MEMS switches were also studied. SU-8 was used as a sealant ring and bonded with a quartz lid to the switch substrate with a flip-chip-bonder. RF performance of the switch was again measured as well as the switching speed.
3.2 RF Characteristics

To produce a high performance MEMS switch at RF and microwave frequencies, it is critical to choose the proper dimensions and material of the switching circuit components. This will ensure that the lowest possible insertion loss and highest possible isolation will be achieved. How isolation and insertion loss are defined is dependent on the type of RF switch, i.e. whether it is a shunt switch or series switch. In the shunt case, isolation occurs when the switch is in the down-state and insertion loss in the up-state, since the signal will be shunted to ground during actuation. The opposite occurs when a series switch design is used. Here, the signal is isolated from the output due to an air gap in the un-actuated switch. Insertion loss applies when the switch is in the down state after actuation. However in both cases, transmission line mismatches are also a contributor to the insertion loss of the switch.

A MEMS shunt switch with capacitive RF coupling will have two distinct capacitive states during switch operation; an up state capacitance when the switch is un-actuated and a down state capacitance in the actuated state. The two capacitance values can be calculated using the parallel-plate model. These are Eq. (2.1) in the down state and

\[
C_{up} = \frac{\varepsilon_0 A}{d + \frac{t_d}{\varepsilon_r}}
\]

in the up state (\(t_d\) represent the dielectric thickness). During the switch design, it is desired to have an up state capacitance as low as possible and generally a down state capacitance that is as high as possible. A low up state capacitance ensures that the least amount of the input signal is leaked to ground through the air gap, therefore minimizing
the contribution to the insertion loss. With a higher down state capacitance, the switch can be operated at lower frequencies and with larger bandwidths. The ratio of the two capacitance value is generally used as a figure of merit for RF shunt switches.

The quality of the isolation of capacitive shunt switches is also dependent on the series resistance of the switch’s transmission lines. The main contributors to this parameter are the series resistances of the MEMS bridge and the CPW transmission line. The transmission line resistance can be calculated from the measured line loss, $\alpha$, in

$$R_{rl} = \alpha^2 Z_0 l,$$

where $l$ is the transmission line length and $Z_0$ represents the characteristic impedance of the line. The series resistance due to the MEMS bridge is more accurately derived from extracted circuit models of the switch in the down state. As in the up state case, series resistance also contributes to the insertion loss of the shunt switch in the down state. This adds to the effects of any reflected signal power due to mismatches between the CPW transmission lines and the switch structures.

A MEMS series switch with capacitive RF coupling operates similarly to the capacitive shunt switch except that the insertion loss and isolation characteristics are reversed in the up and down states. These are usually switches with cantilevers where the isolation occurs due to the air gap between the tip of the cantilever and the dielectric surface of the CPW’s lower center conductor. In this state, the larger the air gap, the higher the isolation. When the switch is actuated, RF coupling is achieved through the dielectric and the insertion loss is dependent on the series resistance of the transmission line and cantilever structure, in addition to any reflection losses.
Generally MEMS series switches use metal-to-metal contact pads to transmit the input signal across the gap when actuated. Isolation in this case is still dependent on the gap height in the up state. In the down state, the insertion loss again is dependent on the CPW and the movable membrane’s series resistance values. But the contact resistance between the two electrodes that results during actuation will also have a significant effect on the insertion loss of the switch. It has been shown this resistance is dependent on the quality of the metal (such as surface roughness, contamination, and surface oxidation), surface area and the applied actuation voltage [47]. Using inert metals such as gold with large areas will help reduce the contribution to the contact resistance. Higher actuation voltages, which result in a larger contact force, can also decrease the contact resistance.

3.3 Switch Design for X-Band Operation

The operational frequency range of the switches studied in this investigation was 8 to 12 GHz. This frequency allocation is also known as the X-Band. However, measurements and simulation were performed over a wider frequency band, generally from 2 to 16 GHz. The switches were designed using high resistivity (\( \rho > 3000 \, \Omega \)) silicon as the substrate with a thickness of about 400 \( \mu \)m. The CPW transmission line component of the MEMS switch was designed as a CPW transmission line with grounded substrate having a 50 \( \Omega \) characteristic impedance, which corresponds to a center conductor width (\( w \)) of 100 \( \mu \)m and a ground-to-center spacing (\( s \)) of 50 \( \mu \)m as illustrated in Figure 3.1. This was calculated using Agilent’s ADS Linecalc transmission line calculation software.
For the shunt and series switch designs with capacitive coupling, the appropriate beam and dielectric dimensions were calculated to ensure optimum operation in the X-Band. RF coupling through the capacitive region of the actuated switch increases with frequency, making capacitive MEMS switch more suitable for application at frequencies above 10 GHz. The capacitive coupling will not increase indefinitely with frequency due to inductive parasitic effects caused by the MEMS beam structure. At some point, a resonance will occur where the capacitance and parasitic inductance will cancel each other, and this behavior can be modeled as a series LCR resonator (see Figure 3.2). By properly choosing the dimensions of the beam, the RF coupling area and dielectric material and thickness, this resonance can be tuned to a desired frequency.
The capacitance of the switch in the down state is a function of the dielectric constant, area and thickness of the dielectric material as was seen in Figure 3.4. Eq. 2.1 can be used to calculate the down state capacitance. Small fringing capacitances also have to be taken into account to calculate the actual coupling capacitance. To calculate the inductive effects of the beam dimension, a simple formula for microstrip line inductors (3.3) was used as a starting point in the switch design [9]. The parameters $W$, $l$, and $t$ are defined in Figure 3.3 with units in µm. The original derivation of this formula was designed for high impedance microstrip lines but by treating the substrate height ($h$) as the conductor to ground spacing, a rough approximation of the line inductance can be made for CPW transmission lines.

\[
L = 2 \times 10^{-4} \left[ \ln \left( \frac{L}{W + t} \right) + 1.193 + 0.2235 \frac{W + t}{l} \right] \cdot K_g \text{ (nH)}
\]

\[
K_g = 0.57 - 0.145 \ln \frac{W}{h}, \quad \frac{W}{h} > 0.05
\]

(3.3)
3.4 Capacitive MEMS Switches

3.4.1 Design and Fabrication Process

Four types of capacitive MEMS switches were designed for X-Band operation: a dual shunt cantilever switch; a series cantilever switch; a shunt switch with dual MEMS beams and inductive shunt line sections; and a shunt switch with a single bridge that also uses inductive line sections. Figure 3.4 shows the layout and description of each switch design.
The overall length of the cantilevers in the shunt and series switch design was 280 μm with a width of 30 μm. The plates at the beam tips of the shunt cantilevers measured 140 x 100 μm. This was also the dimension of the fixed metal plate on top of the dielectric in the series cantilever. For the shunt beam switch designs, the top plate had a width of 100 μm and a length of 200 μm. The fixed-fixed bridges of the shunt switches were 415 μm long and 60 μm wide. The inductive line section in the shunt switch designs were 15 x 100 μm in size.
The MEMS switch designs studied in this chapter were fabricated on 400 μm thick high resistivity silicon (ρ > 3000 Ω). For the switch with a shunt cantilever, first the CPW metal lines were fabricated using 10,000 Å thick thermally evaporated silver and gold, with a thin chromium adhesion layer underneath, and a metal lift-off process. Then the 2800 Å thick dielectric layer (silicon nitride) was deposited over the center conductor by RF sputtering and lift-off processing. Photoresist (SP 1818) was initially chosen as the sacrificial material and was spun on top of the sample to obtain a thickness of 1.5 to 1.8 μm. After the base pads of the cantilevers were patterned on the resist, which was then removed by developing the exposed resist material, a thin (400 Å) Au/Cr seed layer was thermally deposited on top of the sample followed by an application of spun-on photo resist; this was required for the subsequent electroplating process of the MEMS cantilevers. After the cantilever geometry was patterned and the resist developed to provide a plating mask for the cantilevers, electroplating of the seed layer was performed to obtain 1 μm thick beams. After the plating mask was removed, the cantilevers were protected with photoresist, the seed layer was removed using Cr and Au enchants, and the sacrificial layer was removed using MicroChem 1165 photoresist stripper and rinsed in a series of methanol baths. The final release process involved the use of a critical point dryer (CPD) that entails heating the samples above the critical point of the liquid it is immersed in. This process is required to eliminate stiction of the MEMS cantilevers that can be caused by evaporation of the methanol. During the CPD process, the methanol is replaced with liquid carbon dioxide (LCO₂), since LCO₂’s critical point occurs at a much lower temperature and pressure than that of methanol. The fabrication steps are shown in Figure 3.5.
Figure 3.5. Fabrication process of MEMS switches with shunt cantilevers. Only one of the cantilevers is shown for clarity.

Testing of the devices showed that there were reliability issues with this fabrication process resulting in devices that either leaked DC current when a bias was applied, resulted in MEMS structures stuck in the down position after actuation, or showed poor isolation in the down state when switched. It was observed that DC leakage occurred through the dielectric layer covering the edge of the lower metal. This can be explained by the fact that sputtering of the dielectric results in a thinner deposited layer over the metal sides than over the surface, due to the deposition properties of the sputtering process.
After inspection of the devices that had premature failure due to stiction, it was found that photoresist residue underneath the cantilevers and poor gold to chromium adhesion in the cantilever structure were the main causes of this type of switch failure. The photoresist residue resulted from incomplete removal of the sacrificial layer during the MEMS release process since Shipley photoresist was observed to be susceptible to hardening at high temperatures, and therefore can become resistant to resist strippers. Chromium adhesion problems were shown to be caused by undercutting during the seed layer etch step. Here, some of the chromium etchant seemed to leak in between the cantilever’s Cr/Au seed layer, causing the chromium to separate from the gold during actuation such that portions of the chromium layer would remain stuck on top of the dielectric. An image showing this type of stiction failure is shown in Figure 3.6.

Figure 3.6. SEM image showing separated chromium seed layer from underneath a gold plated shunt beam structure.

Tested switches that did not have stiction and DC leakage problems showed poor isolation in the down state. This behavior can be attributed to non-planar contact surfaces of the cantilever tips and is caused by metal stresses that occur during the fabrication
process. This residual stress will result in curved metal structures, especially for those with large surface areas as can be observed in Figure 3.7. Due to the air gap that occurs in between the cantilever and the dielectric in the down state, the effective dielectric constant is decreased. This lowers the capacitance and therefore the RF coupling at the design frequency.

Figure 3.7 Shunt cantilevers with non-planar beam tips.

To increase the reliability and performance of the MEMS switches, the following changes were made to the switch design and the fabrication process: PEVCD silicon nitride was used for the dielectric material; the center conductor area underneath the actuation structures was thinned; the chromium adhesion layer was replaced with titanium; PMMA was used as the sacrificial material; and the beam tip area was reduced while a thin fixed metal pad was placed on top of the dielectric with an area equal to that of the original beam tip.
The PECVD process was used since it gives a much better step coverage than line-of-sight deposition procedures. Combining this with the thinner metal (0.4 µm) underneath the contact area resulted in the elimination of DC leakage in the down state. Also, titanium proved to have much better adhesion to the gold surface metal and no separation of the two metals was observed during testing.

The decision to place a fixed metal pad on top of the dielectric was made to ensure maximum RF coupling though the dielectric in the down state. This configuration is equivalent to having a fixed metal-insulator-metal (MIM) capacitor in the lower electrode. With actuation, capacitive coupling level is now independent of the resulting shape of the MEMS beam; i.e. since there is now metal-to-metal contact with a MIM capacitor that has a fixed capacitance value, the effects of residual stress will not contribute to the effective capacitance value in the down state. The online draw-back to this change is that the isolation is now dependent on the quality of the down state contact resistance between the beam and MIM metal.

Since the center conductor underneath the actuation area is now fabricated in a separate process step from the rest of the CPW transmission line segments, one more processing change was made. This change consisted of fabricating the CPW lines at the same time as the electroplated MEMS structures. By making the CPW lines and the MEMS beams a continuous metal structure, problems with adhesion and contact resistance that could occur at the beam to CPW line interface were eliminated. The fabrication procedure is illustrated in Figure 3.8. Appendix I gives the detailed fabrication procedure.
3.4.2 Measured Results

Measurements were performed on a Wiltron 360 VNA with a Karl Suss probe station and 150 µm pitch G-S-G (ground-signal-ground) probes. S-parameter measurements were collected using Wincal software and the data was imported into Agilent’s ADS [10] microwave circuit simulator. DC biasing was accomplished through the RF probes using bias tees to isolate the DC from the RF signals.

Measured results of the shunt cantilever switch shown in Figure 3.9 can be seen in Figure 3.9 (isolation in the up state) and 3.10 (insertion loss in the down state). The beam
tip’s planarity issues associated with residual stresses is made evident in the isolation plot of the switch. It can be seen that the isolation increases with increasing voltage after actuation, starting at 10 V. A significant increase was seen at 25 V bias but still much lower than the simulated results. This result can be explained by the fact that the beam tip was most likely pulled flat at the higher voltage but still not flat enough to allow optimal capacitive coupling. Higher bias voltages resulted in a break down of the switch due to leaking DC current.

Figure 3.9. Comparison between measured and simulated $S_{21}$ data for the shunt cantilever MEMS switch in the down state. Measured results are shown with increasing bias.
Figure 3.10. Comparison between measured and simulated $S_{21}$ data for the shunt cantilever MEMS switch in the up state.

Results for the series capacitive switch with fixed metal pads shown in Figure 3.11 are illustrated Figure 3.12 and 3.13. Figure 3.12 shows the comparison between the measured and simulated S-parameter data in the down state (the actuation voltage was greater than 35V for these types of switches). Simulated and measured S-parameter data in the up state is shown in Figure 3.13. The high bias voltage can be attributed again to residual stresses in the actuation structures that caused the cantilever to bend slightly upwards after release. This stress also resulted in a higher isolation when compared to simulated date due to the larger air gap in the up state. A gap height of 3 µm was specified in the simulation. An insertion loss of about 0.5 dB at 10 GHz was measured in the down state and an isolation of 25 dB in the up state.
Figure 3.11. Series MEMS capacitive switch with fixed metal pad. The left input port was assigned as port 1 during the measurements, the right side as port 2.

Figure 3.12. Comparison between measured and simulated $S_{11}$, $S_{22}$ and $S_{21}$ data of the series capacitive MEMS switch in the down state. Simulated data is shown with markers and measured data with straight lines.
Figure 3.13. Comparison between measured and simulated $S_{11}$, $S_{22}$ and $S_{21}$ data of the series capacitive MEMS switch in the up state. Simulated data is shown with markers and measured data with straight lines.

It is also apparent from the two plots that the return loss is greater at port 1 than at port 2 due to bigger impedance mismatch of the thin metal beam at port 1 than at port 2.

A capacitive shunt switch with dual beam design (Figure 3.14) was measured and a comparison between these results and simulated data are shown in Figure 3.15 and 3.16. At 10 GHz, the isolation was greater than 45 dB with an insertion loss of less about 0.15 dB at the same frequency. Actuation of the MEMS bridge occurred at about 35 V.
Figure 3.14. MEMS capacitive shunt switch with dual fixed-fixed beams and inductive line section.

Figure 3.15. Comparison between measured and simulated $S_{11}$ and $S_{21}$ data of the shunt capacitive MEMS switch with double beams in the up state. Simulated data is shown with markers and measured data with straight lines.
Figure 3.16. Comparison between measured and simulated $S_{21}$ data of the shunt capacitive MEMS switch with double beams in the down state. Simulated data is shown with markers and measured data with straight lines.

For the single beam shunt switch as shown in Figure 3.17, measurements were performed for switches with straight and meandered shunt inductive line sections. Figure 3.18 and 3.19 show the frequency response of the switch with straight inductive line. Insertion losses of less than 0.3 dB were measured with isolation greater than 35 dB at the resonant frequency. For the single beam, 25 V of bias voltage was needed to actuate the device. A lower actuation voltage for the single beam switch, when compared to that of the double beam device, was achieved since the beam metal of the single beam design was about 1 µm, compared to 1.3 µm for the double beam design.
Figure 3.17. MEMS capacitive shunt switch with single fixed-fixed beam and inductive line section.

Figure 3.18. Comparison between measured and simulated $S_{21}$ data of the shunt capacitive MEMS switch with single beam in the up state. Simulated data is shown with markers and measured data with straight lines.
Figure 3.19. Comparison between measured and simulated $S_{21}$ data of the shunt capacitive MEMS switch with single beam in the down state. Simulated data is shown with markers and measured data with straight lines.

After comparing the performance of the tested MEMS switch designs, it was determined that the MEMS shunt switches with fixed top plates were more suitable for switching applications at the design frequency. The isolation of the switch with the series beam was about 15 and 25 dB lower than that of the single beam and double beam shunt switches, respectively. This can be explained by the fact that there is significant coupling between the air gap of the series cantilever design in the up state. The insertion loss was also higher when compared to those of the shunt switches due to the larger contact resistance in the down state. Also, a cantilever design will be more susceptible to planarity and residual stress issues as well as a higher chance of stiction in the down state.
The latter is due to the inherent lower spring constant of cantilevers, resulting in a lower restoring force, when compared to those of fixed-fixed beams.

A summary of the switching performance for the three types of fabricated and tested switches is given in Table 3.1. The shunt switch with large beam tips is not included due to the poor RF performance during testing.

Table 3.1. Comparison of the RF performance of the series and shunt MEMS switches at 10 GHz.

<table>
<thead>
<tr>
<th>MEMS Switch</th>
<th>Isolation (dB)</th>
<th>Insertion Loss (dB)</th>
<th>Actuation Voltage (V)</th>
<th>20 dB Bandwidth (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Series</td>
<td>0.5</td>
<td>25</td>
<td>35</td>
<td>6-16</td>
</tr>
<tr>
<td>Shunt with Double Beam</td>
<td>0.15</td>
<td>45</td>
<td>35</td>
<td>6.5-16</td>
</tr>
<tr>
<td>Shunt with Single Beam</td>
<td>0.25</td>
<td>35</td>
<td>25</td>
<td>6.5-16</td>
</tr>
</tbody>
</table>

3.4.3 Switch Model

A circuit model of the designed and tested capacitive MEMS switch was developed to extract the various equivalent circuit values, such as up- and down-state capacitances, series resistance, and series and shunt inductances from the measured results. \( C \) represents the up- and down-state capacitance values of the MEMS actuation structures. \( L_s \) and \( R_s \) are the inductance and resistance values, respectively, of the beam structure taken from the base to the coupling area. \( R \) models the contact resistance between beam and the top metal when the switches are in the down state. The series inductance and resistance values of the inductive line sections are represented by \( L_{sh} \) and \( R_{sh} \), respectively. Library models of 50 \( \Omega \) CPW transmission lines were also added to the ports to model the short CPW line sections at the inputs and outputs of the circuit layouts.
used during the measurements. These sections were also used to model any line portions of the fabricated devices not de-embedded during the calibration. The circuit model schematic is shown in Figure 3.20.

![Circuit Model Schematic](image)

Figure 3.20. Equivalent circuit model for capacitive shunt MEMS switches.

The model components were extracted by optimizing the circuit model’s S-parameter data against the measured S-parameter results of the single and double beam shunt switches. Modeling was performed in the up and down states. Figure 3.21 shows the comparison between the magnitude and phase of $S_{11}$ of the double beam shunt switch in the upstate. The $S_{21}$ magnitude and phase comparisons between modeled and measured results are shown in Figure 3.22.
Figure 3.21. Measured and modeled $S_{11}$ (magnitude and phase) for double beam shunt switch in the up state. Modeled data is shown with markers and measured data with straight lines.

Figure 3.22. Measured and modeled $S_{21}$ (magnitude and phase) for double beam shunt switch in the up state. Modeled data is shown with markers and measured data with straight lines.
Figure 3.23 and 3.24 shows the comparison between the measured and modeled $S_{11}$ and $S_{21}$ data, respectively, for the double shunt beam in the down state.

Figure 3.23. Measured and modeled $S_{11}$ (magnitude and phase) for double beam shunt switch in the down state. Modeled data is shown with markers and measured data with straight lines.

Figure 3.24. Measured and modeled $S_{21}$ (magnitude and phase) for double beam shunt switch in the down state. Modeled data is shown with markers and measured data with straight lines.
The modeled and simulated $S_{11}$ responses of the MEMS switch with single fixed-fixed beam in the up state is shown in Figure 3.25. The same comparison, but with $S_{21}$, is illustrated in Figure 3.26. Figures 3.27 and 3.28 repeat this comparison, but with the single beam shunt switch in the down state.

Figure 3.25. Measured and modeled $S_{11}$ (magnitude and phase) for single beam shunt switch in the up state. Modeled data is shown with markers and measured data with straight lines.
Figure 3.26. Measured and modeled $S_{21}$ (magnitude and phase) for single beam shunt switch in the up state. Modeled data is shown with markers and measured data with straight lines.

Figure 3.27. Measured and modeled $S_{11}$ (magnitude and phase) for double single shunt switch in the down state. Modeled data is shown with markers and measured data with straight lines.
Figure 3.28. Measured and modeled $S_{21}$ (magnitude and phase) for single beam shunt switch in the down state. Modeled data is shown with markers and measured data with straight lines.

The extracted equivalent circuit components of the two shunt switches in both the up state and down state are summarized in Table 3.2 and 3.3, respectively. It can be observed from these tables that the up and down state capacitances were very similar for both types of switches. This gives an up state to down state capacitance ratio of about 50. It can also be observed that the series resistance of the single beam switch in the down state is almost twice as high as that of the double beam. This was expected since the metal contact area between the MEMS beam and the fixed metal plate of the double beam switch is twice that of the single beam switch.

The capacitance in the down state can also be calculated using the parallel plate capacitor equation for comparison with the extracted results. Using (2.1) gives a calculated capacitance of about 4.1 pF. The higher extracted value is most likely due to...
some fringing capacitance not included in the theoretical calculation and fabrication errors that could have resulted in a larger capacitive area.

| Table 3.2. Equivalent circuit model components of capacitive shunt switches in the up state. |
|---|---|---|---|---|---|---|
| Switch Type | C (pF) | R (Ω) | Ls (nH) | Rs (Ω) | Lsh (nH) | Rsh (Ω) |
| Double Beam | 0.15 | 0 | 0.095 | 0.61 | 0.039 | 0 |
| Single Beam | 0.11 | 0 | 0.084 | 0.36 | 0.042 | 0 |

| Table 3.3. Equivalent circuit model components of capacitive shunt switches in the down state. |
|---|---|---|---|---|---|---|
| Switch Type | C (pF) | R (Ω) | Ls (nH) | Rs (Ω) | Lsh (nH) | Rsh (Ω) |
| Double Beam | 5.4 | 0.18 | 0.11 | 0.59 | 0.048 | 0 |
| Single Beam | 5.1 | 0.51 | 0.079 | 0.38 | 0.051 | 0 |

3.5 Switching Speed Measurements

Switching speed measurements were performed using the setup shown in the block diagram of Figure 3.29. To test the switching performance of the fabricate MEMS devices, a 10 GHz signal is injected into the input port of the switches. An arbitrary waveform generator, which is amplified to provide the required voltage to actuate the switch, was used to provide the modulating signal. A 35 V square wave with 50% duty cycle was used to drive the switch. The modulated 10 GHz output of the switch was then measured using a diode detector; a device that converts a detected RF signal to a corresponding DC voltage. Both the detected modulation waveform and the driving waveform are compared using an oscilloscope.
Figure 3.29. Block diagram of switching speed test setup.

Figure 3.30 and 3.31 show screen shots of the fastest rise and fall times of the double beam shunt switch, respectively. The quickest actuation speed was seen to be about 900 µs with a slightly faster release time of about 500 µs. It is believed that the relatively slow switching times were caused by the separate beam components of the dual beam structure not being able to actuate and release simultaneously, which could be due to a slight discrepancy in beam dimensions caused by a non-uniform fabrication processes.
Figure 3.30. Rise time of double beam shunt switch.

Figure 3.31. Fall time of double beam shunt switch.
The problem with the actuation and release time delay of the dual MEMS bridges was eliminated with use of the single beam shunt switch design. The rise and fall times decreased significantly to about 80 and 40 $\mu$s, respectively. The fall time of the single beam shunt switch is shown in Figure 3.32 and the rise time in Figure 3.33.

Figure 3.32. Rise time of single beam shunt switch.
It can be concluded that the shunt switch with the double beam design will provide higher isolation than the single beam switch due to the double beam switch’s lower contact resistance in the down state. This would make the double beam design more suitable for applications were very high isolation at the design frequency is required. However if faster switching speed is more important, the single beam design will provide better rise and fall times.

Figure 3.33. Fall time of single beam shunt switch.
3.6 MEMS Switch Packaging

To protect the fabricated switches from moisture and particle contamination, packaging of the MEMS devices in an inert environment with hermetic seals is required. The single beam shunt switches developed in this work were packaged using a quartz lid and Microchem’s SU-8 photoresist as a sealing ring. SU-8 has photo-imagable properties, a low reflow temperature (~90°C) and is resistant to most etchants and solvents.

To fabricate the lids, an array of SU-8 rings were patterned and developed on top of a 500 µm thick quartz wafer. The outside ring dimension measured 400 by 1500 µm with a thickness of 100 µm. The SU-8 was fabricated to a thickness of about 50 µm.

Before bonding to the MEMS devices, the individual lids were separated using a wafer dicing saw. Each piece was then bonded to the MEMS switch with a flip-chip bonder. Both the lid and sample holders were kept at a temperature of 150°C and a 10 N force on the lid was applied for 20 minutes to bond the two surfaces together. An outline of the fabrication process is shown in Figure 3.38 and a detailed description is given in Appendix I. A packaged MEMS switch with quartz lid is also shown in the Figure 3.34.
A packaged MEMS switch’s RF performance was measured before and after lid bonding. The packaged switch had an isolation of about 30 dB at 10 GHz in the down state. The $S_{21}$ plot is shown in Figure 3.35, for both the packaged and un-packaged MEMS switch in the up state. A $S_{21}$ graph showing a comparison of the MEMS witch with and without lid in the down state is shown in Figure 3.36. It can be observed from the plots that there is no significant change in the response of the switch after the quartz lid is bonded to the switch substrate.
Figure 3.35. Comparison of magnitude and phase of $S_{21}$ between MEMS switch with and without packaged lid, with the switch in the up state. Data for switch with lid is shown with solid round markers.

Figure 3.36. Comparison of magnitude and phase of $S_{21}$ between MEMS switch with and without packaged lid, with the switch in the down state. Data for switch with lid is shown with solid round markers.
3.7 Shunt Contact MEMS Switch

3.7.1 Design and Fabrication

A shunt MEMS switch with metal-to-metal contacts was developed that uses actuation structures that have been separated from the RF contact area. In this design, as shown in Figure 3.37, actuation occurs underneath the center of the beam. What differs in this design is that biasing is performed using high resistivity SiCr metals lines. The bias lines run underneath the ground plane to the middle of the CPW center conductor. The center conductor itself wraps around the SiCr line that lies underneath the shunt beam. To avoid DC shorting of the RF signal, air bridges were placed over the SiCr portions that cross both the center conductor and the ground planes.

![Figure 3.37. Shunt metal-to-metal contact MEMS switch with DC bias line.](image-url)
The motivation behind the development of this design was to decouple the DC actuation area from RF coupling in the down state and to increase the usable bandwidth of the switch. The bandwidth increases with this design since the isolation will not be dependent on the down state capacitance due to the metal-to-metal contact when actuated. Generally with metal-to-metal contact shunt beam switch designs, the actuation pads are placed to each side of the RF contact area of the center conductor. However with this design, a single actuation pad is used and directly placed underneath the center of the MEMS beam. This configuration will reduce the actuation voltage since the force needed to pull down the beam is the minimum at its center. And since the majority of the RF signal travels along the outside edge of the center conductor, the gap will not contribute significantly to reflection losses.

Fabrication for these types of switches started with E-beam deposition of SiCr (1000 Å thick) using a metal lift-off technique. This step was followed by thermal evaporation of a 3000 Å thick Cr/Au layer for the part of center conductor that surrounds the SiCr plate underneath the center of the MEMS bridge. PECVD silicon nitride was deposited to a thickness of 2500 Å on top of the SiCr layer. PMMA was then spun-on to a thickness of 1.8 µm and etched pack after patterning with photoresist to define and open up the outlines of the CPW and beam structures. After a thin Cr/Au seed layer deposition, the CPW lines and the shunt beam was gold electroplated to a thickness of about 1.0 µm using patterned photoresist as a mask. After the photoresist mask was removed and the beam and CPW lines protected with another photoresist layer, the see layer was removed and the MEMS switches were released using 1165 and a critical point drying technique. The fabrication process is illustrated in Figure 3.38.
3.7.2 Measured Results

Measurements were performed from 1 to 16 GHz using the same set up that was used for testing the capacitive MEMS switches. The bias was applied through a bias tee between the CPW ground planes and the high resistivity SiCr line. Actuation occurred at about 20 V and the $S_{21}$ responses in the actuated and un-actuated states are shown in Figure 3.39 and Figure 3.40, respectively.
Figure 3.39. Measured and simulated $S_{11}$ and $S_{21}$ of the shunt contact switch in the down state. Measured data is shown with markers and measured data with straight lines.

Figure 3.40. Measured and simulated $S_{11}$ and $S_{21}$ of the shunt contact switch in the up state. Measured data is shown with markers and measured data with straight lines.
From Figure 3.39, it can be observed that the insertion loss of the metal-to-metal contact switch was about 0.8 dB from 6 to 14 GHz. The relatively high loss can be attributed to a high contact resistance between the beam metal and the lower center electrode which could be lowered by increasing the contact area.

3.8 Conclusion

The various switch designs were introduced that include shunt switches and series switches with capacitive coupling as well as a shunt contact switch using separate actuation structures. Initial design iterations were shown that include a dual cantilever shunt switch with a large actuation pad at the tip of the suspended beam structure and series switches, and shunt switches with double beams, where a thin metal pad was fabricated directly on top of the dielectric to improve the RF coupling when the beam is in the down state. Also shunt switches with single beams and with various geometries of thin meander line sections were shown. Finally, a shunt DC contact switch was introduced.

After the switches were fabricated and measured, isolations between 30 and 45 dB were typically measured at their respective resonant frequencies in the off-state. The series capacitive switches showed isolation of about 30 dB measured at 10 GHz. Insertion losses in the on state were measured to be 0.3 to 0.5 dB for most switches. The results were also compared to simulated data. Equivalent circuit models for the MEMS shunt and series switches were derived and circuit component values were extracted from simulated and measured data. Circuit models for both the on-state and off-state cases are
shown. Switching speed measurements were performed and results in the 100 $\mu$s range will be shown. Quartz lids with SU-8 rings were used to package a shunt MEMS switch with little loss to RF performance.
4.1 Introduction

RF MEMS switches, as introduced in earlier chapters, are not only useful as stand-alone devices, but are often used as building blocks for more complicated circuits such as phase shifters, tunable filters and tunable transmission lines [17-21]. Another application is in multi-port switch design where the properties of MEMS switches can be utilized to achieve high performance switching networks at high frequencies. A single-pole double-throw (SPDT) switch is one example of a multi-port switching circuit that has become an integral part in many microwave systems. This includes transceivers, where they are used to switch a device between receiver and transmitter modes, and receiver front ends for diversity operation [48, 49].

In this chapter, a MEMS coplanar waveguide SPDT switch has been designed using the capacitive shunt switch topology introduced in the previous chapter as building blocks. By implementing the shunt switch as a series and shunt pair, separated by a quarter wavelength transmission line, just a single voltage to route the signal between the two output ports is needed, i.e. with no applied bias, the signal flows through one port but gets routed through the other port when a bias is applied. The advantage of this is that
the switch only needs to be powered in one state therefore simplifying the biasing as well as potentially increasing the switching lifetime of the device.

The SPDT 3-port switch was fabricated and measured. Port isolations of about 15 dB and an insertion loss of 1 dB were obtained in the up state. In the down state, 40 dB of isolation with a 1 dB insertion loss were measured. The actuation voltage was 35 V. An equivalent circuit model was derived by first modeling each switch section separately. The models of the two sections were then connected with a quarter wavelength transmission line model and the response was compared to measured results.

4.2 Switch Operation

Conventional SPDT MEMS switches generally route a signal between 2 output ports by alternating the actuation of the MEMS devices that either open one signal path while closing the other, and vice-versa, or alternately shunt one of the signal lines to ground. These methods are illustrated in Figure 4.1. In the latter case, quarter wavelength transmission line segments are needed to provide a virtual open at the output when the switch is closed. What both types of switching methods have in common is that they require each MEMS switch in the circuit to be actuated separately. This can complicate the bias network design, especially when a large number of SPDT switches are required, such as in a steering array for antenna feed networks. Also, these types of switches call for a constant bias voltage that is applied to either switch continuously; i.e. for signal a signal to get routed at least one switch needs to be biased. This means that during operation, power has to be always applied to the circuit. With zero bias, the SPDT switch would either just reflect the input signal, as in the series switches case, or would
act as a power divider, as in the double shunt case. Table 4.1 lists a comparison of both shunt and series SPDT switches, using MEMS technology, that have been reported.

![Diagram of conventional SPDT switches. Configurations are shown with 2 series switches (left) and with 2 shunt switches separated by quarter-wavelength transmission lines (right).](image)

**Figure 4.1.** Diagram of conventional SPDT switches. Configurations are shown with 2 series switches (left) and with 2 shunt switches separated by quarter-wavelength transmission lines (right).

<table>
<thead>
<tr>
<th>Author</th>
<th>Switch Configuration</th>
<th>Insertion Loss (dB)</th>
<th>Isolation (dB)</th>
<th>Frequency (GHz)</th>
<th>Actuation Voltage (V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Park [50]</td>
<td>Shunt</td>
<td>0.45</td>
<td>22</td>
<td>47</td>
<td>35</td>
</tr>
<tr>
<td>Scardelletti [51]</td>
<td>Shunt</td>
<td>0.11</td>
<td>26</td>
<td>26.5</td>
<td>30</td>
</tr>
<tr>
<td>Cho [52]</td>
<td>Series</td>
<td>0.52</td>
<td>36</td>
<td>20</td>
<td>4.3</td>
</tr>
<tr>
<td>Tang [53]</td>
<td>Series</td>
<td>0.5</td>
<td>25</td>
<td>10</td>
<td>20</td>
</tr>
<tr>
<td>Dubuc [54]</td>
<td>Shunt</td>
<td>0.6</td>
<td>21</td>
<td>30</td>
<td>35</td>
</tr>
</tbody>
</table>

The SPDT MEMS switch that has been developed in this investigation is similar to the conventional double shunt switch design but instead of using two identical shunt MEMS switches, one of the switches will be used to couple the signal to a corresponding output port in the down state, i.e. it will either isolate the port or provide a signal path when the switch is actuated. The shunt switch still acts as a RF open when actuated. The advantage of this design is, since the materials and dimensions of the dielectric and actuation structures are identical for both switches, near simultaneous actuation can be accomplished with a single voltage. Figure 4.2 shows the layout of the switch.
Figure 4.2. SPDT switch layout (switch length not drawn to scale).

The operation of the switch can be described as follows: When both switches are in the upstate, the signal flows through port 2 and is isolated from port 3, due to the air gap in both switches. When a bias is applied to the device, causing both switches to actuate simultaneously, the signal will now flow through port 3 due to capacitive coupling in the down state of the series switch. The actuated shunt switch shorts the signal at port 2, which is transformed to an RF open due to the quarter wavelength separation from the series switch. This provides the isolation from port 2 in the down state.

4.3 Switch Design and Fabrication

The geometry and material dimensions of the SPDT switch’s shunt switch section are identical to that of the capacitive single beam shunt switch developed in the previous chapter. The MEMS section that couples the signal to port 3 also consists of identical beam, actuation pad, and dielectric properties and geometry. In both cases, meander inductive line sections are used to lower the resonant frequency to 10 GHz. The CPW
line consists of 100 µm wide center conductors and 50 µm wide slots. The thin metal plates measure 100x200 µm and the actuation beams are 415 µm long and 60 µm wide. The overall dimension of the SPDT switch is 1.4x3.42 mm. Figure 4.3 shows an optical image of the shunt and series switch sections of the fabricated SPDT switch.

![Optical image of a fabricated SPDT switch. The shunt switch section is shown on the right, the series section (with bond wires) on the left. The fabrication procedure in Appendix 1 was followed.](image)

4.4 Measured Results

The SPDT switch was measured using an Anritsu 37397C vector network analyzer and 4-port switching box. The VNA and the 4-port box were controlled using the NISTCAL multi-port calibration and measurement software. The port convention shown in Figure 4.2 was used during the measurements and for the simulations. With no applied bias, i.e. with the MEMS bridges in the up state and the signal getting transmitted between ports 1 and 2, a port 1 to port 3 isolation of 15 dB was measured. This was about the same for the isolation between ports 2 and 3. An insertion loss of 1 dB was measured between ports 1 and 2. Fig. 4.4 shows a comparison between the measured and simulated results of the return losses at each port. Figure 4.5 shows a comparison
between the measured and simulated data for $S_{21}$, $S_{32}$, and $S_{31}$. It can be observed that the isolation was about 5 dB lower for the measured switches since the air gap of the MEMS bridges was lower than the specified gap of $1.8 \ \mu$m of the simulated switches.

![Graph showing the comparison between measured and simulated results for $S_{21}$, $S_{32}$, and $S_{31}$](image1)

**Figure 4.4.** Port 1 ($S_{11}$) and port 2 ($S_{22}$) return losses are shown and compared to simulated results in the un-actuated state. Return loss of the isolated Port 3 ($S_{33}$) is also plotted.

![Graph showing the comparison between measured and simulated results for $S_{31}$ and $S_{32}$ isolation](image2)

**Figure 4.5.** Port 3 to port 1 ($S_{31}$) and port 2 to port 3 ($S_{32}$) isolation are compared to simulated results with the switch in the un-actuated state. Measured and simulated port 2 to port 1 ($S_{21}$) insertion loss is also shown.
The switches were then actuated with a 35 V bias. The return loss at each port in the down state is shown in Figure 4.6. The isolation and insertion loss response of the switch are shown in Figure 4.7. The port 2 to port 1 and port 3 isolation reached nearly 40 dB in the down state. Again, an insertion loss of about 1 dB was measured from port 1 to port 3. Both plots show a comparison of the measured data to corresponding simulated results.

Figure 4.6 Port 1 ($S_{11}$) and port 2 ($S_{22}$) return losses are shown and compared to simulated results in the actuated state. Return loss of the isolated Port 3 ($S_{33}$) is also plotted.
Figure 4.7 Port 3 to port 1 (S_{31}) and port 2 to port 3 (S_{32}) isolation are compared to simulated results with the switch in the actuated state. Measured and simulated port 2 to port 1 (S_{21}) insertion loss is also shown.

4.5 SPDT Switch Model

A circuit model of the designed and tested SPDT switch was developed to extract the various equivalent circuit values, such as up- and down-state capacitances, series resistance, and series and shunt inductances. To help simplify the modeling procedure, the SPDT switch was separated into two parts: the 2-port shunt switch section (Figure 4.8) and the 3-port series switch section (Figure 4.9). Also, each section of the SPDT circuit layout was also simulated separately using ADS Momentum; the series switch as a 3-port device and the shunt switch as a 2-port device. The corresponding model was then used to extract the equivalent circuit values for each section. The two models were finally combined with a quarter wavelength CPW transmission line section and the resulting model was compared against the modeled and measured results of the complete SPDT switch.
Figure 4.8. 2-port model for shunt MEMS switch section.

Figure 4.9. 3-port model for shunt MEMS switch section.
In both models, $C$ represents the up- and down-state capacitance values of the MEMS actuation structures. $L_s$ and $R_s$ are the inductance and resistance values, respectively, of the beam structure taken from the base to the coupling area. $R_c$ models the contact resistance between beam and the top metal when the switches are in the down state. The series inductance and resistance values of the inductive line sections are represented by $L_{sh}$ and $R_{sh}$, respectively. A shunt capacitor, $C_s$, was added to the 3-port model to help take the coupling between the beam and the ground plane opposite to the port 3 output into account. Library models of 50 $\Omega$ CPW transmission lines were also added to the ports to model the short CPW line sections at the inputs and outputs of the circuit layouts used during the simulations. These sections were also used to model any line portions of the fabricated devices not de-embedded during the calibration.

Figure 4.10 shows a comparison between the results of the simulated up state shunt switch section and the circuit model results after the model parameters were optimized to the simulation results. The same comparison of the switches high frequency response, but this time with the beam in the down state, can be viewed in Figure 4.11. Figure 4.12 contains the RF responses of the simulated 3-port switch section and its equivalent circuit model in the un-actuated state after optimization. The plotted simulated and model results of the 3-port section in the down state are shown in Figure 4.13. The extracted equivalent circuit values of each model are listed in Table 4.2.
Figure 4.10. Modeled and simulated response of the 2-port section in the up state. Top shows $S_{11}$ magnitude and phase plots, bottom plot shows $S_{21}$ magnitude and phase plots. Simulated data is shown with markers and modeled data as straight lines.
Figure 4.11. Modeled and simulated response of the 2-port section in the down state. Top shows $S_{11}$ magnitude and phase plots, bottom plot shows $S_{21}$ magnitude and phase plots. Simulated data is shown with markers and modeled data as straight lines.
Figure 4.12. Modeled and simulated response of the 3-port section in the up state. Top shows $S_{11}$ magnitude and phase plots, middle plot shows $S_{21}$ magnitude and phase plots, and bottom shows $S_{31}$ magnitude and phase plots. Simulated data is shown with markers and modeled data as straight lines.
Figure 4.13. Modeled and simulated response of the 3-port section in the down state. Top shows $S_{11}$ magnitude and phase plots, middle plot shows $S_{21}$ magnitude and phase plots, and bottom shows $S_{31}$ magnitude and phase plots. Simulated data is shown with markers and modeled data with straight lines.
Table 4.2. Extracted equivalent circuit values for the 2-port and 3-port section, modeled and simulated separately, in the up- and down-states.

<table>
<thead>
<tr>
<th>Beam Position</th>
<th>C (pF)</th>
<th>R (Ω)</th>
<th>Ls (nH)</th>
<th>Rs (Ω)</th>
<th>Lsh (nH)</th>
<th>Rsh (Ω)</th>
<th>Cs (fF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2-Port up</td>
<td>0.0625</td>
<td>0</td>
<td>0.07</td>
<td>0</td>
<td>0.057</td>
<td>0</td>
<td>NA</td>
</tr>
<tr>
<td>2-Port down</td>
<td>4.4</td>
<td>0.016</td>
<td>0.07</td>
<td>0</td>
<td>0.056</td>
<td>0</td>
<td>NA</td>
</tr>
<tr>
<td>3-Port up</td>
<td>0.045</td>
<td>0</td>
<td>0.066</td>
<td>0</td>
<td>0.265</td>
<td>0</td>
<td>9</td>
</tr>
<tr>
<td>3-Port down</td>
<td>4.25</td>
<td>0.01</td>
<td>0.065</td>
<td>0</td>
<td>0.23</td>
<td>0</td>
<td>9</td>
</tr>
</tbody>
</table>

From Table 4.2, it can be seen that the series resistance contribution of the beam segments and meander lines to the circuit response is very small and was therefore neglected. It can be also observed that the series capacitances values of C, whether actuated or in the up state, are slightly smaller for the 3-port device than those of the corresponding 2-port section. This is due to the position of the beam structure with respect to the lower contact electrode as can be seen in Figure 4.2. Since the shunt beam is centered over the actuation pad while the series beam is located over the edge of its corresponding pad, less fringing capacitance in the latter case will contribute to the overall capacitance in the coupling area. It can also be observed that inductance values in the up- and down–states for both switch sections are nearly identical which gives credibility to the extracted capacitance values.

Once all the equivalent circuit values were extracted for each actuation state, the two and three port models were then combined using a 2.9 mm long CPW line component, to represent the quarter wavelength section, and simulated across the frequency band. Small changes had to be made to some of the circuit values to improve the modeled data fit to the corresponding simulated response. The combined outputs of the simulated circuit and the equivalent circuit model for the SPDT switch in the up state
are shown in Figure 4.14. The same comparison of the simulated and modeled results for the switch in the down state can be viewed in Figure 1.15. The equivalent circuit values of the combined model can be found in table 4.3.

Figure 4.14. Modeled and simulated response of the simulated SPDT switch in the up state. Top shows $S_{11}$ magnitude and phase plots, middle plot shows $S_{21}$ magnitude and phase plots, and bottom shows $S_{31}$ magnitude and phase plots. Simulated data is shown with markers and measured data with straight lines.
Figure 4.15. Modeled and simulated response of the simulated SPDT switch in the down state. Top shows $S_{11}$ magnitude and phase plots, middle plot shows $S_{21}$ magnitude and phase plots, and bottom shows $S_{31}$ magnitude and phase plots. Simulated data is shown with markers and modeled data with straight lines.
Table 4.3. Extracted equivalent circuit values for the combined circuit model in the up- and down-states obtained from the simulated SPDT response.

<table>
<thead>
<tr>
<th>Switch Section</th>
<th>Beam Position</th>
<th>C (pF)</th>
<th>R (Ω)</th>
<th>Ls (nH)</th>
<th>Rs (Ω)</th>
<th>Lsh (nH)</th>
<th>Rsh (Ω)</th>
<th>Cs (fF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2-Port</td>
<td>up</td>
<td>0.0625</td>
<td>0</td>
<td>0.07</td>
<td>0</td>
<td>0.056</td>
<td>0</td>
<td>NA</td>
</tr>
<tr>
<td>2-Port</td>
<td>down</td>
<td>4.4</td>
<td>0.01</td>
<td>0.07</td>
<td>0</td>
<td>0.066</td>
<td>0</td>
<td>NA</td>
</tr>
<tr>
<td>3-Port</td>
<td>up</td>
<td>0.045</td>
<td>0</td>
<td>0.096</td>
<td>0</td>
<td>0.265</td>
<td>0</td>
<td>9</td>
</tr>
<tr>
<td>3-Port</td>
<td>down</td>
<td>4.25</td>
<td>0.012</td>
<td>0.011</td>
<td>0</td>
<td>0.22</td>
<td>0</td>
<td>9</td>
</tr>
</tbody>
</table>

Only small changes had to be made to the some of the inductance values, as seen in Table 4.3, to obtain a good fit of the combined circuit model to the simulated response of the SPDT switch. It also can be observed that the coupling capacitance that were found in the separate section stayed the same in the combined cases.

Equivalent circuit values were also extracted using the same combined model from the measured results of the fabricated SPDT switch. The values for the measured switch in the up-state as well as after actuation are shown in Figures 4.16 and 4.17, respectively. These are listed in Table 4.4.
Figure 4.16. Modeled and measured response of the simulated SPDT switch in the up state. Top shows $S_{11}$ magnitude and phase plots, middle plot shows $S_{21}$ magnitude and phase plots, and bottom shows $S_{31}$ magnitude and phase plots. Measured data is shown with markers and modeled data with straight lines.
Figure 4.17. Measured and simulated response of the simulated SPDT switch in the down state. Top shows $S_{11}$ magnitude and phase plots, middle plot shows $S_{21}$ magnitude and phase plots, and bottom shows $S_{31}$ magnitude and phase plots. Measured data is shown with markers and modeled data with straight lines.
Table 4.4. Extracted equivalent circuit values for the combined circuit model in the up- and down-states obtained from the measured SPDT response.

<table>
<thead>
<tr>
<th>Switch Section</th>
<th>Beam Position</th>
<th>C (pF)</th>
<th>R (Ω)</th>
<th>Ls (nH)</th>
<th>Rs (Ω)</th>
<th>Lsh (nH)</th>
<th>Rsh (Ω)</th>
<th>Cs (fF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2-Port up</td>
<td>0.092</td>
<td>0</td>
<td>0.015</td>
<td>0.22</td>
<td>0.26</td>
<td>0.22</td>
<td>NA</td>
<td>NA</td>
</tr>
<tr>
<td>2-Port down</td>
<td>5.6</td>
<td>33</td>
<td>0.035</td>
<td>0.11</td>
<td>0.04</td>
<td>0.46</td>
<td>NA</td>
<td>NA</td>
</tr>
<tr>
<td>3-Port up</td>
<td>0.07</td>
<td>0</td>
<td>0.11</td>
<td>1.17</td>
<td>0.085</td>
<td>0.28</td>
<td>8</td>
<td></td>
</tr>
<tr>
<td>3-Port down</td>
<td>5.3</td>
<td>38</td>
<td>0.11</td>
<td>1.8</td>
<td>0.11</td>
<td>1.7</td>
<td>12</td>
<td></td>
</tr>
</tbody>
</table>

Table 4.4 shows slightly higher capacitance values for the MEMS coupling structure. This is due to changes in the component dimension of the metal lines and can be attributed to fabrication errors, even though the SPDT was simulated with parameters matching those of the fabricated device, such as dielectric thickness, beam thickness, and air gap. It can also be seen from the table that there is a significant amount of contact resistance in the down state for both actuation structures. This resistance, as was explained in Chapter 3, is related to the force of contact between the actuated beam structure and the thin metal pad of the bottom electrode.

4.6 Conclusion

A SPDT RF MEMS switch, where an X-band signal can be switched between two output ports by applying a single bias voltage, was introduced. A 35 V bias was needed to actuate the switch. For the fabricated SPDT switch, the measured isolation of the isolated ports in the up state was 15 dB with an insertion loss of 1 dB through the transmitted port. In the down state, when a 35 V bias was applied, the isolation of the isolated ports was found to be 40 dB. The insertion loss to the other port was again about 1 dB.
Two separate equivalent circuit models were derived to represent the two switch sections of the SPDT device. Each section was also simulated separately and the results were used to extract the component values of its circuit model. After a good fit was obtained, the two circuit models were combined with a quarter wavelength CPW line model after which the circuit responses of the model and the simulated SPDT switch were compared. It was seen that a good match was obtained with only slight changes needed to the component values. Finally, the same combined model was used to extract the equivalent circuit parameters of the measured SPDT response.

As already stated, the advantage of this new SPDT switch design is that a signal can be routed between two ports using just a single bias. This can be attributed to the use of two capacitive bridge MEMS structures with identical geometries, one in series and one in shunt configuration off-set by a quarter-wavelength line section, in the design. It has been found though from the measured isolation plots, that the isolation is much higher in the down state. This is to be expected since the parasitic coupling capacitance of the series beam to the center conductor underneath can be significant with gaps below 3 µm when the switch is un-actuated. A capacitive shunt switch inherently provides much better isolation at higher frequencies. This helps give the SPDT switch a much better isolation in the down state at 10 GHz.
Chapter 5

FIB Fabricated Nano Devices

5.1 Introduction

The work presented in this chapter addresses the use of focused ion beam (FIB) milling for micro and nano structure fabrication. Due to the FIB’s ability to mill material with high precision, high-aspect-ratio structures with relatively smooth sidewalls can be achieved without the use of a photo mask. Devices such as microfabricated accelerometers [55], microgratings for integrated optics [56], and micromilled trenches [57] have already been demonstrated. This technology could also be applied to fabricate capacitive structures by milling a gap with a width in the tens of nanometers across micro- and millimeter wave transmission line structures. And by applying the fabrication technique to suspended beams, MEMS structures such as switches and varactors can potentially be fabricated, due to the FIB’s ability to produce angle-cuts.

In this chapter, the use of a FIB system to cut submicron wide gaps across the conductors of coplanar waveguide (CPW) transmission lines is discussed. Capacitance values were extracted from measured data in the 1 to 65 GHz frequency range using lumped equivalent circuit models. Besides straight cuts, meander line cuts and gaps milled at a 52 degree angle (angle cuts) have been investigated. Series gap capacitances of 8.5-12 fF were obtained for the straight cut, meander cut and angle cut lines. An
equivalent shunt capacitance of 11 fF was found for a shunt cut CPW line. All devices measured showed negligible parasitic effects over the entire frequency band with estimated series resonant frequencies exceeding 1 THz. The circuit models account for a slight conductance across the gap, most likely caused by gallium contamination of the silicon substrate [58]. This contamination can occur when gallium ions from the focused ion beam are implanted into the milled material.

This fabrication technique was also utilized to develop a MEMS DC contact series switch with gap separation of about 100 nanometers by making an angle cut across the center of a suspended beam structure. This was done to significantly reduce the distance needed to close the gap during actuation and therefore increase the switching speed of the switch. Generally, this gap between MEMS structure and actuation area is in the order of micrometers for conventional MEMS switches and is considered a timing factor on the switching speed performance. However, by reducing the travel distance by a factor of over 10, the switching should also be greatly improved. To date, MEMS switches with switching speeds in the microseconds have been reported [43-46]. The RF characteristics of a fabricated nano-switch were measured from 1 to 50 GHz in the actuated and un-actuated states. An insertion loss of less than 0.3 dB and isolation greater than 25 dB up to 10 GHz were obtained. Switching speed tests showed a switching time of less than 300 ns for the on state and around 1 µs for the off state.
5.2 FIB Milled Capacitive Gaps

5.2.1 Focused Ion Beam

Milling with a FIB system is achieved by focusing a beam of ions down to a submicron area. This beam is accelerated to a high voltage, generally between 5 and 50 keV, and interacts in a well-defined area within the target material. The ion beam is produced from the field ionization of a gallium metal that is coated on a needle tip, usually made of tungsten or platinum, with a radius in the sub micron range. The ionized field (\(>10^8\) V/cm) is created by a high electric field at the needle tip. The ion beam can then be focused to a beam diameter ranging from less than 5 nm up to half a micron by changing the beam current density. This change in the beam diameter is accomplished by controlling the strength of the electrostatic lenses and adjusting the effective aperture sizes. Material removal in the area of interest can be precisely controlled and viewed since the accelerated ions will generate secondary electrons and ions which can be detected much in the same way as in a scanning electron microscope system. Exact etch patterns and depths can be specified in many computer controlled FIB systems. And since FIB milling is generally a sputtering process, undesired Ga ion implementation into the sample substrate can also occur. Figure 5.1 shows the various components, such as lenses and apertures, in a FIB column required to generate an ion beam.
5.2.2 FIB Milled CPW Gaps

Figure 5.2 shows layouts of the CPW lines and three types of FIB cuts tested; straight cut, meander cut, and shunt with straight lines. A fourth type of milled transmission line was cut with the ion beam set at a 52˚ angle with respect to the straight cut (see Figure 5.3). All devices were fabricated on 400 µm-thick, high-resistivity silicon wafers (ρ > 2000 Ohm-cm). The CPW dimensions used were 45 µm for the center conductor width (S) with 27 µm wide conductor to ground gaps (W). Each line segment of the meander cuts was 10 µm long. The shunt segments of the shunt device were 30 µm wide.
Figure 5.2. Layout drawings of the types of cut geometries studied; straight, meander, and shunt. $S=45 \, \mu\text{m}$, $W=27 \, \mu\text{m}$. Shunt lines are 30 $\mu\text{m}$ wide.

The metal lines consisted of a 0.4 $\mu\text{m}$ thick chromium/gold layer fabricated using electron beam deposition. After the CPW transmission line deposition (1 $\mu\text{m}$ thick), a 0.2 $\mu\text{m}$ thick polymethal methacrylate (PMMA) layer was spun onto the sample and cured. This was done so a conductive coating (50 Å of Cr) can be applied to the sample that is needed to eliminate electron charging during the FIB milling process. The fabrication process of the milled capacitors is shown in Figure 3.3 and a cross sectional view of the cut through the layers is shown in Figure 5.4. The addition of the PMMA layer also has the beneficial effect of creating narrower cuts in the transmission lines since the focused ion beam generally tapers down to a point during milling.
Figure 5.3. Fabrication process of FIB milled capacitive devices.

Figure 5.4. Cross sectional view of FIB cut into sample surface. Top shows straight cut and bottom shows the cut at a 52° angle. Not drawn to scale.
The gaps in the CPW lines were milled using the FEI DB235 dual beam FIB system. A 30 keV ion beam was used with a set current of 10 pA. The etch depth was set to 0.6 µm, corresponding to the total metal and PMMA layer thickness. This corresponds to an etch time of about 3 minutes for the straight cut, and 9 minutes for the meander pattern. For the angled cut, an etch depth of 0.9 µm was selected which took about 5 minutes to complete. The straight and angle cuts were made slightly longer than the 45 µm wide center conductor to ensure a complete cut over the entire width. After milling, the Cr and PMMA layers were removed by placing the sample in a heated (80º C) Microposit 1165 photoresist stripper. SEM images of a milled gap are shown Figure 5.5.

Figure 5.5. SEM images showing FIB milled submicron gaps in metal line.
5.2.3 Measured and Modeled Results

Series gaps in a CPW center conductor have already been modeled using a lumped Pi-network consisting of a series capacitor and two fringing capacitors [6]. For this application, a series inductor/capacitor in parallel with a resistor model was used. The parallel resistor ($R_g$) was needed to account for the conductive effects caused by the ion beam milling, as already mentioned. The series resistor ($R_s$) was used to model parasitic resistive effects in the circuit. The shunt model also includes a parallel capacitor/inductor to resistor combination, but this time shunted to ground. Model extraction from the measured data was performed using Agilent ADS. CPW transmission line models were added to include the effects of any transmission lengths not de-embedded during calibration. Figure 5.6 shows a schematic of both circuit models.

![Circuit Models](image)

Figure 5.6. Equivalent circuit models of FIB milled capacitors. Top shows series cut model and bottom shows shunt cut model.
Measurements were performed from 1 to 65 GHz on a Karl Suss probe station with a Wiltron 360 vector network analyzer (VNA). 100 µm pitch 3 prong G-S-G probes were used to measure the devices and the TRL on wafer calibration standards. Figures 5.7 through 5.10 show measured to modeled data $S_{11}$ and $S_{21}$ comparisons for the straight, meander, angle, and shunt cut circuits, respectively.

![Comparison of S-parameters between measured and modeled data for the straight cut FIB device.](image)

Figure 5.7. Comparison of S-parameters between measured and modeled data for the straight cut FIB device.
Figure 5.8. Comparison of S-parameters between measured and modeled data for the meander cut FIB device.

Figure 5.9. Comparison of S-parameters between measured and modeled data for the angle cut FIB device.
Table 5.1 summarizes the extracted equivalent circuit model parameters of each of the tested devices. It can be seen that the gap capacitance is on the order of 10 fF for all the measured circuits. A higher capacitance for the meander cut circuit was expected since the effective length of the meander was over double the length of the straight cut. Only the shunt and meander cut circuits showed significant parasitic inductive effects, which can be attributed to the 30 µm wide shunt lines in the shunt case and to the lengths of the interdigital fingers in the meander case. The values extracted for parallel resistor, $R_g$, clearly shows there is some conductive effect across the gap in all the tested cases. This seems to validate the undesired effect of gallium contamination of the substrate during the ion beam milling process.
Table 5.1. Extracted equivalent circuit values.

<table>
<thead>
<tr>
<th>Type</th>
<th>C (fF)</th>
<th>Rg (kΩ)</th>
<th>Rs (Ω)</th>
<th>L(pH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Straight</td>
<td>8.5</td>
<td>25</td>
<td>11.2</td>
<td>0</td>
</tr>
<tr>
<td>Meander</td>
<td>10.0</td>
<td>26</td>
<td>9.8</td>
<td>1.25</td>
</tr>
<tr>
<td>Angle</td>
<td>12.0</td>
<td>25</td>
<td>10.8</td>
<td>0</td>
</tr>
<tr>
<td>Shunt</td>
<td>11.0</td>
<td>46</td>
<td>1.6</td>
<td>1.33</td>
</tr>
</tbody>
</table>

The low capacitance values across the gaps resulted mainly due to the small surface area of the milled walls (45 μm long and 0.4 μm high), since the capacitance can be roughly modeled as a parallel plate capacitor with an air gap. If the parallel plate capacitor equation (2.1) is used, a capacitance of 2.27 fF for the straight cut capacitor is obtained. Due to fringing capacitances in the air and in the substrate, as shown in Figure 5.11, the effective capacitance will be higher and closer to the values extracted in Table 5.1, and is given as:

\[ C = C_{\text{gap}} + C_{f_{\text{air}}} + C_{f_{\text{sub}}}, \]

where \( C_{\text{gap}} \) is the capacitance across the milled gap and \( C_{f_{\text{air}}} \) and \( C_{f_{\text{sub}}} \) are the fringing capacitances in the air and substrate, respectively.

\[ \text{Figure 5.11. Capacitance components of total of capacitance across FIB milled gap.} \]
Agilent’s HFSS electromagnetic field simulation software was used to simulate the electric field strength across the gap at 40 GHz with a gap of 70 nm. The fringing fields in the air above and the substrate below the gap is clearly illustrated in Figure 5.12. It can be observed that the electric field is concentrated in the air gap. The fringing field is spread over a wider area in the air above the gap than in the substrate.

The RF response form 1 to 40 GHz of the milled capacitor was also obtained using HFSS. Good agreement between measured and simulated data can be seen in the plots of $S_{21}$ and $S_{11}$, as shown in Figure 5.13 (measured plot is shown with markers). Above this frequency, the simulation showed a higher RF coupling than what was observed in the measured results.
From the modeled results, it was shown that very small scale (in the nano-meter range) capacitive structures with low parasitic effects up to mm-wave frequencies can be fabricated using a FIB’s milling capabilities. The advantage of this type of milling technique is that capacitors can be fabricated with a single layer process, without the need of a dielectric layer, and can be used as building blocks for other nano- and micro-fabricated devices. If higher valued capacitances were required, the gap could be filled, by sputtering or chemical vapor deposition, with a dielectric material with high dielectric constant.
5.3 FIB Nano Switch

5.3.1 Switch Design

To develop a MEMS switch with switching speeds in the nano second regime, the milling capabilities of a FIB system have been utilized to cut submicron wide gaps across suspended portions of the center conductors of coplanar waveguide (CPW) transmission lines. The MEMS switch consists of a CPW line where a small portion of the center conductor is suspended over a thin silicon chromite (SiCr) actuation pad shown in Figure. 5.14. The high resistivity SiCr layer is used to electro-statically actuate the suspended cantilevers after the FIB milling process.

![FIB Cut SiCr](image)

Figure 5.14. FIB milled MEMS series switch (not drawn to scale).

The nano milled MEMS switch is based on a CPW series metal-to-metal contact configuration. Milling the suspended section of the center conductor at an angle results in two overlapping cantilevers that can be actuated independently, as illustrated in Figure 5.15. If a bias is applied beneath the lower electrode the gap can be increased to improve the isolation of the nano scale gap. When the bias is changed to the other electrode, the gap will close resulting in a DC contact between the electrodes.
5.3.2 Fabrication Process

The switches were fabricated on 400 µm-thick, high-resistivity silicon wafers ($\rho > 2000 \text{ Ohm-cm}$). The CPW dimensions used were 65 µm for the center conductor width with 27 µm wide conductor to ground gaps. The suspended portion of the center conductor has a width of 60 µm and length of 80 µm. Figure 5.16 summarizes the fabrication process of the MEMS cantilevers. First a 0.1 µm thick SiCr layer is fabricated using E-beam deposition and lift-off technique followed by a 0.1 µm thick PECVD nitride layer to isolate the SiCr from the cantilevers in case of over actuation. Then a 1.5 µm thick PMMA, which acts as the sacrificial layer, was spun on the sample, patterned and etched to define the CPW transmission line geometry. A seed layer of
Ti/Au was then deposited on the PMMA and the open areas for gold electroplating of the metal lines. After masking with photoresist, the metal lines were electroplated to a thickness of 1.1 μm. And finally, after the seed layer was removed, the sacrificial layer was dissolved in photoresist remover and the sample was taken through a critical point drying process to finish the release of the actuated structures.

![Fabrication process of the suspended cantilever before FIB milling.](image)

Figure 5.16. Fabrication process of the suspended cantilever before FIB milling.

The suspended CPW lines were then milled using the FEI DB235 dual beam FIB system. A process similar to that described in section 5.17 was used to mill the narrow gaps. A 30 keV ion beam was used with a set current of 100 pA to mill the gaps. A SEM of the top view at of a milled gap and a close up view of the gap at an angle is shown Fig. 5. The gap width was measured to be approximately 100 nm.
5.3.3 Measured Results

The RF characteristics of fabricated nano milled switches were measured in the un-actuated state as well as when each cantilever was actuated separately. Figure 5.18 shows the insertion losses and return losses, represented by the $S_{21}$ and $S_{11}$ curves, respectively, of a FIB nano switch in the un-actuated and the two actuated states. It can be seen from the figure, the isolation increases slightly when the lower electrode was actuated with a bias of 20 V. A higher isolation was not achieved with increasing voltage due to the small air gap of about 0.5 µm underneath the lower electrode. This means that at 20 V, the lower electrode is probably already making contact with the dielectric on top of the SiCr layer. The insertion loss is shown in the $S_{21}$ plot in Figure 5.19.

* Each cantilever of the switch can be actuated independently. If the upper overlapping cantilever is actuated, the switch will close. If the under lapping cantilever is actuated, the switch opening will increase.
Another switch was measured that appeared to have the two electrodes already in contact after the FIB milling process. This is likely due to residual stresses in the beam structure that can occur during the fabrication process. In this case, only a bias was
applied to the lower electrode and the switch was opened using a bias of 18 V. Figure 5.20 show the insertion loss and isolation of the switch in the actuated and un-actuated states. From this example, it can be observed that even with undesired residual stresses that would generally have detrimental effects to switching performance in conventional MEMS switches, this type of switch structure is still usable with good performance.

![Graph](image)

**Figure 5.20.** $S_{11}$ and $S_{21}$ plots of switch in the un-actuated closed and actuated open positions.

The RF performance of a FIB milled nano-switch was also simulated using HFSS with an open gap of 100 nm. A comparison between the measured and simulated S-parameter in the un-actuated state is shown in Figure 5.21.
Also using HFSS, the electric field distribution was simulated across the milled gap at a frequency of 50 GHz. The field strength is illustrated in Figure 5.22. Again, it can be observed that most of the electric field occurs in between the milled gap with the fringing field spreading in the air above and below the cantilevers.
5.3.4 Switching Speed Results

Switching speed measurements were performed using the same setup that was used to measure the switching performance of the shunt MEMS switches (Chapter 3). The nano milled switch with the open gap in the un-actuated state had a rise time of about 1.3 µs (the time it took for the signal to rise from 10% to 90% of the maximum signal strength) and a fall time of less than 300 ns when the actuation bias was removed. Figures 5.23 and 5.24 show the rise and fall times of the measured switch, respectively. Again, the diode detector outputs a reverse polarity voltage with respect to the detected signal level.

Figure 5.22. Electric field strength across FIB milled nano gap in suspended beam.
Figure 5.23. Rise time of nano switch from open to closed position with bias.

Figure 5.24. Fall time of nano switch from closed to open position with bias removed.
The nano milled switch with cantilevers that were originally in the closed position had rise and fall times of about 1 µs and 600 ns, respectively. The rise and fall time plots are shown in Figures 5.25 and 5.26, respectively. For this switch, the lower electrode was actuated to obtain the signal isolation in the off state.

Figure 5.25. Rise time of nano switch from closed to open position with bias.
Figure 5.26. Fall time of nano switch from open to closed position with bias removed.

It was observed from the switching speed data that the time required to pull down an electrode was longer than the time it took for the cantilever to come back to the un–biased state, regardless of whether the switch was originally in the open or closed position. One possible reason for this effect is that there is a time lag between the time when the bias is applied and when the actuated structure starts to respond to the bias. This time lag is longer than the time difference between the bias removal and the subsequent release of the switch. A possible explanation for this effect is that it takes longer for the static charge to build up to a high enough level to pull down the cantilever when the bias is turned on than it takes for the static charge to dissipate once the bias is removed.
5.5 Conclusion

A simple equivalent circuit model was derived for FIB milled sub-micron gaps in CPW transmission lines with widths of approximately 70 nm. Extracted capacitive values were obtained with these models for various types of FIB cuts and circuits, including a straight cut, meandered cut, angle cut, and straight cuts across shunt lines. Capacitances on the order of 10 fF were extracted from the models. One significant finding was the apparent effect of gallium contamination during the milling process, which was modeled as a parallel resistance across the gap in the circuit model. One possible way to reduce or eliminate this unwanted doping of the substrate would be to precisely control the etch depth to a greater degree without cutting too much into the underlying layer. The etched metal could also be placed on top of a sacrificial layer to be removed after the milling process is complete.

The results for a novel MEMS series switch produced using nano-fabrication techniques were also presented. This switch utilized focused ion beam milling to cut a ~100 nm wide gap in the suspended metal beam of a CPW center conductor. The FIB cut was done at a 52° angle, which results in two overlapping cantilevers. Each cantilever could then be biased separately to either open or close the switch. Measurements of the nano switches in the actuated and un-actuated states were performed from 1 to 50 GHz. Insertion losses of less than 0.3 dB at 10 GHz and less than 0.6 dB up to 50 GHz were measured with the cantilevers in the contact position. The switch isolation was increased by about 8 dB when the lower electrode was actuated. Actuation voltages were 18 and 20 V to close the lower electrode and the upper electrode, respectively. For a nano switch that was already in the on state after fabrication, isolation was also achieved with a
bias of 18 V. The rise and fall times of these switches ranged from less than 200 ns to
1.4µs. With a lower actuation voltage, the contact force between the cantilevers in the
on-state should also be reduced. This could significantly increase the reliability and
difetime of the switch. This fabrication technique could prove to be very valuable in the
design of high performance RF MEMS switches where switching speeds and reliability
comparable to those of solid state switches are needed.
Chapter 6

Reflectenna

6.1 Introduction

Wireless data transmission is becoming increasingly popular as a method to send data from sensors at remote locations, especially if direct physical connection cannot be achieved due to environmental constraints [59, 60]. Therefore it is important that the sensor’s embedded communication system consumes as little power as possible to optimize the usable life-time of the sensor. It is also desirable to reduce the size and complexity of the transmission system to minimize cost. Radio transmitters that are generally used for this application require active devices, such as power amplifiers, as well as up-converting circuitry to transmit the baseband signal [61].

This chapter introduces a simple, quasi-passive wireless telemetry system, similar to a passive transponder, where a continuous wave signal is modulated after reception and then redirected back to the transmission source. This system, called reflectenna, requires only a pair of antennas and a device to modulate the sensor data onto the continuous wave (CW) carrier signal for short range communication. To reduce the complexity of the system even further, on-off keyed (OOK) modulation is used which only requires a switch placed in the transmission path between the two antennas to produce the desired digital output. Since the transceiver that sends the CW and
demodulates the received signal from the sensor needs to differentiate between the two signals, the two antennas are oriented with opposite polarization, which can either be right and left hand circular, respectively, or vertical and horizontal linear, respectively. Figure 6.1 illustrates the operation of the reflectenna. No other active devices such as amplifiers or voltage-controlled oscillators are required on the sensor itself, therefore greatly reducing the energy requirements of the sensor system. This principle is also similar to a retro-directive communication system proposed by Wanselow [62]. Other communication techniques using retro-directive antenna arrays for telemetry applications have also already been demonstrated [63-65].

![Figure 6.1. Illustration of the reflectenna with modulating switch and transceiver.](image)
Since a CW wave still needs to be transmitted, received and demodulated for the reflectenna system to operate, a corresponding transceiver was designed and built. This system will transmit a 10GHz CW carrier signal, which will then be down converted and demodulated in the receiver portion after reflection. A down conversion to 434 MHz was performed which was the operational frequency of the demodulator. A dual polarized high gain reflector antenna was used to transmit and receive, and a switching network was included to eliminate signal inference from the transmitted signal through the dual polarized ports of the reflector antenna. An outline of the required transceiver is shown in Figure 6.2.

![Figure 6.2. Block diagram of reflectenna system with transceiver.](image)

A prototype with micromachined microstrip patch antennas and an off-the-shelf diode SPDT switch was tested using a 10.6 GHz CW signal and 2 kHz square wave baseband and a range of 25 m was achieved. Using computer simulation, it was found that by adding a low power, low noise amplifier (LNA) in the reflectenna, the range can be significantly increased up to 1 km.
6.2 Reflectenna Design

6.2.1 Micromachined Microstrip Patch Antenna

A major component of the reflectenna is the antenna element that is needed to receive the CW signal and retransmit the modulated signal. The requirements for the reflectenna radiating element are small size, relatively good bandwidth and high gain, and ease of fabrication as well integration with the switching circuit. A microstrip patch antenna with micromachined substrate was chosen for this task and designed for X-band operation. The design frequency of about 10 GHz was chosen since this ensures that the antenna size will remain relatively small (in the mm range) and an adequate range without significant signal loss due to the free space loss and atmospheric fading can still be achieved.

It can probably be said that the microstrip patch antenna has been the most widely studied and used antenna for microwave operation. This element is a planar antenna which means it will radiate a plane wave perpendicular to the metal patch when excited with an RF signal. And since microstrip antennas are grounded on the backside, the vast majority of the plane wave will be transmitted in one direction, away from the ground. If a rectangular shape is chosen for the patch design, the frequency at which the patch will radiate can be approximated by

\[ f \approx \frac{c}{2L\sqrt{\varepsilon_r}}, \]

(6.1)

where \( c \) is the speed of light, \( L \) is the length of the radiating dimension of the patch, and \( \varepsilon_r \) is dielectric constant of the microstrip substrate.
From (6.1), it can be seen that if the desired operational frequency is known, the appropriate patch length required to radiate at this frequency will be about half of the guide wavelength of the microstrip circuit, since the wavelength is given by

\[ \lambda = \frac{c}{f \sqrt{\varepsilon_r}}. \] (6.2)

It has been shown that antenna performance of microstrip antennas is dependent on the substrate’s dielectric constant and thickness. Better efficiency is generally achieved with thicker substrates and lower dielectric constants but with a price of requiring larger sized elements. Reducing the thickness and choosing a material with higher dielectric constant will reduce the circuit size but tends to lower efficiency and results in smaller bandwidths. In this design, the same substrate that was used for the MEMS switches introduced previously in this work will be used for the patch antennas; i.e. 400 μm thick, high resistivity silicon. This design choice would allow simultaneous fabrication of the antenna structure and the MEMS switches on the same substrate instead of placing a packaged chip into the circuit. Since microstrip antennas on silicon have an inherently small bandwidth due to the high dielectric constant, micromachining of the substrate will be performed to increase the performance of the antenna [66]. By thinning the substrate through silicon etching techniques, the effective dielectric constant will be reduced which will in turn increase the antenna’s radiation efficiency and bandwidth. A superstrate will also be included in the design, which consists of a 2.5 mm thick dielectric plastic to protect the patch elements and serve as a packaging material for the reflectenna circuit. A cross-sectional view of the patch antenna design is shown in figure 6.3.
The following design parameters were used to obtain the appropriate dimensions of the patch antenna for operation at 10.6 GHz: a grounded micromachined silicon substrate with a 200 \( \mu \text{m} \) thick air gap and a 2.5 mm thick superstrate with a dielectric constant of 4.5. After initially finding the approximate length of the patch antenna using (6.1), simulations were performed using Agilent’s Momentum electromagnetic field simulation software to find the optimum dimensions of the antenna layout. And to further increase the antenna efficiency, an inset microstrip line feed was used to match the input impedance of the patch to that of the line feed (a 650 \( \mu \text{m} \) wide microstrip line, corresponding to a 50 \( \Omega \) characteristic impedance, was used in the simulations). The final antenna dimensions are shown in Figure 6.4. Also included in the figure is a CPW to microstrip transition required to measure the fabricated antenna. The transition uses shorting stubs at the ends of the ground planes to provide an RF short at the design frequency.
Fabrication of the patch antennas to characterize the antenna performance started with a 1 µm thick, thermally evaporated Cr/Ag/Cr/Au layer for the patch material. The silicon substrate was etched using the substrate’s native silicon dioxide layer (~1 µm thick) as an etching mask. The mask was obtained by applying a top and bottom layer of photoresist onto the substrate and patterning the lower resist with the dimensions of the required cavity; the dioxide layer was then etched using a buffered oxide etch. The photoresist layers were then removed and the sample was placed in a TMAH solution to etch the silicon. Finally, the sample was attached to a metal ground plane and the
dielectric lid was placed on top of the antenna during measurement. The fabrication steps are illustrated in Figure 6.5 and given in detail in appendix A.

![Fabrication steps of microstrip patch antenna.](image)

A comparison between the measured and simulated reflection coefficient of the designed patch antenna with CPW to microstrip line feed is shown in Figure 6.6. Return loss measurements were performed from 8 to 12 GHz using a Wiltron 360 vector network analyzer. An approximate 20dB return loss of the overall circuit was achieved at the design frequency. The simulated 10 dB bandwidth of this antenna is roughly 350 MHz and 650 MHz for the simulated and measured antenna, respectively. This corresponds to a 3.3 % bandwidth for the simulated antenna and a 6.1% bandwidth for the measured sample.
Figure 6.6. Measured and simulated return loss of the micromachined patch antenna. Measured data is shown as dashed line, simulated data with a straight line.

The theoretical gain of the patch antenna was also calculated from the ADS Momentum simulation. The gain is shown in Figure 6.7 and illustrates that the patch antenna has a gain of about 8 dB with a half-power bandwidth of about 80°.

Figure 6.7. Calculated gain of the microstrip patch antenna.
6.2.2 Reflectenna Network and Transceiver Design

The reflectenna system consists of a modulating switch and antenna elements to receive the CW signal and passively retransmit the modulated signal. To differentiate between receive and transmit signal at the user end, the signal will be sent and received with linear orthogonal polarizations. This means that on the reflectenna, two patch antennas are used that are oriented at 90° with respect to each other. The two antennas are connected with a CPW transmission line which will hold the modulation switch. Butterfly stubs are used at the CPW-to-microstrip transitions to provide the necessary RF short at the design frequency. Biasing electrodes, which will supply the modulation voltage to the switch, are also connected to the ground and signal lines using butterfly stubs and quarter wavelength long high impedance lines. These components are needed to isolate the DC bias and RF signals. The layout of the reflectenna design is shown in Figure 6.8.
A transceiver was designed to transmit and then demodulate the reflected signal that consists of a high power (1 W) transmitter, a dual polarized and high gain reflector antenna, and a 433 MHz receiver/demodulator (Maxim MAX1473). Since the reflector has a finite isolation between the ports, a switching network had to be added to the transceiver that alternatively switches the receiver and transmitter on and off at a frequency of 10 MHz. In the next section, it will be shown that this is fast enough to yield an effective continuous wave signal to the reflectenna and the demodulator. Before demodulation, the reflected 10 GHz signal is down converted to 433 MHz. A diagram of the transceiver is shown in Figure 6.9.
6.2.3 BASK Modulation and Link Simulation

RF telemetry utilizes modulation techniques to efficiently transmit data by implanting the base-band onto a carrier signal with a much higher frequency. If the base band signal is in digital format, digital modulation techniques such as amplitude shift keying (ASK), where a change in the carrier signal’s amplitude signifies a change in binary state, frequency shift keying (FSK), where each bit corresponds to a different frequency, or phase shift keying (PSK), where a change of phase in the carrier signal represents the different data bits, can be used. Figure 6.10 shows examples of these types of modulation techniques.
Advantages of digital transmission are more efficient modulation, better error control to reduce noise and interference through coding, filtering and equalization, and improved encryption for secure data transmission [67]. Demodulation of the carrier signal is then performed to retrieve the base band signal. The quality of the demodulated signal depends on parameters such as signal-to-noise ratio (SNR) of the receiver, the data rate of transmission, and the signal bandwidth. Table 6.1 shows how each affect the bit error rates (BER) of a demodulated signal.

Figure 6.10. Digital modulation techniques using a CW analog signal.
Table 6.1. Effects of SNR, data rate, and bandwidth on BER.

<table>
<thead>
<tr>
<th>BER</th>
<th>Increase in SNR</th>
<th>Increase in Data Rate</th>
<th>Increase in BW</th>
</tr>
</thead>
</table>

Probably one of the simplest forms of digital modulation to implement is on-off-keying (OOK). This is a type of ASK where a bit is represented either by a detected carrier signal or no detected signal, i.e. the carrier is either on or off. A benefit to this is that the modulator needs to be powered only half of the time to modulate a signal, translating to a 50% reduction in power consumption. OOK transmitters and receivers are also generally easier and less costly to implement than FSK or PSK devices and a properly designed and implemented OOK receiver can be shown to have very high sensitivity as well as good performance in the presence of co-channel interfering signals [67].

The reflectenna’s communication capabilities were simulated using Agilent ADS and its Signal Processing Component libraries. These models include antenna and propagation components and timed components that model the switching network and reflectenna modulation switch. The only part of the transceiver that was not included in the schematic was the 434 MHz demodulator. Figure 6.11 shows the circuit schematic of the component and propagation models. These simulations calculate and display the output of the receiver through the models in the time domain. A detailed explanation of each component is given in Appendix II.
Simulations were performed with a specified range of up to 50 m. Longer distance showed the signal getting lost in the noise floor. The time domain output of the receiver is shown in Figure 6.12. This simulation used a 5 kHz modulation frequency and a distance of 50 m.
By including an LNA in the reflectenna circuit, it was seen through the simulations that the communication range could theoretically be increased by a factor of up to 20. Figure 6.13 shows the time response of the receiver output when the range was increased over 500 m and with a 30 dB LNA included in the reflectenna circuit.
A modified antenna range equation was also used to calculate the theoretical power levels received over the specified range with a given transmitted signal level and is given as follow:

\[ P_r = P_t + 2G_p + 2G_r + L_r + FSL + G_a, \]  \hspace{1cm} (6.3),

where \( P_r \) is the received signal, \( P_t \) is the transmitted signal, \( G_p \) is the gain of the parabolic reflector antenna, \( G_r \) is the gain of the patch elements on the reflectenna, \( G_a \) is the receiver amplifier gain, \( L_r \) is the switch and reflectenna mismatch losses, and \( FSL \) is the free space loss for a 10 GHz signal. All values are given in decibel format.

In the calculations, a 30 dBm transmitted signal with reflector and reflectenna gains of 35 and 8 dB, respectively, was used, as was a 15 dB LNA in the receiver network. The reflectenna loss was assumed to be about 1 dB and the free space loss is dependent on the distance of the RF link and is given by

\[ FSL = 32.4 + 20 \log(f) + 20 \log(d), \]  \hspace{1cm} (6.4)

where \( d \) is the twice the distance of the RF link in km (since the complete roundtrip of the signal needs to be taken into account) and \( f \) the operational frequency in MHz. Results showing the received signal levels with comparisons to corresponding simulation outputs are shown in Table 6.2. Above 50 m, the 30 dB LNA in the reflectenna circuit was included in the calculations and simulations to be able to detect the signal above the noise floor.

<table>
<thead>
<tr>
<th>Received Power (dBm)</th>
<th>25 m</th>
<th>50 m</th>
<th>100 m (with reflectenna LNA)</th>
<th>500 m (with reflectenna LNA)</th>
<th>1 km (with Reflectenna LNA)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Calculated</td>
<td>-30.7</td>
<td>-42.7</td>
<td>-24.8</td>
<td>-52.7</td>
<td>-64.8</td>
</tr>
<tr>
<td>Simulated</td>
<td>-35.6</td>
<td>-47.5</td>
<td>-28.6</td>
<td>-60.8</td>
<td>-73.9</td>
</tr>
</tbody>
</table>

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6.3 Reflectenna Assembly and Measurements

A reflectenna with an IC switch was assembled and tested. The switch used in this demonstration was a SPDT p-i-n diode switch (Ma-Com MA4GAGSW1) with an insertion loss of 1.2 dB and isolation of 40 dB at 10 GHz. Silver epoxy was used to attach the switch to the reflectenna circuitry before the die was wire bonded to connect the transmission lines. The reflectenna was then attached on top of the metal side of a FR4 PCB. This gives the reflectenna circuit the required ground plane and good stability. The PCB also contains electrodes and connector pads to connect the reflectenna’s bias lines to the PCB. Finally, the dielectric lid, with a milled cavity that covers the circuit components, is attached on top of the reflectenna using a silicon sealant. A schematic of the reflectenna assembly and the packaged device are shown in Figure 6.14.

Figure 6.14. Reflectenna assembly (left) and packaged reflectenna (right).
Testing of a fabricated reflectenna without the LNA was performed using a 2 kHz square wave to modulate the reflectenna switch. The 434 MHz BASK demodulator was able to detect the reflected signal up to a range of 25 m. The TTL output of the demodulated square wave is shown in Figure 6.15.

![Figure 6.15. TTL output of the demodulated reflected signal using a 2 KHz square wave for modulation at a range of 25 m.](image)

A reflectenna was also tested with the MEMS shunt switch with single beam design introduced in chapter 3. A TTL level output was observed using a modulation frequency of 100 Hz at a range of 20 m. However, due to reliability issues with the MEMS switch during testing, the data could not be captured in time before the failure of the MEMS switch.
6.4 Conclusion

A wireless and quasi-passive line-of-sight telemetry system with very simple circuitry that modulates and then reflects a carrier signal, called the reflectenna, has been introduced. This device can be applicable to field deployable and remote sensors that require low power and small size for data transmission. The advantage of this system is that high power components, such as power amplifiers, as well as complex radio circuitry do not need to be embedded within the remote sensors, since they are included with the transceiver and data processing unit. A reflectenna was fabricated and tested using a 2 kHz OOK modulated signal up to a 25 m range. Simulations show that the range could be increased significantly by adding a LNA to the reflectenna circuit. This can be greatly reduced using lower power devices.
Chapter 7

Summary and Recommendations

7.1 Summary

The work presented in this dissertation demonstrated the potential of micro and nano fabricated devices as tuning elements for use at micro and millimeter wave applications. Electro-statically actuated MEMS and nano-scale switches were introduced for switching applications and showed good RF performance. This included shunt and series MEMS switches which showed isolations between 25 dB and 45 dB at a design frequency of 10 GHz. Insertion loss was measured below 0.25 dB at the design frequency. A shunt metal-to-metal contact switch was also introduced that uses high resistive metal lines for extrinsic biasing.

The capacitive shunt MEMS switch design was implemented as a building block for a SPDT switch. The switch uses a single bias to route a signal between two ports. This has the advantage of greatly reducing complex biasing networks generally needed for array designs. The SPDT switch consists of a series and a shunt MEMS capacitive switch and showed an isolation of 15 dB in the un-actuated state, with an insertion loss of 1 dB. In the actuated state, isolation and insertion loss was measured at 40 dB and 1 dB, respectively. Modeled results were also shown and good agreement was shown with measured data.
The milling capabilities of a focused ion beam system were used to mill submicron wide gaps in the center conductor of CPW transmission lines. Various capacitive geometries were milled and measured. Gaps as thin as 70 nm were obtained and an equivalent circuit model was developed to extract the capacitances and parasitic effects of the milled devices. Capacitances of around 10 fF were extracted at frequencies up to 65 GHz. This process was also used to fabricate nano-MEMS switches with slanted cuts across a suspended center conductor bridge of a CPW line. The resulting cantilevers were then actuated separately to open and close the signal path. Switching speeds in the 100’s of ns were obtained with an isolation greater than 25 dB below 10 GHz and an insertion loss of less than 0.3 dB at the same frequency.

Finally, a low power RF telemetry system, called the reflectenna, that uses MEMS switching devices for modulation, was introduced. This system can be used for one way, remote data transmission were a signal is reflected back with a modulated signal. This quasi-passive communication device does not require an RF front end for data transmission and uses two orthogonally polarized patch antennas to differentiate between the send and receive signals. A transceiver was designed that alternates the operation of the transmitter and receiver sections and eliminates signal leakage through the transceiver antenna. The patch antennas were designed at 10 GHz and showed good performance due to their micromachined substrate. A reflectenna with diode switch was tested at a range of 25 m with a data rate of 2 kHz.
7.2 Recommendations for Future Work

Continued research into increasing the reliability of MEMS switches can be made. This can include investigation into the use of other metals for the MEMS structures as well as different dielectric material and actuation mechanisms. Other packaging techniques can also be researched, such as using liquid crystal polymer (LCP) as the lid material due to its good electrical properties at microwave frequencies, low water absorption rate, and also low re-flow temperature for bonding applications.

Optimization of the FIB milling process can also be performed to increase the yield of working nano-switches. Re-deposition of milled material needs to be investigated and characterized as well as the effect of undesired doping of the substrate underneath the in the milled area.

Reflectenna research can be continued, which can include testing the maximum range of communication by adding a LNA to the reflectenna circuit, using a single patch with dual polarized feeds to reduce the size of the system, and implementing the reflectenna into a retro-directive array. This type of array gives the reflectenna the capability to retransmit the incoming CW signal back in the direction of the transceiver without needing prior knowledge of its location.
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Appendices
Appendix 1

Photolithography and Micromachining Procedures

1.1 Introduction

This Appendix gives the procedures for the fabrication of the MEMS switches that were investigated in this chapter. First the procedure for a negative tone (lift-off) process is given, followed by a procedure for positive tone processing. Finally, the procedure used to fabricate the MEMS tunable inductor is presented.

1.2 Negative Tone Process (Lift-Off)

1) Clean sample with Acetone and IPA.

2) Bake on a hot plate at 150 °C for 5 minutes to remove excess moisture from the sample. Let cool for 10 minutes but don’t place the sample on a cold surface!

3) Spin coat the sample Futurex NR 3000 PY photoresist at 3000 rpm for 30 seconds.

4) Soft bake on a hot plate at 155 °C for 60 seconds.

5) Expose the sample for 17 seconds in a mask aligner. Properly align the sample first in case more than one layer is used.

6) Hard bake on hot plate at 110 °C for 60 seconds.

7) Develop the photoresist for 25 seconds in Futurex RD6 developer batch and rinse with DI-H₂O. Dry with Nitrogen.
8) Deposit desired material by high vacuum deposition technique of choice.

Place in a beaker of Acetone or heated Futurex RR4 resist remover until the unwanted metal layer has lifted off the sample surface.

1.3 Positive Tone Process

1) Follow steps one and two of the preceding procedure.

2) Spin coat the sample with HMDS, followed by Shipley 1827 positive tone photoresist. Spin at 3000 rpm for 30 seconds.

3) Soft bake on a hot plate at 110 °C for 60 seconds.

4) Expose sample in a mask aligner for 13 second.

5) Hard bake on a hot plate at 130 °C for 60 seconds.

6) Develop the photoresist in Microchem MF-319 developer for 60 seconds.

1.4 Capacitive Shunt MEMS Switch Fabrication Procedure

1) Start with negative tone process to fabricate thin center conductor section of the CPW lines using thermally evaporated Cr/Au (100/2000 Å)

2) Deposit Silicon Nitride layer using LPCVD process (performed at Center for Ocean Technology’s clean room facility) to a thickness of 2800 Å.
Appendix 1 (Continued)

3) Apply Shipley 1827 photoresist using positive tone process to pattern tech mask for SiNi layer, but skip the hard bake. Etch back SiNi layer using a reactive ion etcher with the following recipe: use a 40:1 CF2/O2 gas ration with a partial pressure of 100 mT; 100 W RF power; for 15 minutes.

4) Remove resist by flood exposing the sample with UV light for 60 s and rinse in MF 319.

5) Use negative tone process to fabricate thin metal top plate using thermally evaporated Cr/Au layer (100/1000 Å).

6) Spin on PMMA A9 layer at 3000 rpm and 45 seconds. Bake sample on hotplate at 180ºC for 75 seconds. Use positive tone process to mask PMMA for plasma etching. Skip hard bake step.

7) Etch PMMA layer in reactive ion etcher using the following recipe: O2 gas with partial pressure of 55 mT; 250 W of RF power; for 8 minutes.

8) Remove resist by flood exposing the sample with UV light for 60 s and rinse in MF 319.

9) Flood evaporate a 100/1000 Å Ti/Au layer using E-beam deposition on the sample surface for electro-plating seed layer fabrication.
Appendix 1 (Continued)

10) Use positive tone process for electroplating mask during the plating step of the beam and CPW lines. Electroplate using Techniq E-25 gold plating solution to a thickness of about 1 µm. Plate bath is kept at 60°C with plating times of 10 minutes each until desired thickness is reached. Plating current is set between 0.18 to 0.25 mA and is adjusted to approximate plating rate of 0.5 Å per 10 minutes of plating time.

11) Remove resist by flood exposing the sample with UV light for 60 s and rinse in MF 319. Use positive tone process to protect beam and CPW lines during seed layer removal. Skip hard bake step.

12) Etch seed layer using Transene gold etchant for about 30 seconds for the gold layer. Etch Ti layer using a 10:1 DI H2O:HF mixture for 5 to 10 seconds. Etch PMMA layer in reactive ion etcher using the following recipe: O2 gas with partial pressure of 55 mT; 250 W of RF power; for 12 minutes.

13) Place samples in heated (90°C) Microchem 1165 resist stripper for 5 minutes, then in second bath of 1165 for at least 6 hours to remove PMMA sacrificial underneath MEMS structure. Rinse in 3 batch if DI H2O and then 3 baths of Methanol.

14) Place sample in critical point dryer to finalize the release process.
Appendix 1 (Continued)

1.5 Metal-to-Metal Contact Shunt MEMS Switch Fabrication Procedure

1) Start with negative tone process to fabricate SiCR line using E-beam evaporation (1000 Å).

2) Use negative tone process to fabricate thin center conductor section of the CPW lines using thermally evaporated Cr/Au (100 Å/2000 Å).

3) Follow steps 2-4 and 6-14 of 1.4.

1.6 MEMS Switch Packaging Using SU-8

1) Follow steps 1-2 of 1.2 to clean quartz wafer.

2) Spin SU-8 50 onto sample at 500 rpm for 10 seconds with a ramp set to 100 rpm/second. Then spin at 2000 rpm for 30 seconds with ramp set to 300 rpm/second.

3) Pre-bake sample on hot plate for 6 minutes at 65°C. Then increase temperature to 95°C with sample still on hot plate. Once temperature reaches 90°C, leave sample to bake for 20 minutes.

4) Let sample cool down for at least 5 minutes. Expose bonding rings, using mask, in mask aligner for 50 seconds. Then increase temperature to 95°C with sample still on hot plate. Once temperature reaches 90°C, leave sample to bake for 20 minutes.
Appendix 1 (Continued)

5) Post-bake sample on hot plate for 1 minutes at 65°C. Then increase
temperature to 95°C with sample still on hot plate. Once temperature reaches
90°C, leave sample to bake for 5 minutes.

6) Place sample in SU-8 Developer and agitate sample for 6 minutes. Rinse with
DI H₂0 and blow dry with nitrogen gas.

7) Dice quartz sample to obtain individual lids

8) Bond lids to MEMS switches using flip-chip bonding. Bonding temperature
is 150°C with a bonding force of 10 N for 10 minutes.

1.7 FIB Milled Capacitor Fabrication Procedure

1) Follow steps 1-2 of 1.2 to clean silicon wafer.

2) Use negative tone process to fabricate thin metal top plate using thermally
evaporated Cr/Au layer (100/4000 Å).

3) Use negative tone process to fabricate thin metal top plate using thermally
evaporated Cr/Ag/Cr/Au layer (100/8000/100/1000 Å).

4) Spin on PMMA A4 layer at 6000 rpm and 45 seconds. Bake sample on
hotplate at 160°C for 75 seconds.

5) Thermal evaporate 50 Å of Cr on top of PMMA.

6) Mill FIB cuts using a beam strength of 30 kV and a beam current of 10 pA.
Set milling depth to 1 μm.

7) Place sample in heated (90°C) Shipley 1165 until PMMA layer is removed.
1.8 FIB Milled Switch Fabrication Procedure

1) Follow steps 1-2 of 1.2 to clean silicon wafer.

2) Use negative tone process to fabricate SiCR line using E-beam evaporation (1000 Å).

3) Follow steps 2-4 from 1.3 for nitride layer fabrication.

4) Spin on PMMA A7 layer at 4000 rpm and 45 seconds. Bake sample on hotplate at 160ºC for 75 seconds. Use positive tone process to mask PMMA for plasma etching. Skip hard bake step.

5) Etch PMMA layer in reactive ion etcher using the following recipe: O2 gas with partial pressure of 55 mT; 250 W of RF power; for 5 minutes.

6) Remove resist by flood exposing the sample with UV light for 60 s and rinse in MF 319.

7) Follow steps 9-14 form 1.3 for beam and CPW line plating and release process.

8) Mill FIB cuts using a beam strength of 30 kV and a beam current of 100 pA. Set milling depth to 6 µm. Pause milling very few seconds to check alignment of cut. Re-align if necessary.
1.9 Patch Antenna Fabrication Procedure

1) Start with negative tone process to fabricate CPW line and microstrip patch antenna using thermally evaporated Cr/Ag/Cr/Ag (100/8000/100 /1000 Å)

2) Spin on 1827 photoresist at 3000 rpm for 30 seconds to cover top surface of sample. Then use positive tone process to pattern underside of the wafer for etch mask fabrication. Etch exposed SiO2 layer in BOW for 20 minutes. Remove resist by rinsing with Acetone and Methanol.

3) Etch exposed Si in heated TMAH (75°C) for about 6 to 8 hours. Check etch depth periodically until a depth of about 180 μm is achieved. Then check return loss of antenna and keep etching until resonance at the desired frequency is achieved.

4) Bond plastic lid on antenna substrate using tape or glue.
Appendix 2

ADS Signal Processing Library

2.1 Introduction

This Appendix gives a description of Agilent’s ADS digital signal library models used during the reflectenna link simulations.

2.2 Library Components

\textit{N\_Tones} - RF tones generator

Generates an RF (complex envelope) timed signal output and used to create the 10 GHz CW signal with 0 dBm output.

\textit{AntBase} - Base Station Stationary Antenna Model

Linear polarized antenna where the gain, position in XY coordinates, and height above ground can be specified. For the reflectenna, the gain of each patch was set to 8 dBi. For the parabolic antenna, the gain was set to 35 dBi for the send and receive cases.
Appendix 2 (Continued)

*PropFlatEarth* - Direct and Reflected Ray Propagation Model

Models the sum of a direct and reflected ray propagation channel model based on polarization and flat earth properties. Polarization type (linear vertical or horizontal), earth’s average permittivity, and earth’s average conductivity can be specified. During simulation, permittivity was set to 25 and conductivity to 0.02.

*RxAntTempK* - Add noise to input signal due to the receiver antenna temperature.

Models the receiver antenna noise temperature by adding white Gaussian noise to the input signal. Input and output resistance as well as the noise temperature can be specified. Noise temperature was set to 290 K
Appendix 2 (Continued)

*GainRF* - Complex gain with gain compression.

![GainRF diagram](image)

Used to model an amplifier with nonlinear gain compression. Input resistance, output resistance, input noise figure, and complex voltage gain were specified during the simulations. For the simulations, the gain was set to 30 dB with a noise figure of 0 dB.

*SplitterRF* - RF signal splitter.

![SplitterRF diagram](image)

Input signal in two output signals, each with the same signal level as the input signal.
Appendix 2 (Continued)

*SummerRF* - RF signal summer.

This component models a summer with two inputs and was used to combine the reflectenna return signal with the leaked signal form the transmitter.

*MatchedLoss* - Power loss referenced to matched source and load resistors.

Was used to model the isolation of the dual polarized antenna feeds of the reflector antenna. Loss in dB can be specified. The loss was set to 55 dB to correspond to the isolation of the parabolic reflector antenna’s feed lines.

*SwitchSPST* - RF single pole, single throw switch.

This component can be used to model a non-ideal switch. The non-idealities modeled are insertion losses, imperfect isolation, and non-zero delay switching

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times. For the reflectenna switch, the loss was set to 1 dB, isolation to 25 dB, and rise and fall times of the switch to 10 µs each. For the transceiver witches, the insertion loss was set to 2.5 dB, the isolation to 90 dB, and rise and fall times to 30 ns.

Clock - Clock generator

This source generates a baseband timed signal output and the duty cycle can be set to drive the switching cycles of the modeled switches. For the reflectenna switch, the period was set to 10 µs, the delay to 0 s, and the duty cycle to 50%. For the transmitter switch, the period was set to 100 ns, the delay to 0 s, and a duty cycle of 40%. For the receiver switch, the period was set to 100 ns, the delay to 50 ns, and a duty cycle of 40%.
Appendix 2 (Continued)

TimedSink – Timed Data Collector

This component collects timed (baseband or complex envelope) data from the devices output. Output of the modeled transceiver was viewed in the time domain using this component.

$DF$ - Data Flow Controller

DF controller is used to control the flow of mixed numeric and timed signals for all digital signal processing simulations. It was used to set the length of the simulation as a function of time. In the simulations, the time was set to run for 100 µs.
About the Author

Thomas Ketterl received a Bachelor of Science in Ocean Engineering degree from the Florida Atlantic University in 1994. In 2000, he received the Master of Science in Electrical Engineering degree at the University of South Florida. While in the Ph.D. program, he was also employed fulltime developing MEMS switches and communication systems for the Center for Ocean Technology at the University of South Florida.