Chapter 6

Various applications of RF MEMS

6.1 Introduction

As seen in chapter 4, the idea of using a virtual air substrate for reducing loss and improving the quality factor is more appropriate for devices having a length higher than the probing area. In such cases, the line capacitance due to the probe cannot fully dominate the effect of the virtual air substrate. For most applications of RF MEMS switches, the length of the device is in the range of 1-10mm. An attempt has therefore been made to utilize the advantage of having a virtual air substrate in tunable filter and phase shifter applications.

6.2 Application in tunable filter

The development of tunable filters is of great commercial and military interest [1]. Tunable filters can help in substantially reducing the size of analog front end systems from multi band applications and can be used to dynamically reject large signal interferers. The design of tunable filter requires easy control of the dynamically tuned resonance frequency with a high tuning range. In general, a 10-15% tuning range is suitable for such applications.

6.2.1 Conventional RF MEMS Filters to Metamaterial Filters

Artificial materials with unique electromagnetic properties like negative permeability, negative permittivity and negative refractive index that are not readily observed in nature were theoretically hypothesized by V. Veselago as metamaterials [2]. Several theoretical and experimental verifications were performed for the existence of such materials over years [3-5]. These researches
explored different regimes of applications which usually range from superlenses, invisibility cloaking, solar cells, zeroth order resonators, oscillators, antennas etc. including filters for microwave and millimeter wave applications. Among these, planar metamaterial based filters have been of major interest because of their compact size and their ability to exhibit a Left Handed (LH) behavior around the resonance frequency as compared to the conventional microwave filters [6].

### 6.2.2 Difference between metamaterial filter and conventional filter

Let us consider a Composite Right Left Handed (CRLH) transmission line as illustrated in Figure 4.5. Considered over all frequencies from $\omega = 0$ to $\omega \rightarrow \infty$, the CRLH network is obviously band-pass filter. When $\omega \rightarrow 0$, $|Z|\rightarrow 1/(\omega C_L)\rightarrow \infty$ and $|Y|\rightarrow 1/(\omega L_L)\rightarrow \infty$, and therefore we have a stop band due to the high-pass nature of the LH elements; when $\omega \rightarrow \infty$, $|Z|\rightarrow \omega L_R\rightarrow \infty$ and $|Y|\rightarrow \omega C_R\rightarrow \infty$, and therefore we have a stop band due to the low-pass nature of the RH elements. Between these two stop bands, a perfectly matched pass band can exist under the balanced condition. Thus, an LC network implementation of a CRLH TL is a band-pass filter, in the sense that it has LH high pass low-frequency and RH low pass high-frequency stop bands [2].

![CRLH transmission line](image)

**Figure 6.1: CRLH transmission line**
Because CRLH transmission lines may look at the first sight similar to conventional filters, let us now point out the essential distinctions existing between these two types of structures:

1) A Left Handed (LH) metamaterial structure exhibits a specific phase response, leading to LH transmission at lower frequencies and RH transmission at higher frequencies. Conventional filters are generally designed to meet magnitude specifications and do not exhibit a LH range.

2) A metamaterial structure is constituted of unit cells satisfying the homogeneity condition $|\Delta \phi| < \pi/2$. Conventional filters do not generally satisfy this condition; they may have node-to-node phase shifts larger than $\pi/2$. In fact, a metamaterial with $\Delta \phi << \pi/2$, or $\Delta z << \lambda_g/4$, would be an ideal (perfectly homogeneous) material, in the same manner as conventional dielectrics, made of molecules with dimensions many orders of magnitude smaller than wavelength. The only reason while $\Delta z$ may be close to $\lambda_g/4$ at some frequencies in today’s metamaterials is the unavailability of more effective structures.

3) A metamaterial structure can be 2D or 3D and behave as bulk media, whereas conventional filters are 1D and behave as electric circuits.

4) A metamaterial structure can be made of identical cells, whereas in a conventional filter each cell has generally different LC values to match the specifications of a given prototype.

6.2.3 Planar Metamaterial

The planar version of metamaterial is realized over a transmission line and is found to be compatible with modern Microwave Integrated Circuits (MICs). Caloz and Itoh put forward the first practical Left Handed Transmission Line
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(LHTL) which is a microstrip based structure in 1D [7] and 2D [8]. Planar metamaterial can be categorized into two ways viz.

(a) **1D metamaterial or CL loaded metamaterial transmission line**

These planar or transmission line (TL) metamaterials can be realized by loading the host transmission line by interdigital series capacitors and shunt stubs shorted to ground plane shown in figure 6.2. They are implemented by lumped circuit elements. The equivalent circuit of a standard Right Handed (RH) line represented by $L_RC_R$ and that of a Left Handed (LH) line represented by $L_LC_L$ are presented in a Composite Right Left Handed transmission line (CRLH) as shown in figure 6.1.

(b) **2D metamaterial or resonant type metamaterial transmission line**

These TL metamaterials are implemented by metamaterial structures. These structures have inherent resonance characteristics and are significantly smaller in size (generally less than one tenth of a wavelength) than conventional resonating structures. So in recent years, these structures have attracted a great
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deal of interest for the design of microwave filters with compact dimensions and exceptional features [9-12].

6.2.4 Metamaterial Structures

The novel structures like Split Ring Resonators (SRRs), Complementary SRRs (CSRRs) are used for the implementation of resonant type metamaterials transmission lines as well as 3D bulk metamaterials in some case.

(a) SRR

SRR was first proposed by J. B. Pendry in 2000 [13]. The SRR essentially consists of two concentric metallic split rings as shown in figure 6.3(a). When it is excited by a time-varying external magnetic field directed along the z-axis, the cuts on each ring (which are placed on opposite sides of the SRR) force the electric current to flow from one ring to another across the slots between them, taking the form of a strong displacement current. The slots between the rings therefore behave as a distributed capacitance, and the whole SRR has the equivalent circuit shown in figure 6.3(b) [9]. In this circuit, $C_0$ indicates the total capacitance between the rings, that is,

$$C_0 = 2\pi r_0 C_{pul}$$  \hspace{1cm} (6.1)

where $C_{pul}$ is the per-unit-length edge capacitance.

$C_s$ is the series capacitance of the upper and lower halves of the SRR, that is,

$$C_s = C_0 / 4$$  \hspace{1cm} (6.2)

The inductance, $L_s$, can be approximated by that of average radius $r_o$ and widths $c$ and $d$ where

$$r_o = r_{ext} - c - d / 2$$  \hspace{1cm} (6.3)
The two important characteristics of SRR are its small electrical size and exhibiting negative effective magnetic permeability $\mu_{\text{eff}}$ around the resonance frequency.

(b) CSRR

On the basis of SRR, numerous sub-wavelength resonators with similar topologies have been proposed. One of them is the complementary counterpart of the SRR, the complementary split ring resonator (CSRR). The complementary of a planar metallic structure is obtained by placing the metal parts of the original structure with apertures, and the apertures with metal plates [Figure 6.4(a)]. Thus, whereas the SRR can be mainly considered as a resonant magnetic dipole with negative permeability that can be excited by an axial magnetic field, the CSRR essentially behaves as an electric dipole with negative permittivity (with the same frequency of resonance) that can be excited by an axial electric field. It is obtained from the idea of applying Babinet’s principle to the SRR. According to this principle, in such complementary structures, the electric, magnetic fields, currents interchange their roles. The equivalent circuit of the CSRR can thus be represented as in figure 6.4(b). The values of $L_c$ and $C_c$ are given by
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\[ L_c = 2\pi rL_{pul} \quad (6.4) \]

\[ C_c = 2\pi rC_{pul} \quad (6.5) \]

where \( L_{pul} \) is the per-unit-length inductance of the CPWs connecting the inner disk to the ground, \( r=(r_{ext}-c-d/2) \) is the mean radius of the CSRR and \( C_{pul} \) is the per unit length capacitance along the slot between the rings [9].

![Figure 6.4: (a) Representation of CSRR. (b) Equivalent circuit of CSRR](image)

The formulas can be modified accordingly for different shapes and geometries of SRRs/CSRRs.

### 6.2.5 From 1D metamaterial to 2D metamaterial

A 1D metamaterial represented by a CRLH transmission line [Figure 6.5(a)] is taken up for circuit analysis in order to get the expressions of effective permittivity \( (\varepsilon(\omega)) \) and effective permeability \( (\mu(\omega)) \) both as function of frequency.

\[ \varepsilon(\omega) = \frac{Y(\omega)}{j\omega} = C_R - \frac{1}{\omega^2 L_L} \quad (6.6) \]
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\[
\mu(\omega) = \frac{Z(\omega)}{j\omega} = L_R - \frac{1}{\omega^2 C_L} \tag{6.7}
\]

For 2D implementation of metamaterials, either SRRs (negative permeability) and stubs (negative permittivity) or CSRRs (negative permittivity) and gaps (negative permeability) can be used [14] for maintaining the properties of left handed material with backward wave propagation [9]. The equivalent circuit is presented in figure 6.5 (b) for the illustration in each case.

Figure 6.5(a): Equivalent circuit of a transmission line loaded with SRRs and stubs
Figure 6.5(b): Equivalent circuit of a transmission line loaded with CSRRs and gaps

It is interesting to note that both the equivalent circuits reported in each case will result in band pass filter which may be verified by basic circuit analysis.

The scope of the thesis is mainly to be within the domain of band-stop/tunable band-stop filter. However, a band-pass filter using two band stop actions at different frequencies will also be realized. It is evident that to design band-stop filter we have to use SRRs (without stubs)/ CSRRs (without gaps). It should be noted that SRR loaded transmission line will provide a negative permeability around the resonance frequency while CSRR loaded transmission line will result in negative permittivity around the resonance frequency.

In the designs put forward as the part of the thesis, we have mainly focused on metamaterial structures (generally complementary structures) as frequency selective elements having the value of only one of the effective
parameters negative. So these structures will not result in negative index of refraction and backward wave propagation.

The above discussion leads to a CLR model of a metamaterial structure (CSRR) loaded transmission line as shown in figure 6.6. The CLR tank resonator representing such a metamaterial structure will occur in series with the CPW line capacitance. This is because the metamaterial structure is *etched onto* the signal or the ground planes.

![Equivalent Circuit of complementary metamaterial structure loaded transmission line](image)

**Figure 6.6:** Equivalent Circuit of complementary metamaterial structure loaded transmission line

### 6.2.6 Metamaterial Filters

In order to use metamaterial technology in filters working at microwave frequency, compatibility with planar circuit technology is needed [15]. Metamaterials find its application in filters whether it is realized by CRLH in case of 1D metamaterials or by structures in case of 2D metamaterials. Two approaches – CL loaded transmission line approach and resonant type transmission line...
approach had already been discussed while realizing planar metamaterials. Now this concept is again used for application of metamaterials and to employ planar metamaterial filters.

**Types of metamaterial filter**

Types of metamaterial planar filters based on two approaches are

(a) *Filter realized by CL loaded transmission line approach*

The transmission line loaded with capacitors and inductors behave as a filter having compatibility to that of the conventional filter. But the capacitive effect designed by interdigital capacitor provides a series impedance whereas the inductors in shunt realized with via-hole provides effective shunt impedance. Typical topologies of CL loaded TL model metamaterial filter is shown in figures 6.7 and 6.8 [16-17].

![Figure 6.7: Microstrip Line loaded with grounded stubs and interdigital capacitors](image-url)
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Several models based on this approach were proposed and were validated afterwards [18-20]. The low pass nature of the LH elements and the high pass nature of the RH elements result in a specific filtering performance of a CL loaded 1D metamaterial unlike a conventional filter [21].

(b) Filter realized by resonant type transmission line approach

Different resonant metamaterial structures which had been proposed are employed on transmission lines in order to have a filtering action [Figure 6.9 and Figure 6.10]. This concept represents the integration of a metamaterial structure on the transmission line. The inherent resonance characteristics present in the metamaterial structures lends itself well to the applications of filters [16]. These metamaterial structures acting as resonators can be used in periodic fashion to obtain filtering action.
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6.2.7 Tunability of metamaterial filters with RF MEMS

The usefulness of a metamaterial based filter can be tremendously increased if the metamaterial’s response characteristics can be dynamically tuned.
Thus, one of the major functionality additions is to incorporate tunability in the metamaterial filter. In general, tunable microwave components utilize different physical effects and materials to achieve desired tuning range. Methods may range from using PN junction diode, Schottky junction, Hetero- structure Barrier Varactor (HBV) and transistors in the semiconductor domain, MEMS (Micro Electro Mechanical System) varactor in the mechanical domain and ferromagnetic/ ferrites, ferroelectric materials and liquid crystals in the material domain. Among these, the semiconductor varactors have low Q factor and large leakage current, the magnetic components typically require large control power and are bulky, the liquid crystals although having a low loss offer a very slow tuning speed. But, MEM varactors have low losses and are suitable for use in millimeter wave applications. Thus the use of MEMS technology for achieving the desired tunability seems to be an optimum choice [15, 21, 22]. The choice of silicon as a substrate is justified keeping the MEMS CMOS integration in mind.

6.2.8 Tunable band stop filter using metamaterial structure and MEMS bridges: Device Structure

Figure 6.11 shows the top view of the CUSR embedded CPW with MEMS bridge. As seen a CUSR consists of two U- shaped concentric structures etched out of a metallization (signal line of CPW in this case). A CPW of gold with a configuration corresponding to W/S/W is housed on an oxidized high resistivity silicon substrate (bottom wafer). The CUSR is patterned on the central line of CPW. The length of the CUSR, the width of the CUSR, the width of the rings and the spacing between the rings are ‘L’, ‘T’, ‘c’ and ‘d’ respectively. It is to be noted that the entire silicon is removed from underneath/ above the CPW central line and a portion of the ground planes. As the CPW is an edge coupled device the removal of the silicon from the said volume results in virtual air substrate.
Figure 6.12 shows the two wafers prior to bonding and figure 6.13 shows the cross sectional view of the bonded wafer. The top wafer is patterned with two MEMS bridges realized using gold. As will be seen later, two MEMS bridges are used to improve the quality factor of the band-stop filter. The MEMS bridges are suspended at an air gap of $g_0$ and have length, width and thickness given by $l$, $b$ and $t$ respectively. The removal of silicon in order to create virtual air substrate takes place over a width of $W_{win}$ and length $L_{win}$ for both the substrates. The silicon underneath the remaining portion of the ground planes of the bottom substrate provides mechanical stability to the device. An added layer of thick silicon nitride may also be deposited on the bottom wafer to provide mechanical strength to the central line. A thin silicon dioxide layer is used to prevent the shorting of the
CUSR with the MEMS bridges and Au-AuGe-Au eutectic is deposited on the ground planes of the CPW for bonding the two substrates. Further, a window is opened in the top wafer for probing the device for characterization. The fabrication methodology is detailed in Section 6.2.13.

Figure 6.12: Two wafers prior to bonding
A metamaterial filter essentially consists of a CLR tank resonator which provides a shunt loading on the transmission line. As the metamaterial structure is
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*etched onto* the signal line of the CPW, the CLR tank appears in series with the transmission line shunt capacitance of the CPW [Figure 6.14]. It is well known that a MEMS shunt varactor is used to vary the shunt loading of any given transmission line depending upon its state of actuation [1]. Thus, when a MEMS varactor is placed in shunt across a transmission line embedded with a metamaterial structure, the effective shunt loading of the device changes. This change leads to a shift in the resonance frequency for different states of actuation.

![Figure 6.14: Equivalent Circuit of the tunable band-stop filter](image)

From figure 6.14, it is seen that the contribution of the metamaterial structure to the shunt impedance is fixed. The bridge capacitance varies depending upon its height over the transmission line and alters the total shunt loading of the
transmission line. The impedance in the shunt path after basic circuit analysis results in equation

\[ Z_{sh} = \omega^2 L_c^2 + \left[ R_c \left\{ 1 - \omega^2 L_c (C_{sh} + 2C_{bu/bd}) \right\} \right]^2 \]  \hspace{1cm} (6.8)\

where \( C_{sh} = C_t + C_c \) and \( C_{bu/bd} \) refer to bridge capacitance for up and down states respectively.

Again, as seen in figure 6.14, the tunability is obtained by the bridge capacitance \( C_b \). Therefore to have a high tuning range, \( C_b \) should be dominant over the combined effect of the line capacitance \( (C_t) \) and the resonator capacitance \( (C_c) \). Therefore, the tunability is achieved with the help of two MEMS varactors. At this point, it is necessary to note that the dimensions of the MEMS bridge is mainly determined by the static (actuation voltage) and dynamic (switching time) switching requirements of the ultimate device. Standard aspect ratios of the bridge are in the range of 2-3 providing an upstate capacitance in the range of 30-140fF for bridge heights of 2-3µm over the CPW [1]. Figure 6.15 shows the fractional change of resonance frequency \( (f_{\text{unloaded}}/f_{\text{loaded up}}) \) with capacitive loading of the bridge (up-state) for different values of \( C_{sh} \). Here \( f_{\text{unloaded}} \) denotes the resonance frequency of the metamaterial structure without MEMS varactors, while \( f_{\text{loaded up}} \) refers to the resonance frequency under varactor loaded condition for up-state (maximum \( g_0=3\mu m \)). Further, \( f_{\text{unloaded}} \) and \( f_{\text{loaded up}} \) are given by

\[ f_{\text{unloaded}} = \frac{1}{2\pi \sqrt{L_c C_{sh}}} \] \hspace{1cm} (6.9)\

\[ f_{\text{loaded up}} = \frac{1}{2\pi \sqrt{L_c (C_{sh} + C_{bu})}} \] \hspace{1cm} (6.10)
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Figure 6.15: Fractional change in resonance frequency v/s Bridge UP Capacitance for different values of ‘$\text{C}_{\text{sh}}$’

It is clearly seen that for higher values of ‘$\text{C}_{\text{sh}}$’ the capacitive loading of the bridge will have little effect on the metamaterial loaded transmission line. In order to reduce the contribution of $C_i$ and $C_c$, a substrate with a low value of permittivity (ideally air) has to be chosen. FEM simulations indicate that 15-20GHz down shift in resonance frequency occurs when a capacitance loading of 30-140fF takes place on a metamaterial embedded CPW on low permittivity substrate validating the drawn observation. Therefore, to obtain a tunable band-stop filter in the Ka band with two MEMS varactors, the bridge unloaded metamaterial structure based filter on low permittivity substrate, should resonate at 50-52GHz. Variation of tuning range {$(f_{\text{loaded,up}}-f_{\text{loaded,down}})/f_{\text{loaded,up}} \times 100\%$} with change in capacitive loading ($\Delta C = C_{\text{bd}}-C_{\text{bu}}$) is plotted in figure 6.16. Here,
\( f_{\text{loaded\_down}} \) refers to the resonance frequency under varactor loaded condition for down-state (minimum \( g_0=0.5\mu m \)) and is given by

\[
\frac{1}{2\pi \sqrt{L_c (C_{sh} + C_{bu})}}
\]

It is seen that to obtain high tuning range with minimum capacitance change the combined effect of \( C_t, C_c \) and \( C_{bu} \) should be low. As the value \( C_{bu} \) will also determine the value of \( \Delta C \), a higher tuning range is mainly obtained for low values of \( C_{sh} \). This further justifies the earlier contention to use low permittivity substrates.

![Figure 6.16: Tuning range v/s Change in Capacitive Loading for different values of \('C_{sh}+C_{bu}'\)](image)

6.2.10 Device Design

(a) Design of CPW
The choice of the unloaded CPW impedance is also very critical as 50 ohms characteristic impedance is to be achieved after loading with the CUSR and MEMS bridge(s). As evident from the equivalent circuit, a CUSR loading will increase the line impedance, and the bridge loading will decrease the same. As the CUSR is on a low permittivity substrate (approximately air in this case), the combined capacitive effect of $C_t$ and $C_c$ is much smaller compared to the bridge capacitance $C_b$. In this case, W/S/W corresponds to 10/200/10µm on micromachined silicon substrate resulting in 64 ohms (as calculated by Agilent Line Calc). Therefore, the unloaded CPW characteristic impedance is kept on the higher side of 50 ohms.

To obtain a virtual air substrate, bulk micromachining of silicon is proposed. As the fabrication methodology involves two bonded silicon wafers, micromachining is to be performed for both the wafers.
For calculating the dimensions of the window for micromachining, conformal mapping has been performed on simplified cross sectional view of the proposed device suitable for theoretical formulation [Figure 6.17]. The capacitance offered by the virtual air substrate is strongly dependent on the width of the micromachined window ($W_{\text{win}}$). Figure 6.18 shows the capacitance offered by the virtual air substrate as a function of $W_{\text{win}}$. The dotted line in the figure corresponds to the capacitance offered for a total air substrate. It is seen that for $W_{\text{win}}$ corresponding to 600µm, the virtual air substrate offers a capacitance approximately equal to that offered by the air substrate. The capacitance values are obtained from the equations discussed in chapter 4. The values of the different parameters of the CPW are listed in table 6.1.
Table 6.1: Parameters of CPW

<table>
<thead>
<tr>
<th>C_{air} (pF/m)</th>
<th>C_t (fF)</th>
<th>\varepsilon_{eff}</th>
<th>Z_0 (Ohms)</th>
<th>L_t (pH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>51.92</td>
<td>51.9</td>
<td>1.03</td>
<td>64</td>
<td>213.9</td>
</tr>
</tbody>
</table>

(b) Design of metamaterial structure

In case of CSRRs, the stop band response suffers from a spurious effect at resonance due to the difference in the electrical length of the inner and outer arm of the CSRRs. This is overcome by the CUSRs by equalizing the electrical length of both the resonator arms. After choosing the suitable CPW configuration, the centre frequency (f_{unloaded}) of the metamaterial filter consisting of CUSR has been chosen to be 51GHz [Figure 6.15]. This will result in f_{loaded_up} corresponding to a frequency in Ka band.

The length of CUSR (L) depends on the unit cell length (p) and is found to be less than 1459\mu m as per equation 14 [10].

\[ p = \frac{\lambda_g}{4} \quad (6.12) \]

where \(\lambda_g\) is the guided wavelength and is given by

\[ \lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_{eff}}} \quad (6.13) \]

and \(\varepsilon_{eff}\) is the effective permittivity and is 1.03 for the micromachined silicon substrate.
(c) **Design of MEMS bridge**

In this case, two MEMS bridges, 400µm apart (center to center), are considered in order to have a high Q factor and smaller bandwidth of the filter. The dimensions of the MEMS bridge are presented in Table 6.2. It is seen that MEMS bridge with such dimensions will have a balance between the actuation voltage required and the switching time [1].

**Table 6.2: Optimum dimensions of the MEMS bridge**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values (µm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length of MEMS bridge (l)</td>
<td>220</td>
</tr>
<tr>
<td>Width of MEMS bridge (b)</td>
<td>100</td>
</tr>
<tr>
<td>Initial height of MEMS bridge (g₀)</td>
<td>3</td>
</tr>
</tbody>
</table>

### 6.2.11 Simulations

**RF Simulation**

The length as well as other parameters of the CUSR (width of rings, spacing between rings) are optimized by Ansoft HFSS® to get the required resonance frequency (shown in Figures 6.19, 6.20 and 6.21). For larger lengths of the CUSR both the inductive and capacitive loading of the CUSR increases leading to a lowering of the resonance frequency [Figure 6.19]. For larger values of 'c' the capacitive and inductive loading of the CUSR decreases leading to an increase in the resonance frequency [Figure 6.20]. Similarly, for larger values of 'd' capacitive and inductive loading of the CUSR decreases leading to an increase in the resonance frequency [Figure 6.21]. As the width of the CUSR is mainly determined by the width of the CPW central line/ ground plane(s), parametric
analysis of the width (T) has not been performed. The optimum dimensions are tabulated below in table 6.3.

Table 6.3: Optimum dimensions of the CUSR embedded CPW

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values (µm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>W/S/W of CPW</td>
<td>10/200/10</td>
</tr>
<tr>
<td>Length of CUSR (L)</td>
<td>1000</td>
</tr>
<tr>
<td>Width of CUSR (T)</td>
<td>180</td>
</tr>
<tr>
<td>Width of Rings (c)</td>
<td>10</td>
</tr>
<tr>
<td>Spacing between rings (d)</td>
<td>10</td>
</tr>
</tbody>
</table>

The proposed design is simulated in Ansoft HFSS® for different heights of the MEMS bridge and the results are plotted in figures 6.22 and 6.23. One may note that for smaller values of ‘d’ the spurious resonance at the higher band is eliminated due to the equal resonator arm lengths. This ensures that the band stop filter does not suffer from any out of band response at higher frequencies.
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Figure 6.19: Parametric Analysis of the length of CUSR (L) with c=10µm and d=10µm

Figure 6.20: Parametric Analysis of the width of the rings of CUSR (c) with L=1000µm and d=10µm
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Figure 6.21: Parametric Analysis of the spacing between the rings of CUSR (d) with $L=1000\mu m$ and $c=10\mu m$

Figure 6.22: Simulated $S_{21}$ for three (3) different heights of the MEMS bridge
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Figure 6.23: Simulated $S_{11}$ for three (3) different heights of the MEMS bridge

(b) Mechanical and Electromechanical Simulation

(i) Static Analysis under equilibrium

The choice of the optimum dimensions of the MEMS bridge is very critical for reliable operation. The length of the bridge in this case has to be in the range of 220µm for proper anchoring on the two ground planes of the CPW. As a rule of thumb, the aspect ratios of MEMS bridges correspond to 2-2.5 [1]. Therefore, a width of 100µm has been chosen to obtain a balance between the desired aspect ratio and the capacitive effect of the bridge. An initial height of the MEMS bridge has been chosen as 3µm to obtain a higher tuning range. It is to be noted that bridge height less than 2µm are not considered due to instability of MEMS bridges for low gap heights. On application of a voltage (V) the MEMS
bridge is acted upon by electrostatic deflection force $F_e$ and by an equal and opposite restoring force $F_r$ which leads to

$$kx = \frac{1}{2} \frac{\varepsilon_0 A V^2}{g_e^2}$$

(6.14)

where

$$k = 32Eb \left( \frac{t}{l} \right)^3 \left( \frac{1}{8(pl/l)^3 - 20(pl/l)^2 + 14(pl/l) - 1} \right) + 8\sigma(1-\nu)b \frac{t}{l} \left( \frac{1}{3 - 2(pl/l)} \right)$$

is the spring constant of the bridge, $l$, $b$, $t$ are as defined earlier, $x$ is the displacement of the bridge under equilibrium, $g_e = (g_o - x)$ is the effective gap between the two plates, $p = (l/2) + (S/2)$, $A = wXS$ is the intersection area of the plates forming the switch, $\sigma$ is the biaxial tensile residual stress and $\nu$ is the Poisson’s ratio. The force is taken to be distributed along the overlap of the two plates [1].

Solving for voltage, one gets the pull-down voltage of the MEMS bridge as

$$V_{PD} = \sqrt{\frac{8k g_o^3}{27 \varepsilon_0 b S}}$$

(6.15)

It is interesting to point out that low values of biaxial tensile residual stress is actually desired in MEMS bridges to compensate for the compressive stress (increase the buckling limit) which is generated due to the thermal effects arising from the passage of RF signal [23].
Figure 6.24: Variation in the actuation voltage of the MEMS bridge for different values of mean biaxial tensile residual stresses

Figure 6.24 shows the variation in the actuation voltage of the MEMS bridge for mean biaxial tensile residual stresses of 0MPa, 5MPa and 10MPa using Eqn. 18. It is seen that the tensile residual stress increases the beam stiffness leading to an increase in the pull down voltage. Coventorware® simulation has been performed and it is seen that the values obtained are in close approximation to analytical results (Table 6.4).
Table 6.4: Summary of Coventorware® simulations

<table>
<thead>
<tr>
<th>Tensile residual stress</th>
<th>Pull down voltage</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Analytical</td>
</tr>
<tr>
<td>σ= 0MPa</td>
<td>30.44</td>
</tr>
<tr>
<td>σ= 5MPa</td>
<td>37.13</td>
</tr>
<tr>
<td>σ= 10MPa</td>
<td>42.78</td>
</tr>
</tbody>
</table>

(ii) Transient Analysis

The dynamic response of the system is given by d’Alembert’s principle as

\[ m \frac{d^2 x}{dt^2} + B \frac{dx}{dt} + kx = F_e = \frac{1}{2} \varepsilon_0 AV_s^2 \]

where

\[ m(mass) = 0.35(lwt)\rho, B(dampingcoeff.) = \frac{3}{2\pi} \frac{\mu b S^2}{g_0^3} \]

and \( V_s = 1.4V_{PD} \)

This non-linear differential equation has been solved in Mathematica v5® and a switching time of around 8µs is obtained as shown in figure 6.25. The various physical parameters of the MEMS bridge as in Table 6.5 and analytical results of the transient analysis are summarized in table 6.5.
6.2.12 Extraction of Parameters of the CLR model

As all the resonator element values are unknown, simulation is carried out for three different heights of the MEMS bridge corresponding to 2, 2.5 and 3µm for the subsequent extraction. This will result in three different bridge capacitances leading to three different values of resonance frequency as tabulated in Table 6.6.
Table 6.6: Resonant frequencies of the filter for different heights of the MEMS bridge along with the bridge capacitance

<table>
<thead>
<tr>
<th>Bridge height (g₀) (µm)</th>
<th>Capacitance due to one bridge</th>
<th>Resonant frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Parallel Plate Capacitance (C_{pp}) (fF)</td>
<td>Fringe field Capacitance (C_f) (fF)</td>
</tr>
<tr>
<td>2</td>
<td>88.54</td>
<td>17.71</td>
</tr>
<tr>
<td>2.5</td>
<td>70.8</td>
<td>14.2</td>
</tr>
<tr>
<td>3</td>
<td>59</td>
<td>11.8</td>
</tr>
</tbody>
</table>

The capacitance of the fabricated MEMS bridges is computed analytically using the equation:

\[ C_b = C_{pp} + C_f \] \hspace{1cm} (6.17)

where \( C_{pp} \) denotes the parallel plate capacitance and \( C_f \) denotes the fringing capacitance.

\[ C_{pp} = \frac{\varepsilon_0 A}{g_0} \] \hspace{1cm} (6.18)

where \( \varepsilon_0 = \) free space permittivity = 8.854 \times 10^{-12} \text{ F/m} \n
\( A = \) effective area of MEMS bridge for actuation and \( g_0 = \) height of MEMS bridge

The fringing capacitance may be taken to be 20% of \( C_{pp} \) [1]. It should be noted that as the loading of the bridge is mainly capacitive, its inductance has not been taken into account. For a metal MEMS bridge the series resistance of the bridge is also small and may be neglected. However, these values may be accounted for while performing a more accurate extraction.

Now, equation (6.8) is solved in Matlab® for these three resonance frequencies to obtain three sets of extracted CLR values. Among these, the valid
set (Table 6.7) is chosen by mere observation or by fitting the CLR values into the proposed equivalent circuit. The equivalent circuit with extracted parameters is simulated in Agilent ADS® to match the resonance as well as off resonance performance with the Ansoft HFSS® results. Only a particular set will match both the resonance and off resonance behaviors. The extracted parameters as tabulated in table 6.7 are fitted in the equivalent circuit and simulated in Agilent ADS® to obtain the S-parameters. This is then validated with the HFSS results. Figure 6.26 show that the ADS and HFSS results are in good agreement.

Table 6.7: Extracted parameters of CLR model

<table>
<thead>
<tr>
<th>CUSR inductance ($L_c$) in pH</th>
<th>CUSR Capacitance ($C_c$) in fF</th>
<th>CUSR Resistance ($R_c$) in Kohm</th>
</tr>
</thead>
<tbody>
<tr>
<td>100.38</td>
<td>57.64</td>
<td>73.9</td>
</tr>
</tbody>
</table>

![Figure 6.26: Validation of simulated $S_{21}$ parameters](image-url)
6.2.13 Fabrication

It is clearly seen that to realize the proposed design, a virtual air substrate has to be created for the CUSR embedded CPW. To create this virtual air substrate, Si has to be removed from a large volume of the substrate, so bulk micromachining is the most suitable choice [23-25]. The basic steps for fabricating the tunable metamaterial filter based on RF MEMS are detailed in the flowchart below [Figure 6.27].

![Flowchart showing the Fabrication steps of the RF MEMS based tunable metamaterial filter](image)

Figure 6.27: Flowchart showing the Fabrication steps of the RF MEMS based tunable metamaterial filter
For a single wafer process, MEMS bridges can be fabricated on an already micromachined Si substrate housing CUSR embedded CPW. But as CUSR embedded CPW has already been micromachined, a dummy wafer is needed for any subsequent processing (photolithography, etching etc.). Moreover, to realize the virtual air substrate by bulk micromachining, the gold metallization may get damaged for prolonged etching. To overcome the single wafer process limitations, a novel fabrication methodology involving micromachining of two bonded Si wafers has been carried out. One wafer houses the CUSR embedded CPW (bottom wafer) while the other wafer contains the MEMS bridges (top wafer). The processing of the two wafers (p <100>, 5-10kohm-cm) is discussed below:

(i) The bottom wafer is oxidized (1µm thick) by pyrogenic oxidation and subsequently coated with a layer of silicon nitride (0.8µm thick) by Hot Wire CVD. Both these layers will provide mechanical stability to the CUSR. A thin layer of Cr-Au is used as a seed layer for subsequent gold electroplating till 1µm in KAu(CN)$_2$ to obtain the metallization for the CUSR embedded CPW. After the CUSR geometry has been patterned selective gold electroplating of the ground planes to a height of 3.6µm has been done.

Further, a stack of Au-AuGe-Au (100nm-200nm-100nm) layers is deposited sequentially by electron beam evaporation without breaking the vacuum and patterned on the ground planes of the CPW by lift off technique. The first layer of evaporated Au is meant for reducing the roughness of the electroplated gold while the last evaporated Au layer is used for preventing the oxidation of the AuGe eutectic layer. This stack acts as an adhesive layer between the two wafers with the AuGe eutectic (88% Au; 12% Ge) having a melting temperature around $360^\circ$C. One may note that the differential height between the signal and the ground planes of the CPW correspond to a bridge height of 3µm. A silicon dioxide
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isolator (0.1µm) thick is sputtered on the CPW central to prevent the shorting of the two electrodes on actuation.

(ii) The top wafer is oxidized (1µm), sputtered with Cr-Au and subsequently gold electroplated till 1µm to obtain the metallization for the two bridges which is subsequently patterned. Windows of dimensions corresponding to $L_{\text{win}}$ and $W_{\text{win}}$ are opened on the back side of each wafer for subsequent micromachining. Two wafers are then aligned properly in a wafer bonder [Figure 6.28] for carrying out the bonding at 157°C for 1 hour followed by 400°C for another hour, at a contact force of 1.5kN for 20cm² area. Micromachining is finally performed in 25% wt. TMAH solution. As discussed earlier in chapter 5, the alignment of the top and bottom substrates is very critical and some non aligned devices were obtained. A typical non aligned device is shown in figure 6.29. The microphotograph of the perfectly aligned final tunable filter is given in Figure 6.30.
Figure 6.28: Microphotograph of two wafers prior to bonding
Figure 6.29: Microphotograph of the non aligned tunable filter after bonding of two wafers as seen from top side
Figure 6.30: Microphotograph of the final tunable filter after bonding of two wafers as seen from top side

6.2.14 Measurements

For MEMS bridges fabricated as detailed in Section 6.2.13, the height of the bridges are measured with the help of an optical profilometer and bridge heights in the range of 2.7-3µm are obtained for devices fabricated by a two inch (2”) wafer process. Pull-down characteristics of the fabricated MEMS bridges indicates an actuation voltage of around 40V as opposed to an actuation voltage of 31V without the effect of residual stress (red mark in figure 6.24). As expected a very low value of mean biaxial tensile residual in the range of 7-10MPa is obtained for the fabricated MEMS bridges leading to an almost flat MEMS bridge
without warping. The fabricated device is finally characterized using a Cascade Microtech probe station with Agilent PNA series Network Analyzer. The PNA (E8361A) was calibrated from 18-40GHz. The device is probed with a G-S-G ACP (Air Coplanar Probe) with 150µm pitch. The DC actuation was applied through the internal bias-T of the probe station. Figure 6.31 shows the test results of the fabricated device.

Figure 6.31: RF characterization of the fabricated device

The measured results are in agreement with the results obtained by simulation and analytical modeling. A tunable filter with a tuning range of 13% over 28GHz to 32GHz is realized with a low insertion loss in the pass band and high rejection in the stop band.
6.3 Application in phase shifters

One of the important application areas of RF MEMS includes the development of true time delay phase shifter for electronically scanned arrays. The most commonly used phase shifter; the distributed loaded line type; on silicon substrate suffers from the loss arising from the substrate conductivity. Thus, CMOS-grade low-resistivity silicon substrate is not suitable for high-frequency applications as it attenuates the RF signal. Most of the phase shifters are therefore fabricated on high resistivity silicon. But high resistivity silicon is also associated with parasitic capacitances and resistive losses arising from substrate conductivity at a frequency of 10 GHz or higher. Such frequencies are large enough to drive the silicon substrate into its dissipative dielectric mode [26]. Several polymers such as polymide resin, BCB resin, and Kapton films tens of microns thick were reported as the passivation layer between the CPW transmission line and the low resistivity silicon substrate to minimize the losses. Etching away the silicon substrate to leave only the polymer under the CPW line would further reduce the insertion loss [27]. However, the use of polymers for RF MEMS application still needs intensive research [28].

Despite reported RF MEMS components on Pyrex 7740 glass [29], silicon remains the most suitable substrate for such applications owing to the existing microfabrication techniques and CPW transmission line becomes a natural choice as the MEMS bridges can be easily implemented by planar IC technology. Silicon dioxide 2-3μm thick is still the standard passivation layer in most of the devices. In an attempt to reduce the losses related to the conductivity of silicon the whole device is fabricated on a thick silicon dioxide with a thin silicon membrane (20-50μm) to provide mechanical stability to the system. Micromachining of the low resistivity silicon substrate has been carried out by different authors for microwave transmission line to reduce the substrate loss significantly. Recently, a RF MEMS
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capacitive shunt switch on semi- suspended CPW using low- loss high resistivity silicon substrate has been reported [30] through surface isotropic etching. However for low resistivity silicon substrates this technique would not result in superior performance as the field lines penetrate much deeper than the depth of the silicon etched out by micromachining. In addition the oxide- silicon interface gives rise to a bias dependent parasitic capacitance attributed to the formation of the accumulation region or an inversion region under the transmission line [26]. Thus the underlying thin silicon membrane will continue to play a detrimental role in device performance under biased condition.

Here, a novel design of phase shifter with two movable plates is presented on easily available low resistivity silicon by etching away the entire silicon and oxide from underneath the CPW central line and selectively under the ground planes by bulk micromachining. As the major areas where the field lines are concentrated are etched away it results in a virtual air substrate leading to superior performance. The design has been optimized keeping in mind the RF performance and mechanical behavior of the device.

6.3.1 RF MEMS DMTL phase shifter on low resistivity silicon: Device Structure

Figure 6.32 (a) shows the 3D view of the proposed two movable plate phase shifter (TMPPS), fabricated by bulk micromachining and bonding, prior to bonding. As shown, two silicon substrates are involved which are bonded together. Figure 6.32(b) shows the cross- sectional view of the TMPPS along length L_2 [Figure 6.32(a)], where the proposed two movable plate system is supposed to be formed. The top plate shunt membrane length, width and thickness are L_1, W_1 and T_1 respectively and the dimensions of the released portion of the CPW central line acting as the bottom electrode are L_2, W_2 and T_2 respectively.
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Figure 6.32(a): 3D view of the proposed Two Movable Plate Phase Shifter (TMPPS)

Figure 6.32(b): Cross-sectional view of the TMPPS along length $L_2$
Figure 6.33(a) shows the cross-section of the TMPPS after bonding along length \( L_2 \). Similarly, Figures 6.33(b) and (c) show the cross-section of the TMPPS along lengths \( B \) and \( L_3 \) respectively. Figures 6.33 (d) and (e) represent the CLR model along length \( L_2 \) and \( B \) respectively. As seen in Figure 6.32(a) the entire silicon and the oxide is removed from underneath the CPW central line and the only silicon left is to provide anchorage to the CPW central line (length \( B \)). It should also be noted that the silicon is also etched from an additional portion \( A \) in both the wafers as shown in Figure 6.32(b). This causes the electric field to be confined in the air dielectric which leads to the reduction is losses arising from the conductivity of silicon. Further the removal of the portion \( A \) from the top substrate also leads to the creation of an improper mechanical boundary condition for the shunt membrane. The fabrication methodology and initial fabrication results are presented in Section 6.3.5.
Figure 6.33: (a) Cross-section of the device after bonding along $L_2$, (b) cross-section of the device after bonding along $B$, (c) cross-section of the device after bonding along $L_3$, (d) CLR model along $L_2$, (e) CLR model along $B$.

6.3.2 Reduction in parasitics

As seen in Figure 6.32(b) the effect of the lossy silicon substrate is completely eliminated underneath the CPW central line and a certain portion of the CPW ground planes. The effect of the stray capacitances at the edges of the shunt membrane has been adequately addressed by the presence of $R_{\text{EDGE}}$ and $C_{\text{EDGE}}$ to account for the top substrate [Figure 6.33(d)]. The value of $A$ is selected as $65\mu$m so that the electric field is confined in the air dielectric. For cross-section along
the CPW anchor (length B) the effect due to the lossy silicon substrate is minimal as the value of B is chosen as 20µm so as to provide sufficient anchorage to the CPW central line to maintain mechanical stability. For the remaining portion of the device cross-section (length L3) the value of the parasitic capacitances is hugely reduced due to the lossless air dielectric. The value of \( C_{\text{AIR}} \) comes to 0.014fF/µm which is in accordance with the static capacitance value of 0.012fF/µm obtained using ANSYS Multiphysics®. This causes a significant reduction in the parasitic as compared to the value of 0.15fF/µm for silicon substrate (\( C_{\text{SUB}} \)). Further the losses due to the finite conductivity of silicon are also hugely reduced as it is replaced by air in all the major areas. It should be noted that the effect of the top silicon substrate is somewhat lower than the bottom dielectric due to the post height of 2µm in the upward direction which provides termination for the field lines.

6.3.3 Design and simulation of RF MEMS DMTL Phase Shifter

The design starts by selecting a suitable CPW for obtaining a high value of unloaded impedance (\( Z_0 \)). An unloaded impedance of in the range of 100-120Ω is recommended to low permittivity substrates [1]. The dimensions of W/S/W should be such that the unloaded impedance of the total device considering the effect of the top substrate should be as specified. It is seen that for a W/S/W corresponding to 65µm/100µm/65µm yields a characteristic impedance of 120Ω (considering the capacitive loading of the top substrate). A virtual air substrate has been assumed for the calculations. For reflection loss to be in the range of 8-10dB for the entire Ku band the up state impedance (\( Z_{\text{up}} \)) comes to around 86Ω from the following equation.
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\[ Z_{lu} = Z_0 \sqrt{\frac{1 + \rho}{1 - \rho}} \quad (6.19) \]

where \( \rho \) is the allowable reflection coefficient.

A Bragg frequency \( f_B = 3f_0 = 60 \text{GHz} \) is chosen for obtaining a very good wideband performance. The values of the unit length inductances and capacitances are calculated by considering an effective dielectric constant as per the model proposed.

\[ Z_0 = \sqrt{\frac{L_t}{C_t}} \quad \& \quad C_t = \frac{\varepsilon_{\text{eff}}}{cZ_0} \quad (6.20) \]

where \( \varepsilon_{\text{eff}} \) is the effective dielectric constant and \( c \) is the speed of light in free space. For calculating the effective dielectric constant of the device in the up and down states conformal mapping has been performed as elaborated earlier and as per figure 6.34.

Figure 6.34: Simplified cross-sectional view of the proposed device suitable for theoretical formulation
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The total capacitance per unit length of the unloaded CPW t-line considering the loading of the top and bottom substrates is given by:

\[ C_t = 4C_{\text{AIR}} \]  \hspace{1cm} (6.21)

Now, the effective dielectric constant of the device in the up and down state can be found by using the equation

\[ \varepsilon_{\text{eff, UP/DOWN}} = \frac{C_{\text{TOTAL, UP/DOWN}}}{C_t} \]  \hspace{1cm} (6.22)

where

\[ C_{\text{TOTAL, UP/DOWN}} = \frac{CL_t + NC_{lu/ld}}{L} \]  \hspace{1cm} (6.23)

and N is the number of shunt switches.

Further, the characteristic impedance in the up state and down state is given by

\[ Z_{lu} = \sqrt{\frac{L_t}{C_{\text{TOTAL, UP}}}} = \frac{L_t}{\sqrt{C_t + C_{lu} / s}} \]  \hspace{1cm} (6.24)

\[ Z_{ld} = \sqrt{\frac{L_t}{C_{\text{TOTAL, DOWN}}}} = \frac{L_t}{\sqrt{C_t + C_{lu} C_t / s}} \]  \hspace{1cm} (6.25)

The design values of the up state capacitance \(C_{lu}\), capacitance ratio \(C_t\) and spacing between the bridges \(s\) are obtained from simultaneous equations as \(Z_{lu}\), \(Z_0\), \(\varepsilon_{\text{eff}}\) and \(f_B\) are known. It is seen that if the length of the device is 8mm then the shunt membrane width can be chosen carefully to obtain a capacitive loading yielding the desired impedances in the up and down states.
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It is seen that for \( w_1 = 30 \mu m \), the up state capacitance \( (C_{lu}) \) comes to around 18fF considering the fringing effect. A bridge height of 2\( \mu \)m, silicon dioxide isolator thickness of 0.5\( \mu \)m and slot widths in the dielectric (20\( \mu \)m dielectric/ 20\( \mu \)m slot arranged periodically) are chosen to obtain a capacitance ratio \( (C_r) \) of 4.5 when the two movable plates make contact upon the application of the actuation voltage. The desired value of the spacing between the MEMS capacitors is dependent on the Bragg frequency [1] and the value comes to 500\( \mu \)m for this design. To obtain the required spacing between the MEMS capacitors the value of \( L_2 \) and \( L_3 \) are chosen as 190 \( \mu \)m and 270\( \mu \)m respectively. The entire structure is simulated using Ansoft HFSS® and it is seen that a phase shift of 180\(^\circ\) is obtained at 17.5GHz with an insertion loss of 1.16dB (Figure 6.35) with 13 MEMS capacitors. The reflection loss is better than 10 dB is obtained in both the cases for the entire Ku band and it is clearly seen that the Bragg frequency of the device is around 60GHz (Figure 6.36).

![Figure 6.35: Frequency vs. phase (deg) of the device in up and down states](image-url)
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Figure 6.36: S-parameters vs. frequency of the device in up and down states

To validate the proposed model the short-circuit and the open-circuit impedances in the up and down states of the bridge are calculated analytically and matched with the simulation results. The short circuit and the open circuit impedance of the device in the up and down states can be obtained from the following equations:

\[ Z_{sc} = Z_0 \tan \beta l \] (6.26)

\[ Z_{oc} = -Z_0 \cot \beta l \] (6.27)

Where

\[ \beta = k_0 \sqrt{\varepsilon_{eff}} \quad \text{and} \quad k_0 = \omega \sqrt{\mu_0 \varepsilon_0} \]

Figures 6.37 and 6.38 show that there is a good agreement between the FEM simulation results and the analytical results obtained numerically. Similarly, figure 6.39 shows the short circuit and open circuit impedances in down state obtained numerically.
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Figure 6.37: Short circuit impedance in the UP state

Figure 6.38: Open circuit impedance in the UP state
6.3.4 Improvement in Mechanical Performance

Another advantage of the two movable plate systems is that it is capable of simultaneously improving the actuation voltage and the switching time. As already discussed, two movable plate system can be represented by a single movable plate system given by

\[
m_{\text{eff}} \frac{d^2 x}{dt^2} + b_{\text{eff}} \frac{dx}{dt} + k_{\text{eff}} x = F_e
\]  

(6.28)

where

\[
k_{\text{eff}} = \frac{k_1 k_2}{k_1 + k_2}
\]  

(6.29)
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$$m_{eff} = \frac{m_1 m_2}{m_1 + m_2}$$ (6.30)

$$b_{eff} = \frac{b_1 b_2}{b_1 + b_2}$$ (6.31)

$m_1$, $b_1$, $k_1$ and $m_2$, $b_2$, $k_2$ are the mass, damping coefficient and spring constant of the membrane and the CPW central line respectively. The damping coefficients $b_1$ and $b_2$ are assumed to be constant for the very small volume of trapped air between the plates.

Equation 6.28 is non-linear differential equations which can be solved numerically under the boundary condition:

$$x_1 = 0; x_3 = g_0 and \frac{d x_1}{d t} = 0; \frac{d x_3}{d t} = 0$$

at $t=0$. The above equations are valid till the two plates come in contact.

Optimal device performance occurs when the two plates are matched (comparable masses, comparable spring constants and comparable damping coefficients) with the CPW membrane being marginally stiffer than the shunt membrane. As in the case of phase shifter the device geometry is mainly governed by its RF performance it is difficult to achieve best mechanical performance. For device dimensions of $(L_1, W_1, T_1)$ and $(L_2, W_2, T_2)$ as $(230, 30, 1.1 \mu m)$ and $(190, 100, 1 \mu m)$ respectively, Mathematica v5® was used to solve the non-linear differential equations and the two movable plate system results in a 10% improvement of the actuation voltage compared to the Single Movable Plate System (SMPS) without affecting the switching speed [Figure 6.40]. Alternatively a 20% improvement in the switching time is obtained if the device is actuated with the same voltage as in the single movable plate system [Figure 6.41].
Figure 6.40: Plot of gap height vs. voltage for TMPS and SMPS.

Figure 6.41: Switching characteristics of the TMPS and the Single movable plate system (SMPS) counterpart for different same switching voltages.
6.3.5 Fabrication

The fabrication of the TMPPS is similar to that for fabricating the TMPS. The process involves the use of two bonded wafers followed by micromachining. The bottom substrate is for housing the movable CPW t-line and the top substrate houses the movable shunt membranes. The required thickness of the CPW central line, CPW ground planes and the MEMS bridge is obtained using gold electroplating. The differential height between the CPW central line and CPW ground planes on the bottom substrate is obtained by selective electroplating as discussed in Chapter 5. After the two wafers are patterned with the desired geometry DRIE is to be performed with a suitable metal mask (Pt, Au) to release the CPW central line, portion of the CPW ground planes and the MEMS bridge. Initial fabrication results are however obtained by bulk micromachining of the substrates. Figure 6.42 shows the microphotograph of the device obtained after bonding. The shunt membrane is suspended at a height of 2µm over the movable CPW transmission line.

Figure 6.42: Microphotograph of the phase shifter showing two (2) bridges as seen from top side
6.4 Summary

Two major application areas viz. a tunable filter and phase shifter have been realized on a virtual air substrate. It is seen that to implement dynamic tunability in metamaterial filters, RF MEMS switches may be used. Design principles have been provided based on the equivalent circuit model and a tunable filter on a virtual air substrate having 13% tuning range in the Ka band is achieved. A novel technique of utilizing two silicon wafers has been introduced for the first time to fabricate CUSR embedded CPW and MEMS bridge. Similarly, the realization of the phase shifter on a virtual air substrate results in an insertion loss of 1.16dB for a 8mm long device. This technique makes it possible for phase shifter design on low resistivity CMOS grade silicon substrates making it eligible for future CMOS integration.

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