Microstrip Solutions for Innovative Microwave Feed Systems

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Abstract

This report is introduced with a presentation of fundamental electromagnetic theories, which have helped a lot in the achievement of methods for calculation and design of microstrip transmission lines and circulators. The used software for the work is also based on these theories.

General considerations when designing microstrip solutions, such as different types of transmission lines and circulators, are then presented. Especially the design steps for microstrip lines, which have been used in this project, are described. Discontinuities, like bends of microstrip lines, are treated and simulated. There are also sections about power handling capability of microstrip transmission lines and different substrate materials.

In the result part there are computed and simulated dimensions of the microstrip transmission lines used in the prototype system. Simulations of conceivable loads in the cavity illustrate quantitatively the reflection coefficient. Even practical measurements are made in a network analyzer and are presented in this part.

Suitable materials and dimensions for the final microwave feed transmission line system for high powers are then presented. Since circulators are included in the system a basic introduction to the design of these in stripline and microstrip techniques is also made.

At last conclusions, examinations of the designed system and comparisons to the today’s systems are made.

Keyword

Circulator, Dielectric materials, Isolator, Microstrip, Microwave feed system, Stripline
This report is introduced with a presentation of fundamental electromagnetic theories, which have helped a lot in the achievement of methods for calculation and design of microstrip transmission lines and circulators. The used software for the work is also based on these theories.

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PREFACE

This report is the result of the author’s diploma work performed at the Research and Development Department at Whirlpool Sweden AB and constitutes the final element of the Master of Engineering exam in Electronics Design at the University of Linköping at Campus Norrköping. Most of the work is performed at the Development Department at Whirlpool Sweden AB, during the autumn 2001, under the supervision of Per Törngren, Ulf Nordh and Håkan Carlsson.

I want to thank my supervisors Per Törngren, Ulf Nordh and Håkan Carlsson at the company Whirlpool Sweden AB and my examiner Håkan Träff at the Institution of Science and Technology (ITN) at Campus Norrköping. Without their support this work would never reach this final result. I would also like to thank Gunnar Filipsson, for his advice concerning circulator components, and Roland Ekinge, manager of the Development Department at Whirlpool Sweden AB, for his support. I would also like to take the chance to thank Ida Sahlin, my lovely girlfriend, for her patience and support during the work.

Finally the author hopes that the work presented in this report will be useful for the further work with the evaluation of these patented new solutions.

Magnus Petersson
Norrköping, October 2001
1 INTRODUCTION

The background to this report is a diploma work performed at Whirlpool Sweden AB, where a proposition on transmission lines and components manufactured in microstrip solutions has been ordered. These lines and components are thought to function as microwave feed system and could maybe replace the traditional waveguides in the future. While the power in the proposed future system is quite high, about 200 watts from each generator, the system must be designed in special ways. In this report there is a prototype system presented and there will also be an investigation of the advantages versus the disadvantages of such a system. There will also be a description of the possibilities to expand the prototype system to handle the higher power. The systems are designed for the used ISM-band frequency, which ranges from 2.4 GHz to 2.5 GHz, of 2.45 GHz.

The main task in this work has been to find a suitable design method for microwave feed systems manufactured in microstrip solutions and an evaluation of these. This new technology is patented by Whirlpool and is thought to imply a more efficient system. Further the work has included literary studies, simulations, design and manufacturing of a prototype system and measurements and evaluations of this.

A proposal of such a transmission line microwave feed system is to let the incoming microwave signal pass through a circulator to the first port in the cavity. The reflected energy waves from this first port are then transferred via an isolator to the second port. The idea to transfer the reflected energy to the next port in this way is to achieve a more efficient system, where the losses are low. While there always are different loads in the microwave oven cavity there will be a different reflection rate, caused by different rate of mismatch in the system. This proposal constitutes the patent that is hold by Whirlpool Sweden AB and is investigated in this diploma work.

The introduction part to this report is a theoretical background, which is a result of the previously done literary study, and is made to get an idea of which designs and types of lines that could be used. There will also be a presentation of performed calculations and simulations of the investigated transmission lines and substrates. Final dimensions and materials for the prototype system will be presented and a proposal of a system for the required power would also be made.

The work will include the following tasks:

- Analysis of the problem
- Literary studies
- Designing and simulating a prototype system
- Order materials for the prototype system
- Perform measurements of the prototype system in a network analyzer
- Analysis and evaluation of the prototype system
- Proposals of a suitable system for higher power
2 THEORY

2.1 Theoretical Background

In this part some important relations in electromagnetism theory will be mentioned, which have played a central role in the development of methods for calculations in transmission line analysis. For a more detailed introduction to electromagnetism theory, literature [9] in the reference list is recommended.

Through Faraday’s and Ampère’s laws, which are cornerstones of Maxwell’s theory, we can get the necessary tools to calculate the line parameters R, L, C and G for, for example a transmission line [2]. Therefore Maxwell’s equations are introduced first, followed by Ampère’s and Faraday’s laws. At the end of this section there is a description of the calculation method that is used in the computer program used during this work.

2.1.1 Maxwell’s Equations

To understand the meaning of Maxwell’s equations they are here presented in integral form. This is because it is much easier to illustrate the meaning of integrals in figures than it is to illustrate derivatives. But first Maxwell’s four famous equations will be stated [9]:

\[
\nabla \cdot E = \frac{\rho}{\varepsilon_0} \tag{2.1}
\]

\[
\nabla \cdot B = 0 \tag{2.2}
\]

\[
\nabla \times E + \frac{\partial B}{\partial t} = 0 \tag{2.3}
\]

\[
\nabla \times B - \varepsilon_0 \mu_0 \frac{\partial E}{\partial t} = \mu_0 J_m \tag{2.4}
\]

where,

\( E \) is the electric field intensity in Volts per meter,
\( \rho \) is the total electric charge density in Coulombs per cubic meter,
\( \rho_f \) and \( \rho_b \) are the free and bound charge density respectively,
\( B \) is the magnetic induction in Teslas,
\( J_m \) is the current density in A/m²,
\( \varepsilon_0 \) is the permittivity of free space, \( 8.854187817 \cdot 10^{-12} \) Farad per meter, and \( \mu_0 \) is the permeability of free space, \( 4\pi \cdot 10^{-7} \) Henry per meter.

These four fundamental equations apply to all electromagnetic phenomena in media that are at rest with respect to the coordinate system used for the del operators [9].
Maxwell’s equations generally give a better understanding in integral form. From Gauss’s law, equation 2.1 could be written

\[ \int_S \mathbf{E} \cdot d\mathbf{a} = \frac{1}{\varepsilon_0} \int_{\tau'} \rho_i d\tau' = \frac{Q_i}{\varepsilon_0} \tag{2.5} \]

in integral form. \( S \) is the surface bounding the volume \( \tau' \), \( \rho_i = \rho_f + \rho_h \) is the total charge density and \( Q_i = Q_f + Q_h \) is the total net charge contained within \( \tau' \). The meaning of equation 2.5 is illustrated in figure 2.1 [9].

In a similar way the magnetic induction \( \mathbf{B} \) also can be evaluated in integral form. From equation 2.2 we can write

\[ \int_S \mathbf{B} \cdot d\mathbf{a} = 0 \tag{2.6} \]

where \( S \) is a closed surface. The equation says that the net outgoing flux of \( \mathbf{B} \) through a closed surface \( S \) is zero and this is illustrated by figure 2.2 below [9].
When integrating equation 2.3 over a surface $S$ bounded by a curve $C$ it gets the form as in equation 2.7 [9].

$$\int_{S} \nabla \times \mathbf{E} \cdot d\mathbf{a} = -\int_{S} \frac{\partial \mathbf{B}}{\partial t} \cdot d\mathbf{a} \tag{2.7}$$

This result is obtained from applying Stoke’s theorem. Finally we are going to integrate the fourth one of Maxwell’s equations, equation 2.4, also according to Stoke’s theorem. The integration is over an area $S$ bounded by a curve $C$ and is performed in formula 2.8 [9].

$$\oint_{C} \mathbf{B} \cdot d\mathbf{l} = \mu_{0} \iint_{S} \left( \mathbf{J}_{m} + \varepsilon_{0} \frac{\partial \mathbf{E}}{\partial t} \right) \cdot d\mathbf{a} = \mu_{0} I_{r} \tag{2.8}$$

After this fundamental introduction of Maxwell’s equations there will, in the following parts, be a description of Ampère’s and Faraday’s laws, which lead us in on methods for calculating the line parameters $R$, $L$, $C$ and $G$ for as example a transmission line [2].

### 2.1.1.1 Ampère’s Law

This law says that moving particles with current density vector $\mathbf{J}$, dependent of the electric field $\mathbf{E}$, give rise to a rotational magnetic field vector $\mathbf{H}$ surrounding the charge flow. This relation is shown in equation 2.9 below

$$\oint \mathbf{H} \cdot d\mathbf{l} = \iint \mathbf{J} \cdot d\mathbf{S} \tag{2.9}$$

with the line integral taken along the path characterized by the area of integration in figure 2.3 below [2].

![Figure 2.3: Path of integration [2]](image-url)
2.1.1.2 Faraday’s Law

In Faraday’s law the relationship between magnetic flux and electric fields is described. The time rate of change of the flux density \( B = \mu H \) \( (\mu = \mu_0 \mu_r) \) since a source gives rise to a rotating electric field according to equation 2.10.

\[
\oint_C \mathbf{E} \cdot d\mathbf{l} = -\frac{\partial}{\partial t} \int_S \mathbf{B} \cdot d\mathbf{a} = -\frac{d\Phi}{dt}
\]  

(2.10)

The above equation is rewritten from equation 2.7 by using Stoke’s theorem on the left side and by inverting the operations on the right and letting the surface \( S \) being fixed in space. As we can see from equation 2.7 and figure 2.4 below the electromotance around a closed curve \( C \) is equal to minus the rate of change of the magnetic flux \( \Phi \) linking \( C \) [2, 9].

![Figure 2.4: Voltage induction from magnetic flux [2]](image)

It will sometimes be more comfortable to convert the above equation into a differential form, which is shown in equation 2.3. From that equation we can see that a time-dependent magnetic flux is needed to obtain an electric field. This electrical field creates in turn a magnetic field according to Ampère’s law, since \( \mathbf{J} \) is dependent of the electric field \( \mathbf{E} \) in equation 2.9.

2.1.2 FDTD - Finite Difference Time Domain

In this section there will be a description of the finite-difference time domain (FDTD) method. The reason for this is that the computer program that have been used during this work, QuickWave-3D, uses this approach for calculations.

The main advantage of the FDTD analysis is that an impulse response contains all the information of a system for the whole frequency range, while the method is time domain based. This in turn implies that discontinuities could be characterized if a pulse excitation in the time domain is used. The characterization of the discontinuities could then be obtained by evaluating the Fourier transform of the time domain pulse response. FDTD is based on time and space discretizations of Maxwell’s equations:

\[
\frac{\partial \mathbf{E}}{\partial t} = \frac{1}{\varepsilon_i} \nabla \times \mathbf{H}
\]

(2.11)

\[
\frac{\partial \mathbf{H}}{\partial t} = -\frac{1}{\mu_0} \nabla \times \mathbf{E}
\]

(2.12)
where the index $i$ emphasizes different values of $\varepsilon_r$ to be used in substrate and air regions. The computer program simulates the wave propagation in three dimensions and therefore the spatial node points where different values of $\mathbf{E}$ and $\mathbf{H}$ are to be calculated are arranged in a mesh configuration. In the mesh there are unit cells. One such unit cell is illustrated in figure 2.5 below [1].

![Figure 2.5: One cell of a typical mesh used in the FDTD method [10]](image)

The computational domain consists of a repetitive arrangement of these cells. We can obtain every component of the magnetic field $\mathbf{H}$ by the loop integral of the electric field $\mathbf{E}$. To obtain the values of $\mathbf{E}$ the four surrounding nodal $\mathbf{E}$ values according to Maxwell’s curl equation for $\mathbf{E}$ are used [1].

When implementing a FDTD algorithm it is important to introduce absorbing boundary conditions to confine the computational space and thus keep the memory requirements at a reasonable level [1].

In QuickWave-3D a pulse excitation is used in most of the FDTD applications at one or more ports of the modeled device. To obtain the scattering parameters for a system a comparison of the Fourier transforms of the input and output signals is made when the wave simulation is accomplished. The Fourier transform calculations require, theoretically, an infinite period of time. But, since the exciting pulse has limited duration and the power entering the circuit is being dissipated at the input and output, which are matched, or at the absorbing boundaries, the signal at the ports becomes negligible after a limited period of time and the Fourier transform calculation can be limited to this period without causing significant errors [10].


2.2 Microstrip Solutions

In this section different microstrip solutions will be treated. There will also be a discussion of the advantages and disadvantages for these and a try is made to find out which solutions that are suitable for the purposes for the work at Whirlpool.

2.2.1 General Considerations

It is a common approach to implement different kind of circuits on motherboards with plane conductors. When designing these conductors for RF circuits, the high-frequency behavior of the conducting strips needs to be considered. The conductors are etched on the printed circuit boards (PCBs) [2].

A general microstrip configuration has a conducting strip with thickness \( t \) and width \( w \). The strip is placed on a substrate with the height \( h \) and the relative dielectric constant \( \varepsilon_r \). Below the substrate there is a ground plane. The ground plane below the current carrying conductor sections helps prevent excessive field leakage and thus reduces radiation loss. The configuration is shown in figure 2.6. A microstrip line may be seen as a distorted coaxial line, with the top strip as the center conductor and the ground plane transformed to a flat plane [1, 6].

![Microstrip transmission line](image)

**Figure 2.6: Microstrip transmission line [1]**

Microstrip solutions have however their disadvantages, while the electric field leakage are quite high. How big this leakage becomes depends on the relative dielectric constant. Generally a larger value on the relative dielectric constant results in a lower electric field leakage. For example, when a design needs high board density of the component layout it is important to use a relative dielectric constant that minimizes the leakage. Figure 2.7 qualitatively illustrates the electric field as a function of the relative dielectric constants [2].

![Electric field leakage](image)

**Figure 2.7: Electric field leakage as a function of dielectric constants [2]**
Another way to reduce the radiation losses is to have the conductor inside a substrate and grounded planes on both sides. This is called “sandwich structure” or stripline configuration and is illustrated in figure 2.8. This method is however more complicated to manufacture, at least as a prototype [2].

![Figure 2.8: Embedded transmission line to reduce radiation losses [2]](image)

A structure that is often used for high power applications is the parallel-plate line. In this configuration there are two plates with a substrate between, as illustrated in figure 2.9 below [2].

![Figure 2.9: Parallel-plate transmission line [2]](image)

The different types of transmission lines presented so far have the commonality that the electric and magnetic field components between the conductors are transversely orientated. That means that they are polarized and that they assume to form a transverse electromagnetic (TEM) field pattern, as is illustrated in figure 2.10. The electric and magnetic fields are recorded at a fixed instance in time as a function of space, where \( \hat{x} \) and \( \hat{y} \) are unit vectors in x- and y-directions [2].

![Figure 2.10: Electromagnetic wave propagation in free space [2]](image)
2.2.2 Design of Microstrip Lines

There are different calculation methods used in the literature. Three of them are mentioned and basically explained below [6].

- **Quasi-static analysis.** Is used if an assumption is made that the mode of wave propagation in microstrip is pure TEM.
- **Microstrip dispersion model.** In this method it must be considered that the effective dielectric constant and the characteristic impedance of the microstrip are functions of frequency and these factors are implemented in semi empirical techniques.
- **Exact evaluation.** Here a full-wave analysis of the microstrip is done. One has to introduce time-varying electric and magnetic fields and solve the wave equation. Field analysis without invoking any quasi-static approximation is also known as full wave analysis.

For the purposes of this diploma work the most relevant method is the Quasi-static, or so called Quasi-TEM approximation, which will be introduced in the following section.

2.2.2.1 Quasi-TEM Approximation

When dealing with design of microstrip lines one have to use approximations in the calculations. Also in computer programs approximations are used, even if they are more exact. One technique is to use the so-called quasi-TEM approximation. This technique utilizes the fact that the longitudinal components of the fields for the dominant mode remain very much smaller than the transverse components. This means that the longitudinal components can be neglected. In other words we can use the TEM mode analysis by this approximation. In TEM mode analysis the transmission line is treated as if it was made of two straight metallic conductors embedded in a homogeneous lossless dielectric. The conductors are also infinitely long, parallel to the longitudinal axis z and the cross section of the line and its material parameters are independent of the longitudinal position. This means shifting along the z-direction does not modify the structure. This assumption is called uniform or translation invariant [3]. Another explanation for quasi-TEM is that both the electric field and magnetic field intensities in the direction of propagation are negligible compared to the intensity of the TEM fields [6].

When using the quasi-TEM approximation an effective dielectric constant is calculated. This because of that the inhomogeneous microstrip is replaced by a homogeneous structure. The dimensions of the line though remain the same. The procedure is illustrated in figure 2.11 below. The precision in this approximate method is better than 0.2 % for $0.01 \leq w/h \leq 100$ and $1 \leq \varepsilon_r \leq 128$, where $w/h$ is the ratio width and height of the line [3].

![Figure 2.11: Procedure for the quasi-TEM approximation [3]](image-url)
There are however limitations in the quasi-TEM approximation. At low frequencies the approximation is quite good, but when the frequency increases there will be a tendency that the energy concentrates within the dielectric. This implies that the longitudinal components in the line increase. In the approximation one assumes that these components are insignificant. This problem can somewhat be reduced if a frequency dependent dielectric constant is used [3].

2.2.2.2 Formulas for Transmission Line Systems

For a generic transmission line system the so-called characteristic line impedance, \( Z_0 \), can be written as

\[
Z_0 = \frac{R + j \omega L}{k} = \sqrt{\frac{R + j \omega L}{G + j \omega C}}
\]  

where \( k \) is the complex propagation constant, \( R \) the resistance, \( L \) the inductance, \( G \) the conductance and \( C \) the capacitance of the line. It is important to note that \( Z_0 \) is not an impedance in the usual way. Its definition is based on the positive and negative traveling voltage and current waves. It is common to use \( Z_0 = 75 \, \Omega \) when dealing with very high frequencies and \( Z_0 = 50 \, \Omega \) at ultra high and super high frequencies (microwave frequencies). From the representation in formula 2.13 the input impedance of a terminated transmission line is developed and results in one of the single most important RF equations, namely

\[
Z_{in}(d) = Z_0 \frac{Z_L + jZ_0 \tan(\beta d)}{Z_0 + jZ_L \tan(\beta d)}
\]  

(2.14)

where \( \beta \) is the propagation constant and \( d \) the length of the line. If the load is represented by a short circuit, \( Z_L = 0 \), the following expression for \( Z_{in}(d) \) is obtained:

\[
Z_{in}(d) = jZ_0 \tan(\beta d)
\]  

(2.15)

and if the load is open ended, \( Z_L \rightarrow \infty \), the expression below is obtained [2].

\[
Z_{in}(d) = -jZ_0 \frac{1}{\tan(\beta d)} = -jZ_0 \cot(\beta d)
\]  

(2.16)

Formula 2.13 for \( Z_0 \) takes resistance and conductance losses into account. These quantities are always present, but in RF (radio frequency) and MW (microwave) circuits they do not cause significant errors. So when dealing with lossless lines we can set \( R = G = 0 \) and formula 2.13 simplifies to

\[
Z_0 = \sqrt{\frac{L}{C}}
\]  

(2.17)

and we can see that the characteristic impedance is independent of the frequency in this case.
If we take the parallel-plate line as example, we obtain the following expression for $Z_0$:

$$L = \frac{\mu d}{w} \text{ and } C = \frac{\varepsilon w}{d} \rightarrow Z_0 = \frac{d}{w} \sqrt{\frac{\mu}{\varepsilon}}$$  \hspace{1cm} (2.18)$$

In the above formula $\mu = \mu_0 \mu_r$ and $\varepsilon = \varepsilon_0 \varepsilon_r$ and results in approximately $376.8 \Omega$ in free space ($\mu = \mu_0$ and $\varepsilon = \varepsilon_0$). This is a useful value when dealing with for example antennas. Further the height of the substrate is $d$ and the width of the line is as usual $w$ [2, 4].

### 2.2.2.3 Design Steps for a Microstrip Line

There are a lot of different approximate methods treating microstrip transmission lines in the literature. In this work the approximate method that L. Reinhold and P. Bretchko describe in their book “RF Circuit Design – Theory and Applications” is chosen.

In this method the authors assume that the thickness $t$ of the conductor forming the line is negligible compared to the substrate height $h$. This means that the condition $t/h < 0.005$ must be satisfied. According to this condition we can use empirical formulas that depends only on the dimensions $w$ and $h$ of the line and the effective dielectric constant $\varepsilon_{eff}$. We must however consider two different cases, when $w/h < 1$ and when $w/h > 1$. We can see that $w/h$ is not a continuous function, it has a small discontinuity at $w/h = 1$. This discontinuity introduces only an error of 0.5 %. Since this error is so small we can use the expressions for $Z_0$ and $\varepsilon_{eff}$ in the following formulas [2].

For narrow strip lines, $w/h < 1$, we get the line impedance

$$Z_0 = \frac{Z_f}{2\pi \sqrt{\varepsilon_{eff}}} \ln \left( \frac{h}{4w} + \frac{w}{h} \right)$$  \hspace{1cm} (2.19)$$

where $Z_f = \sqrt{\mu_0/\varepsilon_0} \approx 376.8 \Omega$ is the wave impedance in free space. $\varepsilon_{eff}$ is the dielectric constant and is evaluated according to formula 2.20 [2].

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ \left( 1 + 12 \frac{h}{w} \right)^{-1/2} + 0.04 \left( 1 - \frac{h}{w} \right)^2 \right]$$  \hspace{1cm} (2.20)$$

For wide lines, $w/h > 1$, the formulas for the characteristic line impedance and the effective dielectric constant become [2]

$$Z_0 = \frac{Z_f}{\sqrt{\varepsilon_{eff}} \left( 1.393 + \frac{w}{h} + \frac{2}{3} \ln \left( \frac{w}{h} + 1.444 \right) \right)}$$  \hspace{1cm} (2.21)$$

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + 12 \frac{h}{w} \right)^{-1/2}.$$  \hspace{1cm} (2.22)$$
In figure 2.12 the ratio characteristic line impedance, as a function of $w/h$ is illustrated [2].

![Figure 2.12: Characteristic line impedance as a function of $w/h$][1]

This diagram could be the first of the design steps in some applications. If, for example, a circuit with the line impedance $Z_0$ on a substrate with height $h$ and a given dielectric constant $\varepsilon_r$ will be designed, the diagram could be used to get an approximate value on $w/h$. This value is then investigated, if it is smaller or bigger than two, and the correct formulas are used.

Similar to the diagram for the ratio of characteristic line impedance as a function of $w/h$ in figure 2.12 above, there is a diagram for $\varepsilon_{\text{eff}}$. This diagram describes the effective dielectric constant as a function of $w/h$ for different dielectric constants and is illustrated in figure 2.13 below [2].

![Figure 2.13: Effective dielectric constant as a function of $w/h$ for different dielectric constants][2]
As illustrated in figure 2.11 the effective dielectric constant is viewed as the dielectric constant of a homogeneous material that fills the whole space around the line. This means that dielectric substrate and surrounding air are replaced with an effective dielectric constant. If we have calculated the effective dielectric constant we can also calculate the phase velocity and the wavelength in the microstrip according to

\[ \lambda = \frac{v_p}{f} = \frac{c}{f \sqrt{\varepsilon_{eff}}} \]  

(2.23)

where \( c \) is the speed of light (\( c = 2.99793 \cdot 10^8 \) m/s) and \( f \) is the operating frequency. But when designing microstrip lines it could be good to have some mathematical relations to find the ratio \( w/h \) when the characteristic impedance \( Z_0 \) and the dielectric constant \( \varepsilon_r \) are given. If assuming an infinitely thin line conductor we can write for \( w/h \leq 2 \) [2]:

\[ \frac{w}{h} = \frac{8e^A}{e^{2A} - 2} \]  

(2.24)

where the factor \( A \) is found by

\[ A = 2\pi \frac{Z_0}{Z_f} \sqrt{\frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1} \left( 0.23 + \frac{0.11}{\varepsilon_r} \right)} \]  

(2.25)

and for \( w/h \geq 2 \) we can write

\[ \frac{w}{h} = \frac{2}{\pi} \left( B - 1 - \ln(2B - 1) + \frac{\varepsilon_r - 1}{2\varepsilon_r} \ln(2B - 1) + 0.39 - \frac{0.61}{\varepsilon_r} \right) \]  

(2.26)

where the factor \( B \) is given by

\[ B = \frac{Z_f \pi}{2Z_0 \sqrt{\varepsilon_r}} \]  

(2.27)

If we now want to use these formulas for the design of a microstrip line we can follow the design steps listed below.

1. Determine an approximate ratio of \( w/h \) and choose a curve that satisfies the dielectric constant.
2. Investigate if the ratio \( w/h \) is smaller or equal and bigger than two. Then the formulas with the corresponding interval are used and a value for \( w/h \) according to the formulas 2.24 or 2.26 is calculated depending on the approximate value of \( w/h \). If then for example the width of the trace is unknown it could be found from this relation.
3. The effective dielectric constant could be calculated according to formula 2.20 or 2.22.
4. When numerical values are calculated for the dimensions of the line the results can be checked with the formulas for the characteristic impedance \( Z_0 \) to see if it results in the wanted value.
If the effective dielectric constant is calculated, the phase velocity, $v_p$, can also be calculated and the effective wave length, $\lambda_e$ at the operating frequency according to formula 2.23 [2].

As seen by these above examples and formulas, there will be quite a lot of calculation steps to going through. So in reality it is often better to use a CAD program, but you presumable obtain a better feeling if solving it analytical.

In many cases one can not assume the line thickness $t$ to be zero (or $t/h < 0.005$). One can then use an effective width of the line, which could be written as

$$w_{eff} = w + \frac{t}{\pi} \left( 1 + \ln \frac{2x}{t} \right)$$

(2.28)

where $t$ is the thickness of the strip and the following conditions determines the value of the parameter $x$ [2].

$$x = h \text{ if } w > \frac{h}{2\pi} > 2t$$

$$x = 2\pi w \text{ if } \frac{h}{2\pi} > w > 2t$$

### 2.2.3 Discontinuities on Microstrip Lines

When designing microstrip circuits one will have to take into account the losses that bends, abruptly stopped open circuits, width changes and transitions give rise to. In these cases so-called discontinuities occur in the transmission line. The discontinuities give rise to very small capacitances (usually < 0.1 pF) and inductances (usually < 0.1 nH). The reactance effects are frequency dependent. So when the frequency is not too high, which is the case in this work, these effects becomes very small. For higher frequencies, say 10-20 GHz, there will be a more significant effect [4]. In this part the bend of a microstrip line will be considered.

### 2.2.3.1 Bends on a Microstrip Line

A common discontinuity is the bended microstrip line, often 90 degrees.

The electrical fields are concentrated in the outer corner, which implies that the capacitance increases there. The inductance in the equivalent circuit arises because of current flow interruption. This can be understandable having in mind that most of the current in a microstrip is flowing in the outer edges [3, 4].
To minimize the above effects one can design a matched bend. An example is the so-called Chamfered Bend and is shown in figure 2.14 together with its equivalent circuit. The equivalent is for the region between planes p and p’. Approximately the degree of chamfer is restricted to about $b \approx 0.57 \cdot w$ [4].

![Figure 2.14: A bended microstrip line and its equivalent circuit [4]](image)

### 2.2.4 Design Considerations

A very important issue in the diploma work is to investigate which lines that could be used at high frequencies and powers. The future oscillators are thought to be able to deliver a power of about 200 watts each, which leads to specific requirements on the design. These requirements are, among others, the used substrate material and the dimensions of the line. Other important issues are power handling capability, temperature rise, dielectric and conductor losses etc.

The goal with this section is to get an understanding of which materials and dimensions that are suitable for the wanted purposes.

#### 2.2.4.1 Substrate Materials

Substrates can be divided into five main categories: ceramic, synthetic, composite, semiconductor and ferromagnetic [3].

Aluminum ($Al_2O_3$) is the most commonly used ceramic substrate. This material has good surface quality, low loss etc. Since aluminum is a very hard material it is difficult to process. An intermediate layer of chromium, or another lossy conductor, is required because of that the adhesion between copper and aluminum is poor. Other ceramic materials are Beryllium ($BeO$) and Quartz ($SiO_2$) [3].

A common synthetic material is polytetrafluoroethylene (PTFE Teflon). These materials have low permittivity and quite bad mechanical properties like distortion and temperature dependence [3]. There are however improvements made to these materials. They could for example be impregnated with a ceramic loaded thermo set plastic resin to yield a thermal stable rigid laminate with electrical properties suitable at microwave frequencies. These materials have the advantages that they resemble FR4 in mechanical integrity and can be fabricated like basic FR4 material [19].
In the *composite* materials fiberglass or ceramic fillers increase the mechanical stability of synthetic materials. Even the permittivity can be modified, but the losses then become larger. One application for low permittivity materials is antennas. Materials that can be used at the low end of the microwave band are epoxy-fiberglass boards. This material has however rather large dielectric losses at higher frequency [3].

The substrate losses are larger in *semiconductor* materials than for *dielectrics*. Some common semiconductor materials are silicon (Si) and gallium archenide (GaAs). Semiconductor materials are commonly used to realize complete monolithic microwave integrated circuits (MMICs) [3].

Circulators and isolators are made of *ferromagnetic* materials. These substrates have relative permittivities between 9 and 16 and have generally low dielectric loss [8].

### 2.2.4.2 Power Handling Capability

There are two main phenomenon when dealing with power handling for microstrip solutions. For this work it is the so-called continuously working operation, which means that the major problems and limitations are thermal, that is the most important to take into account. Even if the other condition, pulse operation, is not the dominant part, there will be a study of effects like dielectric breakdown, for this condition too [4].

#### 2.2.4.2.1 Continuously Working Condition

Heat flow and temperature rise that occur due to conductor and dielectric losses control the maximum average power in the continuously working condition. The conductor losses are in most cases the dominant factor. An expression for this temperature rise is shown in formula 2.29

\[
\Delta T = \frac{0.2303 h}{K} \left( \frac{\alpha_c}{w_{\text{eff}}} + \frac{\alpha_d}{2w_{\text{eff}}(f)} \right) \quad \text{°C/W} \quad (2.29)
\]

where \( \alpha_c \) is the loss due to the conductor, \( \alpha_d \) the dielectric loss and are measured in decibels per meter. \( w_{\text{eff}} \) and \( w_{\text{eff}}(f) \) are effective microstrip widths, where \( w_{\text{eff}}(f) \) is frequency dependent. The constant \( K \) is the thermal conductivity of the used substrate. When studying formula 2.29 we can see that the substrate dominates the temperature rise in the case when continuous power is transmitted along the line. If a temperature rise \( \Theta \) above ambient is considered the following expression for the maximum average power can be used [4].

\[
P_{ma} = \frac{\Theta}{\Delta T} \quad (2.30)
\]

Here it is very important to choose a substrate that can handle at least about 200 W. Lines implemented on for example polystyrene or other plastic substrates are restricted to only about 100 W. Improvements can however be made, like using a wider microstrip line with lower characteristic impedance [4]. The design, in the proposition from Whirlpool, has not narrow lines as requirements, so this improvement might be used in this project.
2.2.4.2.2 Pulse Power Handling Capability
In microwave systems there are limitations in the coaxial connectors. They set for example the limit for the peak power. The final portion of the coaxial connector sets a limit because of air breakdown. Generally a transmission line with characteristic impedance $Z_0$ and maximum breakdown voltage $V_{mb}$ has an allowable power given by [4]

$$P_{np} = \frac{V_{mb}^2}{2Z_0}$$

(2.31)

There are two ways to improve the microstrip to be able to handle higher peak power. These are quoted below [4]:

- Since sharp edges intensify the electric field and thus reduce the allowable breakdown voltage, thick, rounded conductors are preferred.
- To avoid air breakdown near the edges, the strip conductor can be painted with a dielectric paint having a permittivity identical to that of the substrate.

2.2.4.2.3 Average Power
As mentioned in the above sections it is the temperature rise in the microstrip and the supporting substrate material that determines the average power handling capability. There are however some more parameters to take into consideration when calculating the average power capability. These parameters are listed below:

- Transmission line losses
- Thermal conductivity of the substrate material
- Surface area of the strip conductor
- Ambient temperature

where the last point, ambient temperature, is the temperature of the medium surrounding the microstrip. The loss of electromagnetic power in the strip conductor generates in turn heat in the strip. The heat distribution in the strip is uniform since it generally has good conductivity. There is also a heat flow from the strip, through the substrate, to the ground plane. This heat flow occurs because the ground plane has ambient temperature [1].

2.2.4.3 Power Losses and Radiation
The approximate radiation conductance $G_r$ could be written (if $h/\lambda_0$ and $w_{eff}/\lambda_0 << 1$) as [4]:

$$G_rZ_0 = \frac{4\pi hw_{eff}}{3\lambda_0 \sqrt{\varepsilon_{eff}}}$$

(2.32)

In the formula above $h$ is the height of the used substrate material, $w_{eff}$ the effective width of the microstrip line, $\lambda_0$ the wavelength in free space and $\varepsilon_{eff}$ is the effective dielectric constant of the used material. The expression is approximate but it gives an understanding of the rate of radiation conductance. The rate of radiation is critical in discontinuities like open circuits and bends. The conductance in the above formula could be transformed to a resistance. This resistance could be seen as a shunted component to the microstrip line and if the current in the line and the resistance value are known one can obtain an approximate value of the radiated...
power. Methods for minimizing the effects that occur by these discontinuities must be applied [4].

The Q-factor for a microstrip configuration is often specified, from which the total attenuation of the circuit can be found [4]:

$$\alpha(l) = \frac{8.686\pi l}{Q\lambda} \quad [\text{dB}] \quad (2.33)$$

In the above formula $\lambda$ is the wavelength in the substrate. Two other attenuation parameters are those due to conductor losses and dielectric losses. The conductor loss, $\alpha_c$, could be found by the following expression:

$$\alpha_c = 0.072\sqrt{\frac{f}{wZ_0}} \lambda \quad [\text{dB / microstrip wavelength}] \quad (2.34)$$

and the dielectric loss, $\alpha_d$, as:

$$\alpha_d = 27.3\frac{\varepsilon_r(\varepsilon_{\text{eff}} - 1)\tan \delta}{\varepsilon_{\text{eff}}(\varepsilon_{\text{eff}} - 1)} \quad [\text{dB / microstrip wavelength}] \quad (2.35)$$

where $\tan \delta$ is the loss tangent [4]. Further the surface wave propagation is important to take into consideration. These waves are in the forms of TE (Transverse Electric) and TM (Transverse Magnetic) modes. At the frequency of this work the dominant part of radiation is $G_r$. There are various techniques to limit the radiation of a system [4]:

- Metallic shielding
- Introduce an absorbing material near discontinuities
- Design the discontinuities for minimized radiation, for example the Chamfered Bend.

**2.2.5 Microstrip Layouts**

There has been an examination of different structures of microstrip solutions in this report. Stripline techniques have their advantages in low electromagnetic leakage and that it is easier to control the impedance of the line. A disadvantage is however that this structure is more complicated to manufacture than a simple microstrip structure. Both microstrip and stripline configurations must be shielded by a metal covering in any way. Since the microstrip configuration is easier to manufacture it will be chosen for this prototype system. It is however not excluded to use the stripline configuration in a future system. A disadvantage with the stripline configuration is that it generally endures a lower power than the microstrip configuration.

**2.2.6 Microstrip Transitions**

In this section the most common transition is described, which will be useable for this work. In the simulation program, *Quick-Wave 3D*, that have been used in this project one can simulate these transitions and in a coming prototype they also will become useful.
2.2.6.1 Coaxial-to-Microstrip Transition

The coaxial-to-microstrip transition principle is shown in figure 2.15 below. This transition is very common and it is the simplest connection and, in most cases broad banded in microstrip solutions, while both media support TEM mode [1]. This kind of transition could be used, for example from the oscillator source to the transmission line in the realized prototype system.

![Figure 2.15: Coaxial-to-microstrip transition [1]](image)

There are two common types of the coaxial-to-microstrip transitions, namely in-plane and right-angled. The center conductor pin is often soldered in the coaxial transition to the microstrip to reach high reliability. It is important that there is no gap between the connector flange and the front-side wall of the fixture. If there is a gap, there will be an additional discontinuity reactance due to long ground plane current flow. Figure 2.16 below shows the equivalent circuit for a coaxial-to-microstrip transition. When designing these transitions one either minimize the reactance or compensate for them by choosing the values so that $\sqrt{L_s / C_s} = 50 \Omega$ or which characteristic impedance one wish to obtain [1].

![Figure 2.16: Equivalent circuit for a coaxial-to-microstrip transition [1]](image)
2.2.7 Circulators

In this section there will be a basic introduction to circulators. The goal with the section is though to get an understanding of the circulators made in microstrip and stripline techniques. These kinds of circulators are then treated more in detail in the result part. The step from conventional circulators to those made in microstrip or stripline techniques is shown to be not so big.

The simplest type of circulator is the symmetrical three-port Y circulator, also called junction circulator. The principle for this type of circulator is illustrated in figure 2.17 below [7]. In this diploma work the circulators are surface mounted, i.e. the three conductor tabs illustrated in figure 2.17 are connected by soldering to the microstrip lines. The input signal transfers to the second tab and the signal from the second tab is transferred to the third tab.

![Image of a three-port junction circulator](image)

*Figure 2.17: Principle for a three-port junction circulator [7]*

When treating this kind of circulator it could be assumed that there are electric walls at the top and bottom and magnetic walls at the sides. In the disc resonator it is also assumed that the substrate thickness and relative dielectric constant are such that there is no variation of the electrical field, which means

\[
\frac{\partial}{\partial z} = 0
\]

(2.36)

where the \( z \) direction is perpendicular to the disc surface. The resonator disc is placed on a non-reciprocal ferrite material, which in turn is placed on a ground plane. For the coordinates in the following calculations the reader is referred to figure 2.18 [13, 15].
The electric field has only a $z$ component according to

$$
E = i_z E_z
$$

(2.37)

where $i_z$ represents a unit vector. We can then express the other field components in terms of the electric field component $E_z$ by Maxwell’s first curl equation, as [8]

$$
H_r = \frac{j}{\sigma \mu_0} \left( \frac{1}{r} \frac{\partial E_z}{\partial \phi} \right)
$$

(2.38)

$$
H_\phi = -\frac{j}{\sigma \mu_0} \left( \frac{\partial E_z}{\partial r} \right)
$$

(2.39)

$$
E_\phi = E_r = H_z = 0
$$

(2.40)

A magnetic DC field, $H_{DC}$, with a direction perpendicular to the disc biases the resonator. As was mentioned above such fields are assumed not to vary. After these assumptions the electric field inside the resonator satisfies the homogeneous Helmholtz equation in cylindrical coordinates [8, 13, 15]:

$$
\left( \frac{\partial^2}{\partial r^2} + \frac{1}{r} \frac{\partial}{\partial r} + \frac{1}{r^2} \frac{\partial^2}{\partial \phi^2} + k^2 \right) E_z = 0
$$

(2.41)

where

$$
k^2 = \omega^2 \varepsilon_0 \mu_{eff} \mu_0
$$

(2.42)

$$
\mu_{eff} = \frac{\mu^2 - \kappa^2}{\mu}
$$

(2.43)
The parameters $\kappa$ and $\mu$ in formula 2.43 are elements of the polder tensor. To obtain a correct degree of magnetization and dimension of the disc radius at resonance we must get the proper connection between $\kappa$ and $\mu$. Since magnetic walls were assumed, which means that it is assumed that the radial component of the surface current must vanish, we get

$$H_\phi(r = R) = 0 \quad (2.44)$$

at the edge of the disc. The above condition is satisfied whenever [8]

$$J'_{n}(kR) = 0 \quad (2.45)$$

The proper connections between $\kappa$ and $\mu$ will then be described by the equations for two mode systems, $+ \text{ mode}$ and $- \text{ mode}$, as $\pm \text{ mode}$ [13]:

$$J_{n+1}(kR) - \frac{n J_n(kR)}{kR} \left(1 \pm \frac{\kappa}{\mu}\right) = 0 \quad (2.46)$$

If $\kappa/\mu = 0$ a standing wave pattern occurs, with maximum amplitude at the input port and equal amplitude at the two other ports. This state indicates demagnetization. While the function of the circulator is to transmit the signal at the input port to the second port, we want a magnetized state that performs this. Such a state occurs when the input port and the second port have the same field amplitude and the third port is isolated with zero field amplitude [13].

The junction circulator can be manufactured in waveguides, microstrip or stripline. At microwave frequencies the circulator in stripline has an isolation of about 20 dB. Losses in the transmission are only a few tenths of a dB. The microstrip circulator design has its advantages in that one can let the whole substrate material be made of ferrite, together with other components and transmission lines. This because of that the static magnetic fields are only directed to the non-reciprocal component [7].

### 2.3 Practical Measurements

To obtain an understanding about how much the simulated parameters vary from reality, practical measurements had also been performed when this was possible. One can for example measure the rate of reflection in the system with a network analyzer, which can represent the results in the Smith Chart and as scattering parameters. This will be done for the prototype system in the result part. The network analyzer performs measurements of RF systems, in which voltages in terms of magnitude and phase are measured. The network analyzer used for this work is the model HP 8714B from Hewlett Packard. In the measurements the frequency is swept over an interval and the results are presented in different ways. Two important ways to get the results are graphically in the Smith Chart and numerical by the scattering parameters. Especially the $S_{11}$ parameter is important because it describes the rate of reflection in the system. For the interested reader, an introduction to the Smith Chart and the scattering parameters are presented in the appendix of this report.
3 RESULT

In this part of the report important results during the diploma work at Whirlpool Sweden AB are presented. To get an idea about which dimensions that are reasonable for the transmission line system, Whirlpool has made a prototype cavity. This cavity is outlined in the following part. The transmission lines used in the design have been simulated in the Quick-Wave 3D software and are also calculated with approximate methods on paper. The FDTD method that the software uses is presented in section 2.1.2 and the approximate calculation methods used are presented in section 2.2.2 – 2.2.2.3. Another program that has been used is AppCAD from Agilent Technologies.

To verify the function of the microwave feed system a prototype on a PTFE substrate will be made. The material used for this prototype can not hold especially high powers, but the losses are similar to those materials that could be used at a higher power. The layout for the prototype is made in the software BOARDMAKER from Tsien. The measurements are performed in a network analyzer and the data is saved and compiled to diagrams, presented in the report.

After this function verification an expansion of the prototype design will be made. A proposal of new dimensions for new materials that can handle the required higher power will be made. Final dimensions and material choices are presented in form of calculations, designs and simulations in the last part of this section. Conclusions and comparisons to the today’s systems are also performed.

There will also be some simulation results in this section that show how the microwaves are transferred through common discontinuities.

3.1 Prototype Cavity

A prototype of the oven cavity is made at Whirlpool, to get an idea of how it could be fed (the space in the oven, where food is placed, is called cavity). The prototype microwave feed system, described later in this section, will be made for the roof section of the cavity. The layout of the air slots at this roof and the dimensions of these are illustrated in figure 3.1 below. The size of the outer area (160×100 mm) is chosen equal to the substrate that is ordered for this prototype. The material is described more in detail in the next section. The dimension of the cavity roof is larger than the size of the used laminate.

Figure 3.1: Dimension of the substrate board at the cavity roof with air slots
The air slots in the roof are made for transferring the energy waves from the microstrip transmission lines to the cavity. Reflected energy is transferred from the first port (to the left in figure 3.1) in the cavity through the system of microstrip lines to the second port. Design aspects here are for example to get a maximum of the magnetic field in the center of the air slots and to get good transfer characteristics. A more detailed description of the prototype system is given in section 3.2.

3.2 Prototype Microwave Feed System

The prototype microwave feed system is designed for a PTFE substrate with height \( h = 0.8 \text{ mm} \) and a copper foil of \( t = 35 \mu \text{m} \) on both sides. This design is made for illustrating the function of the system and is not intended for high power. With such a prototype feed system one gets an understanding of reflection coefficients, losses etc. For higher power another material with other dimensions is needed. These requirements and proposal of material choice are presented in section 3.3.

3.2.1 Calculations and Simulations

In this section important calculation and simulation results are presented. The design steps and final chosen dimensions are presented and the function is verified by simulation results in the Quick-Wave 3D software. A prototype will also be made, with all components included, to be able to perform practical measurements in the network analyzer.

The PTFE-substrate, from the manufacturer ELFA, used in the prototype design has the following properties:

<table>
<thead>
<tr>
<th>Properties</th>
<th>Remark</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dielectric constant</td>
<td>@ 10 GHz</td>
<td>2.75</td>
</tr>
<tr>
<td>Dissipation factor</td>
<td>@ 10 GHz</td>
<td>0.0030</td>
</tr>
<tr>
<td>Dielectric conductivity</td>
<td>Estimated @ 2.45 GHz</td>
<td>0.001122 S/m</td>
</tr>
<tr>
<td>Dielectric breakdown</td>
<td>Estimated</td>
<td>40 kV</td>
</tr>
<tr>
<td>Thermal conductivity</td>
<td></td>
<td>0.25 W/m/K</td>
</tr>
<tr>
<td>Crystalline melt point</td>
<td></td>
<td>327 °C</td>
</tr>
<tr>
<td>Max. operating temperature</td>
<td></td>
<td>260 °C</td>
</tr>
</tbody>
</table>

*Table 3.1: Properties of the PTFE-substrate from ELFA*

The value of the dielectric conductivity in table 3.1 is calculated according to formula 3.6 in section 3.3.1.1 and is estimated at 2.45 GHz.

The first step is to calculate dimensions, which should imply suitable characteristic impedance of the line. By using figure 2.12 in section 2.2.2.3 it could be seen that \( w/h \) gets an approximate value of 3 for a dielectric constant of 2.75. To get a more exact value of \( w/h \) formula 2.28 could be used:

\[
\frac{w}{h} = \frac{2}{\pi} \left( B - 1 - \ln(2B - 1) + \frac{\varepsilon_r - 1}{2\varepsilon_r} \left[ \ln(B - 1) + 0.39 - \frac{0.61}{\varepsilon_r} \right] \right) \approx 2.66
\]
where the factor $B$ is given by

$$ B = \frac{Z_f \pi}{2Z_0 \sqrt{\varepsilon_r}} = 7.138 $$

The above calculations give us a relationship between the width of the line and the height of the substrate, $w = 2.66h$. This would imply a line width of about $2.13 \text{ mm}$. To check if this value seems to be correct, formula 2.21 and 2.22 for the calculation of $Z_0$ and $\varepsilon_{\text{eff}}$ is used ($w/h > 1$):

$$ Z_0 = \frac{Z_f}{\sqrt{\varepsilon_{\text{eff}} \left(1.393 + \frac{2}{3} \ln \left(\frac{w}{h} + 1.444\right)\right)}} = 50.3 \Omega $$

where

$$ \varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + 12 \frac{h}{w}\right)^{-1/2} \approx 2.25. $$

As seen from the above calculations the characteristic impedance becomes a little bit higher than $50 \ \Omega$. The line width is therefore chosen to be $2.2 \text{ mm}$, which implies that the characteristic impedance becomes $49.2 \ \Omega$. The effective dielectric constant does not differ significantly by this modification of the line, so the wavelength in the substrate does not either affect significantly.

The principle for how the cavity is fed by the microstrip transmission line is shown in figure 3.2, where $d$ is the length of the line and $l$ is the distance from the center of the air slot to the termination of the line. The microstrip feed line is chosen to be open-circuited, which implies that there will be a maximum magnetic field at $l = \lambda/4$. The place for the line is chosen to get this value of $l$ of the line over the center of the air slot. Since the input signal assumes to be a transverse electromagnetic (TEM) wave, where the electric and magnetic fields are orthogonal to each other and the propagation is the cross product, or Poynting’s vector, of the fields. There is a maximum voltage and a minimum current at an open-circuited point. This implies that there will be maximum current a quarter of the actual wavelength away from that point. A maximum current gives rise to a maximal induction of the magnetic field at that point, which is wanted.
The outlined frame is the dimension of the board, used for the prototype system. The arrangement in the system could be placed on a side or on the roof of the cavity in a microwave oven.

As seen from the above figure the substrate with the strip is placed over the port, or the air slot, on the cavity roof. This air slot is marked with dashed lines in the figure and has the same dimensions as the openings in the walls and the roof in the microwave oven. The board is positioned direct to the cavity, so the cavity and the ground plane is connected. The copper foil is etched away from the bottom ground plane of the substrate at the dashed region.

To be able to find out which lengths, from the open-circuited side, the lines should have to get a maximum node of the magnetic field in the center of the air slots, the wavelength of the transferred microwaves has to be known. The wavelength in the used PTFE substrate material is shown to be

\[ \lambda = \frac{v_p}{f} = \frac{c}{f \sqrt{\varepsilon_{\text{eff}}}} \approx \frac{3 \times 10^8}{2.445 \times 10^9 \sqrt{2.253}} \approx 81.75 \text{mm}. \]

The characteristics of a TEM sinusoidal wave were illustrated earlier, in figure 2.10. The dimensions of the line should be chosen so that maximum magnetic field occurs in the center of the air slot. If the line is chosen to be open-circuited the distance \( l \), according to figure 3.2, should be a quarter of the actual wavelength. With this distance the magnetic field get maximum value at the center of the air slot. Since the calculated wavelength is about 81.75 mm the distance \( l \) will be about 20.44 mm. Figure 3.3 shows the simulation result when the above dimensions are chosen.
As seen from the above figure there is a maximum magnetic field in the $z$-direction in the center of the air slot. This indicates that the distance $l$ is correct chosen. The transmission line is open-circuited, which implies that there is zero magnetic field, since there is in principle zero current, at the end of the conductor. In the above figure the input is to the left and the termination to the right, where there is in principle zero field. There is however little leakage at the sides of the open-circuited termination. The distance of the strip in the above simulation result is about 70 mm.

In some simulation scenarios there will be small differences between the methods used in this work and in the Quick-Wave 3D software, because of that the software takes more effects into consideration. The program also calculates the effective dielectric constant in another way, with more effects included, and takes the surrounding volume into consideration. However the accuracy for the approximate calculations are enough for the purposes of this work. This implies that some scenarios in the software are modified to get, for example, the same characteristic impedance as in the design steps, so they become comparable with the calculated results. The reason for modifying the parameters in the software is that its method is much harder to control than that for the hand-made calculations. However, the differences are very small and as a third security tool the software AppCad from Agilent Technologies is also used. These calculations do agree very well with the methods for the approximate calculations described in this report.

To obtain an understanding of the rate of reflection coefficient for the prototype material, calculations and simulations have been done. In the simulation program it is only possible to measure the $S_{11}$ parameter, since only one port is used. This is however the most relevant parameter to investigate, since we are interested in how much of the energy that is reflected. The value of the $S_{11}$ parameter for the line depends on how well the characteristic impedance $Z_0$ of the line is matched to the input impedance. Increasing rate of mismatch implies higher reflection coefficient $S_{11}$ according to

$$S_{11} = \Gamma_m = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$$

(3.1)
In the *Quick-Wave 3D* software one can study the impedance of a line in the Smith Chart. The real and imaginary values in the Smith Diagram are normalized to the characteristic impedance of the line. If for example the real part is $r$ and the imaginary part is $x$, the absolute value of the input impedance is calculated by

$$|Z_{in}| = \sqrt{(r \cdot Z_0)^2 + (jx \cdot Z_0)^2}$$  \hspace{1cm} (3.2)

For a future system there will be different rate of reflection depending on which loads that are used in the cavity. Even if the impedance of the load in the cavity was known it varies when it is heated. In other words the impedance in the cavity is always varying and is impossible to control. The investigated system in this report “reuses” this different rate of reflection power, due to the mismatch.

As mentioned before, open-circuited lines will be used in the design. This implies that the load impedance, $Z_L$, has ideally infinite value and that there is in principle zero magnetic field at that point. The voltage at this point has a maximum and the current a minimum, assumed that the open-circuited termination is pure. To check that maximum magnetic field occurs in the center of the air slot, a quarter of the wavelength away, simulations with continuous input sinusoidal at 2.45 GHz has been made. In the simulation result in figure 3.4 below the amplitude of the magnetic field is enveloped during one period. Note that there is in principle zero field at the open-circuited termination. The simulation result is performed at the level just below the air slot in the ground plane. This means that it is this field distribution that the port sends out into the cavity.

![Field Distribution](image)

*Figure 3.4: Magnetic field enveloped just below the air slot*

In section 2.2.2.2 formulas for different terminations were presented. Formula 2.14 is the general expression and 2.16 is valid for open-circuited transmission lines. In the feed system there will be circulators implemented and it is very important that the input impedance of the connected microstrip lines is matched to the characteristic impedance of the lines. This implies low reflections from the lines toward to the circulator. The characteristic impedance in the circulators and the surrounding network of microstrip lines are ideally 50 Ω. The lengths of the lines are dependent of the input impedance, so we want to choose suitable lengths to get as well matching as possible. However, the circulators will meet varying impedance because of the different loads in the cavity.
For the prototype system, with open-circuited microstrip lines, the following calculations are valid.

\[
Z_m(d) = \frac{Z_0}{j \tan(\beta d)} \Rightarrow l Z_m(d) = Z_0 \left\langle \frac{1}{j} \Rightarrow \beta d = -\frac{\pi}{4} + n\pi \right. 
\]

The length \( d \) of the line will therefore be

\[
d = \frac{-\frac{\pi}{4} + n\pi}{\beta} = \frac{-\frac{\pi}{4} + n\pi}{2\pi} = \lambda \left( -\frac{1}{8} + \frac{n}{2} \right)
\]

and values for different \( d \) are listed in the table below.

<table>
<thead>
<tr>
<th>n</th>
<th>d [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>-10.19</td>
</tr>
<tr>
<td>1</td>
<td>30.57</td>
</tr>
<tr>
<td>2</td>
<td>71.32</td>
</tr>
<tr>
<td>3</td>
<td>112.1</td>
</tr>
</tbody>
</table>

*Table 3.2: Values of the line length \( d \) as a function of \( n \).*

The first circulator is fed by the matched input signal to the system. The first output of the circulator (the second port) is matched to the line feeding the port in the cavity, i.e. the circulator feeds the first port in the cavity, via the microstrip line placed over the air slot described earlier. The line length for this purpose is chosen to be about 30.57 mm according to table 3.2. This length gives us approximately \( \frac{\pi}{8} \) equal to \( \frac{\pi}{4} \) of the line, without load. \( Z_m \) is the impedance that the circulator meets at its second port. If we can reach approximately equal impedance the reflection in the transition between circulator and line will be minimized.

From the circulator’s third port the reflected energy is transferred to the input of an isolator. The output of the isolator then feeds the second port in the cavity in the same way as described earlier. The distance of the line over the second port is chosen to be about 71 mm according to table 3.2. For a description, the whole prototype feed system is outlined in figure 3.5.
The circulator to the left in the above figure has a matched input, which means that the length of that line does not matter, for the ideal case. In reality there will be losses in the discontinuities of the line, such as the input coaxial transition. These effects are however very difficult to control, so there will be tests at the prototype system for the best matched input point at that line. For the output port, that feeds the first port in the cavity, there must be a distance at the microstrip line, which should imply that the input impedance for the microstrip line is equal to the characteristic impedance of the circulator. This is however not easy, while the line meets different loads, or different impedance, all the time. But for the test condition, as was mentioned earlier, the distance $d = 30.21 \, \text{mm}$ satisfies this. This configuration has been simulated in the Quick-Wave 3D software. The simulation result, in figure 3.4, illustrates the magnetic field ($H_L$) around the strip at the center of the air slot. We can see from this figure that there is maximum field amplitude at this area. In other words the distance $l$, according to figure 3.2, seems to be correct.

To see how well the microstrip line is matched to the characteristic impedance of the circulator, the reflection coefficient $S_{11}$ is simulated. First the simulation result for the line length of 30.57 mm according to table 3.2 is presented. Ideally $S_{11}$ should be zero, that is the line impedance equal to the characteristic impedance of the circulator. This is however not possible in reality, but we want as low value of $S_{11}$ as possible. To get an example of a reasonable comparison to reality, a special scenario in the Quick-Wave 3D software is used. In this scenario the whole cavity is simulated for each port. The cavity contains objects with set material parameters that simulates the glass plate and a liter of water in the microwave oven. The first case is the straight line that connects the circulator in figure 3.5 to the first port in the cavity. The distance $l$, according to figure 3.2, of the open-circuited line is chosen to be about $20 \, \text{mm}$. As seen from the simulation result in figure 3.6 the parameter $S_{11}$ takes the absolute value $0.2644$ at $2.45 \, \text{GHz}$. 

Figure 3.5: Outline of the prototype system.
In the final system there will however be different loads in the cavity, which will change the input impedance of the system. This also implies that the rate of mismatch in the system will vary. If, for example, the oven is empty there will in principle be total reflection and since the ports to the cavity are orthogonal to each other, most of the reflected energy will return to the sending first port. For the above example we can see that about 26 percent of the input power is reflected. The advantage with a system like this is that the varying rate of reflection will be transferred through the circulator and the isolator to the next port. In this way the reflected energy will be used for the same load a second time, which will increase the efficiency of the system.

The microstrip line that connects the isolator output to the second port in the cavity has a suitable length of about 71 mm. This corresponds to the alternative $n = 2$ in table 3.2 for calculated line lengths. As seen from the outline of the prototype system there must be a 90-degree bend of that line. These discontinuities are considered in section 3.2.3.1, where the chamfered bend is presented. This kind of bend is optimized to have low reflection and is used in this system. The dimensions for the Chamfered Bend are illustrated in figure 3.14. These dimensions are also chosen in the simulation program for the prototype material and the results are shown in figure 3.7. The line length in this simulation is chosen so that there is a good matching between the input impedance and the characteristic impedance.
As illustrated from the above figure the $S_{11}$ parameter has an absolute value of 0.1105 at 2.45 GHz. In other words there is about eleven percent of the input power that is reflected. In this value of the reflection coefficient the mismatch between the input impedance and the characteristic impedance is included. This is however an insignificant part and most of the reflection is due to the bend.

The reflected power from the second port in the cavity is transferred to the third, matched, port on the isolator. This port constitutes a power resistance and could on the used isolator handle a power of 110 watts, which is enough because of that the prototype system is made for lower powers.

### 3.2.2 Circulator and Isolator Components

For the function of the prototype feed system there are circulator and isolator components from the manufacturer RADITEK implemented. These components could handle an average maximum forward power of 300 watts. Maximum reverse average power is 110 watts, which is the maximum power that the isolator could handle. This power is the summary of the reflection through the system. The frequency band is 2.3-2.5 GHz, with insertion loss of 0.25, isolation of 22 dB and a voltage standing wave ratio (VSWR) of 1.19. The ripple in the pass band is 0.05 dB, peak-to-peak [20].

The meaning of the described prototype system in this section, is to evaluate if there is a possibility to realize a system like this, intended for microwave feeding at higher power. The simulated transitions, microstrip lines and air slots are also implemented as a prototype on a substrate material, described earlier in this section, from the company ELFA in Sweden. The procedure for this implementation and performed measurements are described in the next section. The prototype is made to be able to perform practical measurements and investigate how the ideally system differs from reality.
3.2.3 Layout and Measurements of the Prototype Feed System

In this section the procedure for implementing the calculated and simulated dimensions of the microstrip lines on the prototype substrate is described. Also measurements, examinations and conclusions of the system are made. Losses caused by the circulator and isolator are measured for the work frequency of 2.45 GHz of the system.

The distance from the open-circuited termination to the edge of the board is about three millimeters. Even the ground plane is etched away below these terminations to avoid as much of the edge effects as possible. Further the isolator is coupled direct, by soldering, from the circulator to the line that feeds the second port in the cavity. To avoid unnecessary reflections at the input, it will be connected direct with a calibrated coaxial cable from the network analyzer.

In the future microwave feed system there could be oscillators based on semiconductor technology. It is therefore very important that the reflection of the feed system is not too high. This is the main reason why the most important parameter here to measure is the reflection coefficient, or $S_{11}$, of the system. Worst conditions, as for example total reflection toward to the ports, will be investigated.

3.2.3.1 Practical Measurements

As was mentioned above the practical measurements are performed in the network analyzer HP8714B from Hewlett Packard. To get as correct validation of the function of the system as possible the first circulator is fed by a coaxial cable direct from the network analyzer, to avoid reflection at the input. The cable is soldered direct to the input of the system and is connected to the same ground as the other components. Before the measurements this cable is calibrated with known loads for the network analyzer.

Measurements that will be performed in this section are total reflection of the system, impedance magnitude measurements of the microstrip lines and the standing wave ratio (SWR) for some different conditions described below. The measurements are performed in the frequency range of 2.4 to 2.5 GHz and the prototype microwave feed system is placed with the component side against a plate made of a very porous material. This material is chosen because of its similar properties to air, which implies that there is in principle no reflection of the microwaves. This arrangement implies that the air slots radiate out to “free space” and is similar to the scenarios in the simulation program. In the program the ports are simulated to radiate in to absorbing walls, in which there is no reflection of the waves. Four different measurement configurations have been chosen to illustrate the function of the system. These are:

1. Both air slots of the system are opened to “free space”.
2. The first port is closed with a metal plate and the second port is opened to “free space”.
3. The first port is opened to “free space” and the second port is closed with a metal plate.
4. Both ports are closed with a metal plate.
First the reflection of the whole system is considered when both ports of the system is open. At 2.45 GHz, \( rZ_0 = 50.87 \, \Omega \) and \( jxZ_0 = -0.1746 \, \Omega \). This implies that the impedance magnitude will be, according to formula 3.2:

\[
|Z_{\text{in}}| = \sqrt{(r \cdot Z_0)^2 + (jx \cdot Z_0)^2} \approx 50.9 \, \Omega
\]

To illustrate how the absolute value of the input impedance magnitude varies in the chosen frequency interval the data from the network analyzer have been compiled into the diagram in figure 3.8.

![Impedance magnitude vs Frequency](image)

*Figure 3.8: Impedance magnitude as a function of frequency*

From this impedance magnitude the reflection coefficient could be calculated for the work frequency of 2.45 GHz, according to formula 8.16:

\[
S_{11} = \Gamma_{\text{in}} = \frac{Z_{\text{in}} - Z_0}{Z_{\text{in}} + Z_0} = 0.009
\]

In the above formula the input impedance of the system assumes to be 50 \( \Omega \), the input impedance of the circulator. The absolute value of the reflection is 0.009, which means that only about 0.9 percent of the energy is reflected. These measurements are performed with the microwave feed system placed with the component side at a plate made of a porous material with similar parameters to air, because of its low reflection (or high absorption). Further the air slots radiate to the surrounding space. The ways in which these measurements are performed are not ideally, but there will be comparable results in the different measurement configurations. Because of the above low value of the reflection the standing wave ratio (SWR) becomes:

\[
\text{SWR} = \frac{1 + |\Gamma_{\text{in}}|}{1 - |\Gamma_{\text{in}}|} \approx 1.02
\]
The SWR in the chosen frequency interval is illustrated in the diagram in figure 3.9.

![SWR vs Frequency](image1)

*Figure 3.9: SWR as a function of frequency with both air slots opened*

In the second measurement configuration the first port is covered with a metal plate. The reason for setting up this configuration is to simulate total reflection from the first sending port. This is an extreme case and could be compared to an empty cavity without glass plate and load. Since there cannot be too high reflection rate towards to the oscillator in a future system this measurement is performed and illustrated in the diagram in figure 3.10. As seen from the figure there will be values above 1.2 in this range, which is higher than the manufacturer of the circulator states. The circulator and isolator manufacturer RADITEK has stated a maximum SWR of 1.19 in the datasheets. The higher values in figure 3.10 depend on that the input signal to the system from the network analyzer is not ideal. A small reflection rate can not be avoided there and this rate is added to the result. A conclusion from this measurement is that the assumption of total reflection is correct and that the system could keep its function even for this worst case.

![SWR vs Frequency](image2)

*Figure 3.10: SWR as a function of frequency with the first air slot covered*

The higher values of the SWR in the frequency interval are explained by the fact that there will be a higher input impedance magnitude at the input. This is illustrated in figure 3.11.
From the diagram in figure 3.10 it could be seen that the SWR is 1.16 at the frequency 2.45 GHz. This in turn implies that the linear reflection magnitude is about 0.07, or seven percent is reflected. Since the network analyzer is connected to the input of the system this rate of reflection is the resulting field in the whole system. In other words it is the remaining 93 percent of the input signal that transfers to the second port in the system. A conclusion of this second measurement configuration is that the circulator is a limitation for the system, while it determines the reflection rate towards the generator.

In the third measurement configuration the first air slot will be open and the second covered with a metal plate. Since the air slots radiate to free space there will in principle be no difference between this case and the case when both ports are open. This is because there will in principle be no reflection towards the first sending port. In other words it does not matter if the second air slot is covered by metal. The result in figure 3.12 below could therefore be compared to the result in figure 3.9, which is in principle the same.

When both ports are covered with metal plates, the fourth measurement configuration, investigations of mismatch in the system could be performed. The system is designed on the basis of this condition, to get the right impedance of the lines. So, for this case there will ideally be zero reflection because of perfect matching. This is however not possible because of effects in the transitions, losses in the substrate and non-perfect open-circuited terminations. In the
measurements by the network analyzer these effects are taken into consideration. The rate of reflection coefficient in the used frequency interval gives a legible picture of the rate of mismatch in the system. The reflection coefficient is illustrated in figure 3.13 below.

![Reflection vs Frequency](image)

*Figure 3.13: Linear reflection as a function of frequency with both ports covered by metal*

From the above figure it could be seen that the reflection is about 3.5 percent at the working frequency of 2.45 GHz. The impedance magnitude is measured to be about 53.6 Ω at that point.

The conclusions from the above presented measurements are that the microwave feed system functions as it should and that the system is well matched for the working frequency. The function is especially verified in the measurements with the first port covered by metal, since there is about 93 percent of the reflected energy that is transferred to the next port. This in turn would imply a more efficient system than today. A limiting factor in the system is, however, the SWR for the input circulator in the system.

### 3.3 Microwave Feed System for Higher Power

In R. Meredith’s book “Engineers’ handbook of industrial microwave heating” there are suitable substrate materials for high power applications listed. Among these materials are PTFE (Teflon®) and Aluminum. The aluminum substrate can for example carry a power of about $12 \, kW$ at $2 \, GHz$ on a $50-\Omega$ microstrip line, if a temperature rise of $75^\circ C$ above ambient is allowed [1]. However, for this work a design on a PTFE substrate material from ROGERS will be presented. With the dimensions chosen for this work such a system handles about 400 watts. The requirement is about 200 watts for the system, so that material seems to be suitable.

#### 3.3.1 Dielectric Materials

When designing a system on a substrate material one have to take its different properties into consideration. These could for example be power handling capability (section 2.2.4.2), dielectric heating etc.

The heating mechanism, caused by a microwave field, in a dielectric substrate material is primary described by two phenomena. In the first one the dielectric material seems to behave like a poor electrical conductor, with a finite resistivity. This resistivity is measurable at DC and becomes usually constant in the microwave region. In the second phenomena there is a
mechanical explanation. In many dielectric materials the dipolar components of molecules couple electrostatically to the microwave electric field. These components tend to align themselves with this field mechanically and since this microwave field is alternating in time, the dipoles attempts to realign when the field reverses. In this way a constant mechanical oscillation occur at the frequency, at which the microwaves works. This phenomenon will cause heat development due to the frictional forces, within the molecules, that occur. This effect increases with increased frequency because of that the amount of energy dissipated is constant per cycle of applied alternating field. In practice this is not entirely true because it assumes that the mechanical displacement of the dipoles remains constant with different frequencies and this is not the case [11].

For the purposes of this project a dielectric material which does not heat significantly in a microwave field have to be chosen. Examples of such materials, and the most common used in microwave applications, are insulators like PTFE, quartz, aluminum, polycarbonate, polypropylene and polyethylene [11].

### 3.3.1.1 Choice of Material

The first step done to find out which material that could be used in the design is a literature study. One can then get an apprehension of which materials that are suitable. Examples are the materials, mentioned in the above section.

The simulated transmission line has a characteristic impedance of $49.9 \, \Omega$, a height of $3 \, mm$ and the line width is chosen to be $7 \, mm$. If a temperature rise of $75 \, ^\circ C$ over ambient is allowed for this design, it can hold a power of about $400 \, \text{watts}$. This power capability is enough for the application, since a power of about $200 \, \text{watts}$ is required. Data for the PTFE substrate, RO4003 from ROGERS, are listed in table 3.3 [19].

<table>
<thead>
<tr>
<th>Properties</th>
<th>Remark</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dielectric constant, $\varepsilon$</td>
<td>@ 10 GHz / 23°C</td>
<td>$3.38 \pm 0.05$</td>
</tr>
<tr>
<td>Dissipation factor, $\tan \delta$</td>
<td>@ 10 GHz / 23°C</td>
<td>0.0027</td>
</tr>
<tr>
<td>Dielectric conductivity, $\sigma$</td>
<td>Estimated @ 2.45 GHz</td>
<td>0.00124 S/m</td>
</tr>
<tr>
<td>Dielectric breakdown</td>
<td>Estimated</td>
<td>40 kV</td>
</tr>
<tr>
<td>Thermal conductivity, $K$</td>
<td>@ 2.45 GHz</td>
<td>0.64 W/m/K</td>
</tr>
<tr>
<td>Dielectric attenuation, $\alpha_d$</td>
<td>@ 2.45 GHz</td>
<td>0.89 dB/m</td>
</tr>
<tr>
<td>Conductor attenuation, $\alpha_c$</td>
<td>@ 2.45 GHz</td>
<td>0.771 dB/m</td>
</tr>
</tbody>
</table>

Table 3.3: Data for the substrate RO4003 from ROGERS Corporation [19]

In the program, Quick-Wave 3D, materials and its media parameters could be defined by the user. The parameters that are useable are the dielectric constant, $\varepsilon_r$, and the conductivity, $\sigma$. When setting the dielectric constant parameter, the real part of it is used. The definition of the dielectric constant is

$$\varepsilon^* = \varepsilon' - j\varepsilon''$$  \hspace{1cm} (3.3)

where $\varepsilon'$ is the real and $\varepsilon''$ the imaginary part. The imaginary part $\varepsilon''$ is called loss factor.
Another term that relates to the losses of a dielectric is the loss angle. It describes the angle by which the resultant current differs from the ideal 90°-phase angle relative to the voltage. The loss angle could be written \[ \delta = \tan^{-1} \frac{\varepsilon''}{\varepsilon'} \] (3.4)

We can also define the more common used loss tangent as \[ \tan \delta = \frac{\varepsilon''}{\varepsilon'} \] (3.5)

In datasheets from manufacturers it is usually only the loss tangent and the real part of the dielectric constant, which are given. According to formula 3.5 the imaginary part of the dielectric constant could be calculated. If we know the dielectric constant and the loss tangent, we can get the conductivity in the dielectric material by combining formula 3.5 with formula 3.6 [12].

\[
\sigma = \omega \varepsilon_0 \varepsilon_r = 2\pi f \varepsilon_0 \varepsilon'' = 2\pi f \varepsilon_0 \varepsilon' \tan \delta
\] (3.6)

This formula is very useful while both the dielectric constant and the loss tangent are given in most datasheets for materials. For the calculations of the conductivity in this report, formula 3.6 is used. These conductivity parameters are also used in the simulation program when defining the media parameters for different materials.

### 3.3.2 Suitable Dimensions for the Feed System

The structure of the proposal of a system for the required power is the same as for the prototype system described in the previous section. It is the dimensions of the substrate and the microstrip line that differs. The procedure for calculating the parameters is the same and the first step is to find out the relationship \( w/h \) for the line and the substrate. By using figure 2.12 in section 2.2.2.3 it could be seen that \( w/h \) gets a value bigger than two for a dielectric constant of 3.38. To get a more exact value of \( w/h \) formula 2.26 could be used:

\[
\frac{w}{h} = \frac{2}{\pi} \left[ B - 1 - \ln(2B - 1) + \frac{\varepsilon_r - 1}{2\varepsilon_r} \left( \ln(B - 1) + 0.39 - \frac{0.61}{\varepsilon_r} \right) \right] \approx 2.31
\]

where the factor B is given by

\[
B = \frac{Z_f \pi}{2Z_0 \sqrt{\varepsilon_r}} = 6.4388
\]
The previous calculations give us a relationship between the width of the line and the height of the substrate, \( w = 2.31h \). This would imply a line width of about 7 mm. To check if this value seems to be correct, formula 2.21 and 2.22 for the calculation of \( Z_0 \) and \( \varepsilon_{\text{eff}} \) is used:

\[
Z_0 = \frac{Z_f}{\sqrt{\varepsilon_{\text{eff}} \left( 1.393 + \frac{w}{h} + \frac{2}{3} \ln \left( \frac{w}{h} + 1.444 \right) \right)}} = 50.0 \Omega
\]

where

\[
\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + \frac{12h}{w} \right)^{-1/2} \approx 2.67.
\]

The wavelength in this substrate material from ROGERS could be found from

\[
\lambda = \frac{c}{f \sqrt{\varepsilon_{\text{eff}}}} \approx \frac{3 \times 10^8}{2.445 \times 10^9 \sqrt{2.67}} \approx 75.09 \text{ mm}.
\]

How the cavity is fed is illustrated earlier in figure 3.2. Since the waves propagate with a smaller wavelength in this material there will be other values for the dimension \( l \) and \( d \). In this case the distance \( l \) will be 18.77 mm to get a maximum magnetic field at the center of the air slot. The simulation result in figure 3.14 verifies these calculations. In the figure the field distribution in the level between the ground plane and the strip, in the substrate, is illustrated.
The lengths of the microstrip lines also have to be changed because of the smaller wavelengths. To get a correct input impedance following distances could be used according to the formulas for the prototype system in section 3.2.1:

<table>
<thead>
<tr>
<th>n</th>
<th>d [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>-9.39</td>
</tr>
<tr>
<td>1</td>
<td>28.16</td>
</tr>
<tr>
<td>2</td>
<td>65.70</td>
</tr>
<tr>
<td>3</td>
<td>103.2</td>
</tr>
</tbody>
</table>

Table 3.4: Values of the line length \( d \) as a function of \( n \).

Suitable distance for the microstrip line from the circulator to the first port in the cavity is \( 28.16 \text{ mm} \) for this material. For the bended line connecting the output of the isolator to the second port in the cavity the distance \( 65.70 \text{ mm} \) is suitable. For the outline of the system the reader is referred to figure 3.5 in the previous section.

### 3.3.3 Power Handling

In this section approximate calculations are made of how much power the microwave feed system handles. The formulas for this purpose are presented in section 2.2.4.2.1.

If a temperature rise in the substrate of 75°C assumes to be allowed it could handle a power of about 400 watts according to the calculations below. The parameter values for RO4003 used in the formulas below are listed in table 3.3.

\[
\Delta T = \frac{0.2303h}{K} \left( \frac{\alpha_c}{w_{eff}} + \frac{\alpha_d}{2w_{eff}}(f) \right) \approx \frac{0.2303 \cdot 3 \cdot 10^{-3}}{0.64} \left( \frac{0.771}{7 \cdot 10^{-3}} + \frac{0.89}{14 \cdot 10^{-3}} \right) \approx 0.188
\]

\[
P_{mo} = \frac{\Theta}{\Delta T} \approx \frac{75}{0.18753} \approx 400W
\]

### 3.4 Simulated Discontinuities

The behavior of microwaves through discontinuities of conducting strips could quantitative be considered by computer simulations. In this section the scattering parameters for a bended strip, so-called Chamfered Bend, are determined. This kind of 90°-bend is used in this project.

#### 3.4.1 The Chamfered Bend

In the investigated microwave feed system bended microstrip lines can not be avoided, while the ports are placed orthogonal to each other. In this section simulations will be done to illustrate the difference between the Chamfered Bend and the non-compensated bend. The method used is to measure the reflection coefficient for the two different configurations on the same substrate material and with the same dimensions on the microstrip line.

The first simulation result, figure 3.15, shows the reflection coefficient \( S_{11} \) for a Chamfered Bend in Quick-Wave 3D. The substrate material used is PTFE, with a dielectric constant of 2.06 and a conductivity \( \sigma \) of \( 4.2253 \cdot 10^3 \text{ S/m} \). The strip has a width of 10 mm and zero thickness is assumed. Further the strip is supposed to be a perfect electrical conductor, so the substrate material occurs the only losses. This assumption is reasonable, while the conductivity in the metal is much better.
As a comparison to the chamfered bend above there is a non-compensated 90°-bend illustrated in figure 3.16.

As seen from the above figure there is approximately twice as much reflection for a non-compensated bend. The capacitance increases for such a bend and the inductance in the equivalent circuit in figure 2.14 b arise because of the current flow interruption. These effects are though minimized in figure 3.15, where the dimensions are chosen as in section 2.2.3.1, namely $b \approx 0.57 \cdot w$, according to figure 2.14 a.

To measure the scattering parameters of the chamfered bend the ports are chosen to a pulse of spectrum $2 \text{ GHz} < f < 3 \text{ GHz}$, while the characteristics at $2.43 \text{ GHz}$ are interesting. The most interesting parameter to measure is $S_{11}$, or the reflection coefficient, to get an idea of how much of the input that is reflected. The software measures the scattering parameters according to Fourier transforms. This implies that the power balance have to be close to unity for the
system to get a correct result. The conclusions from the simulations are that there are better transfer characteristics for the Chamfered Bend than for a non-compensated bend. Therefore this kind of bend is used in this project.

3.5 Circulators

The theory for circulators, both waveguide and in microstrip and stripline techniques was treated in section 2.2.7. In this section design aspects for circulators made in microstrip and stripline techniques are treated. The design is shown to be rather complex why there are commercial circulators used in the prototype microwave feed system.

3.5.1 Microstrip and Stripline Techniques

As was mentioned in section 2.2.7 the three-port circulator is built up by three symmetrical coupled transmission lines. In the region where these lines are brought together there is a magnetized ferrite. In a similar way a circulator in microstrip or stripline technique is built. A schematic figure is shown below, where a circular disc on a ferrite material, which in turn is mounted on a ground plane, builds the microstrip design. According to the figure the stripline design is built in a similar way.

![Microstrip and Stripline Circulator Designs](image)

*Figure 3.17: Circulator designs in microstrip and stripline techniques*

The transmission ways are from the first to the second port with the third port isolated, from the second to the third port with the first port isolated and from the third to the first port with the second port isolated. To obtain this proper function the resonator must be magnetized, which results in different propagation constants and a rotation of the standing wave pattern (where port two and three have same field amplitude and port one has maximum field amplitude). This rotation shall continue until there is one port that is isolated, which means zero electrical fields in the z-direction, and this results in that the two other ports will have the same field amplitude. The rotation is shown to be 30° to get the electrical field zero at port three (zero voltage), which implies that there is no magnetic field lines that could induce current to port three. This 30° rotation is achieved if the impedance of each mode has an angle of ± 30°. Figure 3.12 illustrates the state when port one and two have equal amplitude and port three is isolated [7, 13, 18].
The rotation of the standing wave pattern could be achieved in two possible ways. The first possibility is to have a DC magnetic field that is lower than that of the material resonance. This state is sometimes referred to as below resonance. For this state the value of the effective $\mu$ is always lower than unity. The second possibility is when the field is rotated by a magnetic field that is higher than the material resonance, referred to as above resonance state. For this state the value of the effective $\mu$ is always greater than one. The sign of the rotation, or splitting, is different for the two possibilities. A result from a magnetic field below the working resonance is that it requires a bigger disc. Above resonance is often suitable because of that low field loss could occur in the one working below resonance [13].

To describe the magnetization mathematically, and get closer to a model for the junction, there are two methods, treated by the authors of [13], namely the singlemode and multimode theory. The singlemode theory though has its drawbacks because of the assumption that the rotation is comparable small. This means that broadband circulators are not very well described by this singlemode theory, while they need a higher degree of the rotation and also wider strips. Another reason for using the multimode model is to get a more correct description of the conditions at the boundary. There will also be a better explanation of the rotation of the field pattern [13].

In this multimode method the input wave impedance is calculated according to Green’s function under the assumption that the field distribution is

\[
H_\phi = \begin{cases} 
  a & -\psi < \phi < \psi \\
  b & \frac{2\pi}{3} - \psi < \phi < \psi + \frac{2\pi}{3} \\
  c & \frac{2\pi}{3} - \psi < \phi < \psi - \frac{2\pi}{3} \\
  0 & \text{elsewhere}
\end{cases}
\]  

(3.7)

around the edge and $a$, $b$ and $c$ are constants. For the coordinates used the reader is referred to figure 2.18 in section 2.2.7. The signification of the intervals in formula 3.7 above is that there

\[\text{Port 1}\]

\[2\pi/3\]

\[4\pi/3\]

\[\Phi\]

\[\text{Port 2}\]

\[\text{Port 3}\]

\[E_z\]
is a constant field at the ports and zero fields elsewhere. We can then get the wave input impedance according to
\[
Z_{in} = -Z_d - \left(\frac{j \cdot 2Z_{eff}}{\pi} \right) \left[ \frac{C_1^3 + C_2^3 + C_3^3 - 3C_1C_2C_3}{C_1^2 - C_2C_3} \right] \]  
(3.8)

where \( Z_d \) and \( Z_{eff} \) are the intrinsic impedance of the surrounding material and of the ferrite. The constants \( C_1 = P + jQ, C_2 = M + jN \) and \( C_3 = M - jN = C_2^* \), where
\[
P = \frac{\psi B_0}{2A_0} + \sum_{n=1}^{\infty} \left( \frac{\sin^2(n\psi)}{n^2\psi} \right) \cdot \frac{A_nB_n}{D_n},
\]
\[
Q = \frac{\pi Z_{d}}{2Z_{eff}},
\]
\[
M = \frac{\psi B_0}{2A_0} + \sum_{n=1}^{\infty} \left( \frac{\sin^2(n\psi)}{n^2\psi} \right) \cdot \frac{A_nB_n}{D_n} \cdot \cos \left( \frac{2n\pi}{3} \right),
\]
\[
N = -\sum_{n=1}^{\infty} \left( \frac{\sin^2(n\psi)}{n^2\psi} \right) \cdot \frac{(n\kappa/\mu sR)B_n^2}{D_n} \cdot \sin \left( \frac{2n\pi}{3} \right),
\]
\[
A_n = J_n(sR), \quad B_n = J_n(sR) \quad \text{and} \quad D_n = A_n^2 - \left( \frac{n\kappa}{\mu sR} \right)^2 \cdot B_n^2 \quad [13].
\]

Finally we can now use the formulas stated in this and section 2.2.7 to determine the design parameters, \( R, \kappa/\mu, w \) and \( h \), for a circulator according to the chosen model. The parameters are chosen to match the surrounding network with the junction, so that \( Z_{in} = Z_{dh} \), where \( Z_{in} \) is a function of \( sR, \Psi \) and \( \kappa/\mu \). The magnetization is appointed by \( \kappa/\mu \) and with
\[
s = \frac{\sigma}{\varepsilon_f} \cdot \sqrt{\mu_{eff}} \cdot \varepsilon_f \quad \text{and while R is known we get w = 2R sin \Psi \quad [13].}
\]

The impedance level of the surrounding network, \( R_j \), is dependent of \( w, h \) and \( Z_r \) according to
\[
R_j = R \left( \frac{w}{h}, \varepsilon(Z_r) \right) \]  
(3.9)

where \( Z_r \) is the real part of the input wave impedance \( Z_{in} = Z_r + jZ_i \). Since we know the width of the junction, we can choose the height of the ferrite in such a way that implies impedance matching between the junction and the surrounding network [13].

The above method describes the design steps for circulators according to the multimode model. Many of the calculations in this method are though very complex, which implies that a computer program that performs the calculations is preferred. The software used in this work could not be used for these purposes, while it does not support non-diagonal matrices. An example here is the permeability matrix.
A limitation of the multimode method is that it is an idealized model and the real field distribution is much smoother than that used in the method. The rotation used in the method can not be explained with just one mode, which is the case in the idealized model. It is possible to describe the boundary conditions in a more exact way with Green’s functions, but this gets much more complicated. A use of computer programs makes it though possible [13].

3.5.1.1 Losses in Y-junction Stripline and Microstrip Ferrite Circulators

Neidert and Phillips [16] have investigated the losses that occur in the Y-junction stripline and microstrip ferrite circulators and made a model that describe these. In this section there will be an overview of these conclusions and a presentation of critical parts in the components. Losses that will be treated are dielectric, conductor and magnetic.

The dielectric losses that occur in the circulator could be described by adding the complex dielectric constant to the ideally scattering matrix. The complex dielectric constant was introduced earlier in the report, in section 3.3.1.1. In this way the lossless theory is modified to take the dielectric losses into account. In the ideally theory for circulator design presented in this work it is assumed that there is no variation of the electric field in the z-direction. This is though not really true while a small tangential component of the electric field exists. This additional field comes up if the conductor in the circulator is not perfect. Another effect due to a non-perfect conductor is that there also will be a small normal component of the magnetic field. The idea with a new model is that a resulting electric field is calculated where these effects are taken into consideration, with the assumption that there is no variation in z-direction. In a similar way as the dielectric losses the magnetic losses are described by introducing complex quantities in formula 2.41. Generally the conductor losses are small compared with dielectric and magnetic losses. But the conductor losses increase with decreasing ferrite thickness, dependent of the increasing electric field component [16, 17].

Losses that occur in the circulator and isolator on the microwave feed system are very difficult to determine in practice. However, it is important to have these losses in mind when designing these components.
4 DISCUSSION

This work is a part of a bigger project, which is performed at Whirlpool, about microwave feed systems and new oscillators according to their patent. For a final evaluation of the whole system there are too many unknown parameters so far. There are for example unknown parameters for the generators such as losses and efficiency and losses in the surrounding networks. The matching network for the output signal, which will serve as input signal to the microstrip microwave feed system presented in this report, probably implies reflection in the transition caused by mismatch. So far it is also unknown how much reflected energy the future generators handle.

However, in this work the microwave feed system is investigated. The main reason that contradicts the use of a microwave feed system manufactured in microstrip or stripline configuration is that the cost is too high. The investigated material in this report could be used for the system that handles the required power is RO4003 from ROGERS. A panel of this material costs about 100 $ and about six systems could be made per panel, depending of chosen board size. This price concerns an order of up to nine panels, after which the price successively decrease. Further circulators and isolators constitute the system. These components are manufactured by RADITEK and are rather expensive, about 60 $ for the isolator and 50 $ for the circulator. Another disadvantage that pertains to these components is that it is difficult to find such that handle high power. The circulator used in the prototype system handles about 200 watts and the isolator only about 110 watts. This implies that they become unusable for the final system, in which at least 200 watts will be transferred. So the remaining task is to find suitable circulators and isolators that can handle the required power. An aspect to be considered here is to investigate if the use of a four-port circulator with a designed load at the fourth port is possible. Then the cost for the isolator would be unnecessary, but it is difficult to find manufacturers of four-port circulators.

The main reason for choosing the microstrip configuration for the presented prototype system is the possibility to be able to surface mount components like circulators and isolators. If a stripline configuration was chosen there must be holes to the conductor, so the components could be hole mounted. Another reason is that it is easier to perform measurements on the microstrip configuration.

If the designed system, with substrate and components as described in the report, is studied the total cost would be estimated to about 130 $ per system. This cost does not include the shielding of the system. This indicates that the system is too expensive for an application like this. The cost only concerns one system and, as stated before, there must be several such systems to reach the required power for a microwave oven. It is, however, recommended to watch the market for decreasing prices of the included components and even order a quotation of bigger volumes. Even if the costs will decrease a bit the question if the customers are ready to pay for the new technology still remains. Advantages of a system that has many feed points are for example that the rotating glass plate could be avoided with the same heating efficiency. Since there are components in the project at Whirlpool that have a long development time, it is recommended to investigate the possibilities to design circulators at the company during this period. This would presumably imply a cheaper system.
The function of the microwave feed system is evaluated by measurements in a network analyzer and resulted in proper function and low losses. These measurements were though performed into “free space” and for a complete investigation of its function and loss, measurements must be performed on a system in microwave ovens. First then a final evaluation of the function could be performed. However according to the measurements in the network analyzer it seems to function, as it should. A disadvantage is though that the efficiency of the system is rather difficult to determine, since transmission measurements are very unreliable in the network analyzer.

Concerning the used microstrip configuration for the prototype system it has disadvantages like radiation to the surroundings. In the report there are proposals of ways presented for shielding this radiation. The cause to use and describe the design procedure of microstrip solutions is that there are surface mounted components included on the system. It is very difficult to make a prototype in stripline configuration. There are however advantages of using stripline solutions, as lesser radiation and easier to control the line impedance. However, the last-mentioned property does not constitute a decisive part for choosing stripline configuration, while the impedance of the lines is varying all the time. Another important parameter that has to be taken into consideration is that the stripline configuration generally can handle lower power. As a comparison the used material RO4003 handles about 400 watts in microstrip configuration and only about 150 watts in stripline configuration. This depends of the better heat conductivity to the surroundings for the microstrip. It is the heat transfer through the substrates that limits the power. In other words a simpler and cheaper material could be used in the microstrip configuration.

If it is assumed that the final design of the microwave feed system will be designed on the RO4003 material, it seems to be the microstrip solution that is most suitable for the required power. The dilemma of shielding, or encapsulation, must though be taken into consideration. The design of a shielding system would more or less be similar to the structure of the traditional used waveguides. It is no coincidence that the dimensions of the encapsulation must be about the same as the dimension of the waveguides, for correct function of the wave propagation. Then the meaning of implementing a feed system in microstrip solutions does not seem so obvious. On the other hand, a system with the function according to the described patent will be very difficult to manufacture in waveguide technique, because of the dimensions. A system on microstrip is more flexible concerning for example bends. The encapsulation could always have a rectangular and simple design. Other important parameters to investigate are the conductor and dielectric losses. For the proposed final system these parameters are stated.
To sum up the discussion topics in this part for obtaining a better overview, important considerations are listed below.

- Suitable material costs about 20 $ per feed system.
- The included circulator and isolator cost about 50 $ and 60 $ respectively.
- A complete system with components and substrate suitable for the required power will presumable become too expensive.

The summed up recommendations for further work are also listed below.

- Watch the market for decreasing prices and ask for a quotation concerning a bigger order.
- Search for circulators and isolators that can handle required power and investigate the possibilities to design these at Whirlpool. An investigation of four-port circulators would also be preferred. This could maybe be the subject to an additional diploma work.
- Perform measurements on the system in microwave ovens with conventional methods for a complete evaluation.
- Evaluate the needed rate of metal shielding and the cost for these purposes.
- Investigate the losses more carefully and compare to the today’s systems.
5 CONCLUSIONS

The work presented in this report has resulted in a proposal of a microwave feed system manufactured in microstrip technique, based on Whirlpool’s patent. Proposals of suitable materials intended for high power and the used frequency are also made. Tests are performed of the prototype system and it is possible that the microwave feed systems could be designed according to the results of this report in the future.

An aspect is that the prototype system described in the report only is evaluated by measurements in a network analyzer. For a complete investigation of these innovative solutions intended for microwave feed systems the author recommends measurements and tests of the systems in microwave ovens to be able to perform comparable conclusions to the today’s systems. Such tests could for example be measurements of heating in a microwave oven and measurements of leakage compared to traditional systems. Not until such investigations are performed the efficiency of the final system could be evaluated. However, by the measurements and simulations performed during this project it could be assumed that the efficiency of the system would increase since the reflected energy transfers via the second port to the cavity again. In the today’s systems all the reflected energy transfers back to the generator, where it is chilled away.

If a microwave feed system is designed in the future the author recommends the microstrip solution and the material for required power described in this report. Today there is however a main disadvantage for a system like this, while the total cost of the system seems to be too high compared to the today’s systems.
6 FIGURES

FIGURE 2.1: THE OUTWARD FLUX OF $\mathbf{E}$ THROUGH A CLOSED SURFACE $S$ IS EQUAL TO $Q_i/\varepsilon_0$ [9].

FIGURE 2.2: THE NET OUTGOING FLUX OF $\mathbf{B}$ THROUGH ANY CLOSED SURFACE $S$ IS ZERO [9].

FIGURE 2.3: PATH OF INTEGRATION [2].

FIGURE 2.4: VOLTAGE INDUCTION FROM MAGNETIC FLUX [2].

FIGURE 2.5: ONE CELL OF A TYPICAL MESH USED IN THE FDTD METHOD [10].

FIGURE 2.6: MICROSTRIP TRANSMISSION LINE [1].

FIGURE 2.7: ELECTRIC FIELD LEAKAGE AS A FUNCTION OF DIELECTRIC CONSTANTS [2].

FIGURE 2.8: EMBEDDED TRANSMISSION LINE TO REDUCE RADIATION LOSSES [2].

FIGURE 2.9: PARALLEL-PLATE TRANSMISSION LINE [2].

FIGURE 2.10: ELECTROMAGNETIC WAVE PROPAGATION IN FREE SPACE [2].

FIGURE 2.11: PROCEDURE FOR THE QUASI-TEM APPROXIMATION [3].

FIGURE 2.12: CHARACTERISTIC LINE IMPEDANCE AS A FUNCTION OF $W/H$ [2].

FIGURE 2.13: EFFECTIVE DIELECTRIC CONSTANT AS A FUNCTION OF $W/H$ FOR DIFFERENT DIELECTRIC CONSTANTS [2].

FIGURE 2.14: A BENDED MICROSTRIP LINE AND ITS EQUIVALENT CIRCUIT [4].

FIGURE 2.15: COAXIAL-TO MICROSTRIP TRANSITION [1].

FIGURE 2.16: EQUIVALENT CIRCUIT FOR A COAXIAL-TO-MICROSTRIP TRANSITION [1].

FIGURE 2.17: PRINCIPLE FOR A THREE-PORT JUNCTION CIRCULATOR [7].

FIGURE 2.18: COORDINATES USED IN THE CALCULATIONS [13].

FIGURE 3.1: DIMENSION OF THE SUBSTRATE BOARD AT THE CAVITY ROOF WITH AIR SLOTS.

TABLE 3.1: PROPERTIES OF THE PTFE-SUBSTRATE FROM ELFA.

FIGURE 3.2: PRINCIPLE OF FEEDING.

FIGURE 3.3: THE MAGNETIC FIELD DISTRIBUTION AT THE SAME LEVEL AS THE MICROSTRIP LINE.

FIGURE 3.4: MAGNETIC FIELD ENVELOPED JUST BELOW THE AIR SLOT.

TABLE 3.2: VALUES OF THE LINE LENGTH $D$ AS A FUNCTION OF $N$.

FIGURE 3.5: OUTLINE OF THE PROTOTYPE SYSTEM.

FIGURE 3.6: $S_{11}$ AS A FUNCTION OF FREQUENCY FOR CHOOSEN LOADS IN THE CAVITY.

FIGURE 3.7: $S_{11}$ PARAMETER FOR A CHAMFERED BEND ON THE PROTOTYPE SUBSTRATE.

FIGURE 3.8: IMPEDANCE MAGNITUDE AS A FUNCTION OF FREQUENCY.

FIGURE 3.9: SWR AS A FUNCTION OF FREQUENCY WITH BOTH AIR SLOTS OPENED.

FIGURE 3.10: SWR AS A FUNCTION OF FREQUENCY WITH THE FIRST AIR SLOT COVERED.

FIGURE 3.11: IMPEDANCE MAGNITUDE AS A FUNCTION OF FREQUENCY WITH THE FIRST AIR SLOT COVERED.

FIGURE 3.12: SWR AS A FUNCTION OF FREQUENCY WITH THE SECOND AIR SLOT COVERED.

FIGURE 3.13: LINEAR REFLECTION AS A FUNCTION OF FREQUENCY WITH BOTH PORTS COVERED BY METAL.

TABLE 3.3: DATA FOR THE SUBSTRATE RO4003 FROM ROGERS CORPORATION [19].

FIGURE 3.14: THE MAGNETIC FIELD DISTRIBUTION IN THE SUBSTRATE AT THE AIR SLOT.

TABLE 3.4: VALUES OF THE LINE LENGTH $D$ AS A FUNCTION OF $N$.

FIGURE 3.15: A CHAMFERED BEND SIMULATED IN QWED.

FIGURE 3.16: A NON-COMPENSATED 90-DEGREE BEND.

FIGURE 3.17: CIRCULATOR DESIGNS IN MICROSTRIP AND STRIPLINE TECHNIQUES.

FIGURE 3.18: THE $E_z$ FIELD AT CIRCULATION.

FIGURE 8.1: REPRESENTATION OF R AND X CIRCLES IN THE SMITH CHART [2].

FIGURE 8.2: INCIDENT AND REFLECTED POWER WAVES DEFINING THE S-PARAMETERS.

FIGURE 8.3: MEASUREMENT OF $S_{11}$ AND $S_{21}$ IN A MATCHED SYSTEM.
7 REFERENCES


[10] QuickWave-3D v.1.9 Electronic Manual, QWED Poland


8 APPENDIX

8.1 The Smith Chart

The Smith Chart is a commonly used graphical design tool for RF applications. In the diagram one can see how the impedance behavior of a transmission line varies as a function of line length or of the frequency.

One of the key ingredients for the Smith Chart as a graphical tool is the transformation from the reflection coefficient \( \Gamma_0 \) to \( \Gamma(d) \). \( \Gamma(d) \), which is the reflection coefficient as a function of line length, can be written as

\[
\Gamma(d) = \Gamma_0 e^{-\gamma dl}
\]

and this expression is valid anywhere along the length \( d \) of the line [2].

The complex impedance \( z_{in} \) has its analogy in another complex plane, the \( \Gamma \)-plane, according to the expression in the formula below [2].

\[
z_{in} = r + jx = \frac{1+\Gamma(d)}{1-\Gamma(d)} = \frac{1+\Gamma_r + j\Gamma_i}{1-\Gamma_r - j\Gamma_i} = \frac{1-\Gamma_r^2 - \Gamma_i^2 + 2j\Gamma_i}{(1-\Gamma_r)^2 + \Gamma_i^2}
\]

The last expression in formula 8.2 results from multiplying the complex conjugate of the denominator with numerator and denominator. The expression for \( z_{in} \) can be separated into

\[
r = \frac{1-\Gamma_r^2 - \Gamma_i^2}{(1-\Gamma_r)^2 + \Gamma_i^2}
\]

and

\[
x = \frac{2\Gamma_i}{(1-\Gamma_r)^2 + \Gamma_i^2}
\]

By these equations one will finally obtain expressions for the normalized \( r \) and \( x \) that describe equations for circles in the \( \Gamma \)-plane. For \( r \) this expression becomes

\[
\left( \frac{\Gamma_r - \frac{r}{r+1}}{r+1} \right)^2 + \Gamma_i^2 = \left( \frac{1}{r+1} \right)^2
\]

and for \( x \) we get [2]

\[
(\Gamma_r - 1)^2 + \left( \frac{1}{x} - \frac{1}{x} \right)^2 = \left( \frac{1}{x} \right)^2
\]
We can then map the different values for $r$ and $x$ to circles in the complex plane, in terms of real and imaginary parts of the reflection coefficients. The mapping in the Smith Chart is one-to-one between the normalized impedance plane and the reflection coefficient plane. To get the final graphical representation in the Smith Chart the parametric representations for normalized resistance and reactance circles are combined. This is illustrated in figure 8.1 [2].

![Smith Chart Diagram](image)

*Figure 8.1: Representation of $r$ and $x$ circles in the Smith Chart [2]*

For the computation of the input impedance of a terminated transmission line, the motion is always *away from the load impedance or toward the generator*. This motion is illustrated in figure 8.1 with an arrow. If making a complete revolution around the unit circle one get

$$2\beta d = 2\frac{2\pi}{\lambda} d = 2\pi$$  \hspace{1cm} (8.7)

where $d = \lambda/2$ or $180^\circ$. Sometimes the quantity $\beta d$ is referred to as the electrical length of the line [2].

In the Smith Chart we can also see the degree of mismatch according to the standing wave ratio (SWR) equation [2]

$$\text{SWR}(d) = \frac{1 + |\Gamma(d)|}{1 - |\Gamma(d)|}$$  \hspace{1cm} (8.8)

We can get the value of $\Gamma(d)$ directly in the Smith Chart and therefore we can get the ratio of mismatch by inserting the value of $\Gamma(d)$ in formula 8.8. We often want to have a system free from reflection, $\Gamma(d) = 0$, which is corresponding to $\text{SWR}(d) = 1$.  

55
8.2 Scattering Parameters

When dealing with RF circuits, the Scattering parameters, or S-parameters, are very important. The S-parameters are a tool to characterize the two-port network description of practically all RF devices without requiring unachievable terminal conditions, as for example short circuit or open circuit conditions [2].

The S-parameters describe the input-output relations of a network. This is in terms of incident and reflected power waves. According to figure 8.2 an incident normalized power wave $a_n$ and a reflected normalized power wave $b_n$ can be defined as:

$$a_n = \frac{1}{2\sqrt{Z_0}}(V_n + Z_0 I_n) \quad (8.9)$$

$$b_n = \frac{1}{2\sqrt{Z_0}}(V_n - Z_0 I_n) \quad (8.10)$$

where the index $n$ refers to port 1 or 2. $Z_0$ is the characteristic impedance for the lines of the input and output side of the network.

After several calculation steps, and according to figure 8.2, the S-parameters can be defined as

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \quad (8.11)$$

where

$$S_{11} = \frac{b_1}{a_1}_{a_2=0} \equiv \text{reflected power wave at port 1} \quad \text{incident power wave at port 1} \quad (8.12)$$

$$S_{21} = \frac{b_2}{a_1}_{a_2=0} \equiv \text{transmitted power wave at port 2} \quad \text{incident power wave at port 1} \quad (8.13)$$
Microstrip Solutions for Innovative Microwave Feed Systems

\[ S_{21} = \begin{bmatrix} b_2 \\ a_2 \end{bmatrix}_{a_2=0} \quad \text{reflected power wave at port 2} \]
\[ \begin{bmatrix} b_2 \\ a_2 \end{bmatrix}_{a_2=0} \quad \text{incident power wave at port 2} \]  

(8.14)

\[ S_{12} = \begin{bmatrix} b_1 \\ a_1 \end{bmatrix}_{a_1=0} \quad \text{transmitted power wave at port 1} \]
\[ \begin{bmatrix} b_1 \\ a_1 \end{bmatrix}_{a_1=0} \quad \text{incident power wave at port 2} \]  

(8.15)

In the expressions for the S-parameters in the formulas above the conditions \( a_2 = 0 \) and \( a_1 = 0 \) imply that no power waves are returned to the network at either port two or port one [2].

The S-parameters can only be determined under conditions of perfect matching on the input or output side. If we for example want to record \( S_{11} \) and \( S_{21} \) we have to get the impedance \( Z_0 \) matched for \( a_2 = 0 \) on the output side. This means that we have to choose the load impedance equal to the characteristic impedance and is illustrated in figure 8.2 [2].

According to figure 8.3 we can compute \( S_{11} \) if we know the input impedance and the characteristic impedance of the system as

\[ S_{11} = \Gamma_m = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \]  

(8.16)

We can also compute \( S_{21} \) as

\[ S_{21} = \frac{2V_2}{V_{G1}} \]  

(8.17)

according to the values in figure 8.3. \( V_2 \) is the voltage over the load side. The parameter \( S_{21} \) describes the transfer characteristics of the system. For example when we transfer microwaves from a source to a load we want to have \( S_{21} \) as close to unity as possible. Similar expressions for \( \Gamma_{ad} \) and \( S_{12} \) can be computed if a generator is applied at the output in figure 8.3, and a load impedance instead of the generator at the input side [2].