Abschlussarbeit im Bachelorstudiengang Physik

Scalable Sample Holder for Experiments with Superconducting Quantum Circuits

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Garching, 8th August 2011

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Chapter 1

Introduction

In 2004, the coupling of a superconducting qubit \[1\] to the quantized mode of a microwave transmission line resonator founded the field of circuit quantum electrodynamics (QED) \[2\]. Meanwhile, many possible applications of such systems have been found in the context of quantum simulation and quantum computing. However, it has turned out that scaling of the involved superconducting microwave technology to large systems is not a trivial task. In particular, the microwave resonators require a relatively large size of the wafers. For a typical resonance frequency of 5 GHz, the resonators have a length of the order of centimetres.

On the one hand, the size of the resonators could be reduced by increasing the relative permittivity $\epsilon_r$ of the sample, but the available materials with low dielectric loss have a maximum $\epsilon_r$ of approximately 10. Furthermore, it is possible to meander the resonator. However, the microwave properties of the waveguide get disturbed for too dense meanders. On the other hand, the size of the substrate could be increased. Unfortunately, large chips with complex layouts tend to exhibit problems with parasitic resonances in the relevant frequency range. One solution to these difficulties is to embed small wafers into a larger printed circuit board (PCB) which is penetrated by metallic interconnects. Such vias suppress the excitation of parasitic standing waves in the PCB. In this manner, it is possible to attach a large number of rather bulky microwave connectors to a large PCB containing one or even more small sample chips. These features represent a substantial benefit for the development of scalable circuit QED systems.

The thesis is structured as follows. First, we give an introduction to microwave theory in Chapter $2$. Three kinds of waveguides – the parallel plate waveguide, the microstrip line and the coplanar waveguide – and microwave transmission line resonators are described. An explanation of the key difficulties for scaling up the number of structures on a chip commences Chapter $3$. Here, the geometry and functionality of sample chips with superconducting circuits (in the following also referred to as "superconducting microchips"), the PCB and the corresponding sample holder are explained. Furthermore, the experimental setup is introduced. In Chapter $4$ the results of impedance and transmission measurements of different PCB structures are described and discussed. Finally, a conclusion and an outlook on future tasks can be found in Chapter $5$. 
Chapter 2

Microwave Fundamentals

A microwave is an electromagnetic wave with frequencies in the range of $300 \text{MHz} \lesssim f \lesssim 300 \text{GHz}$ corresponding to wavelengths $1 \text{m} \lesssim \lambda = \frac{c}{f} \lesssim 1 \text{mm}$. The wavelength of microwaves is therefore comparable to the dimension of the device itself leading to many problems in the design of microwave components.

In Section 2.1 wave propagation on a transmission line, the characteristic impedance and the reflection coefficient is introduced. In Section 2.2 the scattering parameters which allow the characterization of an network are defined. Section 2.3 is about waveguides and their modes which is followed by Section 2.4 where microwave resonators are described.

This chapter closely follows the explanations and figures of [3].

2.1 Transmission Line Theory

2.1.1 Wave Propagation on a Transmission Line

An essential prerequisite for describing circuits by classical circuit theory is that the wavelength of the input signal is substantially larger than the circuit itself. In microwave theory, however, the wavelength varies within the length scale of the device.

Transmission lines are often schematically pictured as two parallel wires as shown in Figure 2.1 (a). In Figure 2.1 (b) the line is formed in an equivalent lumped-element circuit. With Kirchhoff’s voltage and current law and by further taking the limit $\Delta z \to 0$ one obtains wave equations for the voltage $v(z,t) = V(z) \cdot e^{i\omega t}$ and the current $i(z,t) = \ldots$
$I(z) \cdot e^{i\omega t}$ with $\omega$ being the radian frequency. These are solved by:

$$V(z) = V_0^+ e^{-\gamma z} + V_0^- e^{+\gamma z}$$

$$I(z) = I_0^+ e^{-\gamma z} + I_0^- e^{+\gamma z},$$

where $e^{-\gamma z}$ represents wave propagation in the $+z$ direction and $e^{+\gamma z}$ wave propagation in the $-z$ direction. The complex propagation constant $\gamma$ is given by:

$$\gamma = \alpha + i\beta = \sqrt{(R + i\omega L)(G + i\omega C)}.$$ (2.2)

The real part $\alpha$ is called the attenuation constant, whereas the imaginary part $\beta$ is called the propagation constant.

### 2.1.2 Characteristic Impedance

Starting from Equation (2.1) the characteristic impedance $Z_0$ of a transmission line is defined as

$$Z_0 := \frac{V_0^+}{I_0^+} = -\frac{V_0^-}{I_0^-}.$$ (2.3)

As the name suggests it is a characteristic quantity of transmission lines and describes a complex resistance. The real part of $Z_0$ is similar to the electrical resistance $R$ of a conductor, whereas the imaginary part can be considered to be a reactance. For transmission lines $Z_0$ is given by

$$Z_0 = \frac{R + i\omega L}{\gamma} = \sqrt{\frac{R + i\omega L}{G + i\omega C}}.$$ (2.4)

### 2.1.3 Reflection and Transmission Coefficient

At the interface between two media with different refractive indices, light is partly reflected and partly transmitted. Analogously microwaves show the same behaviour in a transmission line at points of varying characteristic impedance $Z_0$.

![Figure 2.2: Reflection and transmission at the junction (z = 0) of two transmission lines with different characteristic impedances.](image)

Figure 2.2 shows two materials with characteristic impedances $Z_0$ and $Z_1$ interfacing at $z = 0$. An incident wave coming from the left with amplitude $V_0^+$ is split into a reflected wave with the amplitude $V_0^-$ and a transmitted wave with amplitude $V_1^+$. 


The reflection coefficient $\Gamma$ is defined as the ratio of reflected wave and incident wave:

$$
\Gamma := \frac{V_0^-}{V_0^+},
$$

(2.5)

It can be shown that [3]

$$
\Gamma = \frac{Z_1 - Z_0}{Z_1 + Z_0},
$$

(2.6)

The transmission coefficient $T = \frac{V_1^+}{V_0^+}$ can be obtained to

$$
T = 1 + R = \frac{2Z_1}{Z_1 + Z_2}.
$$

(2.7)

The theory of reflection and transmission at discontinuities is essential for the direct measurement of the impedance of transmission lines using Equation (2.6).

### 2.2 Scattering Parameter

At microwave frequencies it is difficult to measure the current and the voltage directly because they are strongly time and space dependent. But the measurement of the reflection or transmission coefficient is possible. To describe a network one uses the scattering matrix (S-matrix) $S$. In general, an $n$ port network is described by an $n \times n$ scattering matrix. In our case $S$ has dimension $2 \times 2$

$$
\begin{array}{c}
V_1^+ \\
V_1^-
\end{array}
\begin{array}{c}
1 \\
\text{Network} \\
2
\end{array}
\begin{array}{c}
V_2^- \\
V_2^+\end{array}
$$

Figure 2.3: A schematic two port network

Figure 2.3 depicts a two port network, where an incident wave on port 1 with amplitude $V_1^+$ causes a reflected signal $V_1^-$. For the case that no signal arrives from port 2 ($V_2^+=0$) a scattering parameter can be defined as

$$
S_{11} = \left. \frac{V_1^-}{V_1^+} \right|_{V_2^+=0}
$$

(2.8)

With the definition of Equation (2.6) one can see that $S_{11}$ is exactly the reflection coefficient $\Gamma^{(1)}$ on port 1. The transmission is described by $S_{21} = \left. \frac{V_2^-}{V_1^+} \right|_{V_2^+=0}$. The general definition of the scattering parameters is:

$$
S_{ij} = \left. \frac{V_i^-}{V_j^+} \right|_{V_k^+=0} \text{ for } k \neq j
$$

(2.9)
Thus, for a two port network, we have 4 scattering parameters describing the relations of incident and reflected voltages:

\[
\begin{pmatrix}
V_1^- \\
v_2^-
\end{pmatrix}
= 
\begin{pmatrix}
S_{11} & S_{12} \\
S_{21} & S_{22}
\end{pmatrix}
\begin{pmatrix}
V_1^+ \\
v_2^+
\end{pmatrix}
\]  
\tag{2.10}

A network is said to be symmetric or reciprocal if \( S_{ij} = S_{ji} \forall i,j \) holds. In most applications \( S_{ij} \) is given in dB (decibel) [4]:

\[
S_{ij}(\text{dB}) = 20 \cdot \log_{10}(|S_{ij}|)
\]  
\tag{2.11}

In this context the power level \( L \) with unit dBm, the power ratio in decibels of the measured power \( P \) referenced to 1 mW, is introduced:

\[
L(\text{dBm}) = 10 \cdot \log_{10}\left(\frac{P}{1\text{mW}}\right).
\]  
\tag{2.12}

### 2.3 Transmission Lines

Although the idea of a electromagnetic wave transmission through waveguides came up in the late 19th century by Oliver Heaviside, the first results of experiments were not published until 1936 [3]. Nowadays, many kinds of transmission lines exist. In the following section, the parallel plate waveguide, the microstrip line and the coplanar waveguide are introduced.

#### 2.3.1 Parallel Plate Waveguide

The parallel plate waveguide is probably one of the simplest kind of waveguides. In Figure 2.4 the geometry (a) and the field configuration (b) of the TEM mode [3] of a parallel plate waveguide is shown. This mode is called the parallel plate waveguide mode. It consists of two parallel arranged conducting plates filled with a dielectric substrate. The geometry of a waveguide very important for its field configuration. In the context of microwave electrodynamics the notion of a waveguide mode usually refers to its characteristic field configuration. Contrary to other types of waveguides treated in this work the parallel plate waveguide does not guide the wave in a specified direction.

![Figure 2.4: (a) Geometry of a parallel plate waveguide. (b) Electric and magnetic field lines (TEM mode).](image-url)


2.3 Transmission Lines

2.3.2 Microstrip Line

A microstrip line is a planar transmission line. Its geometry is shown in Figure 2.5 (a). A conductor is situated on grounded layer of dielectric substrate with a relative permittivity $\varepsilon_r$. Figure 2.5 (b) depicts a field configuration of the electric $E$ as well as magnetic field configuration $H$. This wave mode is called the microstrip mode.

When dealing with the propagation of such a wave mode, it is convenient to define the phase velocity via an effective dielectric constant $\varepsilon_{eff}$:

$$v = \frac{c}{\sqrt{\varepsilon_{eff}}} \quad (2.13)$$

Therefore $\varepsilon_{eff}$ can be interpreted as the dielectric constant of an homogeneous medium that replaces the air and dielectric regions of the microstrip [3]. As we are only interested in interference phenomena we only treat the phase velocity here.

2.3.3 Coplanar Waveguide

Just like the microstrip line, the coplanar waveguide (CPW) is also a planar waveguide. The geometry is shown in Figure 2.6 (a). The CPW can be thought of as being a two dimensional sheet of a coaxial cable. It has an inner conductor and a grounded outer conducting plane. In contrast to the microstrip line the ground plane below the dielectric substrate is not necessary to guide a wave. Therefore only the field configuration of a CPW without bottom ground plane is shown in Figure 2.6 (b). This mode is called the even CPW mode, where the electric field lines consistently run from the inner to the outer conductor.

Parasitic Modes in the CPW

Besides the CPW mode there are several other so-called parasitic modes propagating in the waveguide. One of these modes is the odd CPW mode. A CPW can be described as a combination of two parallel slotlines. Each slotline has its own wave mode which can be seen as independent from the other. If the CPW runs around a bend, the wave in the inner slotline has to cover less distance than the wave propagating in the outer slotline. For this reason the phases of the waves are shifted against each others. If this difference in distance half a wavelength, the phase shift amounts $180^\circ$ as shown in Figure...
Chapter 2 Microwave Fundamentals

Figure 2.6: (a) Geometry of a coplanar waveguide. (b) Electric field and magnetic lines without bottom ground plane (Even mode).

As the wavelength is around 2 cm only a small fraction of the wave will be in the odd CPW mode. In comparison to the even mode in Figure 2.6 (b) the field lines run both from the outer conductor to the microstrip and vice versa. A condition for such a field configuration is that the electric potential is different on both sides of the CPW. As a consequence the odd mode has a different phase velocity than the even mode of the CPW. Furthermore, the odd mode has a higher energy because the far-field is not cancelled out as is the case for the even mode.

Figure 2.7: CPW electric field configuration of (a) odd CPW mode and (b) microstrip mode.

Another parasitic mode is the microstrip mode of the CPW in the presence of a ground plane at the bottom as shown in Figure 2.7 (b). As this mode is similar to the parallel plate waveguide mode the probability of leaving the CPW is increased. In this case the wave propagates as a parallel plate waveguide mode between the bottom ground plate and the outer conductor of the CPW.

2.4 Microwave Transmission Line Resonators

Generally, a resonator is a system that oscillates at certain frequencies – the resonance frequencies. Microwave transmission line resonators consist of two capacities which reflect microwaves. For a CPW the capacitance is realized by fabricating a gap into the inner conductor, as Figure 2.8 (a) shows.

A capacitance does not reflect a microwave completely. The transmission – described
by the scattering parameter $S_{21}$ - depends on the frequency $f$ and increases towards higher frequencies.

The resonance frequency $f_n$ of a transmission line resonator correlates with its length $l$. For a $\lambda/2$ resonator $f_n$ is given by the condition $l = n\lambda/2$ and thus

$$f_n = \frac{v}{\lambda} = n \cdot \frac{c}{2l \cdot \sqrt{\varepsilon_{eff}}}, \quad n = 1, 2, \ldots .$$  \hfill (2.14)

A schematic transmission spectrum for a low quality resonator is shown in Figure 2.8 (b). The frequency dependence of scattering parameter $S_{21}$ is well recognizable.

![Figure 2.8](image)

Figure 2.8: (a) CPW $\lambda/2$-resonator of the length $l$. (b) Schematic $S_{21}$ curve for a low quality transmission line resonator with resonance frequencies $f_n = n \cdot \frac{c}{2l \cdot \sqrt{\varepsilon_{eff}}}$.

A characteristic quantity of a resonator is the quality factor (Q-factor) $Q$. It is defined as

$$Q = \frac{2\pi f}{(\text{average energy stored})} \frac{(\text{energy loss/second})}{(2.15)}$$

For a transmission line resonator

$$Q = \frac{\beta}{2\alpha},$$  \hfill (2.16)

where $\alpha$ is the attenuation constant and $\beta$ is the propagation constant as introduced in Equation (2.2).
Chapter 3

Experimental Methods and Setup

This chapter focuses on the experimental part of this work. In Section 3.1 superconducting microchips are described which leads to Section 3.2 where the sample holders are introduced. Section 3.3 discusses printed circuit boards. The measurement setup is described in Section 3.4.

3.1 Superconducting Microchips

A superconducting microchip with a qubit-resonator structure printed on its conducting layer is shown in Figure 3.1 (a). The silicon wafer of the chip has a size of $6\, \text{mm} \times 10\, \text{mm} \times 250\, \mu\text{m}$. On the surface the wafer has a layer of 50 nm thermal oxide (cf. Figure 3.1 (b)). The structure is imprinted on a 100 nm niobium layer.

Since the critical temperature of niobium is 9.2 K the devices can be characterised at helium-temperature in their superconducting state. The high purity silicon reduces energy dissipation in the substrate.

The scalability of the structures on these microchips is limited due to several reasons: the size of the layouts is bound to the properties of the substrate. A smaller design would be possible with an higher $\epsilon_r$. But the available materials with low dielectric loss have a maximum $\epsilon_r$ about 10 (e.g. in ideal silicon: $\epsilon_r = 11.9$).

The frequencies for qubits of about 5 GHz cannot be modelled. This corresponds to a wavelength of about 2 cm. The length of the resonator has to comply with this condition.

The radius of meandered structures has to be about 10 times the width of the transmission line [6]. As a result the structures on the chip cannot be arbitrary small. A meandered transmission line shape can be seen in Figure 3.1 (a) at top left. Therefore, as all the size reduction patterns are already implemented, the only option to scale further is to increase the size of the chip.

3.2 Sample Holder

Superconducting microchips are usually characterised in a metallic box. It shields the inner part against electromagnetic radiation and static electric fields.

3.2.1 The Small Sample Holder

The usual measurement procedure in the WMI is to place a single chip in a small sample holder. In Figure 3.2 (a) such a small box is shown. The box itself is made of oxygen free high conductivity (OFHC) copper with a thin gold coating. In order to connect the
Figure 3.1:
(a) Optical image of the superconducting $\lambda/2$ coplanar waveguide resonator (light blue box). Black rectangles: area shown in b. Red rectangle: area shown in d. b, SEM image of one of the coupling capacitors. d, SEM image of the galvanically coupled flux qubit. [7]
(b) Constituents of a superconducting microchip produced at the WMI.

chip we employ four SMA connectors\(^1\) used in Figure 3.2 (b) two of the four connectors are terminated with a 50-Ohm resistor. The four tapped holes on the top side of the box are used to mount a lid.

Figure 3.2: (a) Small Sample Holder, (b) SMA connector.

In the small boxes the chip is contacted to the SMA connectors using silver glue. A little drop of silver glue (Leitsilber 200) is trickled on the strip. The drop is drawn between the connector strip and the conducting structure, as one can see in Figure 3.3 (a). This type of connection is not reliable because at very low temperatures silver glue becomes fragile and therefore it is not possible to reproduce measurements exactly [4].

\(^1\)32K724-600S5, panel jack, Rosenberger Hochfrequenztechnik GmbH & Co. KG
3.2.2 The Big Sample Holder

In the course of scalability a new, larger sample holder was designed by Max Häberlein, who is a member of the WMI. The box was manufactured in the workshop of the WMI. The essential component is the body, depicted in Figure 3.3 (b) and (c). Similar to the small box it is made of OFHC copper plated with a thin layer of gold ($\sim \mu m$) which preserves the surface from corrosion [8].

The big sample holder has eight ports for SMA connectors. If the measurement does not require all ports, the abundant ports will be closed by gold plated plugs. These plugs are made of OFHC copper and close the SMA through holes tight. Furthermore, it has 16 lugs for bonding wires (Figure 4.2 (c)), four on each side.

![Figure 3.3: (a) Contact with silver glue [4], (b) the body of the big sample holder, (c) Autodesk Inventor 2011 screen shots: left side: Body of the big sample holder including SMA connectors. Red rectangle: shown on the right side: Below the pin a platelet is located in an indentation. Besides the ports there are lugs for bonding wires.](image)

The big sample holder is designed to hold a printed circuit board (PCB). A discussion on PCBs can be found in Section 3.3. In contrast to the silver glue connection in Section 3.2.1, the SMA connectors are connected by employing a pressure connection.

Noticeable at the top surface of the inner part of the box are little indentations, as shown in Figure 3.4. They are located underneath the ports for the connectors. Small platelets are put inside. The platelets are made of copper and are roughly 1.5 mm thick.
With the PCB lying on top of the platelets the SMA connector is inserted and fixed. Now it is possible to raise the platelets via turning screws located at the bottom side and thereby pressing the PCB against the connector pin. This results in a stable connection, even at low temperature measurements.

### 3.3 Printed Circuit Board

A *printed circuit board* (PCB) yields mechanical support and electric connections of circuit components. Within the scope of this thesis high frequency PCBs are used. The geometry is shown in Figure 3.5. Two conducting layers made of copper and a thin tin coating build up the surface of a non-conductive dielectric material. This high frequency dielectric substrate is *Rogers 3010* and is produced by Rogers Corporation[^3]. Examples of PCBs are depicted in Figure 3.9.


### 3.3.1 Via

A *via ([vertical interconnect access](https://en.wikipedia.org/wiki/Via_(electronics)))* is an vertical electrical connection between the different conducting layers of a PCB. A scheme is illustrated in Figure 3.6. A hole is drilled through the PCB and is filled with a conductor. Vias are one of the main aspects of this thesis. Their influence on microwave transmission behaviour is studied and discussed in Chapter 4.

[^1]: fabricated by Hofmann Leiterplatten GmbH
3.3 Printed Circuit Board

3.3.2 PCB Structures

Altogether four PCB CPW-structures – the transmission line, the resonators, the coupled resonators and the chip holder – are analysed in this thesis. They are explained in the following.

CPW Transmission Line

A CPW transmission line can be seen in Figure 3.9 (a). The transmission line has a length $l_{TL} = 54.75$ mm. Other dimensions are shown in Figure 3.7.

\[
\text{Conductor} \quad \text{Dielectric}
\]

With these dimension one can compute $\epsilon_{\text{eff}} = 4.7$ for the CPW using application TXLINE 2003.

CPW Resonator

A CPW resonator can be seen in Figure 3.9 (b). The difference compared to a transmission line is the presence of two gaps at the inner conductor. The width of the gap is $g = 100 \mu m$. With the resonator length $l_{\text{Res}} = 24.8$ mm the theoretical resonance frequencies can be calculated according to Equation (2.14) as:

\[
\begin{align*}
    f_n &= n \cdot 2.79 \text{GHz for } n = 1, 2, \ldots .
\end{align*}
\]  

(3.1)

Coupled CPW Resonators

A PCB with the structure of four CPW resonators is shown in Figure 3.8 (a). They are arranged in a way that there are three different coupling distances. The lengths of the
resonators are the same as the length of the single resonators which are described above. Thus they have the same theoretical resonance frequencies.

Figure 3.8: (a) Coupled CPW Resonators. (b) There are two areas where the via density varies. First, at the surface (green area) and second, the area close to the CPW (red area).

**Chip Holder**

In order to implement superconducting microchips a PCB chip holder are designed. One structure is shown in Figure 3.9(d). The wafers can be embedded into a milling groove which is located in the middle of the PCB. The CPWs function as feeding lines for the chip. The mechanical and electrical connection to the wafers are fabricated by bonding wires (compare also Section 4.2.4).

**3.3.3 Nomenclature of PCB Samples**

Due to a considerable variety of PCB samples we define a common nomenclature.

**Different Structures**

There are four different structures printed on the PCBs which are shown in Figure 3.9 (a)-(d).

(a) CPW transmission lines, abbr. **TL**

(b) CPW resonators, abbr. **RES**

(c) CPW coupled resonators, abbr. **C-RES**

(d) CPW chip holder, abbr. **Chip-H**
Figure 3.9: (a) - (d) different PCB structures and (a) - (f) different via density configuration. (a) CPW transmission line \{0 - 0\}, (b) CPW resonator \{LD - 0\}, (c) coupled CPW resonators \{HD - 0\} and (d) chip holder \{LD - LD\}. Picture (e) \{HD - LD\} and (f) \{HD - HD\} to illustrate the via density. The expressions in curly brackets \{x - x\} label the via density configuration according to nomenclature.

**Different Via Densities**

Besides the structure the PCBs can be classified by the via density. First there are two different areas on the PCB, as marked in Figure 3.8 (b):

(a) **global** of the outer conductor of the CPW

(b) **close to** the CPW

Second there are three different via densities:

(a) zero, abbr. **0**

(b) low density, abbr. **LD**, the distance between two vias is 2mm

(c) high density, abbr. **HD**, the distance between two vias is 1mm

Overall this leads to six configurations as shown in Table 3.1
Chapter 3 Experimental Methods and Setup

Table 3.1: Different via density configurations.

<table>
<thead>
<tr>
<th>Figure 3.9</th>
<th>close to CPW at surface</th>
<th>label</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a)</td>
<td>0 0</td>
<td>0 - 0</td>
</tr>
<tr>
<td>(b)</td>
<td>LD 0</td>
<td>LD - 0</td>
</tr>
<tr>
<td>(c)</td>
<td>HD 0</td>
<td>HD - 0</td>
</tr>
<tr>
<td>(d)</td>
<td>LD LD</td>
<td>LD - LD</td>
</tr>
<tr>
<td>(e)</td>
<td>HD LD</td>
<td>HD - LD</td>
</tr>
<tr>
<td>(d)</td>
<td>HD HD</td>
<td>HD - HD</td>
</tr>
</tbody>
</table>

For example the RES \{LD - 0\} is a CPW resonator with a low density of vias along the CPW and no vias at the surface of the outer conductor (cf. Figure 3.9 (b)).

3.3.4 Advantages of Printed Circuit Boards

Due to the geometry and vias the parasitic modes of a CPW are suppressed as described in the following.

The large substrate thickness of about 1.3 mm suppresses the excitation of the microstrip mode. The reason is that the total field energy of the microstrip modes proportional to the distance between the top and the bottom conducting plane. As the ideal CPW mode is independent from that distance it becomes energetically favoured by increasing the thickness of the substrate.

In order to minimise the odd CPW mode both sides of the CPW have to be connected with a short electrical length. The rise time of a microwave signal is of the order \( \tau = 1/5 \cdot 10^9 s = 200 \text{ps} \). A time dependent potential is approximately matched between two points in space if a signal connects these points in a time shorter than the rise time of the microwave. Assuming a phase velocity of the microwave of \( v = 10^8 \text{ m/s} \) and a contact without inductance this means that the interconnect between the ground planes has to be shorter than \( l_{\text{crit}} = v \tau = 2 \text{ cm} \). The distance of the interconnect via the box can be greater than \( l \). Using vias placed parallel close to the CPW this distance is approximately \( l_{\text{via}} \approx 3 \text{ mm} < l_{\text{crit}} \). Therefore the odd mode is suppressed.

Furthermore, the vias suppress the parallel plate waveguide mode. A qualitative explanation reads that in a PCB - penetrated with vias - the excitation of standing wave
has to meet the same boundary condition as for a periodic lattice:

\[ k = \frac{2\pi}{\lambda} = n \frac{\pi}{d} \text{ for } n = 1, 2, \ldots , \]  

where \( d = n \cdot \lambda / 2 \) is the distance between two vias as shown in Figure 3.10 and \( k \) the wave vector. The recurring vias permit only propagation of microwaves with a wavelength of \( \lambda = 2d/n \) and therefore the overall probability for the excitation of parallel plate waveguide modes is damped.

![Diagram](image)

Figure 3.10: Only standing wave with a wavelength of \( \lambda = 2d/n \) (black curve) can arise. The dashed red curve illustrates a disallowed wave.

### 3.4 Measurement Setup

In this thesis two measurement devices were used – a *time domain reflectometer* and a *vector network analyser* – which are introduced in the following.

![Diagram](image)

Figure 3.11: Scheme of the measurement setup of (a) TDR. A feeding cable connects the sample holder with the TDR. For later discussions in Chapter 4 we define that the feeding cable is part of the measurement device. (b) VNA. Two feeding microwave coaxial cables connect the sample holder to the VNA. The recorded data of the VNA is sent to a computer and processed by LabView.
3.4.1 Time Domain Reflectometer

A time domain reflectometer (TDR) is an instrument used to characterize and locate discontinuities in transmission lines. After sending a defined signal into the sample the TDR measures the reflections which occur at impedance discontinuities and computes the characteristic impedance using Equation (2.6).

A scheme of the TDR measurement setup is shown in Figure 3.11 (a).

The TDR which is used for impedance measurements in this thesis is fabricated by the company Tektronix. The designation is DSA8200 Digital Serial Analyser with an add-on module 80E80 TDR/Sampling Module.

3.4.2 Vector Network Analyser

A vector network analyser (VNA) is used for measurements of scattering parameters (cf. Section 2.2) of a microwave network. Within this work one is only interested in the transmission properties of the sample holder and the PCBs. Therefore only the scattering parameter $S_{21}$ is measured.

A scheme of the TDR measurement setup is shown in Figure 3.11 (b).

The VNA which was used for measurements in this thesis has the designation HP8722D 50MHz - 40 GHz Network Analyser and is fabricated by the company Hewlett Packard.
Chapter 4

Results and Discussion

This chapter treats the network analysis of the big sample holder and its results. At the beginning the pressure connection of the box (cf. Section [3.2.2]) is examined by time domain reflectometry. The second part of the chapter deals with the vector network analysis of different PCB structures.

4.1 Time Domain Reflectometry

From the discussion presented in Section [2.1.3] and Section [3.4.1] it follows that a good microwave transmission through a cable or a whole network requires all waveguide components to be impedance matched. For real devices however, usually impedance mismatches occur at connectors.

This is also the case for the pressure connection, which is introduced in Section [3.2.2]. This connection is a junction between two different kinds of waveguides - coaxial line and CPW.

In order to improve the contact of the pin of the SMA connector to the PCB, in a first experiment two insulators - Teflon ($\epsilon_r \approx 2$) and Rogers 3010 ($\epsilon_r \approx 10$) - are pressed against the top of the pin, as shown in Figure 4.1. This should test the electrical connection between the pin and the PCB.

The TDR results are depicted in Figure 4.1 (a). The cyan graph represents the open end of the supply cable to give a reference where the SMA connector begins. The black curve shows the connector without any modification. The total impedance mismatch here is about $14,0 \Omega$. The red and blue curves refer to the teflon or Rogers 3010 insulators respectively. It is noticeable that the data shows two peaks. Only the second peak is influenced by the dielectrics. Teflon has little influence on the impedance decreasing the impedance mismatch by roughly $2 \Omega$, whereas Rogers 3010 can reduce the maximum connector impedance to $57.5 \Omega$ at 44.2ps. However, as the CPW is still influenced by the dielectric after the peak the impedance exhibits to a minimum of $44 \Omega$ in the case of Rogers 3010.

Due to an error during the fabrication of the box there was a gap between the teflon part of the SMA connector and the metallic box. In a second experiment, we investigated the influence of this gap by filling it with silver glue. The data in Figure 4.1 (b) shows that in this case the first peak at 44.2ps lowers. Also the second peak is shrunk. Overall the maximum impedance at 44.2ps is reduced to $54.9 \Omega$ and the total impedance mismatch to a value of $8.9 \Omega$. According to Equation (2.7) this corresponds to a transmission improvement of 6%. This indicates that a precise fabrication of the holes for the con-
Chapter 4 Results and Discussion

Figure 4.1: Impedance measurement, (a) pressing Teflon and Rogers 3010 onto the pin, (b) the plug hole is filled up with silver glue.

Connector dielectric in the sample holder is of utmost importance for an impedance matched connection.

4.2 Vector Network Analysis

This section deals with the analysis of transmission spectra of PCB structures implemented in the big sample holder. The measurements were made with the VNA as described in Section 3.4.2. The output power of the VNA amounts \(-10\,\text{dBm}\) except for the measurements with the chip holder in Section 4.2.4. Here the power is reduced to \(-40\,\text{dBm}\) because of the heating effect at a temperature of about 4.2K.

In Section 4.2.1, different CPW transmission lines structures are characterised followed by a theoretical approximation for transmission line spectra in Section 4.2.2. Section 4.2.3 treats the CPW resonators and in Section 4.2.4 the behaviour of two superconducting structures on a chip mounted in a chip holder is examined.

Unless otherwise mentioned, all measurements are performed without optimizing the impedance mismatch of the connectors with silver glue (cf. Figure 4.1).
4.2 Vector Network Analysis

4.2.1 Transmission Lines

Transmission Dependency on Ground-Contact

The PCBs need to be grounded for a defined transmission of microwaves. The lower ground planes of all the PCBs are connected to the box ground by the pressure connection. The box ground is in turn connected to the ground of the SMA connectors. Furthermore, for all PCBs with vias the upper ground plane is connected to the lower one. In the next section, we investigate the various grounding configurations with additional ground contacts. In this section, use a PCB without vias and study the following possibilities the ground connection:

(a) SMA connectors, terminated with a short,
(b) SMA connectors, terminated with 50Ω load,
(c) bonding wires (here 48 in total) from the PCB to the box.

Figure 4.2: Variation in grounding connection. (a) Connectors terminated with shorts, (b) connectors terminated with 50Ω resistors, (c) bonding wires from the PCB to the box. The 6 previously used SMA connector ports are replaced by plugs.

In the entire transmission spectrum the characteristic shape of the black (terminated by shorts) and the green (terminated by 50Ω resistors) curve is quite similar. The absorption dips are almost at the same frequencies. They differ in the fact that 50Ω terminated data generally shows less attenuation. Considering the measurement where the connection is realized with 48 bonds, the peaks are shifted towards higher frequencies. Compared to the other measurements the dips are less significant. Therefore, qualitatively we find that bonding leads to the best connection between the PCB ground and the box ground.

Comparison of the Ground Contact with Bond Wires and Pressure Contact

Figure 4.4 shows three transmission spectra of the PCB \(TL = \{LD - 0\}\) with varied number of bonds to connect the PCB to the ground.

One can see that the 16 bonds – compared to the 0 bond curve – considerably suppresses most of the dips. For 48 bonds the spectrum is even flatter. This shows that a better ground contact is obtained even if there already is a pressure contact. However, in the
Figure 4.3: Transmission spectrum of $TL - \{\theta - \theta\}$ varying grounding method.

Figure 4.4: Transmission spectra of the PCB $TL - \{LD - 0\}$ with varied number of bonds.
rest of the measurements (except for those on the chip holder) shorts were used for ground connection.

### Effect of the Via Density

In this section, the correlation between the via distribution and the transmission spectra is studied.

The measurements are shown in Figure 4.5. Comparing the graphs of $TL - \{0 - 0\}$ (black line) and $TL - \{LD - 0\}$ in Figure 4.5(a) one can see how vias located only along the CPW suppress the large absorption maxima and therefore improve the transmission. The shape is almost flattened up to $f \approx 6$ GHz for the highest via density. Interestingly, at higher frequencies in the range of $5$ GHz $\leq f \leq 8.5$ GHz, one recognizes that the shape of $TL - \{HD - 0\}$ (green line) compared to $TL - \{LD - 0\}$ varies more intensively. In general this indicates that more parasitic modes are excited.

In Figure 4.5(b) the transmission spectra of transmission lines with vias throughout the whole ground plane are shown. The reader should note that the transmission is displayed in a modified scale. In contrast to Figure 4.5(a) the graphs are now flattened up to 8.5 GHz, but at higher frequencies the transmission also varies strongly with frequency. Again, a higher via density clearly improves the low-frequency region of the spectra, whereas at higher frequencies, the situation is more complicated.

Although a sound explanation for the above results will require experimental further studies beyond the scope of this work, we can imagine two reasons.

First, the electric field has to change its shape continuously to meet the requirements of the Maxwell equations on the metallic interface of the vias. These state that the field lines thread the conducting interface perpendicularly [10]. An illustration is shown in Figure 4.6.

Secondly the ohmic losses within the penetration depth of the conducting vias can cause a transmission loss. Yet, for thorough analysis more of these problems we need a data with varying via densities.
Chapter 4 Results and Discussion

4.2.2 A simplified Model of the Transmisson in a lossy CPW

In this section we will try to apply the standard transmission line theory \[3\] to the measured transmission spectra. Assuming a low loss transmission line the attenuation constant \(\alpha\) is given by \[3\]:

\[
\alpha \simeq \frac{1}{2} \left( \frac{R}{Z_0} + GZ_0 \right),
\]

(4.1)

The first term \(\frac{1}{2}(R/Z_0)\) describes the ohmic loss whereas the second term \(\frac{1}{2}(GZ_0)\) integrates the dielectric loss. We note that Equation (4.1) was originally derived for a coaxial geometry and adapted to the CPW by adjusting a few geometric constants.

In order to estimate \(R\) Figure 4.7 shows the strongly simplified geometrical situation in a CPW line. In this model \(R\) is approximately calculated by the skin depth \(\delta_S\) and the specific resistance \(R_S\) as:

\[
R = R_S \frac{l}{A} \cdot \frac{1}{l} = R_S \frac{1}{4\delta_S^2},
\]

(4.2)

where \(l\) is the length of the CPW transmission line and \(d\) the height of the CPW conductor.

The skin depth itself is defined as \[3\]:

\[
\delta_S := \sqrt{\frac{2}{\omega \mu \sigma}},
\]

(4.3)

where \(\mu\) is the permeability of the medium and \(\sigma\) the electrical conductivity.

In order to predict the dielectric losses qualitatively we assume the CPW to be similar to a coaxial line. Then we can use \[3\]:

\[
GZ_0 = \omega \epsilon'' \eta
\]

(4.4)

where \(\eta = \sqrt{\frac{\mu}{\sigma}}\) is the intrinsic impedance of the dielectric material filling the space between the conductors. \(\epsilon'\) and \(\epsilon''\) are the real and imaginary part of the complex permittivity \(\epsilon = \epsilon' - i\epsilon''\) and are related by the loss tangent \[3\]

\[
\tan \delta = \frac{\omega \epsilon'' + \sigma \omega \epsilon'' \gg \sigma}{\omega \epsilon'}.
\]

(4.5)
4.2 Vector Network Analysis

Figure 4.7: Estimation of the series resistance $R$ of a CPW. In this model only the opposing boundaries of the CPW are taken into account. In this approximation the field density can be considered to be uniform.

For a CPW $\epsilon' = \epsilon_0 \epsilon_{eff}$ where $\epsilon_0$ is the vacuum permittivity. The conductance $\sigma$ for air and Rogers 3010 is assumed to be negligible.

Supposing a conductor and a dielectric substrate with $\mu = \mu_0$ an approximated expression for $\alpha$ is given by:

$$\alpha \approx \frac{1}{2} \left( \frac{R_s}{4dZ_0} \sqrt{\pi \mu_0 \sigma} \cdot \sqrt{f} + \frac{2\pi \tan \delta}{c} \sqrt{\epsilon_{eff} \cdot f} \right).$$

(4.6)

Here one can see that the electric losses have a square root dependence which is relevant only at low frequencies. At higher frequencies the linearly depended dielectric loss is much higher.

The voltage of a microwave in a transmission line of the length $l$ is given by

$$|V(z = l)| = |V_0| e^{-\alpha l}$$

(4.7)

and thus the attenuation in decibel is

$$10 \cdot \log_{10} \left( \frac{|V(l)|}{|V_0|} \right)^2 = 20 \cdot \log_{10} e^{-\alpha l} = \frac{20}{\ln 10} (-\alpha l)$$

(4.8)

Assuming no reflection in the transmission line this equation yields $S_{21}$.

Using the values of Table 4.1 one gets the expression:

$$S_{21} = \frac{20}{\ln 10} (-\alpha l) = -(a' \cdot \sqrt{f} + b' \cdot f) \ (dB)$$

(4.9)

with $a' = 8.74 \cdot 10^{-6} \frac{1}{\sqrt{s}}$ and $b' = 2.48 \cdot 10^{-11} \frac{1}{s}$.

<table>
<thead>
<tr>
<th>$R_s$ (Ωm)</th>
<th>$d$ (µm)</th>
<th>$Z_0$ (Ω)</th>
<th>$\sigma$ (Ωm)</th>
<th>$\tan \delta$</th>
<th>$\epsilon_{eff}$</th>
<th>$l_{TL}$ (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.7 \cdot 10^{-8}</td>
<td>35.0</td>
<td>50.0</td>
<td>58.0 \cdot 10^6</td>
<td>2.3 \cdot 10^{-8}</td>
<td>4.7</td>
<td>54.75</td>
</tr>
</tbody>
</table>

Table 4.1: Values of the physical quantities.

In Figure 4.8 the transmission spectrum of $TL - \{LD - LD\}$ together with the theoretical prediction of $S_{21}$ Equation (4.9) and a fit up to 3GHz is shown. The fit is made according to Equation (4.9) with an additional offset $c$ ($a, b, c > 0$):

$$S_{21} = -(a \cdot \sqrt{f} + b \cdot f) + c$$

(4.10)
Chapter 4 Results and Discussion

<table>
<thead>
<tr>
<th>a (\frac{1}{\sqrt{s}})</th>
<th>b (\frac{1}{s})</th>
<th>c (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(9.835 \pm 0.996) \cdot 10^{-7}</td>
<td>(4.011 \pm 0.142) \cdot 10^{-11}</td>
<td>0.005 \pm 0.001</td>
</tr>
</tbody>
</table>

\[ a/a' = 0.11 \quad b/b' = 1.62 \]

Table 4.2: Fitting parameters with standard errors.

The computed fitting parameters are listed in Table 4.2. The comparison of these parameters with the predictions of Equation (4.9) – the third column in Table 4.2 – and Figure 4.8 show that despite its approximation the model shows a reasonable agreement with the data. The fit reproduces the data excellently.

![Approximated S_{21} curve and fit up to 3GHz](image)

Figure 4.8: Approximated $S_{21}$ curve and fit up to 3GHz

4.2.3 Resonators

In this section microwave transmission measurements of another PCB structure – the CPW resonators – are analysed.

In Figure 4.9 three transmission spectra of CPW resonators with different via configurations are shown. All resonators show an increasing transmission towards higher frequencies as it is expected for a capacitance. However, there are no clearly visible resonance peaks at the expected frequencies $f_n$ as listed in Table 4.3 and also indicated by the vertical blue lines in the upper area of the diagram. An explanation for this could be that the quality factor $Q$ of the resonator is very low. Furthermore, there exists a periodic structure of the absorption dips (Table 4.4). They cannot be attributed to the transmission line resonator, where peaks are expected. These dips rather represent a loss of energy into another, less obvious resonant structure at these frequencies.

1The fit was made with gnuplot.

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4.2 Vector Network Analysis

\[ f_n = \frac{c}{2 \sqrt{\varepsilon_{\text{eff}}}} \cdot n, n = 1, 2, \ldots \]

Table 4.3: Resonance frequencies of PCB resonators calculated with equation \( f_n = n \cdot \frac{c}{2 \sqrt{\varepsilon_{\text{eff}}}} \), \( n = 1, 2, \ldots \).

<table>
<thead>
<tr>
<th>( n )</th>
<th>( f_n ) (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>2.78</td>
</tr>
<tr>
<td>2</td>
<td>5.58</td>
</tr>
<tr>
<td>3</td>
<td>8.36</td>
</tr>
<tr>
<td>4</td>
<td>11.15</td>
</tr>
<tr>
<td>5</td>
<td>13.93</td>
</tr>
<tr>
<td>6</td>
<td>16.72</td>
</tr>
</tbody>
</table>

Table 4.4: Frequencies of the attenuation dips, as shown in Figure 4.9.

<table>
<thead>
<tr>
<th>( f_{n,\text{absorb}} ) (GHz)</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.95</td>
<td>8.18</td>
<td>12.19</td>
<td>16.74</td>
<td></td>
</tr>
</tbody>
</table>

**Coupled Resonators**

In the course of analysing resonators coupled resonators were studied. As we did not see any resonances in the spectrum of the single CPW resonators we also did not observe any useful information in the spectra of coupled resonators. We therefore omitted to show any data of these structures in this work.

**4.2.4 Chip Holder**

The main goal of the thesis is a successful implementation of superconducting microchip in a PCB. For this purpose the PCB chip holders are designed.

A sample chip with a superconducting CPW-slotline-transition which is shown in Figure 4.9.

![Figure 4.9: Transmission spectra of Resonators. The dashed blue lines are located at the expected resonance frequencies \( f_n \), \( n = 1, 2, \ldots \), which are listed in Table 4.3. At the black dashed lines there are absorption peaks of the red and the green graph. Their values are listed in Table 4.4.](image)
Figure 4.10 (b) was embedded in the milling groove of the CPW. From a theoretical point of view, the structure behaves similar to a capacitance at low frequencies and should turn into a transmission-line-like behaviour (0dB) at a certain frequency [11], as shown in Figure 4.10.

![Figure 4.10](image1)

(a) Calculated and measured insertion loss of a CPW-slotline-transition [11]. (b) A superconducting CPW-slotline-transition is located in the milling groove of a PCB chip holder.

Figure 4.11 depicts the transmission spectra of the CPW-slotline-transition, on the one hand measured in the small sample holder, on the other hand implemented in a PCB with the highest available via density (Chip-H \{HD – HD\}).

The PCB embedded chip shows a similar transmission spectrum as the same structure in the small box up to 8 GHz. In this frequency range, which is very important for circuit QED experiments, the performance is limited only by the properties of the structure itself. For higher frequencies, the chip holder measurements show a more erratic spectrum than the small box measurements. We note that this behaviour is consistent with our other PCB measurements discussed in this work.

In the next step, a sample with a superconducting resonator is mounted into the chip holder. The results are shown in Figure 4.12. The resonance peaks – marked with the dashed lines – can be clearly seen in both graphs. Despite this encouraging result with this more complicated structure, it becomes evident that the performance of the chip holder is still inferior to that of the small box sample holder and requires further optimization. We further note that for both curves the low resonator peak height is not due to the coupling capacitor but to the frequency resolution of the spectrum.

![Figure 4.11](image2)
Figure 4.11: Transmission spectra of a superconducting CPW-slotline transition [12]. The black curve was measured by using the small sample holder, the red curve by the use of a PCB.

Figure 4.12: Transmission spectra of a superconducting CPW resonator. The black curve was measured by using the small sample holder, the red curve by the use of a PCB. The graph of the big sample holder (red line) is shifted +15dB to higher transmission for a better visibility of the data.
Chapter 5
Conclusion

In this bachelor’s thesis the characteristics of a PCB-based sample holder for scalable circuit QED experiments in the microwave regime is investigated. The majority of the measurements is performed using coplanar transmission lines. We experimentally confirm that these can be approximately described using a simplified model for coaxial cables. For CPWs, the importance of vias for the suppression of parasitic modes up to 8 GHz could be verified. Vias are especially substantial if the size of the PCB is of the same magnitude as the wavelength of the frequencies to be measured. Furthermore, a pressure contact between a SMA connector and a CPW imprinted in a PCB is studied and successfully applied at room temperature and at 4.2K. Finally, experiments with superconducting microchips embedded in PCBs reproduce a similar structure to measurements done in a well-known small box setting.

For future measurements, a more systematic analysis of the impact of different via densities promises a more detailed insight in the processes that excite parasitic modes. It may be possible to suppress the nearly equidistant absorption dips by placing the vias in an aperiodic fashion. In our view this would further decrease the propagation of parallel plate modes through the PCB.

There is a huge variety in PCB dielectrics as well as metallic PCB layers. It is promising to compare, for example, conducting layers made from gold-plated copper to the present tin-plated copper. As gold has a lower attenuation constant it might be possible to design resonator structures with a sufficiently high quality factor on the PCB, allowing one to study the impact of the via density on the spectrum of single or even coupled resonator structures.

As there were several imperfections in the layout and fabrication of the big sample holder, a new box could be designed. The radius of the through holes for the SMA connectors has to be very accurate to avoid impedance mismatches. Dielectric bars could be built into the lid of the big sample holder and pressed on the top of the pins to improve the pressure connection. Furthermore, one can apply other connector types such as surface mount device connectors.

Also, a more profound theoretical description of a lossy CPW would allow for a better understanding of the attenuation in transmission lines. These can be compared to combined reflection/transmission measurements which also allow a calculation of the attenuation.

Finally, more than one superconducting microchip can be embedded in a PCB or the number of feeding lines to the chips can be increased. Keeping the background spectrum under control during this step would establish an important precondition to scale up either the number of microwave experiments or the complexity of a single experiment in our sample holder.
Bibliography


Danksagung

An dieser Stelle möchte ich mich bei allen bedanken, die mich während meiner Tätigkeit am Walther-Meißner-Institut und der Anfertigung dieser Arbeit unterstützt haben.

Prof. Dr. Rudolph Gross danke ich, dass er diese Arbeit überhaupt möglich gemacht hat. Er kümmerte sich stets um unser Wohl und hatte immer ein offenes Ohr für Probleme.


Für überaus hilfreiche Diskussionen und Hilfestellungen bezüglich meiner Arbeit möchte ich mich bei Frank Deppe bedanken. Sein Korrekturen führten immer zu erheblichen Verbesserung der Qualität.

Bei Matthias Danner möchte ich mich für seinen theoretischen Beistand und für seine Mühe beim Korrekturlesen dieser Arbeit bedanken. Durch ihn wurde der ein oder andere ungeschickt formulierte Satz verbessert, der vermutlich zur Belustigung aller Leser beigetragen hätte.

An dieser Stelle möchte ich mich auch für die Mithilfe der WMI Werkstatt, ganz besonders bei Robert Müller bedanken. Er stand stets bei technischen Problemen immer mit Rat und Tat zur Seite.


Elisabeth Hoffmann danke ich für die Hilfe bei Messproblemen und für Korrekturarbeiten.

Für die tolle Atmosphäre in unserem Büro möchte ich mich bei Lukas Augsbach, Norbert Kalb, Philipp Summer und Thomas Ramirez bedanken.
