Millimeter-wave MEMS-loaded transmission-line phase shifter

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von

Peter Simon Heeb
aus Altstätten SG, Schweiz
Betreuer / Gutachter:  Prof. Dr.-Ing. Hermann Schumacher  
                           Prof. Dr.-Ing. Ingmar Kallfass  

Amtierende Dekanin:  Prof. Dr. Tina Seufert  

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ABSTRACT

Beam forming and scanning at microwave and millimetre-wave frequencies (1 GHz – 100 GHz) using electronic steerable array (ESA) antennas, requires phase shift and time-delay units. RF path phase shifters provide an attractive solution, which enable RF transceivers of reduced complexity compared to phase shifting in the LO path. RF phase shifters based on ferrimagnetic, ferroelectric or semiconductor materials usually exhibit a high insertion loss, small bandwidth, poor linearity, or high power consumption. RF MEMS phase shifters using dielectric substrates can overcome these drawbacks and compete with state-of-the-art phase shifters fabricated in other technologies. This dissertation addresses the concept of distributed MEMS-loaded transmission-line phase shifters and uses dielectric substrates.

A rigorous analysis of the electro-mechanical switching behaviour gives insight into the mechanics in static and dynamic cases. The results allow estimation of the power handling limitations due to self-actuation, switching time and bouncing versus actuation voltage, and the influences of process tolerances or changing operating conditions. Given an arbitrary actuation waveform, the mechanical and electrical switch responses can be studied. The energy balance shows the model accuracy, and relates the stored and dissipated energy to the total power consumption.

The described fabrication process provides good chemical and thermal compatibility throughout all process steps, reduces process tolerances, prevents contact sticking, and enables realization of uniform and stress-free electroplated films. Proof of process is demonstrated by various fabrication runs carried out over a period of three years. Electrical measurements have been carried out to characterize the MEMS switches as a basic building block of the phase shifter. These comprise: switching time, switching and contact hysteresis, insertion loss, return loss, forward loss and isolation, lifetime predictions as well as the power handling capabilities of the switch. The measurement results are compared to the simulations.

Following a new design procedure, RF MEMS-loaded transmission-line phase shifters have been realized. The common lumped element design equations using element values of prototype filters have been adapted for the synthesis of distributed element networks. Low-pass filters with programmable cutoff and insertion phase have been designed and fabricated using the Chebyshev coefficients for weighting of the loaded-line sections. The sections can either exhibit predominate reactance or susceptance loading, which results in a relative change of the insertion phase and characteristic impedance. Analytical descriptions are derived for S-parameters and T-parameters, which consider the
length of the loaded-line sections and Bragg reflections. The limitations of the analytic approach are illustrated by comparison with Momentum simulation and discussed.

The developed phase shifters are characterized by means of RF measurements. The phase shifter bits 1 to 5 designed for the insertion phase differences of $11.25^\circ$, $22.5^\circ$, $45.0^\circ$, $90.0^\circ$, and $180.0^\circ$ showed a maximum phase error of $3^\circ$ between 60 and 61.5 GHz. In the on state, the corresponding insertion losses of the phase shifter bits were measured to $0.0$ dB, $0.1$ dB, $0.3$ dB, $1.3$ dB, and $2.0$ dB. The return losses between 30 GHz and 62 GHz of the phase shifter bits were better than $26$ dB, $24$ dB, $20$ dB, $22$ dB, $20$ dB.

The estimated insertion loss of a 5-bit phase shifter consisting of the phase shifter units 1 to 5 is below $3.7$ dB. The average phase error over all 32 states within the ISM band is approximately $1^\circ$. 
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CHAPTER 1

Introduction

1.1 RF MEMS on dielectric and semiconductor substrates

While semiconductor technologies continue downscaling as a result of sustained consumer needs for higher integration density, centimetre and millimetre wave circuitry on non-semiconducting low-loss substrates has experienced a comparably conservative technological development over the past decades. Due to shrunk dimensions and the enhancement of new semiconductor technologies, higher and higher transit and maximum frequencies have become state of the art. Since monolithic microwave integrated circuits (MMIC) combine active and passive elements on one single chip at small length scale without the need for separate packaging and assembling of active subcomponents, many hybrid microwave integrated circuits (HMIC) have been replaced by MMICs. The main advantages of today’s microwave integrated circuits (MIC) on dielectric substrates over passive elements in MMICs are their low dielectric and conductive losses. Improved performance by compact monolithic integrated transmission lines has recently gained increasing attention. Potential solutions include metamaterials and composite right-/ left-handed (CRLH) structures [1], [2], CPW lines on thick polyimide layers [3], thin film coaxial lines [4] and thin film microstrip lines (TFMS) compatible with existing BiCMOS and CMOS processes [5]-[7]. However, their reported insertion loss per length significantly exceeds that of standard MICs.

The growing demand for high data rate wireless communication has pushed the development from super high frequency (SHF) 3 GHz - 30 GHz to extremely high frequency (EHF) 30 GHz - 300 GHz systems. At these frequencies parasitic reactances and susceptances due to electrical interconnects and other assembly interfaces can significantly reduce the circuit performance. This requires new types of electrical interconnects and impedance matching components. Other basic components for future mobile wireless communication at EHF are planar antenna arrays and phase shifters, which in combination enable electronic beam steering.

Programmable micro-electro-mechanical systems (MEMS) are an interesting solution to perform signal routing, filtering, impedance matching and phase shifting. However, MEMS switches have often been left out of consideration due to reliability issues, such as charging problems, power handling limitations and contact degradation. Traditionally, RF MEMS components have been realized as MIC on dielectric substrates. Next to the low reliability, the high actuation voltage of capacitively driven MEMS switches and the large area consumption of MEMS programmable
networks, compared to their potential insertion loss improvements below 60 GHz, have made MEMS not very attractive for integration into an MMIC. Recently, however, RF MEMS switches have become available on top of BiCMOS substrates at IHP Microelectronics [8-12]. Other commercially available processes include: the multi-user MEMS processes (MUMPs) of MEMSCAP, the thick epitaxial layer for micro-gyrosopes and accelerometers (THELMA) process of STMicroelectronics, the MEMS integrated design for inertial sensors (MIDIS™) from Teledyne DALSA and the bulk micromachining of AMS [13]. Due to their achievable outstanding performance up to EHF, RF MEMS phase shifters and tuneable RF MEMS filters can become key components for future communication systems [12], [14]-[17]. The choice of a MEMS substrate is the answer to a product-specific and economic question, where integration density, costs per area and fabrication capabilities play a major role.

1.2 Phase shifters and their applications

A fixed phase shifter can be represented by a two-port network which provides a defined phase shift \( \phi(\omega) \) over a desired frequency range \( \omega \) (figure 1.1). The notation “fixed” refers to a network which is not capable of external phase control. This type of phase shifter is frequently used in combination with a power divider (figure 1.2) in order to create two signals which are in quadrature. Subsequently, the two orthogonal amplitude components can be processed independently. 90° phase shifters are found in microwave measurement equipment as a part of an I-Q vector modulator.

![Figure 1.1. Fixed phase shifter](image)

Fixed phase shift networks can also be found in transitions between differential waveguides and microstrip transmission lines. A 180° constant phase shifter makes it possible to transform a balanced to an unbalanced signal.

By combining the functions of a power splitter and a quarter wave phase shifter, we end up with a standard hybrid ring, a branch-line coupler or a directional coupler. Each is a four port network, with one port isolated. These components are building blocks in any coherent radar system and used in balanced amplifiers. Balanced amplifiers are applied in microwave systems for better impedance match. The reflections are absorbed by the isolated ports of the input and output coupler. Due to superpositioning of the anti-phase signals, odd harmonics are supressed [18].

Generally, the interpreted phase shift refers to a reference path or reference state. If we are dealing with a two-port network, the obtained insertion phase difference is a function of the state or time. In
the case of an N-port phase shift network with $N > 2$, the insertion phase difference can additionally become a function of the port number.

In contrast to a fixed phase shifter, a variable phase shifter allows for analogue tuning or digital programming (figure 1.3). This phase shifter can be represented as a two-port network which enables phase shift control or selection by an external signal or trigger $x(t)$. The set phase shift remains a function of frequency but becomes an additional variable in time. The tuning range or numbers of states are characteristics of the phase shifter.

By combining a power divider with an analogue tuneable or digitally programmable phase shifter (figure 1.4), we end up with a coupling system offering tuneable or programmable frequency characteristics. Connecting a power combiner to the ports on the right side of the circuit given in figure 1.4 results in a tuneable comb filter or band switch. If using the Wilkinson power combiner, odd modes are absorbed by the load of $Z_L = 2Z_0$.

Within a given bandwidth, the insertion phase difference is desired to be either linear dependent or invariant with respect to frequency. Phase shifters with a linear dependence are also referred to as true time-delay networks. An array antenna combined with a network of parallel RF paths, each equipped with a controlled phase shifter (figure 1.5) can be used for beam forming and for electronic scanning. Such electronic scanning array (ESA) antennas are found in automotive radar systems [19]-[21].
Depending on the circuitry requirements regarding switching time, power handling capabilities, phase range, bandwidth and frequency of use, phase resolution and accuracy, insertion and return loss, temperature dependence, size and weight, lifetime, etc., different kinds of phase shifter architectures have been demonstrated and find application in today’s communication and measurement systems.

1.3 State-of-the-art phase shifter implementations

Today available phase shifters comprise the ferrimagnetic, ferroelectric and semiconductor phase shifters. The state of the art of these phase shifter implementations is summarized here. The underlying material system and physical principle of the different types are described. Partially, the individual performance and technological limitations are discussed.
The class of semiconductor phase shifters spans a huge variety of implementations including different technologies. Basically, they can be differentiated into active and passive types of phase shifters. Passive phase shifters are considered here exclusively.

1.3.1 Ferrimagnetic phase shifters

The class of ferrite phase shifters make use of a ferrimagnetic material, such as yttrium iron garnet (YIG) or the ferrites of cobalt, lithium, strontium and barium. Typically, ferrimagnetic materials exhibit low electrical conductivity. Exposed to a bias magnetic field, the magnetic dipoles of the spinning electrons precess around the direction of the magnetic field. The resonance frequency of precession $f_0 = \gamma 4\pi M$ is proportional to the magnetization $4\pi M$, with $\gamma$ the gyromagnetic ratio. This effect causes anisotropy in the permeability tensor $\mu$ when interacting with an exterior electromagnetic wave.

If magnetization is reduced close to the demagnetization field, the minimum tunable ferrimagnetic resonance frequency is observed $f_{0min}$. In this low field regime, the ferrite is only partially magnetized and absorption losses become severe. The maximum frequency of operation is limited by the saturation magnetization $M_s$ of the ferrite. This insertion loss consideration typically limits the frequency range to $1.6f_0 < f_{RF} < 5f_0$ [22], [23]. The corresponding frequency of operation for YIG, thus ranges from 8 to 25 GHz. Other ferrites with higher saturation magnetization allow for extension of the frequency range up to 70 GHz.

Traditionally, available ferrimagnetic phase shifters are based on rectangular or circular wave guides. Planar architectures of microstrip meander line have also been demonstrated [22], [24]. Commercially available ferrite phase shifters working at V-band (50 GHz - 75 GHz) have a limited fractional bandwidth of 1 - 3% due to the dispersion of permeability. The insertion loss in this frequency band ranges from 1.6 dB - 3.0 dB. The switching speed of ferrite phase shifters ranges from 10 $\mu$s - 150 $\mu$s. The average power capabilities are below 2 W for frequencies above 50 GHz.

1.3.2 Ferroelectric phase shifters

Ferroelectric materials show a non-linear relation between the electrical polarization and the applied electric field. Hence, their permittivity changes in dependence on an external electrical field. Paraelectric materials differ from ferroelectric materials due to their zero residual polarization in the absence of an electrical field. In contrast to ferroelectrics, the crystal structure of paraelectrics does not allow different stable states of charged atoms within the lattice. Above the Curie temperature, ferroelectric materials such as barium titanate and barium strontium titanate (BST) change from a tetragonal to a cubic crystal structure, and become paraelectric.
Whereas paraelectric materials such as PTFE, BCB, SiO₂, Al₂O₃, TaO₂ and TiO₂ show permittivities ranging from 2.1 to 50, ferroelectrics offer maximum permittivities of 1000 - 5000 at low frequencies (<100 MHz) and zero DC bias [25]. Above 5 GHz, the relative permittivities of barium titanate and BST are lowered to a value of some hundred [25]-[27].

The properties of ferroelectrics are strongly related to fabrication route and processing conditions, which alter the desired crystalline structure and affect the material’s permittivity tensor ε. Extensive investigations on BST thick films have been made by [26]. Slightly better performance with respect to the loss tangent has been published for the frequency range between 10 GHz and 30 GHz [28].

Ferroelectric phase shifters using CPW and loaded microstrip configurations have been demonstrated, working around 10 GHz and 60 GHz, both with insertion losses of approximately 10 dB at their operation frequency [29], [30]. Their tuning voltage ranges peak at 250 V and 320 V. Implementations using transmission lines with concentrated ferroelectric capacitors allow an improvement in insertion loss to 5 dB between 30 GHz and 40 GHz [31]. A phase shifter implementation with moderate tuning voltage operating up to 40 GHz is presented in [32].

1.3.3 Passive semiconductor phase shifters

Passive semiconductor phase shifters make use of integrated or hybrid electronic circuits to perform switching or tuning functions. The semiconductor components can be used as variable resistors or capacitors enabling continuously and digitally controlled phase shift and time-delay units. Despite of the wide variety of different phase shifter topologies and the diversity in semiconductor technologies, basically, passive semiconductor phase shifters make use of the switched-line, loaded-line or switched-filter concept.

The switched-line phase shifter routes between RF paths of different length providing a controlled time delay. This requires switches with a low insertion loss in the closed state and a high isolation in the open state, as provided by PIN diode switches. The PIN diode behaves like a current controlled resistor, which can exhibit an on resistance of 1.6 Ω at a bias current of some 10 mA. Its off-state resistance can exceed 10 kΩ, whereas the off-state capacitance can be reduced to some 10 fF at a reverse bias voltage of 10 V.

In a loaded-line phase shifter, the characteristic impedance and the propagation constant of a transmission-line is manipulated. This is typically realized through cascading sections of transmission-lines which can be switched or variable coupled to a lumped or distributed element. As a result, each section creates a predominant reactance or susceptance loading. For example, susceptance loading can be obtained by using voltage controlled capacitors, realized by connecting the source and drain of the transistor. A CMOS-based phase shifter of this type has been demonstrated in [7]. The phase shifter shows nearly zero power consumption and an insertion loss of about 10 dB at 60 GHz.
The switched-filter phase shifter typically consists of lumped elements arranged in a Π or T network. PIN diodes or FET switches are used to select between the different filter paths. In contrast to PIN diodes, FET switches behave like a voltage controlled resistor. They usually have a somewhat higher insertion loss and consume significantly less power. The isolation of FET switches is limited by the drain-source capacitance with typical values around 100 fF. A switched-filter phase shifter using MOS switches in TSMC 90 nm technology has been presented by [33]. The phase shifter shows an average insertion loss of 15 dB at 57 - 64 GHz. Reflection type switched-filter phase shifters with 6 to 7 dB insertion loss around 60 GHz using varactors in CMOS and BiCMOS technology can be found in [34] and [35]. A switched-filter phase shifter using InGaAs PIN diodes has been presented in [36]. This PIN diode phase shifter shows a maximum insertion loss of 7.8 dB at 26 - 30 GHz.

As a shortcoming of passive semiconductor phase shifters, power consumption noise and intermodulation can limit the system performance. Due to their non-linearity, PIN diode and FET switches produce harmonics and intermodulation. While operating at high current, shot noise dominates in PIN diodes. Next to the above mentioned technologies pHEMTs in GaAs have found application as switches and in phase shifters. Phase shifters in this technology as provided by TriQuint exhibit an insertion loss of 7 dB at 32 - 37 GHz [37].

1.3.4 MEMS phase shifters

MEMS phase shifters have been researched and developed since the 1990s, with the goal of changing the phase with a minimum phase error and insertion loss. Phase shifters with variations in architecture and implementation have been realized and published showing remarkable performance [14], [38]-[44]. MEMS phase shifters have demonstrated insertion losses of 3.0 dB to 5.7 dB, an average phase error of 3° to 13° and a fractional bandwidth of up to 10%. The centre operation frequencies of these phase shifters lay between 60 and 78 GHz. Theses MEMS phase shifters make use of mechanical switches which exhibit a galvanic separation at DC and high isolation up to SHF and EHF. The miniaturization combined with the electro-static actuation mechanism enables transition times of 1 µs to 10 µs and an energy consumption of some µWs per cycle.

The MEMS switch acting as a basic building block for RF phase shifters was the first RF MEMS component requiring qualification [45]. Since the early 2000s, the major challenge with MEMS switches are electrical charging problems with dielectric layers on top of the DC bias electrode which result in a pull-down and restoring voltage drift. The lack of studies concerning the switch lifetime and a low fabrication yield confirmed that MEMS still demand technological improvements. For most applications MEMS switches are required to exceed $10^9$-$10^{10}$ cycles [14]. Many studies and analyses have been undertaken in order to solve the dielectric charging and contact welding problems.
Companies, such as RADANT or MEMtronics with prototype switches started to characterise their devices intensively with regard to their long-term reliability [46].

Charging problems can occur in switches consisting of stacked dielectric layers or in dielectric-passivated switches. During DC actuation at electrical fields as high as 30 V/µm, charges can be injected into the dielectrics and trapped in interfacial, surface and bulk states. Due to their long relaxation times the charge accumulates and causes an actuation voltage drift or self-actuation. Different solutions for the charging problem have been presented. The approaches include bipolar actuation [47], dielectric switches with isolated stoppers [46], as well as dielectric-less switches with isolated stoppers [48]. Unlike the bridge-shaped capacitive switch, the ohmic-contact cantilever switches can be easily implemented in series and shunt configuration without a dielectric passivation. Their discontinuous stiffness constant can prevent a short between the bias electrode and the lever while the contact force ensures reliable low-resistance contact [49].

Today, packaging issues seem to pose lasting difficulties. RF MEMS packaging requires feeding transmission lines (TL), which add to the overall insertion phase. At frequencies above 20 GHz the feed lines make up a significant part of one wave length. Hence, a reflective type RF MEMS switch is approximated by a reconfigurable filter rather than a lumped high isolation switch. Above 20 GHz, single packaged switches are not very attractive due to multiple electrical interfaces and the additional insertion loss [14].

MEMS based phase shifters surpass competitive state-of-the-art phase shifters in terms of low insertion loss, high bandwidth, low bias power consumption and high linearity. The technology can provide a substantial increase in receiver and transmitter performance and reduce the system complexity in ESA antennas. Within the past two decades, typical operation frequencies for communication and radio detection and ranging (RADAR) systems have shifted from super high frequencies (3 GHz - 30 GHz) to extremely high frequencies (30 GHz - 300 GHz). At mm-wave frequencies (>30 GHz), passive power dividers, combiners and SPDT switches can significantly add to the total loss of ESA antenna networks. At ever higher frequencies ohmic contact MEMS switches require larger physical dimensions in order to decrease the open-state capacitance needed to provide comparable isolation. Alternatively, RF switches or first order filters are replaced by higher order frequency selective networks, which results in improved isolation. Programmable or tuneable MEMS can be implemented to form inductive and capacitive loaded transmission-line sections, which enable a modular architecture of filter and phase shift networks.

The MEMS phase shifter presented here uses ohmic contact MEMS switches on a dielectric substrate (figure 1.6). The phase shifter shows outstanding low insertion loss and high-precision phase shift, whereas the implemented MEMS switches enable programmable circuits. The switch architecture avoids charging problems due to the absence of any solid state dielectric between the DC bias electrode and the movable lever. Contact between the bias electrode and the lever is prevented by
a discontinuous stiffness constant. The switching dynamics were studied using a fully parameterized model. A copper lever is used to reduce contact bouncing and to keep fabrication costs low.

The ohmic contact MEMS switches are implemented into reconfigurable loaded-line (RLL) sections, which can be inductively or capacitively loaded. These reconfigurable loaded microstrip sections are useful building blocks for the construction of frequency selective networks operating at EHF.

![Microscope image showing a partial view of a DMTL phase shifter operating at 60 GHz. The network consists of alternating capacitive and inductive loaded RLL sections. The tapering function or the section loading results in a Chebyshev low-pass filter response.](image)

1.4 Content overview

This dissertation comprises a comprehensive documentation of the modelling, fabrication and characterization of RF MEMS components. The components described are: an ohmic contact MEMS switch, reconfigurable loaded microstrip sections and a 60 GHz phase shifter with 5-bit resolution.

The major claims are the rigorous analysis of the electro-mechanical switch behaviour and the phase shifter design procedure.

Chapter 2 covers the theory, giving the definitions and basic relations used throughout the dissertation.

Chapter 3 deals with the ohmic contact RF MEMS switch. This chapter is subdivided into the sections on modelling, fabrication and characterization. Different innovations have been tested by the MEMS switch described here. Among these, the ohmic contacts are formed by the metal pairing of platinum and gold and show low ohmic contact resistance. The combination of a hard, low conductive and a soft, high conductive metal is reported to be preferential with respect to contact tribology and the electrical switching performance [50-52]. The static mechanics of the switch with its discontinuous stiffness constant abandons the use of a dielectric passivation on top of the bias electrode [49]. Further, a light-weight lever with high flexural stiffness is used to suppress contact bouncing.
In chapter 4, the loaded-line theory is presented and the basic network parameters are derived. The concept and implementation of reconfigurable loaded line (RLL) sections is described. Shunt-capacitive and series-inductive loaded microstrip sections are introduced. To the author’s knowledge, MEMS reconfigurable series-inductive and shunt-capacitive loaded microstrip sections are combined here for the first time. The equivalent circuits are given and the loading factors are derived.

In chapter 5, a new phase shifter design procedure is proposed and exemplary applied for the synthesis of a 60 GHz phase shifter with 5-bit resolution. The described phase shifter design procedure follows a rigorous analytical approach and links the element values of a low-pass prototype filter to the phase response. Limitations of the approach are illustrated by comparison to Momentum simulations. The effect of the number of RLL sections on the model accuracy is discussed. Chebyshev and Bragg low-pass prototypes are designed, fabricated and characterized. The effect of the average loading factor on the insertion loss is addressed. The measured Bragg and Chebyshev phase shifters are in good agreement with the intended designs and exhibit exceptional low insertion loss and high phase precision.

In chapter 6 the main results are discussed and conclusions are drawn. An outlook on prospective applications of the presented MEMS technology is given in chapter 7.
CHAPTER 2

Theory

2.1 Principles of phase shift and time-delay networks

The basic function of a phase shifter is to generate a time variable insertion phase between two transversal electromagnetic waves travelling from a terminal T1 to terminal T2. The two terminals can be understood as the source and detector placed at a defined location in space, e.g. T1 at position $z_1$ and T2 at position $z_2$, or alternatively correspond with the ports of a two-port network. The source continuously emits waves at a given frequency. The insertion phase difference can be expressed as the phase lag of the wave which has passed the network placed between the two terminals for the time $t_0 + \tau$ or state $n_m$ and the phase lag of a wave which has passed the network for the time $t_0$ or state $n_0$ (2.1). In practice, a reference state is more convenient than a reference time. Hence the insertion phase difference is expressed in terms of the states (2.2), whereas the insertion phases at the two times $t_0 + \tau$ and $t_0$ correspond to the states $n_m$ and $n_0$. By convention, the insertion phase difference $\Delta \phi_{m,0}$ of the state $n_m$ is expressed relative to the zero or reference state $n_0$, whereas the zero state typically exhibits the lowest negative insertion phase value. The subscript $m$ denotes the state number and is an integer value between $0 \ldots N - 1$, whereas $N$ is the maximum available number of states.

$$\Delta \phi_{t_0 + \tau, t_0} = \phi|_{t_0 + \tau} - \phi|_{t_0} = (\phi|_{z_2, t_0 + \tau} - \phi|_{z_1, t_0 + \tau}) - (\phi|_{z_2, t_0} - \phi|_{z_1, t_0})$$ (2.1)

$$\Delta \phi_{m,0} = \phi|_{n_m} - \phi|_{n_0}$$ (2.2)

The notation used distinguishes between the absolute phase state for a given space and time $\phi|_{z,t}$, and the insertion phase of a two-port network for a given state $\phi|_{n_m}$. The insertion phase difference between two different states occurring at different times is denoted by $\Delta \phi_{m0}$.

Phase shift is time-discrete or continuously controlled by a type of switch or an element with tuneable properties. For a continuous phase shifter with infinite states, we can replace $n_m$ by the tuning coefficient $\xi$ with a tuning range of $\xi_0 \ldots \xi_{\text{max}}$. The reference state can be expressed by $\xi_0$. The performance of tuneable phase shifters can be affected by a change in the physical materials properties resulting in drift of the tuning characteristics. Temperature changes cause thermal expansion and stress within the mechanical structure. Further, it affects the electrical resistance of conductors and matched load terminations. Digital phase shifters are more immune to changes in the physical material properties, since they typically use only few well-defined states.
Given a linear polarized transversal electromagnetic (TEM) wave with a defined frequency travelling in z-direction along a waveguide, its time delay and attenuation is altered by the effective wave number \( k_{z,\text{eff}} \) and the dissipation constant \( \alpha \) of the transmission line. Dependent on the time \( t \) and location \( z \) of detection, the electrical field (2.3) with amplitude \( \hat{E}_0 e^{-\alpha z} \) and argument (2.4) is observed.

\[
\hat{E}(r, t) = \hat{E}_0 e^{-\alpha z} e^{i\phi_y}
\]

\[
\phi = \omega t - k_{z,\text{eff}} z = \omega \left[ t - \frac{\epsilon_{r,\text{eff}} \mu_{r,\text{eff}}}{c_0} z \right]
\]

Using the two measurement terminals T1 and T2 at the positions \( z_1 \) and \( z_2 \), we detect the arguments (2.5). The insertion phase is measured by comparing the two arguments at the same point in time \( t \) (figure 2.1). An error \( dt \) in the detection time of the two ports would cause a phase deviation equal to \( \omega dt \).

\[
\phi|_{z_1,t} = \omega t + \phi_1
\]

\[
\phi|_{z_2,t} = \omega(t - t_d) + \phi_1
\]

\[
\phi|_l = \phi|_{z_2,t} - \phi|_{z_1,t} = \omega \frac{\epsilon_{r,\text{eff}} \mu_{r,\text{eff}}}{c_0} (z_2 - z_1) = \omega \frac{\epsilon_{r,\text{eff}} \mu_{r,\text{eff}}}{c_0} l = \omega(-t_d)
\]

\[\text{Figure 2.1.} \text{ Visualization of the phase states at the two terminals T1 and T2. The black rotated arrows indicate the absolute phase at one time } t_1 = t_2.\]

The insertion phase (2.6) of a real waveguide does not necessarily depend in a pure linear fashion on the frequency. Dependent on the chosen substrate and the transmission-line configuration, dispersion of \( \epsilon_{r,\text{eff}} \) and \( \mu_{r,\text{eff}} \) can severely contribute to the frequency characteristic of the phase.
Within a time $t$ or a state $n_m$, the effective permittivity $\varepsilon_{r,z,\text{eff}}$, the effective permeability $\mu_{r,z,\text{eff}}$, the path length $l = z_2 - z_1$ and hence the group time delay $t_d$ are assumed to remain constant. A change of the insertion phase can be expressed by the total differential $d\phi|_t$ of (2.6). By definition, an error in the detection time is not considered here.

$$d\phi|_t = \frac{\partial \phi}{\partial \omega} d\omega + \frac{\partial \phi}{\partial \varepsilon_{r,z,\text{eff}}} d\varepsilon_{r,z,\text{eff}} + \frac{\partial \phi}{\partial \mu_{r,z,\text{eff}}} d\mu_{r,z,\text{eff}} + \frac{\partial \phi}{\partial l} dl$$

$$= \left(\frac{\varepsilon_{r,z,\text{eff}}}{2c_0} \frac{\partial}{\partial \varepsilon_{r,z,\text{eff}}} \left(\frac{\varepsilon_{r,z,\text{eff}}}{\mu_{r,z,\text{eff}}} \right)\right) l d\omega - \frac{\omega}{2c_0} \frac{\mu_{r,z,\text{eff}}}{\varepsilon_{r,z,\text{eff}}} d\varepsilon_{r,z,\text{eff}}$$

$$- \frac{\omega}{2c_0} \frac{\varepsilon_{r,z,\text{eff}}}{\mu_{r,z,\text{eff}}} d\mu_{r,z,\text{eff}} - \frac{\omega}{c_0} \frac{\varepsilon_{r,z,\text{eff}}}{\mu_{r,z,\text{eff}}} dl$$

$d\omega, d\varepsilon_{r,z,\text{eff}}, d\mu_{r,z,\text{eff}}$ and $dl$ are deviations of the frequency, effective permittivity, effective permeability and the path length. The frequency sensitive part $\partial \phi / \partial \omega$ in equation 2.7 comprises the sum of two terms, which are the first and second order delay time observed by passing a wave guide section of length $l = z_2 - z_1$. The first term gives the delay time of an ideal transmission line. The second term accounts for the additional phase error if the wave undergoes dispersion.

For a pure dielectric waveguide, we can simplify this general description by assuming $\mu_{r,z,\text{eff}} = 1$ and $d\mu_{r,z,\text{eff}} = 0$ leading to equation 2.8, whereas $Z_0$ is the characteristic wave impedance. Since the wave impedance depends on the effective permittivity, deviations in $d\varepsilon_{r,z,\text{eff}}$ can affect the port impedance matching and return loss. A deviation of $dl$, on the other hand, does result in additional time delay only.

$$d\phi|_t = \left(\frac{\varepsilon_{r,z,\text{eff}}}{c_0} - \frac{\omega}{2} \frac{\partial}{\partial \omega} \left(\frac{\varepsilon_{r,z,\text{eff}}}{c_0} \right)\right) l d\omega - \frac{1}{2} \omega \varepsilon_0 Z_0 d\varepsilon_{r,z,\text{eff}} - \omega \frac{\varepsilon_{r,z,\text{eff}}}{c_0} dl$$

(2.8)

The three different influencing variables $\omega$, $\varepsilon_{r,z,\text{eff}}$ and $l$ in (2.8) can be used to realize a change of the insertion phase. The resultant insertion phase difference is given by the indefinite integral (2.9).

$$\Delta \phi = \int d\phi|_t = \frac{l}{c_0} \left(\frac{\varepsilon_{r,z,\text{eff}}}{c_0} - \frac{\omega}{2} \frac{\partial}{\partial \omega} \left(\frac{\varepsilon_{r,z,\text{eff}}}{c_0} \right)\right) d\omega - \frac{1}{2} \omega \varepsilon_0 d\varepsilon_{r,z,\text{eff}} - \omega \frac{\varepsilon_{r,z,\text{eff}}}{c_0} dl$$

(2.9)

While paraelectric materials exhibit a nearly frequency-invariant time delay $\partial \varepsilon_{r,z,\text{eff}} / \partial \omega \approx 0$, ferroelectric materials can produce additional signal distortion due to their frequency-dependence of permittivity.
A transmission line on a ferroelectric substrate providing tuneable effective permittivity can be used to control the differential insertion phase between different states. Along with the change in permittivity, the characteristic wave impedance $Z_c$ changes. Alternatively, a phase shift can be obtained by manipulation of the transmission-line length.

The group delay difference (2.11) can be derived from the insertion phase difference (2.10).

$$\Delta \tau_{m,0} = \tau_{d|m} - \tau_{d|n} = -\frac{\partial \phi|_{n_m}}{\partial \omega} - \frac{\partial \phi|_{n_n}}{\partial \omega} = -\frac{\partial (\Delta \phi_{m,0})}{\partial \omega}$$ (2.11)

In case of a true time-delay network, the time delay is frequency-independent. This requires an ideal phase response of $\phi|_t = a_0 - a_1 \omega$, whereas $a_0$ and $a_1$ are constants. Given a certain separation to the cut-off, the phase response observed in the passband of a filter can be approximated by a linear function. Various approximations such as Bessel, Butterworth, Chebyshev are well-known from common filter design [54]-[57]. The phase difference obtained by a tuneable or programmable filter can be expressed according to (2.12)

$$\Delta \phi_{m,0} = \int_{\phi|_{n_0}}^{\phi|_{n_m}} d\phi|_t = \phi|_{n_m} - \phi|_{n_0} \approx -(a_1|_{n_m} - a_1|_{n_0}) \omega$$ (2.12)

### 2.2 Mechanisms of phase shift controlling

The mechanisms to realize a change of the wave propagation in a two-port network can be categorized depending on whether it is based on a material effect or a switching or tuning architecture. From the above theory for TEM wave phase shifting, we can derive the four different basic principles of phase shifters implemented in transmission lines. These are:

- Manipulation of the line length $\Delta l$.
  - The implementation is called a switched-line phase shifter and it makes use of single pole double throw (SPDT) or single pole multiple throw (SPnT) switches to select transmission lines of different path length.
  - The material strain caused by the elasticity under mechanical stress or via thermo-mechanical coupling allows very small phase changes at EHF. However, this mechanism barely finds application in practice, since the insertion phase difference is usually below 0.3% of the total
insertion phase of the line. A technical implementation using a mechanism for changing the length of a coaxial line is the trombone phase shifter.

Manipulation of the effective line permeability $\mu_{r, x, eff}$.

- Implementations based on this principle are called ferrite phase shifters. Ferrimagnetic materials are used to enable a change in permeability $\Delta \mu_{r, x, eff}$ by a change of the magnetic field applied. These phase shifters are predominantly carried out in configuration of a rectangular waveguide. The planar implementation of a ferrite phase shifter is called a meander line phase shifter and uses a ferrimagnetic substrate.

Manipulation of the effective line permittivity $\varepsilon_{r, x, eff}$.

- The implementation is called a ferroelectric phase shifter, this is if $\Delta \varepsilon_{r, x, eff}$ is enabled by a change of the electrical field applied.
- Alternatively, mechanical actor architectures can change the local substrate permittivity by use of a movable graded composite dielectric. This dielectric can be placed on top of the transmission line or inserted in the slot of a line.

Manipulation of the phase response $\phi|_t$ of a frequency selective network.

- Specific implementations are called switched-filter phase shifter or loaded-line phase shifter. The first concept uses SPDT or SPnT switches to select between different filter paths, which exhibit a fixed frequency response and are separated in space. The other concept uses a tuneable or programmable filter instead. In contrast to the first concept, the phase shifter is one single two-port network, which uses reconfigurable loaded-line sections [58], [41], [42]. The approximated phase response of the phase shifters can be expressed by $\phi = \phi_0 - (\partial \phi/\partial \omega)\omega - \frac{1}{2} (\partial^2 \phi/\partial \omega^2)\omega^2 = a_0 - a_1 \omega - a_2 \omega^2$ for $\omega < \omega_c$. If $a_2$ is negligible, the time delay $t_d = -a_1 - 2a_2 \omega$ becomes frequency-independent, as desired for a true time-delay (TTD) phase shifter.

2.3 MEMS enabled programmable RF circuit architectures

RF MEMS switches are miniaturized mechanical structures. This mechanical structure is deformed by an external DC bias. In case of ohmic-contact cantilever switches, the RF path is separated by an air gap in the open state and galvanically connected in the closed state. The bridge-shaped capacitive-contact switch shorts the RF path in the closed state and provides low insertion loss in the through state.
Ohmic-contact switches can be implemented in series as well as in shunt configuration (figure 2.2), whereas capacitive-contact switches are implemented in shunt configuration, exclusively. The series configuration is typically used to provide high isolation from DC to 20 GHz and a low to moderate insertion loss. The shunt configuration enables high isolation above 10 GHz and low insertion loss. In combination broadband reflective and absorptive switches become feasible.

Figure 2.2. Reflective type switches in series and shunt configuration. Absorptive type switch combining switches in series and shunt configuration.

MEMS switches have long been considered to be a way of improving RF subsystems and circuits in terms of their insertion loss, power consumption, weight and size. Mechanical switches exhibit high linearity and outstanding intermodulation performance. Its miniaturization and implementation within a planar waveguide enables passive, low-loss and high bandwidth circuits to be realized. Single die MEMS switches are desired for routing matrices in T/R modules. Applied in automatic test equipment (ATE), MEMS switches are used for the testing of high speed circuits. Typically, these switches offer a high isolation up to SHF [59].

RF MEMS switches, like the ones presented here, are intended to be integrated in the metallization traces on the substrate. The precise dimensions and defined insertion of the device make it suitable for operation up to EHF5. This approach enables the realization of reconfigurable loaded microstrip sections (RLL) as depicted in figure 2.3. RLL can be used as basic design components for the filter and phase shifter synthesis. The integrated switches enable selection between different digital states in order to provide a time-dependent magnitude and phase response.
Figure 2.3. First order high- and low-pass response networks. Reconfigurable loaded microstrip sections in series and shunt configuration.

Applications of on-chip RF MEMS circuitry are: impedance tuning networks used to match the input and output impedance of antennas, PAs or LNAs working at different frequency bands, reconfigurable antennas for multi-band communication, digitally controlled filters used for band selection in future front-end of mobile communication systems, tuneable filters for adaptive networks and reconfigurable band-stop / band-pass filter adopted for high isolation switching functions [60]-[67], [14], [15], [45].
CHAPTER 3

Electrostatically actuated ohmic contact MEMS switch

3.1 Introduction

This chapter is subdivided into the sections of: modelling, fabrication and characterisation. The model includes a detailed description of the electrostatically actuated ohmic contact MEMS switch. Simulation results are illustrated and explained. The MEMS fabrication process summarizes the essential process steps and describes the process execution. An extensive characterisation of the RF MEMS switch follows at the end of this chapter. The measurement results are illustrated and compared to the simulation results. The fabricated RF switch performs well, in good agreement with the modelled data.

3.2 MEMS switch model

In order to determine the geometry, material and operation condition dependent electro-mechanical behaviour of the MEMS switch, a comprehensive analytical model has been developed [49]. This model describes the MEMS switch in the static and dynamic cases. The parameterized model has been implemented into MATLAB Simulink, which allows for fast calculation of the switch behaviour. Thus changes in the switch dimensions, cantilever material and gas film properties can be considered easily. Moreover, the electrical and mechanical dynamic system response of the electrostatically actuated MEMS switch can be studied with regard to a user-defined actuation signal.

To validate the mathematical model, the energy balance is solved assuming elastic bouncing. Taking into account non-elastic bouncing, a momentum absorption coefficient is introduced which allows the mechanical contact energy to be estimated. Considering the energy transfer and consumption enables an efficient electrostatic actuator to be developed by reducing the dissipated energy in the resistor, the squeezed gas film, and by contact bouncing.

The calculated switching dynamics are confirmed by the electrical characterization of the developed RF switch.
3.2.1 The electro-mechanical system approximation

The proposed mathematical-physical model uses a one-dimensional approximation of the MEMS switch. The model depicted in figure 3.1 accounts for the lumped discontinuous lever stiffness \( k(x_s) \), the lumped non-linear gas film damping \( b(x_s, t) \), and the lumped equivalent resonant mass \( m(x_s, \omega_i) \). The lumped descriptions represent the dynamics of the lever with respect to the load’s centre of acceleration, which is at \( x = x_s \). A detailed description to the symbols applied by the model is given in table 3.1.

![Figure 3.1. MEMS switch model, consisting of the squeeze-film damped cantilever and the bias circuit. The dynamic behaviour of the model can be represented by a one-dimensional mass-spring-dashpot mechanism.](image)

The electro-mechanical model consists of a free-standing cantilever made of copper. The driver circuit comprises the series resistor \( R_b \) and the variable capacitance \( C(t) \) formed by the moveable cantilever and the bias line electrode. These two electrodes are separated by the gas film. Additional solid dielectric layers may be introduced on top on the bias electrode. The switch is subdivided along the x-direction into the lengths \( l_1, l_2 \) and \( l_3 \), which represent sections of different cross-section and loading conditions. Due to the voltage drop across the gap \( g_0 - z(x_s, t) \), the lever bends downward and mechanical contact is obtained at \( x = l_3 \). With increasing contact force a decrease in the electrical contact resistance is observed. By turning down the bias voltage the cantilever is restored to its initial position.

A) The lumped excitation force

In order to define the lumped mass-spring-dashpot mechanism, the exciting force needs to be determined with respect to the load’s centre of acceleration \( x = x_s \). Assuming an electrostatic load \( q(x, z) \) acting between \( x = l_1 \) and \( x = l_2 \) and a linear decreasing gap \( z(x) \) given by (3.1) and figure 3.2, the centre of acceleration is determined by the weight method (3.2).
Figure 3.2. Load’s centre of acceleration $x = x_s$ of a constant sloped cantilever, with a non-linear distributed load $q(x, z)$ acting between $l_1$ and $l_2$.

For the dimensions given in table 3.1, the action due to gravity on the mass load $\rho w h$ becomes negligible compared to the average electrostatic load $\bar{q}(x)$.

$$x(x) = z_0 - \frac{z(l_2) - z(l_1)}{l_2 - l_1} x$$  \hspace{1cm} (3.1)

$$x_s = l_1 + \frac{\int_{l_1}^{l_2}(x - l_1) F_{\text{ext}}(x) \, dx}{\int_{l_1}^{l_2} F_{\text{ext}}(x) \, dx} = l_1 + 0.718(l_2 - l_1)$$  \hspace{1cm} (3.2)

The electrostatic force $F_{\text{ext}}$ can be calculated according the equations (3.3) and (3.4). It represents a lumped force determined with respect to the reference point $x = x_s$. A parallel plate arrangement is assumed to calculate the variable bias capacitance, neglecting fringing fields occurring along the edges of the electrodes. Staked dielectrics may be considered by an additional dielectric passivation layer of thickness $d$, which adjoining the air gap $g_0 - z(x_s, t)$.

$$F_{\text{ext}}(x, t) \big|_{x=x_s} = \frac{dE_{\text{ext}}}{dx} = \frac{d}{dx} \left( \frac{1}{2} C(t)V^2(t) \right) = \frac{w_2(l_2 - l_1)V^2(t)}{2\varepsilon_0\varepsilon_{ra}x^2(x, t)}$$  \hspace{1cm} (3.3)

$$x(z, t) \big|_{x=x_s} = \frac{d}{\varepsilon_0\varepsilon_{rd}} + \frac{g_0 - z(x_s, t)}{\varepsilon_0\varepsilon_{ra}}$$  \hspace{1cm} (3.4)

The non-linear lumped electrostatic force $F_{\text{ext}}(z, t) \big|_{x=x_s}$ at $x = x_s$ is gained from (3.3) and (3.4). If we avoid using an additional dielectric passivation layer (3.5) simplifies by $d = 0$.

$$F_{\text{ext}}(z, t) \big|_{x=x_s} = \frac{w_2(l_2 - l_1)V^2(t)}{2\varepsilon_0\varepsilon_{rd} \left( \frac{d}{\varepsilon_0\varepsilon_{rd}} + \frac{g_0 - z(x_s, t)}{\varepsilon_0\varepsilon_{ra}} \right)^2}$$  \hspace{1cm} (3.5)

B) The lumped stiffness coefficient

The stiffness of the cantilever is best chosen as a trade-off between different target values. Target values are the off-state capacitance, the on-state resistance, and the maximum RF power. At high RF power, the lever can latch in the down state or even be self-actuated. The off-state capacitance is primarily determined by the contact separation and the area of the contact fingers. In order to provide high isolation up to millimetre-wave frequencies, a large contact gap is desired. The on-state resistance
depends on the contact force. The contact resistance is minimized by choosing a highly conductive and soft metal. Hot-switching at high RF power can cause switch latching in the on state, when the DC bias is turned off. The range of hot-switching may be extended by increasing the restoring force.

Considering the above target values as the specifications, the cantilever stiffness coefficient and the actuation voltage become dependent. Typical values of the stiffness constant measured at the cantilever tip are between 30 N/m and 90 N/m. The corresponding actuation voltage ranges from 30 V to 90 V.

The displacement curve $z(x)$ is derived for the cantilever divided into the three sections of $0 \leq x \leq l_1$, $l_1 \leq x \leq l_2$, and $l_2 \leq x \leq l_3$. For static conditions, equation 3.6 relates the load to the displacement curve, where $I_y$ denotes the momentum of inertia. Assuming a plate and small deformation, the Young’s modulus along the neutral axis $\tilde{Y}$ can be expressed by (3.7).

$$\frac{d^2}{dx^2} \left( \tilde{Y} I_y \frac{d^2 z(x)}{dx^2} \right) = q(x) = -F'(x) = -M'_y(x)$$  \hspace{1cm} (3.6)

$$\tilde{Y} = \frac{Y}{1 - \nu^2}$$  \hspace{1cm} (3.7)

Basically, the mechanical properties of the metallization can be altered by impurities, the crystallite orientation distribution as well as the crystallite size. Hence, the deposition equipment, the additive and the deposition parameters can significantly influence the film structure and mechanical properties. The Young’s modulus and Poisson’s ratio of isotropic copper bulk is $Y = 67$ GPa and $\nu = 0.42$, respectively. These values are calculated out of the single crystalline elastic constants of pure bulk copper given in reference [68]. However, the obtained values differ significantly from reported values of electroplated copper. Experimentally determined values for the Young’s modulus of plated copper are published in [69], [70] to be 110 GPa - 130 GPa.

The displacement curve $z(x)$ is given in (3.9) assuming a uniform distributed load of $\tilde{q} = F_{ext}/(l_2 - l_1)$ acting between $x = l_1$ and $x = l_2$. The individual flexural stiffnesses $\alpha_1$ and $\alpha_2$ are given by (3.8).

$$\alpha_1 = \tilde{Y} I_{y1} = \frac{1}{12} \tilde{Y} w_1 h^3$$  \hspace{1cm} (3.8)

$$\alpha_2 = \tilde{Y} I_{y2} = \frac{1}{12} \tilde{Y} w_2 h^3$$
The lumped stiffness coefficient $k(x_s)$ of the cantilever shown in figure 3.3 is derived in (3.10). For constant proportions, the section lengths $l_1$, $l_2$ and $w_1$ can be expressed in terms of the length $l_3$ and width $w_2$ of the cantilever (3.11).

![Figure 3.3](image)

**Figure 3.3.** Cantilever geometry of the micro-mechanical switch. The lever is fixed on the left side.

$$k(l_1 \leq x_s \leq l_2) = \frac{d\dot{q}(l_2 - l_1)}{dx(x_s)} = \frac{\dot{q}(l_2 - l_1)}{z(x_s)}$$

$$= 24(l_2 - l_1) \left[ -\frac{2l_1(l_1 - l_2)(l_2^2 - 3l_1l_2 + 6l_2x)}{a_1} + \frac{(l_1 - x_s)^2(3l_1^2 - 8l_1l_2 + 6l_2^2 + 2l_1x_s - 4l_2x + x_s^2)}{a_2} \right]^{-1}$$

(3.10)

$$k(x_s) = 0.918 \frac{\varrho w_2 h^3}{l_3^3}$$

(3.11)
Table 3.1. Switch dimensions, material properties and characteristic parameters

<table>
<thead>
<tr>
<th>Description</th>
<th>Symbol</th>
<th>Design values</th>
</tr>
</thead>
<tbody>
<tr>
<td>length section 1</td>
<td>( l_1 )</td>
<td>25 µm</td>
</tr>
<tr>
<td>length section 2</td>
<td>( l_2 )</td>
<td>130 µm</td>
</tr>
<tr>
<td>length of cantilever</td>
<td>( l_3 )</td>
<td>165 µm</td>
</tr>
<tr>
<td>centre of acceleration ( x_s = l_1 + 0.718(l_2 - l_1) )</td>
<td>( x_s )</td>
<td>100 µm</td>
</tr>
<tr>
<td>support width</td>
<td>( w_1 )</td>
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</tr>
<tr>
<td>cantilever width</td>
<td>( w_2 )</td>
<td>110 µm</td>
</tr>
<tr>
<td>bias electrode area</td>
<td>( A_e )</td>
<td>12100 µm²</td>
</tr>
<tr>
<td>height of cantilever</td>
<td>( h )</td>
<td>4.0 µm</td>
</tr>
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<td>sacrificial layer thickness (initial actuation gap)</td>
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</tr>
<tr>
<td>initial contact separation at ( x = l_3 )</td>
<td>( z_0 )</td>
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</tr>
<tr>
<td>number of contacts</td>
<td>( N )</td>
<td>2</td>
</tr>
<tr>
<td>contact overlap width</td>
<td>( W_{ov} )</td>
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</tr>
<tr>
<td>contact overlap length</td>
<td>( L_{ov} )</td>
<td>15 µm</td>
</tr>
<tr>
<td>single contact area</td>
<td>( A_c )</td>
<td>225 µm²</td>
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<td>lateral conductor gap</td>
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</tr>
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<td>dielectric passivation thickness</td>
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<td>dielectric constant of passivation</td>
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<td>estimated Young’s modulus of electroplated Cu</td>
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<tr>
<td>estimated Poisson ratio of electroplated Cu</td>
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<td>plain-strain Young’s modulus</td>
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<tr>
<td>density of Cu cantilever</td>
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<tr>
<td>dynamic viscosity of gas</td>
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<td>17.1 µPas</td>
</tr>
<tr>
<td>effective viscosity of squeezed gas film (slip-flow)</td>
<td>( \eta_{eff}(K_a) )</td>
<td>14.2 µPas</td>
</tr>
<tr>
<td>density of gas</td>
<td>( \rho_{g} )</td>
<td>1 g/m³</td>
</tr>
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<td>bias line resistance</td>
<td>( R_b )</td>
<td>2 · 40 kΩ</td>
</tr>
<tr>
<td>actor capacitance in the up state</td>
<td>( C_{up} )</td>
<td>76 fF</td>
</tr>
<tr>
<td>actor capacitance in down state</td>
<td>( C_{down} )</td>
<td>≈ 260 fF</td>
</tr>
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<td>calculated stiffness constant ( k ) at ( x = x_s ) (free condition at ( x = l_3 ))</td>
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<td>221 N/m</td>
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<tr>
<td>calculated stiffness constant ( k^* ) at ( x = x_s ) (contact condition ( x = l_3 ))</td>
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<tr>
<td>equivalent mass</td>
<td>( m )</td>
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</tr>
<tr>
<td>simulated first Eigen frequency</td>
<td>( f_1 )</td>
<td>69.3 kHz</td>
</tr>
</tbody>
</table>

At a certain electrostatic load the switch closes, \( z(l_3) = z_0 \), and mechanical contact is reached. With increasing load \( \bar{q} \) the contact force \( F_c \) rises (figure 3.5) and more energy is stored in the bending moment. Figure 3.4 shows the internal force \( F_{q_e}(x) \), bending moment \( M_{by}(x) \) and displacement curve \( z^*(x) \) for the load \( \bar{q} \) of 4 N/m. The closed-form description of the displacement curve \( z^*(x) \) is given in (3.12).
Figure 3.4. Internal force $F_{qq}(x)$, bending moment $M_{by}(x)$ and displacement curve $z^*(x)$ of the cantilever at the load $\bar{q} = 4.0 \, \text{N/m}$. The contact force is determined to $F_c = -100 \, \mu\text{N}$.

For the load case with $F_c \neq 0$, the stiffness coefficient $k^*(x_s)$ is calculated by (3.15). In order to distinguished between the two load cases with $F_c = 0$ and $F_c \neq 0$, the superscript asterisk is introduced. Since a switching cycle comprises both load cases, the stiffness coefficient exhibits a discontinuity with a significant increase of the stiffness coefficient when in the down state. In the event of contact the stiffness coefficient exceeds the derivative of the electrostatic load with respect to the $z$-axis, thus electrode discharge is prevented and a dielectric passivation layer becomes
superfluous. Dielectric passivation layers on top of the bias electrode can result in an actuation voltage drift, self-actuation and electrode sticking due to charges trapped in the dielectric stack.

![Graph showing force and voltage relations](image)

**Figure 3.5.** Contact force $F_c(l_3, z = z_0)$, extrapolated restoring force $F_R(l_3, \bar{q} = 0)$ and applied voltage $V(\bar{q})$ versus average electrostatic load $\bar{q}$. The force becomes dependent on $\bar{q}$ when contact is established at the lever tip.

The resultant contact force $F_c(\bar{q})$ and the restoring force $F_R(\bar{q})$ at $x = l_3$ are given by (3.13) and (3.14). The force value at the intersection of $\bar{q} = 0$ with the extension of the contact force curve represents the restoring force $F_R(l_3, \bar{q} = 0)$ of -45 $\mu$N acting on the lever contact tip if the bias voltage is turned off. In figure 3.6, the stiffness function $k(x_s, \bar{q})$ and the displacement $z(x_s)$ is depicted in dependence of the average load $\bar{q}$ applied.

$$F_c(q, l_3, z = z_0) = \frac{1}{8(l_1 - l_3)^2 \alpha_1 - 8l_1(l_1^2 - 3l_1l_3 + 3l_3^2)\alpha_2} - \frac{2l_1(l_1 - l_2)(l_1^2 - 3l_1l_2 + 6l_2l_3)\alpha_2}{24z_0\alpha_1\alpha_2}$$

(3.13)

$$F_c(l_3, z = z_0) = -0.2193l_3\bar{q} + 0.1704z_0 - \frac{\bar{q}w_2h^3}{l_3^3}$$

$$-F_R(l_3, \bar{q} = 0) = -\frac{3z_0\alpha_1\alpha_2}{(l_1 - l_3)\alpha_1 - l_1(l_1^2 - 3l_1l_3 + 3l_3^2)\alpha_2}$$

(3.14)

$$-F_R(l_3, \bar{q} = 0) = 0.1704z_0 - \frac{\bar{q}w_2h^3}{l_3^3}$$

$$k^*(l_1 \leq x_s \leq l_2) = \frac{d\bar{q}(l_2 - l_1)}{dx(x_s)} = \left(\frac{dx(x_s)}{d\bar{q}}\right)^{-1} (l_2 - l_1)$$

(3.15)

$$k^*(x_s) = 9.333 - \frac{\bar{q}w_2h^3}{l_3^3}$$

- 31 -
C) The lumped damping coefficient

Generally, micro-electro-mechanical systems, such as accelerometers, gyroscopes, membrane based pressure sensors, acoustic transducers and switches interact with the fluid environment. Their dynamic system response is significantly altered by the fluid density and viscosity. Typically, sensors and actors should provide a fast step response and a short settling time. Thus, system damping is required. Most miniaturized sensors and actors make use of a kind of squeeze-film damping.

The squeeze-film damping primarily depends on the cantilever velocity and the fluid film thickness. It exhibits a non-linear behaviour with decreasing gap \( g_0 - z(x, t) \). Assuming isothermal conditions, low pressure variation \( dp(x, y)/dt \approx 0 \) within the fluid and small relative film thickness changes, the Reynolds equation can be simplified to the heat transfer equation (3.16).

\[
\text{div} \left[ \frac{(g_0 - z(x, t))^3}{12 \eta_{\text{eff}}(T)} \nabla p(x, y) \right] = \frac{g_0 - z(x, t)}{p} \frac{dp(x, y)}{dt} + \dot{z}(x, t)
\]

(3.16)

Finite element simulations can be used to extract an exact damping coefficient in dependence of the squeeze-film thickness and cantilever velocity. For simplicity, the switch is approximated by a circular disc with the radius of \( R = w_z/2 \), which allows an analytical expression of the pressure field \( p(x, y) \) as a function of the velocity \( \dot{z}(x, t) \). By integration over the pressure field, the damping coefficient (3.17) can be derived [71].

\[
b(x_s, t) = \frac{dF}{dv} = \frac{3 \eta_{\text{eff}}}{2} \frac{\pi R^4}{(g_0 - z(x_s, t))^2}
\]

(3.17)
A Knudsen number $K_n$ between 0.03 and 0.04 indicates the slip-flow regime. As characteristic to this regime, the viscosity reduces due to lower intermolecular interaction \[72\]. Thus, the effective viscosity $\eta_{eff}$ is used instead \[3.18\]. The dynamic viscosity and the corresponding effective viscosity for the chosen switch geometry can be found in table 3.1.

D) The mechanical Eigen frequency

Generally, the first natural Eigen frequency limits the dynamic range of a sensor or actor. The restoring and actuation time of a mechanical sensor or actor can be reduced by a higher Eigen frequency, which is essential for MEMS switches. Considering the mechanical switch as a transversal vibrating Euler-Bernoulli beam, equation (3.19) can be applied \[73\].

\[
\frac{d^2}{dx^2} \left( \frac{q(y)}{I_y} \frac{d^2z(x,t)}{dx^2} \right) + \rho wh \frac{d^2z(x,t)}{dt^2} = q(x,t) \tag{3.19}
\]

For the initial conditions $z(x, t = 0) = z_0$ and $z'(x, t = 0) = 0$, the solutions to (3.19) are given by (3.20), with the angular Eigen frequency expressed by (3.21). Equation 3.20 describes the free oscillating cantilever when changing from the down state to the up state. Since the one-dimensional model proposed in this chapter is required to provide an approximate description of the dynamic switch response, its mechanical resonance frequency $\omega_i$ is equal to that of the beam calculated in (3.21).

\[
z_i(x, t) = z_i(x) z_i(t) \\
\quad = z_0 \left\{ \cos \left( \frac{x}{l} \lambda_i \right) - \cosh \left( \frac{x}{l} \lambda_i \right) \right. \\
\quad \left. + \cos(\omega_i t) \right\} \left[ \sin \left( \frac{x}{l} \lambda_i \right) - \sinh \left( \frac{x}{l} \lambda_i \right) \right] \left\{ \sin(\omega_i t) \right\} \tag{3.20}
\]

\[
\omega_i = \frac{\lambda_i^2}{l^2} \sqrt{\frac{qI_y}{\rho wh}} \tag{3.21}
\]

For a beam with a constant cross-section $w_2 h$, the first Eigen frequency is given by $f_i = 98.3$ kHz. The Eigen frequency of the exact switch geometry depicted in figure 3.3 was calculated by finite element simulation using COMSOL. The simulation settings and results are given in table 3.2.
<table>
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<th>Description</th>
<th>Values</th>
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</tr>
<tr>
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<td>Eigen frequency, Eigen frequency</td>
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<td>element type</td>
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</tr>
<tr>
<td>first Eigen frequency $f_{xc}$</td>
<td>69.2 kHz, 365.2 kHz</td>
</tr>
</tbody>
</table>

**E) The lumped equivalent mass**

In order to end up with a system exhibiting the lumped stiffness coefficient $k(x_s)$ and the mechanical Eigen frequency $\omega_1$ as described in the previous sections, the equivalent mass $m(x_s)$ is determined by (3.22).

$$m(x_s, \omega_1) = \frac{k(x_s)}{\omega_1^2}$$  \hspace{1cm} (3.22)

**3.2.2 The mechanical subsystem**

The mechanical subsystem is described by the inhomogeneous differential equation of motion (3.23). It uses the lumped equivalent mass, damping coefficient, stiffness coefficient and excitation force. In the case of a critical damped system with a constant damping coefficient the system is balanced by the ratio $b(x_s)/k(x_s) = 2$.

$$m(x_s, \omega_1)\ddot{z}(x_s, t) + b(x_s, t)\dot{z}(x_s, t) + k(x_s)z(x_s, t) = F_{ext}(x_s, t)$$  \hspace{1cm} (3.23)

The solution to this equation describes the time-dependent displacement of the cantilever at the load’s centre of acceleration $x = x_s$. The displacement at the cantilever tip $x = l_3$ can be determined using the transformation factor $\beta$ defined in (3.24). The transformation factor shows proportionality for the displacement between the two different positions along the lever. Proportionality is given if $F_c = 0$ and the damping action does not influence the lever displacement curve.

$$\beta = \frac{\ddot{z}(l_3, q)}{\ddot{z}(x_s, q)}_{z(l_3) < z_0}$$

$$\beta = \frac{\ddot{z}(l_3, q)}{\ddot{z}(x_s, q)}_{z(l_3) < z_0} = \frac{1.100 \mu m}{0.592 \mu m} = 1.857$$  \hspace{1cm} (3.24)

In the event of contact $t = t_c$, the momentum due to the moving equivalent mass is reflected and the cantilever velocity changes its sign. This bouncing effect is well-known from conventional mechanical relays. Accounting for the bouncing effect, the non-harmonic parametric condition (3.25) is applied during solving (3.23).
\[ \dot{x}(t = t_{ci}) = -\kappa \dot{z}(t = t_{ci}) \tag{3.25} \]

With diminishing velocity due to squeeze-film damping and momentum absorption, bouncing is followed by sustaining mechanical contact. The cantilever can start a vibrating mode, whereas the lever tip remains in contact with the lower metallization, and finally changes to its static equilibrium. The static contact force is built up while the vibrating mode fades.

### 3.2.3 The electrical subsystem

The electrical subsystem models the bias network and is coupled to the mechanical subsystem by the electrostatic excitation force. The response of the mechanical subsystem is linked back to the time-dependent capacitance of the electrical subsystem. The advantage of modelling both parts including coupling, allows the influence of different actuation waveforms to be studied in detail. The applied waveform can be also considered as transmitted by the RF path while the bias pad is on DC ground, thus the quality factor and the system resonance frequency can be estimated.

The bias line is modelled by a bias line resistor \( R_b \) and a capacitor \( C(t) \). The voltage at the actuation electrode \( V(t) \) is given by \( V(t) = V_{ext}(t) - R_b i(t) \). The total bias current \( i(t) \) comprises the current components due to the capacitance change \( \dot{C}(t) \) and the change of the electrode potential \( \dot{V}(t) \), as shown in (3.26).

\[
i(t) = \frac{dQ(t)}{dt} = \frac{d(V(t)C(t))}{dt} = V(t)\dot{C}(t) + C(t)\dot{V}(t) \tag{3.26}\]

\[
\dot{V}(t) = \dot{V}_{ext}(t) - R_b \frac{di(t)}{dt} \tag{3.27}
\]

Using (3.26) and (3.27), equation (3.28) is gained. The feedback can be attributed to the coefficients of \( C(t) = w_2(l_2 - l_1)/\chi(t), \dot{C}(t) \) and \( \ddot{C}(t) \).

\[
\ddot{V}(t) = -\frac{1}{R_bC(t)} \left[ \ddot{V}_{ext}(t) - R_b\dot{C}(t)V(t) - \left( 1 + 2R_b\dot{C}(t) \right) \dot{V}(t) \right] \tag{3.28}
\]

### 3.2.4 Energy balance

In order to validate the mathematical model, the error is determined by means of the energy balance. The energy balance is given in (3.29). It comprises the total electrical energy applied to the system \( E_{ext}(t) \), the kinetic and potential energy of cantilever \( E_{kin}(t), E_{pot}(t) \), the energy dissipated in the squeeze-film and the resistor \( E_{dmp}(t), E_{res}(t) \) and the energy stored in the variable capacitor \( E_{cap}(t) \). Optionally, the amount of energy absorbed by contacts can defined by \( E_{abs}(t) \) using a
momenum absorption coefficient $\alpha > 0$. Considering momentum absorption allows tuning the decay of the bouncing independently of the transition time.

$$E_{\text{ext}}(t) = E_{\text{kin}}(t) + E_{\text{pot}}(t) + E_{\text{damp}}(t) + E_{\text{cap}}(t) + E_{\text{res}}(t) + E_{\text{abs}}(t)$$

$$\int V_{\text{ext}}(t) i(t) dt = \frac{1}{2} m \dot{x}^2(t) + \frac{1}{2} k x^2(t) + \int b \dot{x}^2(t) dt + \frac{1}{2} C(t) V^2(t) + \int R_b i^2(t) dt + E_{\text{abs}}$$

(3.29)

The energy absorbed by the contacts $E_{\text{abs}}(t)$ can be calculated by integration of the absorbed momentum (3.30) with the momentum reflection coefficient defined as $\kappa = (1 - \alpha) = -p_{\text{ref}}/p_{\text{in}}$.

Alternatively, the energy balance can be solved for $E_{\text{abs}}(t)$ at different values of $\alpha$.

$$E_{\text{abs}}(t) = \sum_{i=0}^{N} E_{\text{abs},i}(t) = \sum_{i=0}^{N} \int_{t_i}^{t_{i+\Delta t_i}} \alpha \frac{dp(t)}{dt} \, dt = \sum_{i=0}^{N} \int_{t_i}^{t_{i+\Delta t_i}} \alpha \left( \frac{m \dot{v}(t)}{dt} \right) \, dt$$

(3.30)

### 3.2.5 Simulation results

The above described mathematical model was implemented in MATLAB Simulink. The simulation results discussed below, refer to the micro-electro-mechanical switch given in figure 3.3 and defined by the parameters listed in table 3.1.

Subsequently, an actuation test signal with a pulse width of 30 $\mu$s and amplitude of 50 V is used. The rising and trailing edge of the test signal are $\pm 50$ V/$\mu$s.

### A) Dynamics of movement

The switch dynamics is primarily described by the time-dependent cantilever deflection $z(x, t)$ at the position $x = x_s$ (figure 3.7). The switch on and switch off can be understood as the step responses of the system. Using the transformation factor $\beta$ given in (3.24) allows to determine the corresponding deflection at the cantilever tip $x = l_3$. The two displacement curves plotted in figure 3.7 show that contact bouncing is followed by static equilibrium. In the static case, the cantilever bends downward at $x = x_s$ providing the static contact force.

In figure 3.8 the lever velocity at the position $x = x_s$ is depicted. When changing from the up-state to the down-state position, damping decelerates the lever prior to the event of contact. Subsequently, the contacts bounce. The bouncing diminishes due to squeeze-film damping and the momentum absorption by the contacts. When changing from the down-state to the up-state position, the cantilever exhibits free oscillation damped only by the squeezed film.
Figure 3.7. Time-dependent lever deflection at the positions of the contact tip and the load’s centre of acceleration. The transition time is 8 µs. The square wave test signal is switched on at \( t = 0 \) µs and switched off at \( t = 30 \) µs. The bias line resistance is \( R_b = 50 \) kΩ, the momentum absorption coefficient is \( \alpha = 0.2 \). A steady contact force is observed after 14 µs.

Figure 3.8. Time-dependent lever velocity at the load’s centre of acceleration, showing the system oscillation in the down state and the up state, including damping caused by the squeeze-film and the energy absorption by the contacts.

The time-dependent bias current (3.26) is plotted in figure 3.9. The current peaks during switch-on and switch-off. The peak magnitude is determined by the voltage change and the capacitance in the up state and down state, respectively. When the bias voltage is at its high level, charge oscillation is caused by the time-variable capacitance.

For the given bias line resistance and switch capacitance, the current shows a low time constant \( \tau = R_b C \) resulting in nearly ideal rectangular-shaped current peaks. The mechanical time-constant can be extracted from figure 3.7 and is approximately one decade larger.

If using multiple switches, the capacitance and hence the time constant is increased by the number of parallel switches driven by the same resistive bias line. Since the RF path needs to be decoupled...
from the DC paths, the resistance of the bias line is kept large. A trade-off is made in exchange between RF isolation and actuation speed. Along with switch speed, squeeze-film damping and the energy consumption can be reduced.

**Figure 3.9.** Time-dependent current. The current peaks are caused by the voltage increase \(i_{\text{on}} \cong C_{\text{up}} \dot{V}(t)\) and decrease \(i_{\text{off}} \cong C_{\text{down}} \dot{V}(t)\), when switched-on and switched-off, respectively. The saw tooth wave in between the two current peaks is caused by mechanically oscillating charges \(i_{\text{osc}} \cong V_{\text{on}}\dot{C}(t)\).

**B) Dynamics of energy flows**

The energy consumption and switch efficiency is shown in figure 3.10 and figure 3.11. The stored and dissipated energy components in the electrical and mechanical subsystem are plotted. For the given actuation signal and the switch dimensions summarized in table 3.1, 30% of the applied energy is dissipated, whereas 70% of the applied energy is stored and released again. The capacitively and mechanically stored energy reaches its maximum at 15 µs. After 60 µs, the applied energy equals the sum of the dissipated energy portions.

The energy loss in the bias line resistor of the electrical subsystem primarily depends on the actuation signal spectrum and the cut-off frequency of the electrical subsystem \(\omega_c = 1/(R_b C)\). For the chosen actuation signal and the value of \(R_b C\), the attenuation in the resistor makes less than 5% of the total dissipated energy. The dissipation in the mechanical subsystem is dominated by the squeeze-film damping. This damping effect is proportional to the velocity and exhibits nonlinear dependence on the thickness of the fluid film layer. The energy dissipation in the squeeze-film can thus be reduced in a trade-off against switching speed, which can be accomplished by reducing the voltage actuation signal bandwidth or magnitude.
Figure 3.10. Time-dependent energy components. About 30% of the applied energy is dissipated in a single switching cycle. For the given actuation signal, energy dissipation is primarily caused by squeeze-film damping.

Figure 3.11. Energy dissipated due to resistive attenuation makes up about 1% of the total input energy. The energy consumed by the contacts is below 1%, assuming $\alpha = 0.2$.

Assuming elastic bouncing with $\alpha = 0$ and $E_{abs} = 0$, the energy balance in (3.29) is used to determine the mathematical error and model uncertainty. For $t \leq 30 \mu s$, the error is below $\pm 0.1 \cdot 10^{-6} \mu Ws$ (figure 3.12). For $\alpha \geq 0$ the energy balance is used to determine the energy transferred to the contacts. In the special case of $\alpha = 0.2$ and a copper lever, the total energy per cycle transferred to the contacts is $1.5 \cdot 10^{-6} \mu Ws$. The time-dependence of the energy transfer is plotted in figure 3.12. The error at $t \geq 30 \mu s$ is caused by the discontinuous stiffness and the finite time-stepping while solving the coupled non-linear differential equations. Especially in the event of switch-off the discontinuities can cause a significant numeric error.
Figure 3.12. Mechanical energy consumed by the contacts, assuming $\alpha = 0.2$, and absolute numerical error, assuming $\alpha = 0$. The error for $t > 30 \mu s$ is caused by the discontinuous stiffness and the finite time-stepping.

The occurring forces during the switching cycle are plotted in figure 3.13. The forces are not in equilibrium unless the cantilever moves at constant velocity. A change in the moving direction is observed when the damping force at that time exhibits a zero crossing.

Assuming steady-state conditions, the electrostatic force and load in the down state are $F_{\text{ext}} = 225 \ \mu N$ and $\bar{q} = 2.05 \ \text{N/m}$, respectively. This corresponds to a contact force of $F_c = -29 \ \mu N$ as shown by figure 3.5. The excess force of $F_b(x_s) = k^*(x_s)z(x_s) = 97 \ \mu N$ is caused by cantilever bending in the down state (figure 3.14), while $F_c \neq 0$. By comparison of figure 3.14 with figure 3.5, the contact force magnitude is about 30% of the bending force.

Figure 3.13. Time-dependent force components, showing the external electrical force, the restoring force and the damping force. The forces are given at the centre of acceleration.
**Figure 3.14.** Time-dependent bending force $F_b(x_s) = k'(x_s)x(x_s)$. The curve shows the instant at which the bouncing period is completed and the contact force is ramped up. By comparison with figure 3.5, the contact force magnitude is about 30% of the bending force.

In figure 3.15, the system-specific response is plotted against the voltage at the pulse amplitude of 50 V. For this purpose, the voltage signal applied has a rise time of 1 ms, a pulse width of 0.1 ms and a decay time of 1 ms. The release voltage is found to be around 39 V. Iterative simulations showed that pull-in is achieved at pulse amplitudes exceeding 41 V. For the pull-in estimation the rise time is set to 1 $\mu$s.

**Figure 3.15.** Displacement-voltage trajectory calculated for the actuation pulse amplitude of 50 V. The release voltage of 39 V can be extracted from the static force equilibrium at a displacement $z(x_s) = 0.59 \mu$m. Attraction forces between the contacts are assumed to be zero.

### 3.2.6 RF power dependencies

The power handling capabilities of a MEMS switch express the operation limits in terms of a specific reliability function dependent on the RF power. The function maybe represented by the
dependent variable of: the number of cycles-to-failure in case of hot-switching and cold-switching, the probability for self-actuation or latching in the down state, or the normalized capacitance modulation. The switch lifetime is limited by the number of cycles the switch is capable of carrying out at a given level of RF power. Switch lifetime can be extended, if cold-switching conditions are provided. Sequencing can then be performed using a variable attenuator.

In the open state, the RF contacts form a metal-air-metal structure. This variable capacitor is formed by the overlapping area $NW_{ov}L_{ov}$ of $N$ contacts and the contact separation $z(l_3)$. At high power signals, capacitance modulation alters the frequency response dependent on the RF power, resulting in non-linearity and potential AM to PM distortion. At even higher RF power the switch tends to self-actuation and latching in the down state.

The schematic (figure 3.16) illustrates the RF grounds, the terminal loads, the bias capacitance $C$ and the contact capacitance $C_s$. In (3.31) the two capacitances are expressed in terms of the displacement $z(l_3)$, which is linked to $z(x_s)$ by the value of $\beta$. For the chosen load state, the transformation factor $\beta$ becomes dependent on the load ratio $F(l_3)/F(x_s)$ but is independent of Young’s modulus or the thickness of the cantilever. For $F(l_3) \leq F(x_s)$, the proportionality of $\beta \approx 1.857$ is a good approximation and shows less than 2% error.

\[
C_s = \frac{NW_{ov}L_{ov}}{\varepsilon_0\varepsilon_{rd}(z_0 - z(l_3))}
\]

\[
C = \varepsilon_0\varepsilon_{ra}\frac{w_2(l_2 - l_1)}{\varepsilon_0\varepsilon_{rd} - \beta z(l_3)/\beta}
\]

For an electrically short device, the voltage can be approximated to be constant along the cantilever. High isolation switches provide an open circuit at terminal T2, which means that the input impedance $Z_{in2}$ at terminal T2 and the voltage standing wave ration (VSWR) become close to infinity.
(3.32). The incident and reflected voltage wave superimpose and double in magnitude at the cantilever tip. Thus, the RF voltage \(V_{RF}\) is replaced by the value of \(V_{max} = V_{RF}(1 + |S_{11}|)\).

Assuming the load of \(Z_{02} = -j/(\omega C_s)\) connected to the lever tip, the reflection coefficient \(|S_{11}|\) equals \(1/\sqrt{1 + (2/\omega r_s)^2}\), with \(\omega r_s = (Z_{02} C_s)^{-1}\).

\[
V_{SWR} = \frac{V_{max}}{V_{min}} = \frac{V_{RF}(1 + |S_{11}|)}{V_{RF}(1 - |S_{11}|)}
\]

The voltage difference between the actuation electrode and the cantilever depends on the capacitively induced current, and the resistance of the bias line \(R_b\) and the DC source \(R_s\). The voltage across the air gap is given by \(V = V_{max}/(1 + j\omega/\omega r_s)\), where \(\omega r_s = 1/[(R_b + R_s)C]\). For frequencies above \(\omega r_s\), \(|V|/|V_{max}| \approx 1\). At frequencies below \(\omega r_s\), \(|V|/|V_{max}| \approx 1\). High voltage signals with spectral portions below \(\omega r_s\) can cause self-actuation. Thus, a low bias line resonance frequency allows self-actuation to be suppressed. However, the time constant \(\tau = R_b C\) can also limit the switching speed if it exceeds 1 \(\mu s\), which is close to the typical switching time of a MEMS switch [62]. Thus, the electrical resonance frequency \(f_{rs}\) is typically above 1 MHz and below 100 MHz.

The maximum power level which can be handled by the switch without significant change of the switch deflection is determined by the impedance \(Z_{01}\) at terminal T1, the stiffness of the cantilever as well as the bias and contact capacitance. The values given in the following equations refer to the root mean square of the signal voltage and power. Equation 3.33 and 3.34 show the two forces at the contact and at the bias electrode in the steady-state. The steady-state assumption is true as long as the cantilever mechanical resonance frequency is well below the RF frequencies. In order to describe the power handling for ohmic and capacitive contact switches, a solid-state dielectric of thickness \(d\) and permittivity \(\varepsilon_{rd}\) is introduced between the two contacts. There is no dielectric layer on top of the bias electrode.

In the absence of a DC bias \(V_{DC} = 0\), the average RF voltage \(\bar{V}_{RF}\) can be factored out.

\[
\bar{q}(l_2 - l_1) = k(x_s)\bar{z}(x_s) = \frac{w_2(l_2 - l_1)\bar{P}^2_{RF}(1 + |S_{11}|)^2}{2\bar{x}\varepsilon_{rd}\left[\frac{\bar{q}(l_2 - l_1)}{\bar{z}(x_s)}\right]^2} \left|_{\omega = 0} \right.
\]

A high bias line resistance is required to reflect the RF signals at terminal T3 and to limit the voltage drop on the measurement ports 1 and 2 in the event of a short circuit between the lever and the bias electrode. The bias line at port 3 can be represented by the impedance \(Z_{03} = R_b + R_s - j/(\omega C)\). Since \(R_b \gg Z_{in3}\), terminal T3 acts like an open circuit for RF signals.

In case of a capacitive contact switch in series configuration, a solid-state dielectric layer of thickness \(d\) and permittivity \(\varepsilon_{rd}\) is assumed between the two contacts.
\[ F(l_z) = k(l_z)z(l_z) = \frac{NW_{av}\omega_{av}}{2\varepsilon_0\varepsilon_{ra}} \left( \frac{\omega}{\omega_{av}} \right)^2 (1 + |S_{11}|)^2 \left( 1 + \left( \frac{\omega}{\omega_{av}} \right)^2 \right) \]

(3.34)

From the displacement curve (3.12), the displacement of the upper contacts can be derived (3.35).

\[
z(l_z) = \frac{q(l_z - l_1)}{24} \left\{ \frac{2l_1(l_z^2 - 3l_1l_z + 6l_2l_3) + (l_z - l_2)^2(3l_1 + l_z - 4l_3)}{\alpha_1} + \frac{F(l_z)l_z(l_z^2 - 3l_1l_z - 3l_1l_3 + (l_z - l_2)^2l_z}{\alpha_1} - \frac{(l_z - l_2)^3}{\alpha_2} \right\}
\]

(3.35)

In case of a port impedance \( Z_{01} = Z_{02} = Z_0 \), the maximum power \( \tilde{P}_{RF} = \tilde{V}_{RF}^2/Z_0 \) can be calculated (3.36). Figure 3.17a shows the maximum voltage and power versus the relative displacement. In figure 3.17b the contact capacitance is plotted for a given RF power applied. The results correspond to the switch definitions given in table 3.1.

\[
\tilde{P}_{RF} = z(l_z) \left\{ \frac{Z_0}{1 + (\omega/\omega_{av})^2 \varepsilon_{ra}} \left( \frac{\varepsilon_{0}\varepsilon_{ra}}{\varepsilon_{0}\varepsilon_{ra}} \right)^2 \left[ \frac{2l_1(l_z^2 - 3l_1l_z + 6l_2l_3) + (l_z - l_2)^2(3l_1 + l_z - 4l_3)}{\alpha_1} + \frac{F(l_z)l_z(l_z^2 - 3l_1l_z - 3l_1l_3 + (l_z - l_2)^2l_z}{\alpha_1} - \frac{(l_z - l_2)^3}{\alpha_2} \right] \right\}
\]

(3.36)

For the given switch configuration (3.16), significant cantilever deflection and thus capacitance modulation occurs at RF power levels exceeding 30 dBm. At an RF power of 43 dBm the switch becomes instable and changes from the up state to its down state.

Theoretically, the 1 dB compression point of the switch in the up state is found, if \( |S_{11}(P_{rf})| \) is lowered by 1 dB compared to its small-signal value as a result of increasing incident RF power \( P_{rf} \). The calculated 1 dB compression point corresponds to the contact capacitance value of 13.5 fF. However self-actuation occurs prior, at the contact capacitance of 5.5 fF, and thus at lower RF power (figure 3.17).
Figure 3.17. (a) Time average power and voltage versus normalized displacement. (b) Corresponding contact capacitance at a given average RF power. The results refer to the ohmic-contact switch, with the definitions given in table 3.1. The port impedance is $Z_0 = 50$ Ω, the capacitances $C$ and $C_s$ are given by (3.31). The operation frequency is 60 GHz.

3.3 MEMS fabrication process

In the course of this thesis, RF MEMS switches have been frequently fabricated and tested with regard to fabrication reliability and mechanical as well as electrical performance issues. Different sacrificial layer materials and release methods have been investigated in order to optimize the reliability and reproducibility of the electro-mechanical switches. Suitable contact material pairings have been evaluated and tested. Further, the influence of the cantilever material and process variations on the switching response have been modelled and characterized. Accounting for the findings in the modelling, the fabrication process given in [49] has been adapted using a low density cantilever material offering a high flexural stiffness. Besides the impact on switching performance of using copper instead of gold, copper plating is more cost-efficient. According to the contact studies [50]-
a contact pairing is best made of a highly resistive and hard material and a ductile highly conductive material. Moreover, the surface of the contact should initially provide a minimum of roughness. In the process described below, gold and platinum were chosen as the contact pairing.

The final fabrication route is illustrated by the figures 3.18-3.23 and shows the states of the MEMS between two process steps. The process executed between the states was as follows.

![Figure 3.18](image)

**Figure 3.18.** Patterning of bias line and bias electrode. The chromium alignment layer acts as an alignment layer for the subsequent mask sets. As an alternative to Ti35 ES, AZ 2035 was used for the lift-off patterning procedure.

For the bias line, a highly resistive layer of indium tin oxide (ITO) was deposited using DC magnetron sputtering in absence of reactive oxygen background pressure (figure 3.18). Sputtering was performed in a BAS 450. All films were deposited on cleaned fused silica surfaces. The as-deposited indium tin oxide ($\text{In}_2\text{O}_3$;SnO$_2$) with composition of 90:10 as percentage by weight showed an amorphous microstructure with metallic character. Crystallite growth of $\text{In}_2\text{O}_3$ was observed by XRD analysis after annealing of the amorphous ITO thin films for 20 min at 200°C, 300°C and 400°C. This caused a decrease of the initial sheet resistance by a factor of 3 to 10, dependent on the film thickness. XRD measurements indicated complete crystallization after 20 min at 300°C in a 20 nm thin film. Hence, the decrease of the electrical conductivity correlates with grain growth. Similar behaviour is reported by [74].

In order to align subsequent mask sets to the pattern of the transparent conductive oxide (TCO), a 40 nm thick chromium layer was deposited prior to the ITO deposition. Ti35 ES image reversal resist was used to pattern the chromium by wet etching. Ceric ammonium nitrate with perchloric acid [75]
was used as an etchant. To activate the metal surfaces and hence reduce the wetting angles of aqueous etchant solutions, oxygen plasma was applied prior to wet etching. The under-etch of the chromium was 1.0±0.5 μm. The remaining resist pattern was used for lift-off of the ITO. The resist profile was processed to show an undercut of 2.5±0.5 μm at a thickness of 3.0 μm. Ti35 ES resist cross-linking was performed in a low power oxygen plasma for 15 min at 500 W and 150 sccm O₂-flow, in order to increase its mechanical strength. After deposition and prior to resist removal, the ITO was oxidized for 20 min at 90°C in an air oven. Oxidation is necessary to obtain the desired chemical stability for subsequent processing. Complete organic removal was obtained after 30 min in Technistrip Ni555 at 70°C.

Figure 3.19. Lift-off patterning sequence. Deposition of the 25 nm Ta adhesion layer and 120 nm Pt forming the lower contacts.

The lower contacts are formed by platinum (figure 3.19). Ti35 ES resist was used as a mask to etch the chromium layer and to provide the desired profile for lift-off of the metallization. 30 nm tantalum and 100 nm platinum were evaporated in a BAK 550. Deposition and lift-off processes were
conducted in a manner to prevent metal wing-tips. Therefore, film strain on the resist, vapour
divergence and substrate temperature were controlled precisely. The tantalum adhesion layer showed
excellent adhesion on the fused silica and the indium tin oxide. With completion of the lift-off, the
chromium alignment layer was removed.

Figure 3.20. Two step patterning sequence of the W:Ti sacrificial layers. The contact separation is given by the thickness of the second W:Ti coating. In the actuation area, the sacrificial layer thickness can be increased by the thickness of the first W:Ti layer.
As a sacrificial layer, a tungsten-titanium-alloy W:Ti (90wt-%:10wt-%) was sputtered in a two-step procedure (figure 3.20). The deposition was performed in a BAS 450 sputtering system. Wet etching was used for film patterning. The observed etch profile was isotropic, which results in a minimum under-etch. For wet-etching W:Ti-Etch 200, a solution containing ammonium fluoride [76] was used. Subsequently a short dip in H₂O₂, with a concentration of 30% was performed. The dimples dimension, and hence the contact separation is controlled precisely by this two layer technique.

After patterning, the W:Ti sacrificial layer the back side metallization was sputtered. The metallization consists of a 50 nm thick chromium adhesion layer and a 500 nm thick gold layer. The gold layer was etched in potassium iodide [77] and the chromium was etched in ceric ammonium nitrate with perchloric acid. The under-etch of the 500 nm thick gold layer was 3.0±0.5 μm.

The upper contacts consist of a 100 nm thick sputtered gold layer (figure 3.21). The sputtering process allowed for excellent edge coverage. Moreover, the upper contacts exhibited a minimum of roughness, comparable to the surface roughness of the fused silica substrate. The patterning of the gold was performed using wet etching. The under-etch of the 100 nm thick gold film was 1.5±0.5 μm. The back side metals were protected by a layer of the image resist S1828.

Figure 3.21. Patterning of the upper contacts using a 100 nm thick sputtered gold layer and subsequent wet etching.
The cantilever was formed by the electroplating of copper (figure 3.22). As an advantage, electroplating allows for steep side walls and thick films exceeding 2.0 μm, while remaining cost-efficient. In contrast to former versions of MEMS, this cantilever was formed of copper instead of gold, which excels in better switching dynamics due to the lower density at a higher Young’s modulus as well as due to fabrication cost considerations. A sputtered 200 nm copper layer was used as a seed for electroplating. AZ9260 with 7.8 μm thickness was used for patterning prior to coating. A short plasma activation was followed by pickling in sodium-persulphat-sulphuric acid at room temperature. Electroplating was performed in InterVia Cu8500 (Bath Formulation I) from Rohm and Haas [78]. The current density resulting in stress reversal films was 1.5 A/dm². Additive and carrier were differentially adjusted relative to the start-value of 5 ml/l in order to control the stress level, thickness uniformity and surface roughness. The normalized growth rate was determined to be \( \frac{0.19 \, \mu \text{m/min}}{(A/\text{dm}^2)} \). The ratio of the charge transfer to deposition volume is \( \frac{Q}{V} = 31.7 \, \text{C/mm}^3 \). The end criterion for a desired plating thickness is given by the cumulative charge transferred. The cumulative charge was measured by a DC Am RITTAL charge meter. After electroplating, the coated substrates were stored for two days at room temperature, where segregation and recrystallization took place [79]. Segregation and recrystallization cause a tensile strain component in the copper film after deposition. The initial stress after deposition can be adjusted by adding carrier and additive or by manipulation of the current density. Compressive stress increases with a reduction of the current density or by increasing the additive and carrier concentration [80].

After storage, the copper seed layer was etched in W:Ti-Etch 200. A short dip in a phosphoric-acetic-acid is applied at room temperature to remove copper oxide residuals [81].
The sacrificial layer underneath the cantilever was removed using H$_2$O$_2$ (30%) at 50°C (figure 3.23). A firm release method using tert butanol (TBA) was applied. The tert butanol was sublimed at reduced pressure in order to avoid triple phase boundaries, which would cause capillary forces and thus sticking. Sublimation was performed according to figure 3.24. The dynamics of evacuation of the chamber needs to be controlled carefully. If pressure is reduced too fast, residual IPA and water expand. The expansion of residual IPA and water underneath the fragile mechanical structure can cause severe plastic deformation. The transfer of the cooled TBA covered MEMS to the
vacuum system was carried out in a solid metal case, which provides a heat sink during transfer and prevents water condensation on the wafer surface.

Figure 3.23. Removal of the W-Ti sacrificial layers in hydrogen peroxide, rinse in TBA and subsequent sublimation drying.
3.4 MEMS switch characterization

The modelled and fabricated MEMS switch has been characterized in terms of its electro-mechanical function and its performance up to 100 GHz. Therefore, the MEMS switch was implemented in a finite ground coplanar waveguide (FGCPW), depicted in figure 3.25. The dimensions, material properties and characteristic parameters of the switch are given in table 3.1. The FGCPW has the dimensions of a conductor gap $s = 16 \, \mu m$, conductor width $w = 110 \, \mu m$, conductor thickness $h = 4.2 \, \mu m$ and terminal separation $l = 260 \, \mu m$. The characteristic line impedance and electrical length at 60 GHz is calculated to be $50.7 \, \Omega$ and $27.7^\circ$.

Figure 3.24. Pressure-temperature diagram showing the equilibrium vapour pressures of isopropyl alcohol (IPA), tert butyl alcohol (TBA) and water ($H_2O$). The ideal sublimation cycle is indicated by the arrows. In reality, the dynamics of pressure reduction causes the molecules at the surface to expand into the gaseous state, while the remaining molecules of the solid are undercooled.

Figure 3.25. RF MEMS switch implemented in a finite ground coplanar waveguide (FGCPW).
The time-dependent, bias voltage-dependent, frequency-dependent and RF power-dependent functions of the electro-mechanical switch have been investigated and are summarized in the subsequent by the switching time, the switching hysteresis, the S-parameters in the open state and close state and the switch power handling capabilities.

### 3.4.1 Switching time

The switch-on time is the delay between the instant when the bias signal is at 50% of its maximum value and the instant when the envelope of the RF signal (3 kHz-300 GHz) is at 90% of its maximum. The determined switching time (figure 3.26) is in good agreement with the modelled data. The time required for switching can be divided into separate sections: the bias capacitance charging, the lever transition from its up state to its down state, contact bouncing and contact force ramp up. The time constant for capacitance charging is determined by \( \tau = R_b C \) and is below 10 ns. Contact bouncing can delay the ramp-up of the contact force and thus increase the switching time significantly [49].

![Figure 3.26](image)

**Figure 3.26.** Experimentally determined switch-on time. The signal voltage amplitude is set to 1.0 V, the signal frequency is 1.1 MHz and the probe impedance 1 MΩ. The bias signal rise time is 1 µs.

The model and experiments showed that the ratio of Young’s modulus to density of the cantilever is the basic parameter increased to minimize bouncing at the same switching time. The damping energy and the energy transferred to the contacts are expected to be the major sources of potential switch degradation in the case of cold switching and moderate RF power. The DC bias of 60 V and 42 V are suggested as the maximum and minimum ratings. Actuation close to the minimum ratings is inherent with slower switching speed but reduces potential aging effects and may result in a longer lifetime.
The test setup for transient characterisation used a Keithley 2400 voltage source, which powered the collector path of a 2SC2911 npn-transistor from SANYO. The transistor base was controlled by a HP 3312A function generator, providing a frequency variable square signal with a high level of 4.0 V and a low level of -13.0 V [82]. The rise time of the voltage at the emitter of the npn-transistor, which biases the DUT, was below 1 µs. The signal path of the DUT was excited by a sine wave with an amplitude of 1.0 V at a frequency of 1.1 MHz oscillating around ground potential. The sine signal was supplied by a HAMEG 8030. At the switched end of the signal line, the voltage was probed by a TDS 2012 oscilloscope with the probe impedance set to 1 MΩ.

The switch-off time is the delay between the instant when the bias signal is at 50% of its maximum value and the instant when the envelope of the RF signal is at 10% of its maximum. The comparison of experimental data (figure 3.27) with modelled data (figure 3.7) states that mechanical contact is interrupted immediately if the bias voltage is set to 0 V at moderate RF power. Nevertheless, an ionic current can be present in air as long the field strength is exceeded. The switch-off time can be divided into the times required for capacitor discharging, and the lever transition from its down state to the contact displacement at which breakdown voltage is undershot. A linear dependence can be found between the bias voltage and the switch off time (figure 3.27). The larger the ratio between the actuation voltage and voltage at which restoring occurs, the longer lasts the electrical contact. Additionally, limited charge transfer slows down the switch during movement from on state to off state.

![Figure 3.27](image)

**Figure 3.27.** Experimentally determined switch-off time. The signal voltage amplitude is set to 1.0 V, the signal frequency is 1.1 MHz and the probe impedance 1 MΩ. The bias signal fall time is 1 µs.

### 3.4.2 Switching hysteresis

The switching hysteresis describes the different characteristics if the voltage is increased stepwise (forward cycle) and if the voltage is decreased stepwise (reverse cycle). In figure 3.28, the measured
and simulated hysteresis loops are plotted. The pull-in voltage and the restoring voltages can be read out from the hysteresis loop and are \( V_p = 36 \, V \) and \( V_r = 31 \, V \) for the measured switch. The results are in good agreement with the modelled data if using Young’s modulus of 75 GPa.

Assuming Young’s modulus of 124 GPa as described in table 3.1, pull-in is expected above 42 V and restoring at 39 V. For the Young’s modulus of 75 GPa the pull-in voltage of \( V_p = 34 \, V \) and release voltage of \( V_r = 31 \, V \) are calculated by the model. The Young’s modulus of 75 GPa is close to the expected values determined by the single crystalline elastic constants of pure copper [68]. For Young’s modulus of 75 GPa, the steady-state model exhibits zero contact force at 32 V. The estimated contact forces, at \( V_{ext} = 36 \, V \), \( V_{ext} = 40 \, V \) and \( V_{ext} = 50 \, V \) are \(-10 \, \mu N\), \(-20 \, \mu N\), and \(-80 \, \mu N\).

![Switching hysteresis](image)

**Figure 3.28.** Switching hysteresis, showing the forward cycle and the reverse cycle. The measured pull-in voltage and the restoring voltage threshold are \( V_p = 36 \, V \) and \( V_r = 31 \, V \). The simulated curves are given for Young’s modulus of 124 GPa and 75 GPa. Assuming Young’s modulus of 75 GPa shows good agreement with the measurements.

The insertion loss in the on state (figure 3.29) is primarily caused by the contact resistance. In the forward cycle, the contact resistance decreases with increasing contact force. In the reverse cycle, the contact resistance remains at its minimum value, until the voltage undershoots the pull-in voltage. The contact hysteresis may be related to the processes of contact forming, partial welding and contact separation.
Contact hysteresis. The insertion loss is primarily caused by the contact resistance. In the forward cycle, the contact resistance decreases with increasing contact force. In the reverse cycle, the contact resistance remains at its minimum value, until the voltage undershoots the pull-in voltage.

### Insertion loss and return loss

The insertion loss is defined as \( L_A = -20 \log_{10}(|S_{21}|) \) and measured in the on state. The return loss is defined as \( L_R = -20 \log_{10}(|S_{11}|) \). Both values can be directly read off the S-parameters (figure 3.30) if plotted in dB scale. The insertion loss of the RF switch in FGCPW configuration is between 0.3 dB and 0.4 dB for frequencies up to 70 GHz. The return loss increases with increasing frequency from 40 dB at 30 GHz to 30 dB at 70 GHz. The total insertion loss can be divided into the effects: conductor skin losses, dielectric losses, surface currents and leakage along the DC bias path. Thin, narrow and high permeability conductors as well as surface oxides can significantly contribute to the skin losses.

The switch contacts form a distinct constriction, where the conductor cross-section is strongly reduced leading to a lumped series resistor. Additional skin losses may be introduced by a thin copper oxide layer on top of the microstrip line. The conductivity of copper oxide has been reported to be around 10 S/m compared to the conductivity of copper, which is \( 6 \cdot 10^7 \) S/m [83]. RF leakage along the DC bias path is suppressed by a pair of 40 k\( \Omega \) resistors.
Figure 3.30. Switch in down state. The actuation voltage is 50 V DC. The S-parameters show the return loss and the insertion loss.

3.4.4 Isolation and forward loss

The switch isolation is defined as $L_A = -20 \log_{10}(|S_{21}|)$ and measured in the off state. Since the switch is implemented in series configuration, it acts as high-pass element. The isolation is reduced with increasing frequency from -22 dB at 15 GHz to -10 dB at 70 GHz (figure 3.31). An asymmetry is observed in the reflection parameters $S_{11}$ and $S_{22}$, which is attributed to the asymmetry of the switch geometry inserted between the two measurement terminals (figure 3.25). The forward loss in the off state $L_F = -20 \log_{10}(|S_{11}|)$ is below 1.2 dB at port 1 and below 2 dB at port 2 up to 60 GHz.

Figure 3.31. Switch in up state. The actuation voltage is 0 V DC. The S-parameters show the forward losses and the isolation loss.
3.4.5 Power handling capability

The switch power handling was investigated with the intention of identifying power-related limitations occurring in the voltage-driven capacitive coupled MEMS switch. Potential effects of the RF power on the switching behaviour have been described in section 3.2.6. Further, the failure mechanisms of contact welding and contact finger degradation are of interest.

The measurement setup used to characterize the power dependencies is depicted in figure 3.32. The signal generator Agilent E8254A was used to drive the power amplifier HP83020A [84], which provides a maximum output power of 33 dBm at 11 GHz. The amplifier power gain was \( G_p = 37 \) dB, below its 1 dB-compression. To avoid reflected power entering the amplifier and to provide a 50 \( \Omega \) impedance to the directional coupler KRYTAR 1821, a ferrite circulator was combined with a 50 \( \Omega \) load to form an isolator [85]. The directional coupler was used to access the forward and reflected power by the use of two power detectors [86]. A 10 dB attenuator was inserted so as not to exceed the measurement range of the detector. To measure the transmitted power, a spectrum analyser HP8563E was used. The cable and probe arrangement, with a THRU in place of the DUT, was used to determine the insertion loss of the cable and probe arrangement.

![Figure 3.32. Measurement setup for power handling characterization of the RF MEMS switch implemented in FGCPW architecture.](image)

The investigations show a linear behaviour of the RF MEMS switch throughout the available power range at the frequency of 11 GHz. The switch isolation represented by the transmitted power lift due to switch-on is determined to be 23 dB. It is invariant up to the maximum power of 29 dBm available at the terminal T2 of the DUT.
As discussed in section 3.2.6 capacitance modulation and self-actuation are expected above 30 dBm for frequencies well-exceeding the bias line resonance of $f_{rs1} = 26$ MHz. The lower electrode, which is capacitive coupled to the RF line, is then expected to exhibit a potential comparable to that of the lever. This is in agreement with the measurements, which do not show significant capacitance modulation or self-actuation below 30 dBm at 11 GHz.

The subsequent power dependent switching behaviour and failure modes could be observed:

(i) At hot-switching conditions, the mechanical lever restoring force became insufficient after some hundred cycles at an incident RF power of 28 dBm. At zero DC bias, the time-average RF voltage across the actuation electrodes exerts attraction forces on the lever. Once in the down state, the carrying RF signal latched the switch and prevented it from moving to the up state unless the RF power was turned down.

(ii) Contact welding at cold-switching conditions occurred after a few actuation cycles at 29 dBm. This indicates that Joule heating and electrostatic discharge can severely contribute to contact degradation.

(iii) At an RF power of 33 dBm on terminal T2 available at a frequency of 4 GHz, contact evaporation was observed. Evaporation occurred while the switch was in the down state. The actuation was performed in absence of RF power. The cross-section area of the two lower contact fingers is $A_{lc} = 4.5$ µm$^2$, which allows an equivalent current density in the fingers of 4.4 MA/cm$^2$ to be estimated.

3.4.6 Switch lifetime

The switch lifetime was determined by means of cycles-to-failure. In order to use SMA connectors as the measurement terminals, the RF MEMS switch die was bonded onto a fused silica support substrate. The support substrate comprises a taper to connect the contacts of the SMA to the contacts of the RF MEMS switch. The taper was carried out as a CPW with a characteristic impedance of 50 $\Omega$. The SMA connectors are flanged to the substrate using a metallic mounting stub. Die bonding was carried out using U300-2 Underfill epoxy. The SMA connectors are joined to the support substrate using H20E, a silver filled epoxy. The switch was electrically connected to the metallization on the support substrate by means of wedge-wedge wire bonding.

RF measurements were performed using a DG8SAQ VNWA 3 from SDR-Kit. The forward scattering transmission parameter $S_{21}$ was logged within a frequency range of 1075 MHz - 1125 MHz. The sweep repetition time was approximately 200 ms. The RF switch actuation signal was controlled by a function generator and a transistor amplifier, providing a symmetric triangle voltage signal varying between 0 V and 50 V at a frequency of 1.00 Hz.
As characteristic of the presented electrostatic actuation, the voltage at switch on exceeds the voltage at switch off, as shown by the hysteresis loop in figure 3.28. This difference between switch on and switch off delays the RF signal relative to the actuation signal (figure 3.33). For this experiments, with the voltage slope of 200 $V/s$, the hysteresis of $\Delta V = 4 V$ results in a time delay of 20 ms.

The control voltage $V_{cr}(t)$ shown in figure 3.33 has been mathematically delayed relative to the bias voltage $V_b(t)$ in order to exhibit equal phase compared to the switched RF signal, thus the average value between the switch on and switch off voltage is determined. In figure 3.34 this average value is depicted by the voltage threshold. A lower threshold voltage corresponds to a longer duration of contact. In the quasi-static case, the shift of the threshold voltage is explained by a shift in the force balance. Possible changes can be caused by mechanical fatigue due to material creep in the lever or a change of the contact tribology. Material creep in the lever can result in a displacement or reduction of the initial stiffness coefficient. A change in the contact tribology affects the contact resistance at a given contact force.

![Figure 3.33](image_url)

**Figure 3.33.** The first curve shows the switch insertion loss / isolation. The second graph illustrates the corresponding bias voltage, the pull-in and restoring voltages. The third curve $V_{cr}(t)$ exhibits a time delay relative to $V_b(t)$. The delay is chosen the pull-in and restoring voltages to coincide.

Figure 3.34 shows the signal transmission versus the control voltage $V_{cr}(t)$. Each plot shows a sequence of 4 min. The first sequence is taken after $10.8 \cdot 10^3$ cycles, the second sequence is taken after $66.6 \cdot 10^4$ cycles, the third sequence is taken after $12.8 \cdot 10^5$ cycles, and the fourth sequence is taken after $34.7 \cdot 10^6$ cycles. In order to obtain the corresponding test results exceeding the number $5 \cdot 10^6$ cycles within suitable time, the bias signal frequency was set to 50.0 Hz.
The average of the pull-in and release voltage has shifted during the lifetime test from initial 33 V to 22 V. The isolation in the off state has decreased from 41 dB to 39 dB at 1.1 GHz.

The shift of the voltage threshold may be explained by fatigue of the copper lever, which results in a decrease of the cantilever stiffness coefficient. The decrease of the isolation may be explained by stress relaxation due to material creep and recrystallization occurring while the electrostatic force is pulling on the lever. This combination can result in a decrease of the initial contact separation.

The hysteresis loop measurements were repeated after $42.0 \cdot 10^6$ cycles (figure 3.35), showing that the hysteresis has changed from the initial value of 5 V to 3 V after cycling. This indicates the pull-in and the restoring are affected by the fatigue process.

![Figure 3.34](image_url)

Figure 3.34. Forward scattering parameter $|S_{21}|$ in dependence of the control voltage. The control voltage has been mathematically delayed in order to exhibit equal phase compared to the switched RF signal, thus the hysteresis cancels. The curves represent a sequence of 4 min. The switching frequency during data acquisition was 1.00 Hz.

![Figure 3.35](image_url)

Figure 3.35. Hysteresis plot after 40 and $42.0 \cdot 10^6$ switching cycles. The RF signal power was set to $-17$ dBm at the frequency of 1.1 GHz.
Electrostatic discharge (ESD) can cause strong attraction forces pulling the lever in contact to the bias electrode, which potentially can cause plastic deformation at the switch anchor. However due to the absence of a dielectric passivation layer an overvoltage will short the two electrodes and prevent a further increase of the electrostatic force or severe plastic deformation of the anchor.

Agglomerations of contact debris and electromigration can narrow the contact gap and reduce the bias voltage required for switching. Due to the absence of a packaging, it cannot be definitely ruled out that particles stemming from the environment affect the switching functions. Figure 3.36 shows the switch after cycling in the up state and the down state. The Newton’s rings allow an indication of the deformation shape.

Figure 3.36. Microscope images of the switch (a) in the up state and (b) the down state. The picture shows the view through the fused silica substrate and the transparent ITO electrode.
CHAPTER 4

Reconfigurable loaded-line (RLL) sections

4.1 Introduction

Electrostatically actuated MEMS can be used to load a transmission-line section of finite length with an excess of reactance or susceptance. This way, the characteristic impedance and electrical length of the transmission-line section can be continuously tuned or discretely programmed, depending on the MEMS architecture. A cascaded network of such reconfigurable loaded-line (RLL) sections can perform the function of a programmable filter or phase shifter.

In this chapter a general model is presented, describing the loaded-line sections in terms of their ABCD-parameters, S-parameters and T-parameters. The ABCD-parameters offer a comprehensive representation of the wave guide characteristics, whereas S-parameters are primarily used to express insertion loss, return loss and insertion phase. T-parameters are preferred for the rapid frequency response calculation of cascaded RLL structures. The model links the influence of the line loading by geometric discontinuities to the characteristic properties of the transmission-line section using closed-form expressions. Practical simplifications are introduced and the analytical description compared with the exact values extracted from Momentum simulations using Advanced Design System (ADS).

Reconfigurable loaded microstrip sections in series and shunt configuration have been fabricated and examined. In the subsequent, the equivalent circuits are given for the low-pass reconfigurable loaded-line sections. The equivalent circuit calculations are performed in order to estimate parasitic resonances and to predict a suitable frequency range of use for microstrip discontinuities with a given set of dimensions. Finally, closed-form expressions for the loading factors of the low-pass RLL sections are derived and the ABCD-parameters of the sections are expressed in dependence on the loading factor.

4.2 General model of loaded-line sections

In practice a variety of different transmission-line architectures are common. These are the microstrip (MS), the coplanar waveguide (CPW), stripline, slotline, et cetera. Descriptions for these types of transmission lines can be found in [53], [54], [87] and [88]. In the subsequent, the microstrip line is used by way of example.
The characteristic impedance $Z_c$ and the effective permittivity $\varepsilon_{r,\text{eff}}$ of a microstrip line with a conductor width $W$, substrate permittivity $\varepsilon_r$ and substrate height $h$ can be expressed as a function of $\varepsilon_r$ and the ratio $W/h$. The expressions (4.1) and (4.2) hold for $W/h \geq 1$, where $\eta_0 = 1/(c_0\varepsilon_0)$ is the free space impedance [89]. More accurate closed-form expressions are presented by [90].

\[
Z_c = \frac{\eta_0}{\sqrt{\varepsilon_{r,\text{eff}}}} \frac{W}{h} + 1.393 + 0.677\ln\left(\frac{W}{h} + 1.444\right)^{-1} \tag{4.1}
\]

\[
\varepsilon_{r,\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + 12 \frac{1}{W/h}\right)^{-1/2} \tag{4.2}
\]

Given a $W/h$ ratio which can be varied and a fixed substrate permittivity, the resultant characteristic impedance $Z_{\text{c,lt}}$ and effective permittivity $\varepsilon_{r,\text{eff},\text{lt}}$ of the loaded line can be linearized around the nominal $W/h$ of the unloaded line. In (4.3) and (4.4), the loaded-line characteristics are expressed in dependence on the first derivatives (4.5) and (4.6).

\[
Z_{\text{c,lt}} = Z_c \left(1 + \frac{\partial Z_c}{\partial (W/h)} \bigg|_{W/h} \delta(W/h)\right) = Z_c \left(1 + \frac{\delta Z_c}{Z_c}\right) \tag{4.3}
\]

\[
\varepsilon_{r,\text{eff},\text{lt}} = \varepsilon_{r,\text{eff}} \left(1 + \frac{\partial \varepsilon_{r,\text{eff}}}{\partial (W/h)} \bigg|_{W/h} \delta(W/h)\right) = \varepsilon_{r,\text{eff}} \left(1 + \frac{\delta \varepsilon_{r,\text{eff}}}{\varepsilon_{r,\text{eff}}}\right) \tag{4.4}
\]

The relative effective permittivity change $\delta \varepsilon_{r,\text{eff}} / \varepsilon_{r,\text{eff}}$ increases with increasing substrate permittivity and decreasing $W/h$ ratio of the unloaded line. This can be read off figure 4.1 and from the sensitivity $\partial \varepsilon_{r,\text{eff}} / \partial (W/h)$ given in (4.6). In order to obtain high phase resolution, low permittivity substrates and a high $W/h$ ratio are desired, whereas for large phase shift, high permittivity substrates and a low $W/h$ ratio are preferred.

\[
\frac{\partial Z_c}{\partial (W/h)} = -\eta_0 \left[2\left(\frac{W/h + 2.121}{1.444 + W/H}\right) \varepsilon_{r,\text{eff}} + \frac{\partial \varepsilon_{r,\text{eff}}}{\partial (W/h)} \left(\frac{W}{H} + 1.393 + 0.677\ln\left(\frac{W}{H} + 1.444\right)\right)\right] \tag{4.5}
\]

\[
\frac{\partial \varepsilon_{r,\text{eff}}}{\partial (W/h)} = \frac{3(\varepsilon_r - 1)}{\left(\frac{W}{H}\right)^2 \left(1 + 12 \frac{1}{W/h}\right)^{3/2}} \tag{4.6}
\]
Figure 4.1. First derivatives of the effective permittivity and the characteristic impedance of a microstrip on a fused silica substrate. The 50 Ω microstrip line has a nominal ratio of \( W/h = 2.296 \).

Due to a change in the effective permittivity, the phase change \( \delta \theta / \theta = \sqrt{\delta \varepsilon_{r,eff} / \varepsilon_{r,eff}} \) is obtained. Both the relative change in the characteristic impedance and the relative change in the phase are of the same order of magnitude. For \( \delta (W/h) > 0 \) the shunt susceptances per length and the characteristic admittance increase (figure 4.2). For \( \delta (W/h) < 0 \) the series reactance per length and the characteristic impedance increase. Additionally, for \( \delta (W/h) > 0 \) the phase shift per length increases, whereas for \( \delta (W/h) < 0 \) the phase shift per length decreases (figure 4.3).

Figure 4.2. Relative changes in the characteristic impedance \( \delta Z_c / Z_c \) and characteristic admittance \( \delta Y_c / Y_c \). The values refer to a 50 Ω line with \( W/h = 2.296 \) and \( \varepsilon_r = 3.72 \).
Figure 4.3. Relative changes in the effective permittivity and electrical length. The values refer to a 50 \( \Omega \) line with \( W/h = 2.296 \) and \( \varepsilon_r = 3.72 \).

### 4.2.1 ABCD-parameters

The transfer matrix of an unloaded transmission-line section is best expressed by the ABCD-parameters (4.7), which in contrast to the T-parameters are independent of the port impedance or termination. In the case of a reciprocal network, the transfer matrix is expressed by three independent parameters: \( A_{12} \), \( A_{21} \) and \( A_{11} \). The supplemental parameter in the transfer matrix is \( A_{22} = (1 + A_{12}A_{21})/A_{11} \). In the subsequent, the letter \( A \) denotes the ABCD-parameters. In case of a symmetrical, reciprocal network the two independent parameters, \( A_{12} \) and \( A_{21} \), exist. All other parameters, \( A_{22} = A_{11} = \sqrt{1 + A_{12}A_{21}} \), can be expressed in terms of \( A_{12} \) and \( A_{21} \). The general form of the transfer matrix for a loaded symmetric line is given in (4.8). For a lossless transmission line, \( \sinh(j\theta) = j\sin(\theta) \) and \( \cosh(j\theta) = \cos(\theta) \) are pure imaginary and real functions.

\[
\mathbf{A}_u = \begin{bmatrix} A_{u,11} & A_{u,12} \\ A_{u,21} & A_{u,22} \end{bmatrix} = \begin{bmatrix} \cos(\theta_u) & Z_u \sinh(j\theta_u) \\ \frac{1}{Z_u} \sinh(j\theta_u) & \cosh(j\theta_u) \end{bmatrix}
\] (4.7)

\[
A_{II} = \begin{bmatrix} A_{u,11} + \frac{\delta A_{11}}{A_{u,11}} & A_{u,12} + \frac{\delta A_{12}}{A_{u,12}} \\ A_{u,21} + \frac{\delta A_{21}}{A_{u,21}} & A_{u,22} + \frac{\delta A_{22}}{A_{u,22}} \end{bmatrix} = \begin{bmatrix} A_{u,11} \left( 1 - \frac{\delta\theta}{\cos(\theta_u)} \right) & A_{u,12} \left( 1 + \frac{\delta Z}{Z_u} + \frac{\delta\theta}{\tan(\theta_u)} \right) \\ A_{u,21} \left( 1 - \frac{\delta Z}{Z_u} + \frac{\delta\theta}{\tan(\theta_u)} \right) & A_{u,22} \left( 1 - \frac{\delta\theta}{\cot(\theta_u)} \right) \end{bmatrix}
\] (4.8)

The linear approximation (4.8) representing the transfer matrix of the loaded line shows that \( \delta A_{12}/A_{u,12} = \delta A_{21}/A_{u,21} \). If furthermore, the loaded transmission line is a reciprocal network with \( 1 + A_{li,12}A_{li,21} = A_{li,11}A_{li,22} \), the relation \( (\delta Z/Z_u)^2 = (\delta\theta/\sin(\theta_u))^2 \) must be satisfied. Hence, the relative change in the characteristic impedance can be expressed in terms of the phase change (4.9).
A) Loading Factor

A loaded line section can exhibit predominant reactance or susceptance loading. The loading factor $\xi$ accounts for the relative change in reactance or susceptance (4.10), where the subscripts X and B are assigned to the reactance and susceptance respectively. In the case of reactance loading, relative impedance change and phase change have equal signs $\delta Z/Z_u = +\delta \theta / \sin(\theta_u)$. In contrast, susceptance loading exhibits opposite signs for impedance change and phase change $\delta Z/Z_u = -\delta \theta / \sin(\theta_u)$.

$$
\xi_X = \frac{\delta A_{12}}{A_{u,12}} = \frac{\delta Z}{Z_u} + \frac{\delta \theta}{\tan(\theta_u)} = \frac{\delta \theta}{\sin(\theta_u)}(\cos(\theta_u) + 1)
$$

$$
\xi_B = \frac{\delta A_{21}}{A_{u,21}} = -\frac{\delta Z}{Z_u} + \frac{\delta \theta}{\tan(\theta_u)} = \frac{\delta \theta}{\sin(\theta_u)}(\cos(\theta_u) + 1)
$$

Basically, the four types of RLL sections depicted in figure 2.3 can be used. Those showing low-pass response in their loaded state exhibit a phase increase $\delta \theta > 0$ compared to the unloaded state, whereas the RLL sections offering a high-pass response in their loaded state exhibit a phase decrease $\delta \theta < 0$.

4.2.2 S-parameters

S-parameters are typically chosen to illustrate the loss and phase characteristics of a device. The plot in dB scale allows easy reading of the insertion loss, return loss. Here, the S-parameters are used to express the cut-off frequency and the insertion phase difference $\delta S_{21} = 4 S_{ll,21} - 4 S_{u,21}$ of the RLL sections. In equation 4.11, the forward transmission parameter $S_{ll,21}$ of the loaded-line section is shown.

$$
S_{ll,21} = \frac{2}{A_{ll,11} + A_{ll,22} + \frac{A_{ll,12}}{Z_0} + A_{ll,21} Z_0}
$$

$$
= \frac{2}{2 \cos(\theta_u) \left(1 - \frac{\delta \theta}{\cot(\theta_u)}\right) + j \sin(\theta_u) \left[\frac{Z_u}{Z_0} \left(1 + \frac{\delta Z}{Z_u} + \frac{\delta \theta}{\tan(\theta_u)}\right) + \frac{Z_u}{Z_0} \left(1 - \frac{\delta Z}{Z_u} + \frac{\delta \theta}{\tan(\theta_u)}\right)\right]}
$$

(4.11)

If the line characteristic impedance $Z_u$ is designed to match the port impedances $Z_{01} = Z_{02} = Z_0$, equation 4.12 is valid for the reactance as well as the susceptance loaded-line sections.
In (4.13) the insertion phase $\delta S_{L,21}$ of the loaded-line element is determined. For small phase changes $\delta \theta$, the expression in (4.13) can be simplified to (4.14).

$$
\delta S_{L,21} = \arctan \left( \frac{-3m(S_{L,21})}{\Re(S_{L,21})} \right) = -\arctan(\tan(\theta_u) + \delta \theta)
$$

(4.13)

$$
\delta S_{L,21} \approx -(\theta_u + \delta \theta)
$$

(4.14)

The insertion phase difference $\delta \delta S_{21}$ is defined as the insertion phase of the loaded line subtracted by the insertion phase of the unloaded line. For the unloaded line $\delta S_{u,21} = -\theta_u$ is true. If the phase change $\delta \theta$ is small compared to the electrical length $\theta_u$, the approximation $\delta \theta \approx -(\delta \delta S_{21})$ can be used. Under the above assumption, the characteristic impedance of an ideal reciprocal symmetrical loaded line can be expressed in terms of the insertion phase difference $\delta \delta S_{21}$.

The magnitude response $|S_{L,21}|$ of the RLL section (4.15) is derived from equation 4.12, using the assumption $Z_0/Z_u \approx 1$. According to 4.15 the signal forward transmission $|S_{L,21}|$ is lowered by -0.1 dB for an insertion phase difference of $\delta \delta S_{21} = -9^\circ$.

$$
|S_{L,21}| = \frac{1}{\sqrt{3m(S_{L,21})^2 + \Re(S_{L,21})^2}} = \frac{1}{\sqrt{\cos(\theta_u) \left( 1 - \frac{\delta \theta}{\cot(\theta_u)} \right) + \sin(\theta_u) \left( 1 + \frac{\delta \theta}{\tan(\theta_u)} \right)}}
$$

(4.15)

$$
\approx \frac{1}{\sqrt{1 + (\delta \theta)^2}} \approx \frac{1}{\sqrt{1 + (\delta \delta S_{21})^2}}
$$

This equation states that a single RLL section cannot provide high transmission and a large phase shift at the same time.

4.2.3 T-parameters

In order to calculate cascaded networks, T-parameters are mainly used. Due to their compact exponential form, T-parameter computing is more straightforward than using A-parameters. The matrices $T_{X,i}$ and $T_{B,i}$ represent the T-parameters of a reactance and susceptance loaded transmission-line section. The description (4.16) exactly corresponds to the A-parameter description in (4.9). The port impedances $Z_0$ of the two reference terminals are assumed to equal the characteristic impedance $Z_u$ of the unloaded line.

The analysis of cascaded networks, e.g. a filter or phase shifter, can be performed by the alternate repetition of reactance and susceptance loaded T-parameter matrices for the elements $i = 1 \ldots n$, where $n$ determines the total number of cascaded RLL sections. The loading factor is directly linked to the
phase change (4.10). In the case of electrical short RLL sections, the loading factor can be approximated by $\zeta_i \approx 2\delta\theta_i / \theta_i$. However, using the expression of the phase change $\delta\theta_i$ instead of the loading factor is more convenient.

$$T_{X,i} = \begin{bmatrix} (1 - j\delta\theta_{x,i})e^{-j\theta_{x,i}} & j\delta\theta_{x,i} \\ -j\delta\theta_{x,i} & (1 + j\delta\theta_{x,i})e^{j\theta_{x,i}} \end{bmatrix} = \begin{bmatrix} 1 + (\delta\theta_{x,i})^2 & \delta\theta_{x,i}e^{j\pi/2} \\ \delta\theta_{x,i}e^{-j\pi/2} & 1 + (\delta\theta_{x,i})^2 \end{bmatrix} e^{-j(\theta_{x,i}+\delta\theta_{x,i})}$$

$$T_{B,i} = \begin{bmatrix} (1 - j\delta\theta_{b,i})e^{-j\theta_{b,i}} & -j\delta\theta_{b,i} \\ j\delta\theta_{b,i} & (1 + j\delta\theta_{b,i})e^{j\theta_{b,i}} \end{bmatrix} = \begin{bmatrix} 1 + (\delta\theta_{b,i})^2 & \delta\theta_{b,i}e^{-j\pi/2} \\ \delta\theta_{b,i}e^{j\pi/2} & 1 + (\delta\theta_{b,i})^2 \end{bmatrix} e^{j(\theta_{b,i}+\delta\theta_{b,i})}$$

(4.16)

Equation 4.16 applies to electrically short and long transmission-line sections with moderate line loading.

4.3 Reconfigurable shunt-capacitive loaded microstrip section

Susceptance loading is gained by applying an RLL section in shunt configuration (figure 2.3). In order to ensure configurability, a switch or tuning component is required. A high-pass or low-pass response can be obtained if a short-circuited or open-circuited stub is connected to the microstrip line.

Figure 4.4 shows an electrostatically actuated MEMS switch which connects and detaches an open-circuited stub to the microstrip line. Since a real switch is a non-ideal device, it is described by its non-zero up-state capacitance $C_{ss}$ and its on-state resistance $R_s$. The ohmic contact cantilever switch in shunt configuration is comparable to the variable capacitance switches typically realized by a MEMS bridge in CPW architecture.

**Figure 4.4.** SEM image of an RLL section in shunt configuration. Capacitive loading is accomplished by an electrostatically actuated ohmic contact MEMS switch and an open-circuited stub. The RLL section is inserted between the two terminals T1 and T2.
4.3.1 Equivalent circuit representation

An approximate description of the RLL section can be represented by its equivalent circuit which includes the predominant element values. The equivalent circuit model allows the loading factor and the occurring resonances due to line loading to be estimated. Further the geometry is linked through the element values to the loading factor and the frequency response.

The equivalent circuits of the reconfigurable microstrip section with the switch in off state and on state are depicted in figure 4.5. The MEMS switch adjoined to the open-circuited stub forms a switched capacitor with the two states $C_{\text{opn}}$ and $C_{\text{clsd}}$. The open-state capacitance can be calculated according to (4.17), where $C_{ss}$ denotes the switch series capacitance (4.18) and $C_{o.c.stb}$ the capacitance of the open-circuited stub. The parallel plate capacitance $C_c$ for $N = 2$ contacts with an area of $(15 \, \mu m)^2$ each and a contact separation of $z_0 = 1.1 \, \mu m$ is of the value $C_c = 3.6 \, fF$. The lateral coupling of the narrow microstrip gap with a separation of $s = 25 \, \mu m$, a conductor width of $W = 110 \, \mu m$ and a substrate thickness of $h = 250 \, \mu m$ is calculated to be $C_g \approx 4.3 \, fF$.

Microstrip gaps can be modelled by an equivalent $\pi$-network using shunt and series capacitances [53], [54]. The models summarized by R. K. Hoffmann are reported to provide good approximations up 20 GHz and are used here for estimation.

The capacitance of an open-circuited stub is described by (4.19). The fringing fields at the open end of a microstrip can be calculated by (4.20) according to [53]. In order to account for the exact non-linear frequency dependence, FEM simulation and measurement within relevant bandwidth are required.

\[
C_{\text{opn}} = \left( \frac{1}{C_{ss}} + \frac{1}{C_{o.c.stb}} \right)^{-1} \quad (4.17)
\]

\[
C_{ss} = C_c + C_g = \frac{\varepsilon_0 \varepsilon_{\text{eff}} A_0}{z_0} + C_g \quad (4.18)
\]

\[
C_{\text{clsd}} = C_{o.c.stb} = \frac{Y_{hs}}{\omega} \tan \left( \frac{\omega \sqrt{\varepsilon_{\text{eff}} hs}}{c_0} \right) + C_{ff} \quad (4.19)
\]

\[
C_{ff} = \frac{1}{2} \left( \sqrt{\frac{\varepsilon_{\text{eff}} hs}{c_0 z_0}} - \frac{\varepsilon_0 \varepsilon_{\text{eff}} W}{h} \right) W \quad (4.20)
\]

The unloaded microstrip line exhibits the characteristic impedance of $Z_u$ along its length $l$. To account for the length of the microstrip a symmetrical network extension is made.
Figure 4.5. Equivalent circuit of the reconfigurable shunt-capacitive loaded microstrip section with the MEMS switch (a) in the open state and (b) in the closed state.

Equation 4.21 and 4.22 represent the cascaded A-matrices corresponding to the equivalent circuits in the open state and the closed state, which are depicted in figure 4.5. The matrix $A_u$ denotes the unloaded line used for network extension (4.23). The A-parameters of the reconfigurable capacitive loaded microstrip section for the switch in the open state and the closed state are calculated in (4.24) and (4.25). The parameters show the effect of the loading by the term $\cosh[(\alpha_u + jk_u)l] + 1$.

\[
A_{\text{open}} = A_u \begin{bmatrix} 1 & 0 \\ j\omega C_{\text{open}} & 1 \end{bmatrix} A_u \tag{4.21}
\]

\[
A_{\text{closed}} = A_u \begin{bmatrix} 1 & 0 \\ R_s - \frac{1}{j\omega C_{\text{closed}}} & 1 \end{bmatrix} A_u \tag{4.22}
\]

\[
A_u = \begin{bmatrix} \cosh[(\alpha_u + jk_u)\frac{l}{2}] & Z_u \sinh[(\alpha_u + jk_u)\frac{l}{2}] \\ \frac{1}{Z_u} \sinh[(\alpha_u + jk_u)\frac{l}{2}] & \cosh[(\alpha_u + jk_u)\frac{l}{2}] \end{bmatrix} \tag{4.23}
\]

\[
A_{\text{open},11} = \cosh[(\alpha_u + jk_u)l] + \frac{j\omega C_{\text{open}} Z_u}{2} \sinh[(\alpha_u + jk_u)l]
\]

\[
A_{\text{open},12} = \frac{j\omega C_{\text{open}} Z_u^2}{2} (\cosh[(\alpha_u + jk_u)l] - 1) + Z_u \sinh[(\alpha_u + jk_u)l]
\]

\[
A_{\text{open},21} = \frac{j\omega C_{\text{open}} Z_u^2}{2} (\cosh[(\alpha_u + jk_u)l] + 1) + \frac{1}{Z_u} \sinh[(\alpha_u + jk_u)l]
\]

\[
A_{\text{open},22} = A_{\text{open},11}
\]

\[
A_{\text{closed},11} = \cosh[(\alpha_u + jk_u)l] + \frac{j\omega C_{\text{closed}} Z_u}{2 + 2j\omega C_{\text{closed}} R_s} \sinh[(\alpha_u + jk_u)l]
\]

\[
A_{\text{closed},12} = \frac{j\omega C_{\text{closed}} Z_u^2}{2 + 2j\omega C_{\text{closed}} R_s} (\cosh[(\alpha_u + jk_u)l] - 1) + Z_u \sinh[(\alpha_u + jk_u)l]
\]

\[
A_{\text{closed},21} = \frac{j\omega C_{\text{closed}} Z_u^2}{2 + 2j\omega C_{\text{closed}} R_s} (\cosh[(\alpha_u + jk_u)l] + 1) + \frac{1}{Z_u} \sinh[(\alpha_u + jk_u)l]
\]

\[
A_{\text{closed},22} = A_{\text{closed},11}
\]
The losses expressed by the real part \( \alpha_u \) of the propagation constant are summed up by conductor loss and dielectric loss \( \alpha_u = \alpha_c + \alpha_d \). The conductor losses can be calculated according to the relations (4.26-4.28).

\[
\alpha_c = \frac{20\,\text{dB}}{\ln(10)} \frac{R_{\text{skin}}}{WZ_u} \tag{4.26}
\]

\[
R_{\text{skin}} = \frac{1}{\sigma_{\text{skin}}} \tag{4.27}
\]

\[
t_{\text{skin}} = \sqrt{\frac{2}{\omega \sigma_{\mu_{\mu_r}}}} \tag{4.28}
\]

The dielectric loss in a microstrip on a substrate with the loss-tangent of \( \tan(\delta) \), and permittivity \( \varepsilon_r \) can be calculated according to (4.29).

\[
\alpha_d = \frac{20\,\text{dB}}{\ln(10)} \frac{\omega}{2} \left( \frac{\varepsilon_{\text{eff}} - 1}{\varepsilon_r - 1} \right) \frac{\varepsilon_r}{\varepsilon_{\text{eff}}} \frac{\tan(\delta)}{c_0} \tag{4.29}
\]

Given a 50 \( \Omega \) copper microstrip line with conductor width \( W = 574 \, \mu\text{m} \), conductivity \( \sigma_{\text{Cu}} = 5.8 \cdot 10^7 \, \text{S/m} \), the skin depth and skin resistance for 60 GHz signals is \( t_{\text{skin}} = 270 \, \text{nm} \) and \( R_{\text{skin}} = 0.064 \, \Omega/\text{sq} \). The attenuation can be expressed by \( \alpha_c = 0.02 \, \text{dB/mm} \). The copper line is deposited on a 250 \( \mu\text{m} \) thick fused silica substrate with \( \varepsilon_r = 3.72 \), \( \varepsilon_{\text{eff}} = 3.12 \) and the loss tangent of \( \tan(\delta) = 0.0007 \) for 60 GHz, which results in a dielectric loss in the range of \( \alpha_d = 0.007 \, \text{dB/mm} \).

For a lossless transmission line \( \alpha = 0 \) and \( R_s = 0 \), the loading factor is denoted as \( \zeta_{\text{B,open}} = C_{\text{open}}/C_{\text{pr}} \) and \( \zeta_{\text{B,closed}} = C_{\text{closed}}/C_{\text{pr}} \) in the open and closed state, respectively. The parasitic capacitance \( C_{\text{pr}} \approx \sqrt{\varepsilon_{\text{eff}}/l/(Z_u c_0)} \) and the electrical length is \( \theta_u = \omega \sqrt{\varepsilon_{\text{eff}}/l}/c_0 \). For this special case the equations 4.24 and 4.25 can be rearranged (4.30):

\[
A_{11} = \cos(\theta_u) - \frac{1}{2} \zeta_{\text{B}} \sin^2(\theta_u)
A_{12} = j Z_u \sin(\theta_u) \left[ 1 + \frac{1}{2} \zeta_{\text{B}} (1 - \cos(\theta_u)) \right]
A_{21} = j Z_u \sin(\theta_u) \left[ 1 + \frac{1}{2} \zeta_{\text{B}} (1 + \cos(\theta_u)) \right]
A_{22} = A_{11} \tag{4.30}
\]

If the transmission line is electrically short, \( \theta < \pi/6 \), the relative error occurring for the approximation of \( \cosh(\theta) \) at \( \theta < \pi/6 \) is \( < 12\% \) and \( < 4\% \) respectively. For the above assumption and a moderate loading factor \( \zeta_{\text{B}} \), the parameters \( A_{12} \) and \( A_{21} \) can be simplified (4.31).

Given symmetry \( A_{11} = A_{22} \) and reciprocity \( A_{11}A_{22} - A_{12}A_{21} = 1 \), the approximations of the parameters \( A_{11} \) and \( A_{22} \) can be expressed by \( \sqrt{1 + A_{12}A_{21}} = \sqrt{1 - \theta_u^2 (1 + \zeta_{\text{B}})} \).
The relation $\zeta_B \theta_u^2 = \omega^2 / \omega_{rs}^2$ shows that a microstrip section of short electrical length and a low loading factor $\zeta_B$ exhibits $\omega^2 / \omega_{rs}^2 \ll 1$ and hence $A_{11} = A_{22} \approx 1$. An increase in the loading factor corresponds to a decrease in the resonance frequency $\omega_{rs} = 1 / \sqrt{(\zeta_B C_{pr})L_{pr}}$.

The loading factor is related to the switch geometry, and the type and dimensions of the loading pattern. Thus, in practice the loading is limited by a minimum and a maximum value which correspond to the minimum and maximum values of the shunt capacitance feasible. The presented technology allows for the minimum stub lengths of 190 $\mu$m limited by the lateral dimensions of the MEMS switch. The upper limit of the stub length is given by the frequency-response of the RLL section, which offers a resonance at $\omega_{rs}$ and a second one if the stub length is equal to a quarter wavelength.

For the RLL section depicted in figure 4.4 with a stub length between 110 $\mu$m and 500 $\mu$m, a loading factor ranging from 0.7 to 3.0 can be calculated at 60 GHz. Ideally, the open-state loading factor would be close to zero. However, due to the non-zero value $C_{ss}$, an offset is observed. For example, a stub with length of 440 $\mu$m exhibits a loading factor of 0.47 in the open state. In order to provide a certain change $\delta \zeta_B = \zeta_{B,clsd} - \zeta_{B,opn}$, the open-state loading factor must be taken into account.

### 4.3.2 Simulation and extraction of RLL characteristics

The presented equivalent circuit model is limited in accuracy due to simplifications. Finite element simulations allow to access the frequency response of specific circuit layouts and to extract the line characteristic properties. Therefore electromagnetic wave simulations have been performed using Momentum, a part of Advanced Design System (ADS). Momentum applies the “method of moments” and is capable to solve 2.5D electrodynamic problems [91]. Hence the simulation object is represented by a 2-dimensional layer structure.

The relative changes of reactance and susceptance values in a RLL section are calculated in (4.32) by solving the set of equations given in (4.8).

\[
\begin{align*}
\frac{\delta A_{12}}{A_{u,12}} &= \frac{\delta Z}{Z_u} + \frac{\delta \theta}{\sin(\theta_u)} \left[ \cos(\theta_u) - 1 \right] \\
\frac{\delta A_{21}}{A_{u,21}} &= -\frac{\delta Z}{Z_u} + \frac{\delta \theta}{\sin(\theta_u)} \left[ \cos(\theta_u) + 1 \right] = \zeta_B
\end{align*}
\]
Significant capacitive loading $\zeta_{B, \text{opn}} > 0$ can be present in the off state, if the stub is detached from the microstrip circuit. Thus the parameters $A_{u,11}, A_{u,12}, A_{u,21}, A_{u,22}$ of the unloaded line can differ from those of the line in the off state. The differences $\delta A_{11}, \delta A_{22}, \delta A_{12}$ and $\delta A_{21}$ are found by Momentum simulations of the reconﬁgurable loaded microstrip section in the on state and off state. The associated A-parameters of a shunt-capacitive loaded microstrip section are plotted in figure 4.6. The exact geometry simulated is depicted in figure 4.4. The simulations are performed assuming lossless dielectrics and conductors.

![Graphs showing the parameters $A_{u,11}$, $A_{u,12}$, $A_{u,21}$, and $A_{u,22}$ for the connected and detached open-circuited stub.](image)

**Figure 4.6.** The figure shows the parameters $A_{u,11} = A_{22}, A_{u,12}$ and $A_{21}$ for the connected and detached open-circuited stub. In the plot on the lower right side, the corresponding changes between the on and off state are shown for $\delta Y/Y_u$ and $\delta \theta/tan(\theta_u)$. The shunt-capacitive loaded microstrip section has a length of 110 µm. The open-circuited stub has a width of 110 µm and a length of 340 µm. The dielectrics and conductors are considered to be lossless. FEM simulations are performed using ADS Momentum.

Reorganizing the set of equations (4.32) gives (4.33). The relative changes of the characteristic admittance and the phase between loaded and unloaded line are shown in the plot on the lower right side. In the case of a negligible change $\delta A_{12} \approx 0$, the two equations yield $\delta Y/Y_u = - \delta Z/Z_u \approx \delta \theta/tan(\theta_u)$. The non-linear frequency dependence of the loaded-line characteristic impedance is primarily caused by the open-circuited stub.
\[
\frac{\delta \theta}{\tan(\theta_u)} = \frac{1}{2} \left( \frac{\delta A_{12}}{A_{12}} + \frac{\delta A_{21}}{A_{21}} \right) \approx \frac{1}{2} \delta A_{21} \bigg|_{\delta A_{12}=0} \frac{1}{2 A_{21}} \delta A_{12}
\]

\[
\frac{\delta Z}{Z_u} = \frac{1}{2} \left( \frac{\delta A_{12}}{A_{12}} - \frac{\delta A_{21}}{A_{21}} \right) \approx \frac{1}{2} \delta A_{21} \bigg|_{\delta A_{12}=0} \frac{1}{2 A_{21}} \delta A_{12}
\]

This set of equations, solved for the impedance change and the change of the phase argument, can be very useful to design sections which contain discontinuities.

4.4 Reconfigurable series-inductive loaded microstrip section

Next to the RLL sections in shunt configuration, sections in series configuration can be used to obtain high-pass and low-pass response circuit elements. In figure 4.7, a series-inductive loaded RLL section is depicted. Between the two terminals T1 and T2, two parallel lines with \( Z_h \) and \( \theta_h \) and with \( Z_{hs} \) and \( \theta_{hs} \) are inserted. The latter is formed by a MEMS switch anchored on terminal T2. In the open state, the switch forms a capacitance close to the terminal T1, which results in a current constriction in the vicinity of the terminals. In the closed state, the switch bridges the constriction with a low resistance contact.

![Figure 4.7. SEM image of an RLL section in series configuration. Inductive loading is obtained by a conductor constriction between the terminals T1 and T2. The MEMS switch enables the constriction to be shorted out.](image)

4.4.1 Equivalent circuit representation

In order to express the loading factor and to estimate the resonances caused by the loading, an equivalent circuit model for the series-inductive loaded microstrip section is developed. The equivalent circuit of the RLL section depicted in figure 4.7 is drawn in figure 4.8. The discontinuities
can be described by a series inductance $L_{dsc}$ and a shunt capacitance $C_{dsc}$. The inductance is caused by an increase in the current density in the adjacent regions outside of the microstrip section. Relation (4.34) is fulfilled for $Z_u < Z_h$. The factor $\delta = 2$ is used for a symmetric discontinuity and $\delta = 1$ for an asymmetric discontinuity [53].

$$L_{dsc} = \frac{2\mu_0 h}{\delta \pi} \ln \left[ \sin^{-1} \left( \frac{\pi Z_u \sqrt{\varepsilon_{r,eff,u}}}{Z_h \sqrt{\varepsilon_{r,eff,h}}} \right) \right] \quad (4.34)$$

The capacitance is caused by the scattering field along the edge of the insert. Alternatively, the capacitance can be replaced by a virtual conductor length extension as described in [53]. The relevant substrate and microstrip properties are summarized in table 4.1.

### Table 4.1. Summary of the effective permittivity and characteristic impedance of the microstrip line elements.

<table>
<thead>
<tr>
<th>Description</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>substrate dielectric constant at 60 GHz, 25°C</td>
<td>$\varepsilon_r$</td>
<td>3.72</td>
</tr>
<tr>
<td>substrate thickness</td>
<td>$h$</td>
<td>250 $\mu$m</td>
</tr>
<tr>
<td>MS line $Z_u = 50$ $\Omega$, conductor width</td>
<td>$w$</td>
<td>574 $\mu$m</td>
</tr>
<tr>
<td>MS line $Z_u = 50$ $\Omega$, effective dielectric constant</td>
<td>$\varepsilon_{r,eff}$</td>
<td>3.12</td>
</tr>
<tr>
<td>MS line $Z_{hs} = 110$ $\Omega$, conductor width</td>
<td>$w_{hs}$</td>
<td>110 $\mu$m</td>
</tr>
<tr>
<td>MS line $Z_{hs} = 110$ $\Omega$, effective dielectric constant</td>
<td>$\varepsilon_{r,eff,hs}$</td>
<td>2.72</td>
</tr>
</tbody>
</table>

Since the microstrip constriction is mainly filled by the cantilever, additional capacitance due to scattering fields is not expected, and thus $C_{dsc} \approx 0$. The MEMS switch in the open state is considered to be a separate parallel path with different characteristic impedance. In the closed state, the equivalent circuit simplifies to that of an unloaded microstrip line, with additional loss due to the contact resistance $R_c$ of the MEMS switch.

![Figure 4.8](image-url)  
**Figure 4.8.** Equivalent circuit of the reconfigurable series-inductive loaded microstrip section with the MEMS switch (a) in the open state and (b) in the closed state.
The cascaded A-matrices of the microstrip section with the switch in the open state and closed state can be written as (4.35) and (4.36). The upper and lower path, with the switch in the open state, can be calculated separately. For small gaps below 30 μm, the capacitance $C_{pe}$ is negligibly small compared with $C_{ss}$ (4.29) and the parasitic line capacitance $C_{pr,hs}$.

$$A_{opn,upr} = \begin{bmatrix} 1 & \frac{1}{j\omega L_{sc}} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \cosh((\alpha_u + jk_u)l_h) & Z_h\sinh((\alpha_u + jk_u)l_h) \\ Z_h\sinh((\alpha_u + jk_u)l_h) & \cosh((\alpha_u + jk_u)l_h) \end{bmatrix}$$

$$A_{opn,lwr} = \begin{bmatrix} 1 & \frac{1}{j\omega L_{sc}} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \cosh((\alpha_u + jk_u)l_h) & Z_h\sinh((\alpha_u + jk_u)l_h) \\ Z_h\sinh((\alpha_u + jk_u)l_h) & \cosh((\alpha_u + jk_u)l_h) \end{bmatrix}$$

(4.35)

$$A_{cst} = \begin{bmatrix} 1 & R_{cst} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \cos((\alpha_u + jk_u)l) & Z_u\sin((\alpha_u + jk_u)l) \\ Z_u\sin((\alpha_u + jk_u)l) & \cosh((\alpha_u + jk_u)l) \end{bmatrix}$$

(4.36)

$Y$-parameter conversion (4.37) is used to add the two parallel sub-networks by $Y_{opn} = Y_{opn,upr} + Y_{opn,lwr}$.

$$Y = \begin{bmatrix} A_{22} & \frac{A_{12}A_{22}}{A_{11}} \\ \frac{A_{12}}{A_{11}} & \frac{A_{12}^2}{A_{11}} \end{bmatrix}$$

(4.37)

For a lossless transmission-line network, $\alpha_h = \alpha_{hs} = \alpha_{st} \approx 0$ and $R_s \approx 0$, the resultant $Y$-parameters are given by (4.38):

$$Y_{11} = \frac{1 - \omega L_{dc} + \frac{1}{Z_h}\tan(\theta_h)}{\cos(\theta_h) \left[ j\omega L_{dc} \left( 2 - \omega L_{dc} \frac{1}{Z_h}\tan(\theta_h) \right) + jZ_h\tan(\theta_h) \right] + \cos(\theta_h) \left[ \frac{1}{j\omega L_{ss}} + jZ_h\tan(\theta_h) \right] + \frac{1}{\omega L_{ss} Z_h}}$$

$$Y_{12} = \frac{1}{Z_h^2} \sin(\theta_h) + \frac{1}{Z_h^2} \sin(\theta_{hs}) - \cos(\theta_h) \left[ \left( 1 - \omega L_{dc} \frac{1}{Z_h}\tan(\theta_h) \right)^2 \right] - \cos(\theta_h) \left[ 1 + \frac{1}{\omega L_{ss} Z_h} \frac{1}{Z_h}\tan(\theta_{hs}) \right]$$

$$Y_{21} = \frac{1}{j\omega L_{dc} \left( 2 - \omega L_{dc} \frac{1}{Z_h}\tan(\theta_h) \right) + jZ_h\tan(\theta_h) + \cos(\theta_h) \left[ \frac{1}{j\omega L_{ss}} + jZ_h\tan(\theta_{hs}) \right] + \frac{1}{\omega L_{ss} Z_h}}$$

$$Y_{22} = \frac{1}{j\omega L_{dc} \left( 2 - \omega L_{dc} \frac{1}{Z_h}\tan(\theta_h) \right) + jZ_h\tan(\theta_h) + \cos(\theta_h) \left[ \frac{1}{j\omega L_{ss}} + jZ_h\tan(\theta_{hs}) \right] + \frac{1}{\omega L_{ss} Z_h}}$$

(4.38)

For small phase arguments, the trigonometric functions can be simplified to $\sin(\theta_h) \approx \theta_h$, $\sin(\theta_{hs}) \approx \theta_{hs}$, $\cos(\theta_{hs}) \approx \cos(\theta_h) \approx 1$, $\tan(\theta_h) \approx \theta_h$, $\tan(\theta_{hs}) \approx \theta_{hs}$. Subsequently, the frequency can be scaled relative to the resonances $\omega_{rs1} = 1/\sqrt{L_{dc} C_{pr,hs}}$ and $\omega_{rs2} = 1/\sqrt{C_{ss} L_{pr,hs}}$. 

- 79 -
present in the two branches. The simplified Y-parameters are given in equation (4.39). The parasitic capacitance and inductance in the upper path are \( C_{pr,h,s} \approx \theta_{hs}/(\omega Z_{hs}) \) and \( L_{pr,h,s} \approx Z_{hs}\theta_{hs}/\omega \). The parasitics in the lower path are \( C_{pr,h} \approx \theta_h/(\omega Z_h) \) and \( L_{pr,h} \approx Z_h\theta_h/\omega \). For \( \omega^2/\omega_{rs}^2 \ll 1 \), the lower path behaves like a lumped inductance. For \( \omega^2/\omega_{rs}^2 \ll 1 \) the upper path, comprising the MEMS switch, behaves like a lumped capacitance.

\[
Y_{11} = \frac{1 - \omega^2}{j\omega L_{dsc} \left( 2 + \frac{L_{pr,h}}{L_{dsc}} - \frac{\omega^2}{\omega_{rs}^2} \right)} + \frac{1}{1 - \frac{\omega^2}{\omega_{rs}^2}}
\]

\[
Y_{12} = \frac{\theta_h + \theta_{hs}}{Z_h} - \frac{1 - \omega^2}{j\omega L_{dsc} \left( 2 + \frac{L_{pr,h}}{L_{dsc}} - \frac{\omega^2}{\omega_{rs}^2} \right)} + \frac{1 + C_{pr,h}}{1 - \frac{\omega^2}{\omega_{rs}^2}}
\]

\[
Y_{21} = -\frac{1}{j\omega L_{dsc} \left( 2 + \frac{L_{pr,h}}{L_{dsc}} - \frac{\omega^2}{\omega_{rs}^2} \right)} + \frac{1}{1 - \frac{\omega^2}{\omega_{rs}^2}}
\]

\[
Y_{22} = \frac{1 - \omega^2}{j\omega L_{dsc} \left( 2 + \frac{L_{pr,h}}{L_{dsc}} - \frac{\omega^2}{\omega_{rs}^2} \right)} + 1 + \frac{C_{pr,h}}{1 - \frac{\omega^2}{\omega_{rs}^2}}
\]

Equation (4.39)

In order to better compare the characteristics of the loaded microstrip section, we convert the Y-parameter back to the A-parameters (4.40). The A-parameters (4.41) are obtained, where the angular resonance frequency \( \omega_{rs} \approx 1/\sqrt{L_{dsc}C_{ss}} \) represents the parallel resonance between the upper and lower path.

\[
A = \begin{bmatrix}
-Y_{22} & -1 & -Y_{22} & -1 \\
Y_{21} & Y_{11} & Y_{21} & Y_{11} \\
Y_{12} + Y_{11} & -Y_{11} & Y_{12} & -Y_{11} \\
Y_{21} & Y_{11} & Y_{21} & Y_{11}
\end{bmatrix}
\]

Equation (4.40)

\[
A_{opn,11} \approx 1 + \frac{\omega^2}{\omega_{rs}^2} \left( \frac{2 + \frac{L_{pr,h}}{L_{dsc}} - \frac{\omega^2}{\omega_{rs}^2}}{1 - \frac{\omega^2}{\omega_{rs}^2}} \right) C_{pr,h} \left( \frac{1 - \frac{\omega^2}{\omega_{rs}^2}}{1 - \frac{\omega^2}{\omega_{rs}^2}} \right)^{-1}
\]

Equation (4.41)

\[
A_{opn,12} \approx j\omega L_{dsc} \left( 2 + \frac{L_{pr,h}}{L_{dsc}} - \frac{\omega^2}{\omega_{rs}^2} \right) \left( \frac{1 - \frac{\omega^2}{\omega_{rs}^2}}{1 - \frac{\omega^2}{\omega_{rs}^2}} \right)^{-1}
\]

\[
A_{opn,21} \approx j\frac{1}{Z_h} \theta_{hs} + \frac{1}{Z_h} \theta_h \left( \frac{\omega^2}{\omega_{rs}^2} \left( \frac{A_{opn,11} - \theta_{hs} \left( 1 + \frac{C_{pr,h}}{C_{ss}} \right)}{1 - \frac{\omega^2}{\omega_{rs}^2}} \right) - \frac{\omega^2}{\omega_{rs}^2} \left( 2 + \frac{L_{pr,h}}{L_{dsc}} - \frac{\omega^2}{\omega_{rs}^2} \right) \right)
\]

Equation (4.44)
The loading factor of the inductive loaded microstrip section is calculated by (4.42). Dependent on the series and parallel resonances in the upper and lower path, the loading factor may simplify. The maximum possible loading factor can be increased by reducing the parasitic inductivity $L_{pr,u}$ and hence reducing the terminal separation.

For microstrip sections with small electrical length, the approximation $A_{opn,22} \approx 1$ can be used, which leads to (4.43). The loaded microstrip can be expressed in terms of the loading factor $\xi_X$ (4.42). For longer electrical length, the symmetry assumption $A_{opn,11} \approx A_{opn,22}$ may fit better. Thus, the $A$-matrix for the open and closed state is given in (4.44). Further, $\theta_h / Z_h + \theta_{hs} / Z_{hs} \approx \theta_u / Z_u$ is a good approximation.

$$\xi_X = \frac{A_{opn,12}}{jZ_u \sin(\theta_u)} - 1 = \frac{(2L_{dc} + L_{pr,h}) - L_{pr,u}}{L_{pr,u}}$$  \hspace{1cm} (4.42)

$$A_{opn} \approx \begin{bmatrix} 1 - \frac{\theta_h}{Z_h} + \frac{\theta_{hs}}{Z_{hs}} & jZ_u \theta_u(1 + \xi_X) & jZ_u \theta_u(1 + \xi_X) \\ j \frac{\theta_h}{Z_h} + \frac{\theta_{hs}}{Z_{hs}} & 1 \end{bmatrix}$$  \hspace{1cm} (4.43)

$$A_{opn} \approx \begin{bmatrix} \sqrt{1 - \theta_u^2[1 + \xi_X]} & jZ_u \theta_u[1 + \xi_X] \\ j \frac{\theta_u}{Z_u} & \sqrt{1 - \theta_u^2[1 + \xi_X]} \end{bmatrix}$$  \hspace{1cm} (4.44)

### 4.4.2 Simulation and extraction of RLL characteristics

Similar to section 4.3.2, the impedance change and the change of the phase argument are determined here for the series-inductive loaded microstrip section. The parameters $A_{11}, A_{12}, A_{21}$ and $A_{22}$ are extracted from the finite element simulations performed using Momentum 2.5D. As shown in figure 4.7, the inductively loaded microstrip section is constricted by a metallization cut-out, which can be shunted out by a MEMS switch. Given reciprocity, the relation $(\delta Z / Z_u)^2 = (\delta \theta / \sin(\theta_u))^2$ must be fulfilled. In fact, the inductive constriction cannot be directly compared to a uniform microstrip line. Given the stepped conductor width, and thus the current density increase in the adjoined microstrip, $\delta \theta$ will be positive if $W / h$ is locally decreased. Thus, we write $\delta Z / Z_u = + \delta \theta / \sin(\theta_u)$ in order to provide reciprocity.

$$\frac{\delta A_{12}}{A_{12}} = \frac{\delta Z}{Z_u} + \frac{\delta \theta}{\tan(\theta_u)} = \frac{\delta \theta}{\sin(\theta_u)}[\cos(\theta_u) + 1] = \xi_X$$  \hspace{1cm} (4.45)

$$\frac{\delta A_{21}}{A_{21}} = - \frac{\delta Z}{Z_u} + \frac{\delta \theta}{\tan(\theta_u)} = \frac{\delta \theta}{\sin(\theta_u)}[\cos(\theta_u) - 1]$$
Since in practice the line can exhibit a certain loading in the off state, the values of the unloaded line, $A_{u,11}, A_{u,12}, A_{u,21}$ and $A_{u,22}$ are not representative for the off state. FEM simulations in the off state and on state allow the changes $\delta A_{12}$ and $\delta A_{21}$ to be calculated. Reorganizing the set of equations (4.45) we obtain (4.46). For the special case of $\delta A_{21} \approx 0$, $\delta Z / Z_u \approx \delta \theta / \tan(\theta_u)$ is a good approximation.

\[
\frac{\delta \theta}{\tan(\theta_u)} = \frac{1}{2} \left( \frac{\delta A_{12}}{A_{u,12}} + \frac{\delta A_{21}}{A_{u,21}} \right) \approx \frac{1}{2} \frac{\delta A_{12}}{A_{u,12} |_{\delta A_{21} = 0}}
\]

\[
\frac{\delta Z}{Z_u} = \frac{1}{2} \left( \frac{\delta A_{12}}{A_{u,12}} - \frac{\delta A_{21}}{A_{u,21}} \right) \approx \frac{1}{2} \frac{\delta A_{12}}{A_{u,12} |_{\delta A_{21} = 0}}
\]

(4.46)

**Figure 4.13.** The figure shows the parameters $A_{11}, A_{12}, A_{21}, A_{22}$ for the open and closed state. In the open state, the constriction acts like a series inductance. In the closed state, the switch bridges the gap and forms a 50 Ω microstrip line. $A_{11}$ and $A_{22}$ are plotted for the open state, since in this configuration the network becomes asymmetric. In the plot on the lower right side, the corresponding changes between the on and off state are shown for $\delta Z / Z_u$ and $\delta \theta / \tan(\theta_u)$. The series-inductive loaded microstrip section has a length of 195 µm. The insert has a width of 190 µm and a length of 190 µm. The dielectrics and conductors are considered to be lossless. FEM simulations are performed using ADS Momentum.

Figure 4.13 shows the A-parameters of the inductive loaded microstrip section in the open state and the closed state. In the open state, the constriction acts like a series inductance. In the closed state, the switch bridges the gap and forms a 50 Ω microstrip line. The relative changes in the characteristic
impedance and the phase angle are depicted in the plot on the lower right side. Compared to the shunt-capacitive loaded microstrip section, the series-inductive loaded section exhibits minor frequency dependence of the characteristic impedance and phase difference.
5-bit phase shifter using reconfigurable loaded microstrip sections

5.1 Introduction

This chapter describes a new phase shifter design procedure. The procedure allows to synthesize cascaded networks of multiple low-pass reconfigurable loaded-line (RLL) sections. The approach relates the layout and dimensions of the RLL sections to the element values for low pass prototype filters. Thus the benefits of choosing a certain prototype can be gained. The conflict between low passband ripple and high insertion phase range or cutoff is illustrated for the Chebyshev, tapered and non-tapered Bragg filters. The effects of the periodic arrangement, the element loading, and the taper function are described. Exemplarily, the design procedure is applied to synthesize a phase shifter required to provide an insertion phase difference of -45° at 60 GHz.

The T-parameter description of the RLL sections is used for rapid frequency response calculation and facilitates the comparison between simulation and measured data. The simulation data for the 5-bit states of the designed phase shifter are summarized. The limitations of the analytic model are discussed by comparison with the simulation.

RF phase shifter characterization is performed using on-wafer probing in combination with probe-to-microstrip transitions (P-to-MS). The comprehensive analysis of the measurement data comprises the S-parameters of the P-to-MS transition, and those of the 5-bit states of the phase shifter. Moreover, the phase responses and the phase errors for the phase shifter states 1 to 5 are illustrated.

5.2 Phase shifter synthesis

The proposed synthesis procedure can be applied to design phase shifters based on the distributed MEMS transmission-line (DMTL) architecture. Here, millimetre-wave phase shifters are synthesized using reconfigurable loaded transmission-line sections in series and shunt configuration. As shown in the previous chapter by the equations 4.7 to 4.10 the loaded-line sections provide predominant reactance or susceptance loading. The loading (5.1) can be linked to the element values $g_l$ of common prototype filters, which enables a network with a desired frequency response to be synthesized.

$$\delta A_{12,x,t} = jZ_o g_l \frac{\omega}{\omega_{ef}}$$  \hspace{1cm} (5.1)
The values $\delta A_{21,B,i}$ and $\delta A_{21,X,i}$ denote the series reactance $jX$ and shunt susceptance $jB$ loading of a low-pass section. The symbol $\delta$ indicates that the filter coefficients are related to the change of reactance or susceptance between the on state and off state of the RLL section. Using the definition (4.10), the reactance or susceptance change can be determined according to (5.2)

$$
\delta A_{21,B,i} = \zeta_{B,i} A_{21,B,i} = jY_{\omega_{ctf}} \theta_{B,i} \left[ \cos(\theta_{B,i}) + 1 \right]
$$

(5.2)

If assuming the cut-off frequency $\omega_{ctf}$ to be invariant, proportionality between $g_i \omega$ and $\delta \theta_i \left[ \cos(\theta_i) + 1 \right]$ is expected. Thus, for a phase shifter with an $n^{th}$ order low-pass response using multiple cascaded RLL sections of similar electrical length, the insertion phase difference of each loaded-line element $\delta \Delta S_{21,i} \approx -\delta \theta_i$ can be derived from the total desired insertion phase difference $\delta \Delta S_{21,csc} \approx \sum_{i=1}^{n} \delta \Delta S_{21,i}$ and the element values of the low pass prototype (5.3). Subsequently, the obtained design values can be substituted in the $T$-parameter description (4.16).

$$
\delta \Delta S_{21,i} = \frac{g_i}{\sum_{i=1}^{n} g_i} \delta \Delta S_{21,csc}
$$

(5.3)

The loading factor of each element is proportional to the relative value of the filter coefficient. Similarly, the change in the reactance and susceptances is expressed by the relative value $g_i / \sum_{i=1}^{n} g_i$. In the case of sections with equal electrical length in the their off state, the cut-off frequency exhibited by an ideal lumped element filter is determined by the sum of the element values of the low pass prototype and the total desired insertion phase difference of the phase shifter (5.4). Although the relative cutoff is not necessarily representative for distributed filters, the equation states that the relative cut-off frequency and thus the phase range increases with increasing filter order.

$$
\omega_{ctf} = \frac{\sum_{i=1}^{n} g_i}{[\cos(\theta_i) + 1] \delta \Delta S_{21,csc} \omega_0} \approx \frac{1}{2} \frac{\sum_{i=1}^{n} g_i}{\delta \Delta S_{21,csc} \omega_0}
$$

(5.4)

The architectural limitation on the width of the inserts and the length of the stubs needs to be considered when choosing the prototype filter. The realization of the proposed RLL sections is feasible only for a certain range of filter coefficients. In this thesis, the Chebyshev and Bragg prototypes are used. This is because their filter coefficients vary less compared to the Bessel or Butterworth coefficients.
5.2.1 Taper function

The simplest structure of a DMTL phase shifter uses uniform distributed loading, which can result in high passband ripples [42]. Weighted loading using common element values of prototype filters allows reducing pass-band ripples over a large bandwidth. The weighted loading typically results in a taper-like layout as shown in figure 5.1. For better visualization, the shunt-capacitive and series-inductive loading patterns are assigned by their corresponding filter coefficients $g_i$.

![Figure 5.1. Phase shifter layout. The analytic design is based on the Chebyshev filter prototype, with a passband ripple of $L_{Ar} = 0.01$ dB. The filter comprises 9 RLL sections, labelled with its corresponding filter coefficient.](image)

As an example, the synthesis procedure is carried out here for a phase shifter, required to provide an insertion phase difference of $-45^\circ$ at 60 GHz. The phase shifter design is based on a Chebyshev low-pass filter of the 9th order. The corresponding element values are listed in table 5.2. Similarly, the element values of the tapered and non-tapered Bragg filters are given in table 5.1. The phase shifters use series-inductive and shunt-capacitive RLL sections arranged in an alternating order.

<table>
<thead>
<tr>
<th>Table 5.1. Element values of the Bragg prototype filters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bragg filter</td>
</tr>
<tr>
<td>$g_i = \begin{cases} g_1 = g_n = 2.0\gamma \ g_2, ..., g_{n-1} = 2.0 \end{cases}$</td>
</tr>
</tbody>
</table>

In figure 5.2 and figure 5.3, the calculated S-parameters of an ideal lossless phase shifter are plotted. The magnitude and phase responses are those of a Chebyshev, with pass-band attenuation $L_{Ar} = 0.01$ dB, a tapered Bragg with $\gamma = 0.5$ and a non-tapered Bragg filter. In order to calculate the S-parameter, the design values given in table 5.2 are inserted into equation 4.16. Subsequently the
matrix multiplication is performed for the 9 sections. The figures show that the passband ripple can be reduced at a minimum expanse of bandwidth.

Figure 5.2. Forward scattering magnitude $|S_{21}|$ of a Chebyshev, tapered Bragg and non-tapered Bragg filter.

Figure 5.3. Insertion phase difference $\delta S_{21} = 4S_{21,LL} - 4S_{21,UL}$ of a Chebyshev, tapered Bragg and non-tapered Bragg filter.

The synthesis is based on the loaded-line model described in section 4.2, which uses a linear approximation. Thus, the model is accurate for frequencies close to the target value $f_0 = 60$ GHz. Higher order modes need to be considered above 70 GHz, since for this frequency the microstrip width is close to a quarter wavelength. Figure 5.4 shows the comparison between the analytic calculation and the Momentum simulation of the exact layout. A deviation of $3^\circ$ can be found between the analytic solution and the simulation. This deviation is caused due to the coupling between adjacent RLL sections, which is not taken into account by the analytic approach. In the following section 5.2.3 the limitation of the presented design procedure is discussed.
Figure 5.4. Insertion phase of the phase shifter layout depicted in figure 5.1 with the dimensions defined in table 5.2. Comparison of the analytic model with the simulation results from ADS Momentum.

The linkage between the insertion phase difference and the layout-specific dimensions of the RLL sections is done using ADS Momentum. In table 5.2, the calculated stub length \( l_{stb} \) and insert width \( w_{ins} \) are listed for the targeted design values \( \delta \theta_i \). These dimensions can be used as the starting values for optimization, which is performed in Momentum for the complete network.

Table 5.2. Element values of the Chebyshev prototype filter, insertion phase difference, and corresponding dimensions of the RLL sections 1-9.

<table>
<thead>
<tr>
<th>( i )</th>
<th>( g_i )</th>
<th>( \delta \theta_i )</th>
<th>( l_{stb,i} )</th>
<th>( w_{ins,i} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1, 9</td>
<td>0.8145</td>
<td>-2.73</td>
<td>232 ( \mu m )</td>
<td></td>
</tr>
<tr>
<td>2, 8</td>
<td>1.4271</td>
<td>-4.78</td>
<td></td>
<td>173 ( \mu m )</td>
</tr>
<tr>
<td>3, 7</td>
<td>1.8044</td>
<td>-6.05</td>
<td></td>
<td>326 ( \mu m )</td>
</tr>
<tr>
<td>4, 6</td>
<td>1.7125</td>
<td>-5.74</td>
<td></td>
<td>188 ( \mu m )</td>
</tr>
<tr>
<td>5</td>
<td>1.9058</td>
<td>-6.39</td>
<td></td>
<td>333 ( \mu m )</td>
</tr>
</tbody>
</table>

In the case of the shunt-capacitive loaded RLL section, the insertion phase difference \( \delta S_{21} = \delta S_{21, on} - \delta S_{21, off} \) depends on the switch contact capacitance and the length of the open-circuited stub. The insertion phase in the off state can exhibit an offset compared to the insertion phase of the ideal unloaded transmission line. In the case of the contact separation \( z_0 = 0.55 \, \mu m \) and stub length \( l_{stb} = 430 \, \mu m \), the insertion phase offset is -1.5° at 60 GHz. The phase offset becomes negligible at 60 GHz for contact separation exceeding \( z_0 = 1.1 \, \mu m \) and the same stub length \( l_{stb} \). In figure 5.5 the insertion phase difference of the RLL section is plotted versus the stub length at switch contact separation of 1.1 \( \mu m \).
Figure 5.5. Insertion phase difference $\delta \angle S_{21} = \angle S_{21,\text{on}} - \angle S_{21,\text{off}}$ versus stub length $l_{\text{stb}}$. The ADS Momentum simulation uses the terminal separation of $l = 112 \, \mu m$ and the contact separation of $z_0 = 1.1 \, \mu m$. The phase values are taken at the frequency of 60 GHz.

In the case of the series-inductive loaded line, the insertion phase difference depends in a similar manner on the separation $z_0$. In figure 5.6, the insertion phase difference is plotted versus the insert width $w_{\text{ins}}$ at the contact separations of 0.5 μm, 1.1 μm and 2.0 μm.

Figure 5.6. ADS Momentum simulation results showing the insertion phase difference $\delta \angle S_{21} = \angle S_{21,\text{on}} - \angle S_{21,\text{off}}$ versus the insert width $w_{\text{ins}}$ at different contact separations $z_0$. The terminal separation is set to $l = 197 \, \mu m$ and the insertion phase is given for the frequency of 60 GHz.
5.2.2 Bragg reflection

In periodic structures, Bragg reflection is observed if interaction between the wave and the structure takes place. For the proposed concept of loaded-line filters and phase shifters using low-pass and high-pass response RLL sections in shunt or series configuration, the smallest periodic element consists of at least two different sections. In figure 5.7 a cascaded network of a low-pass filter is depicted together with its equivalent S-parameter flow graph. The smallest periodic element is repeated \( n/2 \) times while the filter comprises a total number of \( n \) RLL sections.

The forward reflection parameter \( S'_{11} \) is calculated according to (5.5). Using the substitution (5.7) and (5.8) we obtain the form given by (5.6). The suscript \( l \) and \( k \) are odd and even numbers respectively, where \( k < n \) and \( l = k + 1 \). Equation 5.6 shows that the reflected waves returning from different discontinuities sum up. If the phase argument of each reflected wave differs from each other by a multiple of 2\( \pi \) the reflections superimpose constructively, and hence a low return loss is observed at the input port.

\[
S'_{11} = \left| \frac{b_1}{a_1} \right| \left| S_{11} \right| \exp(-j\delta S_{11}) + \left| S_{21} \right| \left| S_{12} \right| \exp(-j\delta S_{21}) + \left| S_{31} \right| \left| S_{13} \right| \exp(-j\delta S_{31}) + \cdots
\]

\[
S_{11} = \left| S_{11} \right| \exp(-j\delta S_{11}) + \left| S_{21} \right| \left| S_{12} \right| \exp(-j\delta S_{21}) + \left| S_{31} \right| \left| S_{13} \right| \exp(-j\delta S_{31}) + \cdots
\]

\[
S'_{11} = S_{11} + \frac{S_{21}S_{12}}{1 - S_{22}S_{33}} S_{33} + \frac{S_{21}S_{34}S_{43}S_{12}}{1 - S_{44}S_{55}} S_{55} + \cdots + \frac{S_{n1}S_{1n}}{1 - S_{nn}S_{0n+1}} S_{0n+1}
\]

\[
S'_{11} = S_{11} + \frac{S_{33}}{1 - S_{22}S_{33}} \left[ S_{21} S_{12} (1 + S_{34} S_{43} + \cdots) \right] + \frac{S_{n1}S_{1n}}{1 - S_{nn}S_{0n+1}} S_{0n+1}
\]

\[
S'_{11} = S_{11} + \frac{S_{21}}{1 - S_{22}} \left[ S_{21} S_{12} (1 + S_{34} S_{43} + \cdots) \right] + \frac{S_{n1}S_{1n}}{1 - S_{nn}S_{0n+1}} S_{0n+1}
\]
Since the reflection coefficient \( \Gamma_{kll} \) is repeated together with the periodic units, the vector sum in (5.6) is maximized if the additional path \( 4S_{kl} + 4S_{lk} \) due to an additional element is equal to \( m2\pi \).

For small \( |S_{11}| \) and \( |\Gamma_{n0n+1}| \), and with an increasing number of repeated units, the effect of the reflections returning from the input and output terminal \( T_1 \) and \( T_n \) becomes negligible.

The high insertion loss at the Bragg resonances can disrupt the passband \( \omega < \omega_{ctf} \). The Bragg reflection of a periodic network comprising more than two periodic units can be determined by (5.9), where \( \theta_i \) denotes the electrical length of the sections in their unloaded state and \( \delta \theta_i \) denotes the phase change due to the line loading. The number \( n \) is the filter order, whereas the number of periodic elements is \( n/2 \) if the filter consists of an even number of sections. Bragg reflections occur for any integer value of \( m \). If all RLL sections exhibit the same electrical length and phase change (5.9) is reduced to \( \theta_i + \delta \theta_i \cong \pi/2 \).

\[
\frac{1}{n} \left[ 2 \sum_{x,i} (\theta_{x,i} + \delta \theta_{x,i}) + 2 \sum_{b,i} (\theta_{b,i} + \delta \theta_{b,i}) \right] \cong (2m-1)\pi
\]  

In equations 5.10 and 5.11 the average propagation constant and phase velocity of the loaded line is shown. The first Bragg reflection frequency is determined by \( \omega_{Bragg,1} = \tilde{\nu}_{ph,1} \pi / (l_x + l_B) \). The return loss minima and the insertion loss maxima are given at \( \tilde{\beta}_{ll}(l_x + l_B) = (2m-1)\pi \).

\[
\tilde{\beta}_{ll} = \sum_{i}^{n} \frac{(\theta_i + \delta \theta_i)}{l_i} \approx \sum_{i}^{n} \frac{\theta_i (1 + \frac{1}{2} \zeta_i)}{l_i}
\]

\[
\tilde{\nu}_{ph,1} = \frac{\varepsilon_0}{\sqrt{\varepsilon_{r,eff,1}}} \approx \frac{\varepsilon_0}{\sqrt{\varepsilon_{r,eff,u}}} \left[ l_x \sum_x (1 + \frac{1}{2} \zeta_x) + l_B \sum_{b} (1 + \frac{1}{2} \zeta_b) \right]
\]

For the example in section 5.2.1, the first Bragg reflection \( f_{Bragg,1} \) is estimated to 243 GHz, which coincides with the sharp decrease in the insertion loss above 160 GHz (figure 5.2) and limits the linear phase response. In contrast, the cut-off frequency of the idealized lumped element filter is estimated above \( f_{ctf} = 500 \) GHz (5.4).
5.2.3 Simulation and limitation of the design procedure

Compared to the analytic approach, computational numerical calculation allows the design accuracy to be improved. Thus simulations have been performed using Advanced Design System (ADS), which computes solutions to electromagnetic problems applying the method of moments (MoM). The underlying 2.5D model comprises a stacked dielectric body with interjacent conductive layers.

Momentum simulation was used to tune the dimensions of the stub $l_{stub}$ and the insert $w_{ins}$ in order to obtain the desired insertion phase difference. In Table 5.3 the final simulation results for the different phase shifter states are summarized. Chebyshev and tapered Bragg filter prototypes are used. The latter are preferred for low order filters. The table shows the insertion phase difference between the on state and off state. Further, it lists the insertion phase difference in relation to the length $l_{DMTL}$ of the phase shifter. The ratio $\delta \alpha_{21}/l_{DMTL}$ is proportional to the average loading factor.

Since the loading factor and relative insertion phase change are frequency-dependent, dispersion and hence a frequency-dependent group delay is observed. The targeted low insertion loss and the linear phase response within the phase range of use both require a high bandwidth. This limits the maximum loading of the single RLL section and determines the minimum filter order.

Table 5.3. Summary of Momentum simulation results for $\omega_0 = 2\pi \cdot 60$ GHz. Optimization was performed for each state. For low filter orders the element values can vary in a wide range, which results in a non-feasible design. In that case the Bragg filter was chosen. Since dispersion is present, the time delay is not completely frequency-invariant. The value $\delta \alpha_{21}/l_{DMTL}$ is proportional to the average loading factor.

<table>
<thead>
<tr>
<th>state</th>
<th>n</th>
<th>$\left(\alpha_{21,on} - \alpha_{21,off}\right)_{\omega_0}$</th>
<th>$\delta t_{\omega_0}$</th>
<th>$l_{DMTL}$</th>
<th>$\delta \alpha_{21}/l_{DMTL}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>3</td>
<td>-11.25°</td>
<td>5.29 ps</td>
<td>0.49 mm</td>
<td>23.0%/mm</td>
</tr>
<tr>
<td>4</td>
<td>5</td>
<td>-22.5°</td>
<td>11.75 ps</td>
<td>0.71 mm</td>
<td>31.7%/mm</td>
</tr>
<tr>
<td>3</td>
<td>9</td>
<td>-45°</td>
<td>22.91 ps</td>
<td>0.131 mm</td>
<td>34.4%/mm</td>
</tr>
<tr>
<td>2</td>
<td>17</td>
<td>-90°</td>
<td>45.37 ps</td>
<td>0.251 mm</td>
<td>35.9%/mm</td>
</tr>
<tr>
<td>1</td>
<td>35</td>
<td>-180°</td>
<td>92.4 ps</td>
<td>0.529 mm</td>
<td>34.0%/mm</td>
</tr>
</tbody>
</table>

The limitations of the proposed design approach are illustrated in Table 5.4. The table compares the final simulation results of the complete network to the results obtained by the simplified design procedure. In contrast to the complete network simulation, the simplified approach cascades single RLL sections and doesn’t considers the near field coupling between adjacent sections. Hence, the difference between the insertion phase of the complete network $\alpha_{21,ntw}$ and the insertion phase of the idealized network $\alpha_{21,cas}$ describes the effect of coupling between the RLL sections due to forward and reverse scattered fields. For the phase shifter states 1 to 5 in the off state, the difference $\left(\alpha_{21,ntw} - \alpha_{21,cas}\right)_{\omega_0}$ is not caused by the loading factor.
Table 5.4. Effect of the near field coupling between RLL sections on the insertion phase at \( \omega_0 = 2\pi \cdot 60 \text{ GHz} \). \( \Delta S_{21, \text{ntw}} \) denotes the insertion phase of the complete network simulated by ADS Momentum, whereas \( \Delta S_{21, \text{cas}} \) denotes the insertion phase of a cascade of single RLL sections, neglecting near field coupling between adjacent sections.

<table>
<thead>
<tr>
<th>State</th>
<th>n</th>
<th>RLL in on state</th>
<th>RLL in off state</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>3</td>
<td>+1.25°</td>
<td>+0.15°</td>
</tr>
<tr>
<td>4</td>
<td>5</td>
<td>+0.60°</td>
<td>-1.37°</td>
</tr>
<tr>
<td>3</td>
<td>9</td>
<td>-3.00°</td>
<td>-3.98°</td>
</tr>
<tr>
<td>2</td>
<td>17</td>
<td>-15.6°</td>
<td>-8.66°</td>
</tr>
<tr>
<td>1</td>
<td>35</td>
<td>-37.6°</td>
<td>-16.6°</td>
</tr>
</tbody>
</table>

The comparison (table 5.4) shows, that the presented design procedure can provide good results if coupling is negligible. For the phase shifter states 3 to 5 which consist of up to 9 RLL sections with moderate loading the design procedure gives acceptable starting values for subsequent optimization.

For the phase shifter states 1 and 2, consisting of 35 and 17 cascaded RLL sections, additional simulation work of the complete network is required previous to optimization.

5.3 Phase shifter characterization

5.3.1 On-wafer probing

On-wafer probing (figure 5.8) allows the convenient testing of RF networks at wafer level before the wafers are singularized in individual dies. On-wafer probing is primarily performed using differential (GS) of coplanar (GSG) probes, which can be positioned with micrometre precision by manually driven positioning stages. The probes are typically guided with the help of a stereo microscope and adjusted to lithographically defined alignment features.

Figure 5.8. Standard probing setup for on-wafer RF measurements. Calibration substrate on metallic wafer support chuck and positioning heads with mounted RF probes.
Since the stripline and coplanar waveguides carry signal and ground traces on the top side of the substrate they can be directly connected by the GS or GSG probes. However, this is not the case if the DUT is implemented in microstrip configuration. In order to use the microstrip configuration in combination with on-wafer probing the ground conductor needs to be transferred to the backplane. Therefore a probe to microstrip transition is required. To access the microstrip network exclusively, the S-parameters of the probe-to-MS transition need to be determined by means of measurement and fixture de-embedding performed. In contrast to other test fixtures, on-wafer probing allows the reference plane to be placed directly at the input and output port of the circuit with micrometre precision.

5.3.2 Calibration, sweep settings and probe placement

The calibration of the measurement equipment was performed using short, open, load, and thru (SOLT) reference standards on a test substrate, provided by GGB Industries Inc.. GSG probes with a pitch of 150 µm were used to connect the standards. The calibration and measurement settings of the vector network analyser (VNWA) are summarized in table 5.5.

<table>
<thead>
<tr>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Start / stop frequency</td>
<td>13.4 GHz / 100.0 GHz</td>
</tr>
<tr>
<td>Number of points</td>
<td>1601</td>
</tr>
<tr>
<td>IF bandwidth</td>
<td>1 kHz</td>
</tr>
<tr>
<td>Sweep time</td>
<td>1.479 s</td>
</tr>
<tr>
<td>Power</td>
<td>-17 dBm (default)</td>
</tr>
<tr>
<td>Attenuator position</td>
<td>0.0(0) mm</td>
</tr>
<tr>
<td>Averaging during calibration / measurement</td>
<td>5 / 1</td>
</tr>
<tr>
<td>Probes</td>
<td>Picoprobe® GSG-150 microns, from GGB</td>
</tr>
<tr>
<td>Test substrate</td>
<td>Picoprobe® CS-5 calibration, from GGB</td>
</tr>
<tr>
<td>Calibration kit</td>
<td>Cal-Kit-42/ID-CS-5-150</td>
</tr>
<tr>
<td>Standards</td>
<td>Shorts, Open, Loads, Thru, Isolation (not calibrated)</td>
</tr>
</tbody>
</table>

The phase repeatability for on-wafer measurements on the same pattern was determined to be ±2.5° at 60 GHz. The deviation occurs due to differences in the contact force $F_c = k_{zz} \delta z + k_{zx} \delta z$, which is exerted by the over travel $\delta z$, ensuring reliable electrical contact. The stiffness tensor of the probes $k_{ij}$ determines the required displacement for the necessary contact force. Friction relates the contact force to the friction force $F_f = k_{xx} \delta x + k_{zx} \delta z = F_c / \mu$. Hence, the displacement of the port terminals is calculated by $\delta x = (\mu k_{zz} - k_{xz}) / (k_{xx} - \mu k_{zx}) \delta z$ and reveals its dependence on the over travel $\delta z$ and the friction coefficient $\mu$. Higher accuracy and better repeatability of the probe placement is achieved using mechanical alignment features (figure 5.9b). Figure 5.9a shows the lateral probe displacement $\delta x$ due to the over travel as well as the contact imprints.
As a control, the network symmetry of the calibration standards is controlled and the imprints are inspected after measurements with help of the stereo microscope. Occuring tilt errors are corrected by probe rotation around the x-axis.

5.3.3 **Fixture de-embedding and equivalent circuit description**

Mounting fixtures can be connectors, transitions or probes, which are used as an interface between the test ports of the measurement equipment and the device under test (DUT). These interfaces are designed to guide the electromagnetic waves from the typical coaxial test ports to the reference planes on the DUT. In order to account for the physical separation of the measurement terminals of the test equipment and the reference planes of the DUT, de-embedding techniques are applied. Network de-embedding calls for well-defined fixtures and measurement standards. The short-open-load-thru (SOLT) and the thru-reflect-lines (TRL) set of standards are widely used [92]. Throughout this thesis, SOLT calibration is used to calibrate for the probe terminals. The reference test substrate and the probe type are specified in table 5.5.

To perform on-wafer probing to microstrip networks, probe-to-microstrip (P-to-MS) transitions are required, transferring the coplanar mode at the probe terminal to a microstrip mode (figure 5.10). Possible architectures of various wideband P-to-MS transitions have been published in [88], [93]-[95]. Here, P-to-MS transitions have been implemented with TRL structures in order to extract the S-parameters of the transitions by subsequent de-embedding. Once the S-parameters of the P-to-MS transition are determined, the transitions can be de-embedded from any cascaded network measured, making use of the same transitions.
The proposed P-to-MS transition has the following architectural features and restrictions: (i) The ground stubs on the top side are electromagnetically coupled to the back-plane. (ii) The microstrip on the top side and the back-plane are considered to be two coupled lines. The connected microstrip is supposed to conduct odd-mode, whereas any even mode is rejected due to the absence of a parallel ground line on the top side. (iv) The back-plane is patterned in order to prevent cross-talk with adjacent microstrip lines. Moreover, the back-plane is isolated by a 15 mm air gap from any subjacent metallic parts, such as the wafer support chuck shown in figure 5.8.

Under the above restrictions, the equivalent circuit (figure 5.11) can be used to represent the electrical behaviour of the P-to-MS transition. The equivalent circuit emphasizes the dominant coupling occurring in the coupled line arrangement. Moreover, it allows the effects of geometrical deviations to be separated. The simplified circuit can be used for analytical calculation and simulation using standardised building blocks (5.12).
Figure 5.11. Equivalent circuits of a lossless P-to-MS transition: (a) directional coupling between the ground and the back-plane, (b) approximated equivalent expressed by the impedances $Z_1$ and $Z_2$.

\[
Z_1 = \frac{Z_c[A_{11}\cos(\theta_c) - \cos(\theta_u)] + j\left(1 - \frac{A_{11}A_{22}}{A_{21}}\right)\sin(\theta_c)}{Z_cA_{21}\cos(\theta_c) - jA_{22}\sin(\theta_c)}
\]

\[
Z_2 = \frac{Z_c[A_{22}\cos(\theta_c) - \cos(\theta_u)] - jA_{21}Z_c^2\sin(\theta_c)}{Z_cA_{23}\cos(\theta_c) - jA_{22}\sin(\theta_c)}
\]

\[
Z_u = \frac{jZ_c\sin(\theta_u)}{Z_cA_{23}\cos(\theta_c) - jA_{22}\sin(\theta_c)}
\]

In the figures 5.12 and 5.13, the simulated S-parameters are compared with the measured S-parameters. The measured insertion loss of one transition is about 1 dB higher than initially calculated by Momentum. This can be caused by the oxidized copper surface, which increases attenuation due to the skin effect. The phase $4S_{11}$ at the probe terminal and the return loss magnitude at the microstrip terminal $|S_{22}|$ differ from the simulation. The deviation may stem from layout-specific responses (5.3.4) of the TRL structures used.
Figure 5.12. De-embedded S-parameter magnitudes of the P-to-MS transition. Port 1 is the probe terminal. Port 2 intersects the microstrip in a distance $l_c = 622 \, \mu m$ from the abrupt stepped conductor width at $(T_2-T_2')$. The lines with the circular markers represent the simulation results.

Figure 5.13. De-embedded S-parameter phase angles of the P-to-MS transition. Port 1 is the probe terminal. Port 2 intersects the microstrip at a distance $l_c = 622 \, \mu m$ from the abrupt stepped conductor width at $(T_2-T_2')$. The simulation results are indicated by the circular markers.

The two series impedances $Z_1$, $Z_2$ and the characteristic impedance $Z_{\text{ui}}$ of the coupler depicted in the equivalent circuit (figure 5.11) are calculated according to (5.12) using the A-parameters extracted from Momentum simulations and RF measurements. The numerical values of the equivalent parameters are plotted in figure 5.14. The series capacitances $C_1$ and $C_2$ account for the fringing fields present at the open ends of the lines. At 55 GHz, $Z_2$ exhibits a series resonance due to the stepped width of the centre conductor on the top layer (figure 5.10). The discontinuity of a stepped conductor width is described by (4.34), with $\delta = 2$ accounting for a symmetrical constriction [53]. The fringing fields and the associated capacitance of the discontinuity are included in the series capacitance $C_2$. The characteristic impedance of the conductor backed CPW is somewhat below 50 $\Omega$. It could be shown that an increase in the conductor backed CPW characteristic impedance can extend the bandwidth of
the coupler by an additional 10 GHz. The real part of the impedances can be ascribed to radiation at the discontinuity of the stepped conductor [53].

![Graph](image)

**Figure 5.14.** Equivalent circuit parameters of the de-embedded P-to-MS transition: the coupler length is equal to \( l_u + \Delta l_u = 707 \mu m + 94 \mu m \), where \( \Delta l_u \) accounts for the scattering field length extension. The length of the adjoining microstrip is \( l_c = 622 \mu m \).

### 5.3.4 Network efficiency and cross-talk

The network efficiency expresses the fraction of the input power, which finally contributes to a desired function, such as radiation, transmission, coupling, heating, et cetera. Typically, some portion of the input power is lost to undesired power distribution. For a fixture, transmission is the desired function. Radiating patterns can be a source of cross-talk or cause additional transmission losses. Joule heating increases the operation temperature of the circuit, and undesired coupling results in cross-talk. In short, the advantages of an efficient circuit are a minimum of insertion loss, immunity against cross-talk and distortion.

Equation 5.13 states the energy balance of an N-port network with respect to the incident power \( P_{in,k} \) at port \( k \). The integral \( \oint ds \) denotes the closed surface, which envelops the device under test (DUT) with the exception of the cross-section area of the waveguides \( i \) flanged at port 1 to N carrying power \( P_{out,i} \) out of the system. Moreover, electric power \( P_{out,(\oint ds)} \) can escape the system boundaries in the form of near and far field radiation, reactive coupling and Joule heating.

\[
\sum_i^N P_{out,i} + P_{out,(\oint ds)} \left/ P_{in,k} \right. = \sum_i^N |S_{ik}|^2 + |S_{(\oint ds)k}|^2 = |S_{1k}|^2 + |S_{2k}|^2 \ldots + |S_{Nk}|^2 + |S_{(\oint ds)k}|^2 = 1 \quad (5.13)
\]

For a two-port network, \( N = 2 \), the forward and reverse efficiencies can be express by (5.14). The forward efficiency is determined by an incident power \( P_{in,1} \) at port 1, whereas the reverse efficiency is
determined by an incident power $P_{\text{in},2}$ at port 2. The efficiencies show the amount of power escaping the system by a path, other than through port 1 and port 2. Since the return loss is summed up in the energy balance, impedance matching does not directly contribute to the efficiency. In figure 5.15, the forward and reverse efficiencies are plotted versus frequency for different line lengths.

$$
\begin{align*}
    e_{\text{fwd}} &= 10\log_{10}\left( \frac{\sum P_{\text{out},i}}{P_{\text{in},1}} \right) = 10\log_{10}(|S_{11}|^2 + |S_{21}|^2) = 10\log_{10}\left( 1 - |S_{1\text{fwd}}|^2 \right) \\
    e_{\text{rev}} &= 10\log_{10}\left( \frac{\sum P_{\text{out},i}}{P_{\text{in},2}} \right) = 10\log_{10}(|S_{22}|^2 + |S_{12}|^2) = 10\log_{10}\left( 1 - |S_{2\text{rev}}|^2 \right)
\end{align*}
$$

(5.14)

**Figure 5.15.** Measured forward and reverse efficiencies of two P-to-MS transitions with a different length microstrip inserted. The listed electrical lengths of the inserted microstrip lines are given at 60 GHz. No mathematical de-embedding has been performed to the plotted efficiencies.

The line length of the inserted microstrip alters the frequency-dependence of the network. Dependent on the microstrip length, the constructive and destructive resonances of the transmitted and reflected power shifts along the frequency axis. Basically, radiation, skin losses and cross-talk with adjacent devices contribute to the efficiency decrease. Far field radiation dominates below 30 GHz, and is caused by the patterned coupled back-plane. Skin losses and radiation at discontinuities increase with increasing frequency and dominate above 30 GHz. Cross-talk occurs by reactive coupling or near field radiation. The frequency dependence of cross-talk is related to the geometry – e.g. the resonator length. The ripple in figure 5.15 shows the resonances, at which cross-talk peaks for a given length of microstrip inserted. If we assume cross-talk and radiation to be the dominant length-dependent dissipation mechanism above 30 GHz, the amount of dissipated energy can be extracted by the difference between the maximum and minimum envelope of the measurements plotted in figure 5.15. Hence, the maximum near field radiation occurring in the networks accounts for approximately 0.5 dB above 55 GHz. The better the matching of the transitions and probes, the lower is the cross-talk power.
compared to the transmitted power. This can be seen by comparing figure 5.12 and figure 5.15. The uncertainty due to layout-specific radiation and cross-talk exceeds the uncertainty or the calibration using SOLT reference standards. The standards showed an uncertainty of 0.05 dB.

5.3.5 Insertion loss and return loss

The individual states of the phase shifter have been measured using the sweep settings given in table 5.5. Since all phase shifter states are symmetrical in their layout, the proper function can be verified by the symmetry of the S-parameters. The phase shifter characteristics are given exclusively by applying de-embedding of the P-to-MS transitions. As described in section 5.3.3, the insertion loss uncertainty of 0.5 dB is introduced by the TRL de-embedding. In the following, this uncertainty can cause a $|S_{21}|$ higher than zero dB. Nevertheless, the differential comparison can be performed with the uncertainty of the SOLT calibration, which is measured to ±0.05 dB.

The devices depicted in the figures 5.16, 5.18, 5.20, 5.22 and 5.24 exhibit two separate DC bias pads and offer four states per unit. The four states can be differentiated into an unloaded microstrip, a periodic series-inductive loaded microstrip, a periodic capacitive-loaded microstrip, and an alternating inductive and capacitive loaded microstrip. Subsequently, the phase responses of the alternating loaded microstrip and the unloaded microstrip line are assigned as the on and off state.

A) Phase shifter using 3 RLL sections

Figure 5.16 shows the phase shifter state 5 in between two P-to-MS transitions. In the layout, the ground pads are circular, whereas the DC pads are square-shaped. This phase shifter state provides the minimum phase shift resolution of -11.25° at 60 GHz. The phase shifter consists of two inductive RLL sections and one capacitive RLL section, with each section using one MEMS switch.
Figure 5.16. Layout and microscope image of the state 5. The phase shift state 5 is designed to provide an insertion phase difference of -11.25° at 60 GHz. The black layer represents the thick copper metallization; the dark grey layer shows the high resistive bias line network. The pads for the DC needles and the RF probes are coated with a thin platinum and gold layer. The light grey layer forms the patterned metallization on the back plane.

Figure 5.17 shows the measured and simulated S-parameters of the phase shifter depicted in figure 5.16. The ripple of the forward scattering parameter $|S_{21}|$ can be explained by the uncertainty of ±0.5 dB. The insertion loss difference between the loaded line and unloaded line is negligible. Moreover, state 5 exhibits negligible insertion loss, and a return loss better than 24 dB up to 65 GHz. The measured S-parameters show good agreement with the S-parameter results obtained by Momentum simulation.

Figure 5.17. Measured and simulated S-parameters, showing the insertion loss and the return loss. The S-parameters are those of a symmetric and reciprocal network. For better visualisation only the input reflection and forward scattering are plotted. For the measurement the DC bias was set to 60 V.
B) Phase shifter using 5 RLL sections

State 4 uses five RLL sections (figure 5.18). The phase shifter is carried out in the form of a tapered Bragg. Three capacitive loaded microstrip sections are actuated simultaneously. The bias electrodes of the capacitive coupled switches are isolated by a network of high-resistive lines. The number of simultaneously driven switches is limited by the required electrical time constant of the RC-bias network.

![Figure 5.18. Phase shifter using 5 RLL sections. The targeted phase shift is -22.5° at 60 GHz.](image)

The S-parameters of state 4 are depicted in figure 5.19. The insertion loss of the loaded line exceeds the insertion loss of the unloaded line by less than 0.1 dB. The return loss is better than 24 dB up to 65 GHz in the loaded and unloaded state. Both, the difference in the insertion loss as well as the return loss match with the simulation results.

![Figure 5.19. Measured S-parameters of the phase shifter state 4.](image)
C) Phase shifter using 9 RLL sections

State 3 uses nine RLL sections (figure 5.20). The phase shifter is carried out in the form of a tapered Bragg filter. A maximum of three capacitive loaded microstrip sections are wired by one resistive DC and GND bias line. The number of five resistive lines can be reduced in a trade-off against switching time.

Figure 5.20. Layout of the phase shifter state 3, providing a phase shift of -45° at 60 GHz.

Figure 5.21 shows the S-parameters of the phase shifter state 3. The difference of the insertion loss between the loaded line and the unloaded line is less than 0.5 dB up to 63 GHz, which is slightly higher than the value predicted by Momentum simulation. The return loss exceeds 24 dB in the on state and 20 dB in the off state up to 63 GHz.

Figure 5.21. Measured S-parameters of the phase shifter state 3.
D) Phase shifter using 17 RLL sections

The phase shifter state 2 comprises 17 RLL sections (figure 5.22). Its design is based on the Chebyshev filter prototype. A maximum of three loaded microstrip sections are wired by the same resistive DC and GND bias line. The time constant of the bias network is $\tau \approx 0.25 \mu s$.

![Diagram of phase shifter state 2](image)

**Figure 5.22.** Layout and microscope image of the state 2. The phase shifter state 2 is designed to provide an insertion phase difference of -90° at 60 GHz.

Figure 5.23 shows the S-parameters of state 2. The insertion loss of the loaded line exceeds the insertion loss of the unloaded line by less than 1 dB up to 60 GHz. The return loss is better than 22 dB in the on state and better than 24 dB in the off state, up to 60 GHz. The simulation of the off state is in good agreement with the measurement. Minor deviation between simulation and measurements are observed for the phase shifter in the on state. The measured insertion loss in the on state is approximately 1 dB higher than expected by simulation.
E) Phase shifter using 35 RLL sections

State 1 uses 35 RLL sections (figure 5.24) to provide a phase shift of -180° at 60 GHz. The phase shifter is carried out in the form of a Chebyshev. In contrast to the phase shifter states 4, 3 and 2, this state starts with an series-inductive loaded line section.

Figure 5.24. Layout of the phase shifter state 1, providing a phase shift of -180° at 60 GHz.

Figure 5.25 shows the S-parameters of state 1. The insertion loss of the loaded line exceeds the insertion loss of the unloaded line by less than 1 dB up to 60 GHz. The return loss is better than 24 dB
up to 57 GHz. The measured insertion loss exceeds the calculated insertion loss by approximately 1 dB. This extra attenuation can be caused by skin loss due to the oxidized copper. Apart from that, simulation and measurements match well for frequencies up to 60 GHz.

5.3.6 Insertion phase difference

Figure 5.26 shows the simulated and measured insertion phase difference for the phase shift states 1 to 5. The measured and the simulation curves fit together. The phase response is that of a low-pass filter and shows good linearity up to 60 GHz. The dispersion limits the bandwidth of use and causes an increase in the time-delay difference with increasing frequency.

Figure 5.26. Summary of the insertion phase differences for the phase shifter states 1 to 5. The measurement data are taken without de-embedding of the P-to-MS transitions and are plotted in colour. The curves obtained by Momentum are plotted in grey.
Figure 5.27 shows the measured phase errors of the states 1 to 5 between 60 and 61.5 GHz. The maximum phase error within this frequency band is 3°. Within the ISM band (61.0-61.5 GHz) the average phase error is calculated to less than one degree.

Figure 5.27. Measured phase error of the states 1 to 5 and calculated average phase error for the 32 digital states.

A comparison between the network simulations using ADS Momentum and the measurements using on-wafer probing is given in table 5.6. The insertion phase difference measurements of the phase shift states 1 to 5 are in excellent agreement with the simulation data. The phase shift states 3 to 5 show low insertion loss and high return loss. Slightly higher insertion loss is obtained for the states 1 and 2.

Table 5.6. Summary of Momentum simulation results and RF measurements for the states 1 to 5 at $\omega_0 = 2\pi \cdot 60$ GHz. The insertion phase difference per length is proportional to the loading factor. The effect of the loading factor and number or RLL section $n$ on the insertion loss per length is shown.

| state | n  | Momentum simulation $(4S_{21, on} - 4S_{21, off})$ | measurements | $\delta S_{21}$ | $|S_{21, on}|$ | $|S_{21, off}|$ |
|-------|----|---------------------------------|--------------|----------------|---------------|---------------|
| 5     | 3  | -11.25°                         | -11°         | 22.4% mm       | -0.0 dB/mm    | -0.0 dB/mm    |
| 4     | 5  | -22.5°                          | -23°         | 32.4% mm       | -0.14 dB/mm   | -0.0 dB/mm    |
| 3     | 9  | -45°                            | -43°         | 32.8% mm       | -0.23 dB/mm   | -0.0 dB/mm    |
| 2     | 17 | -90°                            | -89°         | 35.5% mm       | -0.52 dB/mm   | -0.08 dB/mm   |
| 1     | 35 | -180°                           | -178°        | 33.6% mm       | -0.38 dB/mm   | -0.15 dB/mm   |

The listing in table 5.6 points out, that the insertion loss per single RLL section and the insertion loss per length increases with increasing loading factor, which is in agreement with equation 4.15.
An RF MEMS process has been developed to fabricate ohmic contact MEMS switches. The developed MEMS switch avoids any solid-state dielectric on top of the actuation electrode. Thus, it is immune to charging problems. Protection against direct contact between the bias electrode and the cantilever is given by the discontinuous stiffness constant. The MEMS switch uses copper to form the mechanical flexural part, which ensures a high mechanical Eigen frequency and reduces the effects of contact bouncing observed with cantilevers made of gold. Moreover, the copper MEMS process allows for cost-efficient fabrication.

A comprehensive model is given in this thesis, providing the reader with all the necessary considerations for designing a voltage-driven capacitively coupled MEMS switch. The static description relates the contact force and restoring force to the bias voltage and the cantilever properties. The switching dynamics is modelled by a fully parameterized model, which returns all relevant system values within a few seconds. The model is capable of handling arbitrary bias waveforms. Moreover, it can be used to predict influences of fabrication process tolerances or changing operation conditions such as temperature and pressure changes. The model is validated by means of the energy balance and electrical measurements. The calculations are in good agreement with the measurements.

The electrical characterization of the RF MEMS switch comprises the switching time, switching hysteresis, switch lifetime, contact hysteresis as well as the RF performance. The RF performance is given by the switch S-parameters in the open and closed state. In addition, transmission and reflection measurements are performed to characterize the power handling capabilities of the RF switch. The maximum power handling of the switch is given by 28 dBm at 11 GHz. The limiting effects are related to restoring force reduction and contact degradation. A single RF MEMS switch inserted into a waveguide can be categorized as a first order filter. Its isolation of -11 dB at 60 GHz can be improved over a wide bandwidth by combination of reconfigurable high-pass and low-pass elements. Due to the switch insertion loss of 0.3 dB at 60 GHz, the reconfigurable loaded-line (RLL) architecture is preferred above the switched line architectures. Integrated into the presented RLL section, a MEMS switch only routes a fraction of the signal power, which results in lower insertion loss and better potential power handling.
Low-pass reconfigurable loaded-line sections in series and shunt configuration have been realized. An analytical model of the RLL sections is presented, which describes the RLL sections in terms of their ABCD-parameters, S-parameters and T-parameters. The network parameters are set in relation to the dimensions of the RLL sections.

By cascading the RLL sections, a 5-bit state phase shifter has been implemented. A new design procedure is used, which links the element values of common low-pass prototype filter to the phase and magnitude responses of the phase shifter. This approach allows reducing the passband ripples and can facilitate the phase shifter synthesis. The limitations of the design procedure are discussed for the phase shift states 1 to 5 and set in relation to the loading factor and number of sections. The proposed simplified design procedure excludes near field coupling between adjacent RLL sections and provides accurate results for the phase shift bits 1 to 3. With an increasing number of sections, which is the case for the phase shift bits 1 and 2, near field coupling significantly affects the frequency-response and must considered.

The realized phase shifters show excellent performance comparable with the state-of-the-art of equivalent designs (table 6.1). At 60 GHz, the insertion phase difference of the individual bits is measured to be 11.0°, 22.5°, 43°, 89° and 178°. The measured phase responses are in good agreement with the Momentum simulation. The estimated total insertion loss of the phase shifter varies between 1 dB in the reference state and a maximum of approximately 3.7 dB for the 348.8° state.

| Description               | $f_0$ | $\text{max. } |I|/|f_0|$ | $\text{max. } |S_{21}|/|f_0|$ | $|\delta S_{21}|$ | $l_{DMT}$ | Ref. |
|---------------------------|------|----------------|----------------|----------------|-------|------|
| 3-bit switched-line MS    | 76.5 GHz | 5.7 dB | -315° | 13.4° | 3.5 mm | [43] |
| 4.25-bit loaded-line CPW  | 75 GHz | 3.5 dB | -270° | 2.6° | 5.5 mm | [96] |
| 3-bit loaded-line CPW     | 78 GHz | 3.2 dB | -315° | 3° | 5.0 mm | [40] |
| 2-bit loaded-line CPW     | 60 GHz | 3.0 dB | -285° | 8° | 6.3 mm | [39] |
| 5-bit loaded-line MS      | 60 GHz | *3.7 dB | *-349° | *3° | *10.3 mm | this work |

* The expected insertion loss is estimated for the measurements of the phase shifters for the states 1 to 5 given in table 5.6

The measured phase error of the phase shift states 1 to 5 is below 3°. The estimated maximum phase error sums up to 4.7° for the 348.8° state. Considering the phase shifter operating in the ISM band (61-61.5 GHz), an average phase error of approximately 1° results for the 32 states.
CHAPTER 7

Technology outlook

The presented MEMS fabrication process is optimized for RF applications. It enables reconfigurable filters, phase shifters, couplers, antennas, inductors et cetera to be realized. The strategy of using low-loss dielectric substrates and low-loss metallization traces makes this MEMS process especially attractive for passive reconfigurable components and networks. Moreover, surface micro-mechanical structures with air gaps or cavities of 400 nm to 3000 nm fabricated on glass substrates using transparent electrodes allow for optical switch state detection.

The continuation of technical process development includes resistive layers with low tolerances and low thermal coefficient of resistance. This allows the realization of absorptive switches, power divider and port terminations. In addition, a low-loss dielectric layer enables metal-insulator-metal capacitors to be integrated.

Material creep and recrystallization is expected to be the lifetime limiting mechanism taking place in the presented micromechanical switch. Thus, the electroplating process demands modification in order to improve the mechanical stability. Fabrication yield will gain in importance, once the MEMS process is defined and quality control is implemented. Yield improvement can be achieved by automatizing challenging process steps, such as the release step.

Demonstrator realization and long term characterization are planned for the technology to win recognition through comprehensive presentation of demonstrator performance and illustration of the technology benefits.

A first packaging stage using organic sealing techniques is planned in order to ensure realistic operation conditions for long-term characterization. From the technological and economic viewpoint it is expected that RF MEMS will require a second packaging stage, which can be conventional housing using available hermetic sealing techniques on the mm-scale.
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