24 GHz RF-MEMS Phase Shifter with Non-Galvanic Electromagnetic Coupling Fabricated in Silicon-Bulk Technology

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Abstract. This contribution presents a Distributed MEMS Transmission Line (DMTL) phase shifter designed for 24 GHz and fabricated in Silicon-Bulk Micromechanics Technology. Using this technology enables to commonly suspend all capacitive loads on one movable plate and allows full range analog and homogeneous gap adjustment. The first available prototypes of the phase shifter are characterized to provide 5.4 °/mm differential phase shift at 24 GHz and to show 0.1 dB/mm insertion loss. An asymmetric coplanar signal coupler with a center frequency of 24 GHz and 10% bandwidth is used to contact the signal through the substrate of the phase shifter chip and back to a supporting printed circuit board. The normalized measured insertion loss of the coupling structure after simulated optimization is 0.13 dB - 0.17 dB.

1. Introduction

The utilization of micro electro mechanical systems (MEMS) instead of conventional semiconductor devices can obtain fundamental improvements in microwave circuits with respect to signal attenuation and DC power consumption. Roadmaps and studies predict a wide field of application for RF-MEMS devices. The development of civil radar systems for automotive applications would strongly gain from the advantageous characteristics of RF-MEMS. Mechanical antenna scanning can be replaced by low loss electrical beam steering. MEMS Technology has great potential for the fabrication of low loss 360° phase shifters required for this application. The main focus of numerous international research teams has been the development of actual MEMS functional elements, whereas less attention has been paid to comprehensive concepts considering assembly and packaging issues as well [1]. Conventional wire bond techniques cause matching problems at microwave frequencies while soldered flip chip connections introduce high mechanical stress that might cause undesirable deflection and bending of the movable MEMS structure. A novel impedance matched asymmetric coplanar signal coupler connects the MEMS phase shifter to the printed circuit board non-galvanically. Epoxy adhesive mounting is used to fabricate self-contained RF-MEMS devices easy to integrate in many circuits.

A Distributed MEMS Transmission Line (DMTL) phase shifter consists of a high impedance transmission line which is periodically loaded with lumped adjustable capacitors. These loads affect the line impedance and phase velocity. Optimization [2] and fabrication [3] of the DMTL phase shifter concept has been reported in surface MEMS technologies before. Initially the distributed approach was investigated to achieve analog phase shift using varactor diodes [4]. DMTL phase shifters are...
considered to show better performance than semiconductors with respect to loss and power consumption. As a drawback the semiconductor’s capability of analog phase shift was hard to realize with individual MEMS bridges [5]. Hence, the surface micromechanic DMTL concept has been developed towards digital phase shifters containing sections of capacitive loads to achieve different states of phase shift.

2. Realization of a DMTL Phase Shifter in Silicon-Bulk Technology

For the design of an analog tunable DMTL phase shifter the Silicon-Bulk Technology offers advantages compared to surface technologies. The capability of etching structures into the wafer allows the use of the extended tuning range concept described in [6] and to suspend all capacitive loads on one movable plate. This leads to wide range and homogeneous analog tunability [7].

The top view in Fig. 1 and the corresponding cross section in Fig. 2 show the two wafer concept of the phase shifter. Six beam springs suspend a structured movable plate forming 25 capacitive loads. In contrast to the referenced approaches [2] and [3] these loads are capacitively coupled to the ground plane of a high impedance coplanar waveguide (CPW) structure which is situated on the bottom wafer. The coupling capacitors change with the moving bridges. As can be seen in Fig. 2 the actuation electrodes on the bottom wafer are placed in a cavity to enhance the tunability of the functional bridge gap. The CPW is suspended on a 30 μm silicon membrane to reduce dielectric loss and to achieve a high impedance despite the high permittivity of silicon. Eqn. (1) shows the linear dependence of the phase shift $\Delta \phi$ from frequency $\omega$ and unloaded line impedance $Z_0$.

$$\Delta \phi = \frac{\omega \cdot Z_0 \cdot \sqrt{\varepsilon_{\text{eff}}}}{c_0} \left( \frac{1}{Z_{\text{max}}} - \frac{1}{Z_{\text{min}}} \right) \left[ \text{rad} \right]$$

The design was optimized with respect to phase shift per line length. For the dimensions listed in Tab. I Method of Moment (MoM) and Finite Difference Time Domain Simulations (FDTD) predict a phase shift of 17 °/mm line length at 24 GHz by tuning the bridge gap from 4 μm to 2 μm. This corresponds to a return loss of >25 dB caused by the impedance mismatch. Due to the wide range tunability of the bridge gap this phase shift can be increased if one can tolerate more return loss. The designed maximum mechanical tuning range of the bridge gap is from 6 μm to 1 μm. By applying the
DC bias voltage the movable plate can be deflected in this range. The optimum impedance range around the match, which occurs close to 3 μm gap, can be adjusted. The fabricated phase shifter consists of 25 bridges with an spacing of 300 μm and a width of 100 μm creating a line length of 10 mm. The maximum actuation voltage is designed to be 45 V with the constraint that the movable plate’s maximum displacement due to gravity is 5% of the bridge gap. Considering the bulk conductivity \( \rho = 3000 \, \text{Ωcm} \) of the 30 μm thick silicon membrane the simulated insertion loss for the coplanar waveguide is 0.8 dB/cm at a frequency of 24 GHz.

| Parameter              | Value    |
|------------------------|----------|
| bridge gap             | 100μm    |
| bridge spacing         | 300μm    |
| center conductor width | 175μm    |
| ground gap             | 175μm    |
| tunable bridge gap     | 6μm – 1μm |
| number of bridges      | 25       |

### 3. Coplanar Electromagnetic Signal Coupler

Since the design is made for the 24 GHz ISM band the dimensions of the phase shifter unit are approximately 10 mm × 5 mm. For respectively large MEMS structures like this the mechanical stress caused by soldering or standard adhesive mounting leads to difficulties like bias deflection and poor parallelism. An approach to overcome this challenge is the novel coplanar electromagnetic signal coupler which has been implemented with the phase shifter. It replaces soldering by adhesive mounting using few stress optimized adhesive spots. As can be observed in Fig. 1 the RF signal path is provided by electromagnetic couplers at each port of the phase shifter and therefore has no ohmic contact outside the chip. For the presented phase shifter the coupling concept is considered to allow low stress adhesive mounting with good RF performance.

Electromagnetic signal coupling is suitable for frequencies where the structure dimensions get into the range of the signal wavelength. The general concept is presented in [8] while an actual surface to surface transition was investigated in [9]. A modified even and odd mode analysis of symmetric coupled CPW structures has been applied. It uses an electric and magnetic wall in between the coupled CPW’s and leads to an even and odd mode impedance and phase velocity respectively. The open circuit boundary condition for two diagonal opposite ports according to Fig. 3 leads to a symmetric transmission matrix (ABCD-matrix) presented in [10]. Using the image impedance concept the trans-
mission characteristics can be described. According to Eqn. (2) [11] this yields an equivalent transmission coefficient $\gamma_{eq}$. At an electrical length of $\theta = \pi/2$ the real and imaginary part of $\gamma_{eq}$ are $\text{Re}(\gamma_{eq}) = 0$ and $\text{Im}(\gamma_{eq}) > 0$. This is equivalent to a lossless transmission line with frequency dependent impedance. As can be seen in Fig. 5 the symmetric image impedance at port 1 and port 2 reaches a maximum at this wavelength. According to Eqn. (3) the image impedance maximum of $(Z_{0,e} - Z_{0,o})/2$ at $\theta = \pi/2$ has to be higher than the system impedance $Z_{0,s}$ to achieve a matched condition. Electromagnetic simulations show that this condition can not be fulfilled for practical dimension which makes $\gamma/4$ impedance transformers necessary [12].

$$\lambda_{eq} = \cosh^{-1}\sqrt{AD}$$  \hspace{1cm} (2)

$$Z_{0,s} \leq \frac{Z_{0,e} - Z_{0,o}}{2}$$  \hspace{1cm} (3)

The actually implemented signal coupler is not symmetric because it has no compensation layer on top of the structure which has the same permittivity $\varepsilon_r$ as the suspending substrate. Hence, it is an asymmetric one displayed in Fig. 4. For this case according to Eqn. 4 and Eqn. 5 the image impedances for port 1 and port 2 are different because the matrix elements A and D are not equal [13] like in the symmetric case. According to Fig. 6, the image impedance maximum for port 1 occurs at a lower electrical length and for port 2 the image impedance has a singularity at $\theta = \pi/2$.

$$Z_{i,1} = \sqrt{\frac{AB}{CD}}$$  \hspace{1cm} (4)

$$Z_{i,2} = \sqrt{\frac{BC}{AC}}$$  \hspace{1cm} (5)

Nevertheless, the investigation [12] showed that it is possible to match the structure to the system impedance because the impedances at both sides are real in the proximity of the maximum at port 1. With an appropriate design the image impedance at port 2 matches the system impedance while the image impedance at port 1 reaches its maximum. Hence, the impedance transformer at port 2 can be omitted and bandwidth is maximized. For a center frequency of 24 GHz the designed structure has a center conductor width of 150 μm and a length of 850 μm. A bandwidth of 10% with a return loss of better than 20 dB was determined by electromagnetic simulations using the Method of Moment (MoM) and the Finite Difference Time Domain (FDTD) Method.

Fig 5: Image impedance at port 1 and port 2 of the symmetric coupler

Fig 6: Image impedance at port 1 and port 2 of the asymmetric coupler
4. Fabrication Technology

The phase shifter is fabricated by bulk micromachining using anisotropic wet etching, reactive ion etching and surface passivation with SiO₂ and Si₃N₄. The gold metallization is sputtered and structured by sputter mask and photolithography. The two wafers are joined by silicon fusion bonding. The wafer material is 3000 Ωcm silicon.

Fig. 7 shows the schematic processing sequence of the bottom wafer. It is a standard wet etching process with a buried mask due to the requirement of two depth levels. The reason is that in the final chip the CPW is suspended by a 30 μm membrane and the whole bottom layer is only 100 μm thick which is not possible to be handled through processing. Later the thicker frame is sawed off during the separation of the chips. The metallization is sputtered on an oxide layer for isolation and stress compensation.

The process scheme of the top wafer can be retraced in Fig. 8. Using a dry etch process the movable structure is shaped and it is released from the backside by wet etching. The Si₃N₄ is removed before the release to avoid the exposure of the movable structure to this removal process. In order to compensate the residual stress of the gold an oxide layer is placed underneath the gold. This is done by burying a structured oxide layer before. At the current prototypes the silicon fusion bonding did not work due to a contamination of the bond surface. Therefore the chips were manually joined using epoxy glue. Hence, the desired bridge distance has not been achieved. Nevertheless, the prototypes have been characterized while the bonding process for this RF-MEMS project is currently under investigation.

Fig. 8 shows the schematic processing sequence of the top wafer.
5. Measurement Results
The fabrication of several prototype designs allows a separate characterization of the phase shifter and
the signal coupler. Chips with the phase shifter only having on-wafer probe pads can be used to char-
acterize the phase shifter separately from the couplers. The signal coupler is characterized with chips
only containing a single coupling structure or two surface to surface transitions in back to back con-
figuration.

5.1. Phase Shifter
The prototype shows a continuously tunable phase shift and a promising RF-performance at different
bias voltages over the full measurement range up to 40 GHz. The measured phase shift for a bias volt-
age of 60 V is 5.4°/mm. Fig. 9 shows the measured insertion loss of the phase shifter unit. It shows
characteristic oscillations over frequency. This and the deviation of the phase shift from the expected
value is caused by the mentioned difficulties with the wafer bonding process resulting in non perfect
parallelism and planarity between the CPW transmission line and the capacitive loads. Considering the
measured phase shift and the insertion loss for the matched condition occurring periodically in Fig. 9
the figure of merit would be about 50° phase shift per 1 dB insertion loss at 24 GHz. This is, yet, a
very good value compared to competing structures and principles. Nevertheless, it is expected that cur-
rent investigations, in order to improve the fabrication process, yield lower and non-periodic insertion
loss and higher differential phase shift which will lead to a significant higher figure of merit.

Fig 9: Measured and simulated insertion loss of
the phase shifter

Fig 10: Measured S-parameters of the coupler and
modified simulation results approximating the air gap

5.2. Signal Coupler
According to Fig. 10, the fabricated coupler shows the expected transmission behavior. Due to initially
not exactly known effective air gap between chip and PCB carrier, the center frequency of about
28 GHz and the impedance match deviate from the simulation. Also the impedance match is affected
by this air gap resulting in an insertion loss of about 0.5 dB. By comparing the measurement results
with modified simulation models (Fig. 10) the air gap has been determined to be \( d_{\text{air}} = 5 \mu m \). Hence,
by considering a reproducible residual air gap achieved by an optimized mounting technique it is pos-
sible to meet the specified center frequency and to achieve a loss per transition of 0.28 dB - 0.43 dB.
Using the method of normalized losses the attenuation caused by the coupling structure is
0.13 dB - 0.17 dB.

6. Summary
The DMTL phase shifter principle was successfully implemented in Silicon-Bulk Technology. Using
the advantages of this technology it has been shown that it is possible to fabricate high performance
phase shifters for microwave frequencies. The figure of merit increases with frequency and the design
is suitable to be scaled to higher frequencies. Particularly for stress sensitive RF-MEMS structures the
signal coupling with adhesive mounting offers a good approach for the integration into RF-circuits
used in communication and radar.
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