Final Project

NON-RADIATING WAVEGUIDE WITH HIGH PERMITTIVITY DIELECTRIC AND ANTENNAS BASED ON IT

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ABSTRACT

In this thesis the propagation characteristics of non-radiating dielectric waveguides (NRD guide) and its feasibility for end-fire and leaky-wave antenna structures in the W frequency band were studied. To obtain the propagation constant, Marcatili’s approximation method was extended to be used with NRD guides and the outcome was verified with simulation results. In order to connect the NRD guide to standard metal waveguides a new direct connection transition was developed and a prototype was manufactured and measured. Based on the concept of a dielectric rod waveguide antenna an NRD guide end-fire antenna is presented and verified experimentally based on the the same experimental NRD guide used for the transition. Furthermore a concept for a new type of periodic leaky-wave antenna based on the NRD guide is presented, where high impedance surfaces allow the radiated electromagnetic wave to propagate out of the waveguide structure.
# TABLE OF CONTENTS

**LIST OF ABBREVIATIONS**  

**LIST OF SYMBOLS**  

## 1 INTRODUCTION  

## 2 NON-RADIATING DIELECTRIC WAVEGUIDES  

<table>
<thead>
<tr>
<th>Section</th>
<th>Title</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>Basic principles</td>
<td>3</td>
</tr>
<tr>
<td>2.1.1</td>
<td>MAXWELL’s equations</td>
<td>3</td>
</tr>
<tr>
<td>2.1.2</td>
<td>Standard metal waveguides</td>
<td>8</td>
</tr>
<tr>
<td>2.1.3</td>
<td>Dielectric rod waveguides</td>
<td>10</td>
</tr>
<tr>
<td>2.1.4</td>
<td>Non-radiating dielectric waveguides</td>
<td>12</td>
</tr>
<tr>
<td>2.2</td>
<td>NRD guide model</td>
<td>15</td>
</tr>
<tr>
<td>2.3</td>
<td>MARCATILI’s method for non-radiating dielectric waveguides</td>
<td>19</td>
</tr>
<tr>
<td>2.4</td>
<td>Standard metal waveguide to NRD guide transition</td>
<td>43</td>
</tr>
<tr>
<td>2.4.1</td>
<td>Design</td>
<td>43</td>
</tr>
<tr>
<td>2.4.2</td>
<td>Simulation results</td>
<td>44</td>
</tr>
<tr>
<td>2.4.3</td>
<td>Measurement results</td>
<td>49</td>
</tr>
</tbody>
</table>

## 3 NON-RADIATING DIELECTRIC WAVEGUIDE ANTENNAS  

<table>
<thead>
<tr>
<th>Section</th>
<th>Title</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.1</td>
<td>Basic principles</td>
<td>54</td>
</tr>
<tr>
<td>3.1.1</td>
<td>Antenna parameters</td>
<td>54</td>
</tr>
<tr>
<td>3.1.2</td>
<td>Leaky-wave antennas</td>
<td>57</td>
</tr>
<tr>
<td>3.1.3</td>
<td>High impedance surfaces</td>
<td>60</td>
</tr>
<tr>
<td>3.2</td>
<td>NRD guide end-fire antenna</td>
<td>62</td>
</tr>
<tr>
<td>3.2.1</td>
<td>Design</td>
<td>62</td>
</tr>
<tr>
<td>3.2.2</td>
<td>Simulation results</td>
<td>63</td>
</tr>
<tr>
<td>3.2.3</td>
<td>Measurement results</td>
<td>69</td>
</tr>
<tr>
<td>3.3</td>
<td>NRD guide leaky-wave antennas</td>
<td>77</td>
</tr>
</tbody>
</table>
LIST OF ABBREVIATIONS

AUT .......... Antenna Under Test
DRW .......... Dielectric Rod Waveguide
DRWA ........ Dielectric Rod Waveguide Antenna
DUT .......... Device Under Test
FEM .......... Finite Element Method
FF .......... Far-Field
FSS .......... Frequency Selective Surface
HFSS .......... High Frequency Structure Simulator
HIS .......... High Impedance Surface
LSE mode ..... Longitudinal Section Electric Mode
LSM mode ..... Longitudinal Section Magnetic Mode
MMIC .......... Monolithic Microwave Integrated Circuit
MVNA .......... Millimeter Vector Network Analyzer
NF .......... Near-Field
NRD guide ..... Nonradiating Dielectric Waveguide
PEC .......... Perfect Electric Conductor
PMC .......... Perfect Magnetic Conductor
RBC .......... Radiation Boundary Condition
RWG .......... Rectangular Waveguide
TE mode ...... Transverse Electric Mode
TEM mode ...... Transverse Electromagnetic Mode
TM mode ...... Transverse Magnetic Mode
VNA ............ Vector Network Analyzer
LIST OF SYMBOLS

\(\alpha\) ............... attenuation constant \(\text{Np/m}\)
\(\beta\) ............... phase constant \(\text{rad/m}\)
\(\Delta s\) ............ scan step size \(\text{mm}\)
\(\Delta \theta\) ........... beam width \(\text{deg}\)
\(\gamma\) ............... propagation constant –
\(\lambda\) ............... wavelength \(\text{m}\)
\(\lambda_0\) ............ wavelength in free space \(\text{m}\)
\(\lambda_g\) ............ guide wavelength \(\text{m}\)
\(\mathbf{B}\) ............ magnetic flux density vector \(\text{T}\)
\(\mathbf{D}\) ............ electric displacement field vector \(\text{C/m}^2\)
\(\mathbf{E}\) ............ electric field vector \(\text{V/m}\)
\(\mathbf{e}\) ............. versor of the cartesian coordinate system –
\(\mathbf{E}_t\) ........... transverse electric field vector \(\text{V/m}\)
\(\mathbf{H}\) ............ magnetic field vector \(\text{A/m}\)
\(\mathbf{H}_t\) ........... transverse magnetic field vector \(\text{A/m}\)
\(\mathbf{J}_s\) ............ electric surface current density vector \(\text{A/m}^2\)
\(\mathbf{M}\) ........... magnetization vector \(\text{A/m}\)
\(\mathbf{M}_s\) ........... magnetic surface current density vector \(\text{V/m}\)
\(\mathbf{n}\) ............. normal vector –
\(\mathbf{V}\) ............. vector of the cartesian coordinate system –

\(\lambda\) ............... wavelength

\(\lambda_0\) ............ wavelength in free space

\(\lambda_g\) ............ guide wavelength

\(\mathbf{B}\) ............ magnetic flux density vector

\(\mathbf{D}\) ............ electric displacement field vector

\(\mathbf{E}\) ............ electric field vector

\(\mathbf{e}\) ............. versor of the cartesian coordinate system

\(\mathbf{E}_t\) ........... transverse electric field vector

\(\mathbf{H}\) ............ magnetic field vector

\(\mathbf{H}_t\) ........... transverse magnetic field vector

\(\mathbf{J}_s\) ............ electric surface current density vector

\(\mathbf{M}\) ........... magnetization vector

\(\mathbf{M}_s\) ........... magnetic surface current density vector

\(\mathbf{n}\) ............. normal vector

\(\mathbf{V}\) ............. vector of the cartesian coordinate system
P  ............ polarization density vector  C/m²
\( \mathbf{E} \)  ............ time-dependent electric field vector  V/m
\( \mathbf{H} \)  ............ time-dependent magnetic field vector  A/m
\( \nabla \)  ............ vector differential operator  –
\( \nabla_t \)  ............ transverse vector differential operator  –
\( \omega \)  ............ angular frequency  s⁻¹
\( \phi \)  ............ azimuth angle  deg
\( \Psi \)  ............ vector field  –
\( \theta \)  ............ zenith angle  deg
\( \theta_m \)  ............ beam angle  deg
\( \theta_{\text{max}} \)  ............ maximum vertical scanning angle  deg
\( \varphi_{\text{max}} \)  ............ maximum horizontal scanning angle  deg
\( C \)  ............ capacitance  F
\( D \)  ............ antenna directivity  dB
\( D_0 \)  ............ maximum antenna directivity  dB
\( d_m \)  ............ AUT - probe measurement distance  mm
\( E \)  ............ electric field component  V/m
\( e \)  ............ antenna radiation efficiency  –
\( e_c \)  ............ conductive efficiency  –
\( e_d \)  ............ dielectric efficiency  –
\( f \)  ............ frequency  Hz
\( G \)  ............ antenna gain  dB
\( G_0 \)  ............ maximum antenna gain  dB
\( H \)  ............ magnetic field component  A/m
\( k \) \ldots \) wavenumber \( \text{m}^{-1} \)
\( k_0 \) \ldots \) wavenumber in free space \( \text{m}^{-1} \)
\( k_x \) \ldots \) wavenumber component in \( x \)-direction \( \text{m}^{-1} \)
\( k_y \) \ldots \) wavenumber component in \( y \)-direction \( \text{m}^{-1} \)
\( k_z \) \ldots \) wavenumber component in \( z \)-direction \( \text{m}^{-1} \)
\( L \) \ldots \) inductance \( \text{H} \)
\( s_{11} \) \ldots \) s-parameter reflection coefficient \( \text{dB} \)
\( s_{21} \) \ldots \) s-parameter transmission coefficient \( \text{dB} \)
\( t \) \ldots \) time \( \text{s} \)

**Constants**

\( \mu \) \ldots \) permeability \( \text{H/m} \)
\( \mu_0 \) \ldots \) permeability of free space \( \text{H/m} \)
\( \rho \) \ldots \) charge density \( \text{C/m}^3 \)
\( \varepsilon \) \ldots \) permittivity, dielectric tensor \( \text{F/m} \)
\( \varepsilon_0 \) \ldots \) permittivity of free space \( \text{F/m} \)
\( \varepsilon_r \) \ldots \) relative permittivity \( – \)
\( c \) \ldots \) speed of light in free space \( \text{m/s} \)
\( n \) \ldots \) refractive index \( – \)

**Dimensions**

\( A \) \ldots \) antenna diameter \( \text{mm} \)
\( a \) \ldots \) width of dielectric rod \( \text{mm} \)
\( b \) \ldots \) height of dielectric rod \( \text{mm} \)
\( d \) \ldots \) distance of metal plates \( \text{mm} \)
\( dl \) \ldots \) length of discontinuity \( \text{mm} \)
\[ dw \ldots \text{width of discontinuity} \quad \text{mm} \]
\[ g \ldots \text{gap between dielectric and metal plate} \quad \text{mm} \]
\[ h \ldots \text{height of the NRD guide} \quad \text{mm} \]
\[ l \ldots \text{length of NRD guide section} \quad \text{mm} \]
\[ m \ldots \text{length of taper section} \quad \text{mm} \]
\[ p \ldots \text{period of discontinuities} \quad \text{mm} \]
\[ R \ldots \text{FRAUNHOFER distance} \quad \text{mm} \]
\[ s \ldots \text{metal plate offset} \quad \text{mm} \]
\[ t \ldots \text{metal thickness} \quad \text{mm} \]
\[ u \ldots \text{inner height of standard metal waveguide} \quad \text{mm} \]
\[ w \ldots \text{inner width of standard metal waveguide} \quad \text{mm} \]
\[ x_{\text{max}} \ldots \text{maximum horizontal scan movement} \quad \text{m} \]
\[ y_{\text{max}} \ldots \text{maximum vertical scan movement} \quad \text{m} \]
1 INTRODUCTION

The non-radiating dielectric waveguide (NRD guide) is a waveguide for the millimeter wave frequency range which offers interesting properties. It can be considered as a standard metal waveguide with the upper and lower plates removed, and the sidewalls extended and placed closer to each other. Its structure is similar to that of an H-guide, however it is capable of suppressing radiation loss, which is a problem with dielectric waveguides.

Using high permittivity dielectric materials it is possible to design dielectric rod waveguides (DRW) for high frequencies with small dimensions, enabling transitions to the standard metal waveguide. By choosing high permittivity dielectric materials with a low dielectric loss tangent for the dielectric material in the NRD guide, the NRD guide can be used in the desired W frequency band ranging from 75 to 110 GHz and transitions to other waveguides can be introduced so that the NRD guide can be integrated in millimeter wave circuits. The research for this thesis was carried out during a study visit at the Department of Radio Science and Engineering¹ at Aalto University² School of Science and Technology³ (formerly known as Helsinki University of Technology) in Espoo, Finland. Based on previous research on dielectric rod waveguides in the W frequency band at the department, the propagation characteristics of NRD guides and its feasibility for both end-fire and leaky-wave antenna structures were studied in this thesis.

MARCATILI’s method is an approximation method used for rectangular shaped dielectric waveguides to estimate the propagation constant. In this thesis MARCATILI’s method was extended to be used for NRD guides. The knowledge of the propagation constant of the NRD guide mode of operation is crucial for the design of periodic leaky-wave antennas. After theoretically deriving the approx-

¹http://radio.tkk.fi/en/
²http://www.aalto.fi/en/
³http://www.tkk.fi/en/
imination method based on the assumptions used in MARCATILI’s method, the results were compared to the dispersion characteristics obtained by simulating the NRD guide structure with the finite element method solver Ansoft\(^4\) HFSS (High Frequency Structural Simulator).

In order to connect an NRD guide to a standard metal waveguide, a proper transition is needed. This transition should offer low reflection, good transmission characteristics and should be easy to manufacture. Therefore different types of transitions were studied. A experimental prototype of the transition was then assembled and measured. Previously certain DRW end-fire antennas were presented at the Department of Radio Science and Engineering, also for the W band. Based on that concept, and NRD guide end-fire antenna was developed. The prototype of the transition was then used for the measurement of the radiation characteristics of the NRD guide end-fire antenna. The experimental end-fire antenna was therefore measured in the anechoic antenna chamber of the department. To verify the radiation characteristics, the results were compared to the ones obtained from simulating the structure within HFSS. As mentioned before, the NRD guide suppresses radiation out of the structure while operating in the mode for which the electric field is parallel to the metal plates. In this thesis a concept for periodic leaky-wave antennas based on NRD guides is presented. In order to allow radiation, caused by introduced discontinuities, out of the waveguide structure, the properties of the sidewalls need to be changed. High impedance surfaces (HIS) can be placed on the inner surface of the metal plates to alter its characteristics and make the antenna structure leaky. HIS are frequency selective surfaces (FSS), however at previous research at the Department of Radio Science and Engineering a voltage controlled HIS was presented. This offers interesting possibilities also for the proposed leaky-wave antenna concept.

2 NON-RADIATING DIELECTRIC WAVEGUIDES

In this chapter the propagation characteristics of non-radiating dielectric waveguides (NRD guide) in the W frequency band ranging from 75 to 110 GHz are discussed and MARCATILI’s method was extended to be used for NRD guides. The results of the calculated propagation characteristics were then verified by simulation results using the finite element method (FEM) solver Ansoft HFSS. Furthermore a novel metal waveguide to NRD guide transition is presented. The simulated results of the transition were verified with an experimental NRD guide.

2.1 Basic principles

2.1.1 MAXWELL’s equations

The MAXWELL equations are a set of four differential equations giving the relations between the electric and magnetic field and their sources. Their differential forms are given as [1]

\[
\begin{align*}
\nabla \cdot \mathbf{D} &= \rho \\
\nabla \cdot \mathbf{B} &= 0 \\
\n\nabla \times \mathbf{E} &= -\frac{\partial \mathbf{B}}{\partial t} \\
\n\nabla \times \mathbf{H} &= \mathbf{J}_f + \frac{\partial \mathbf{D}}{\partial t}
\end{align*}
\]

with (2.1a) being GAUSS’s law, (2.1b) GAUSS’s law for magnetism, (2.1c) FARADAY’s law and (2.1d) AMPERE’s circuital law with MAXWELL’s correction where \( \mathbf{D} \) is the electric displacement field, \( \mathbf{B} \) the magnetic flux density field, \( \mathbf{E} \) the electric field and \( \mathbf{H} \) the magnetic intensity field.
The electric displacement field is related to the electric field with the polarization density \( P \):

\[
D = \varepsilon_0 E + P
\]  
(2.2)

where \( \varepsilon_0 \) is the permittivity of free space. In a linear, homogeneous and isotropic dielectric equation (2.2) is

\[
D = \varepsilon E
\]  
(2.3)

where \( \varepsilon = \varepsilon_r \varepsilon_0 \) is the permittivity and \( \varepsilon_r \) the relative permittivity of the material.

The magnetic field intensity and the magnetic flux density are related to each other with the magnetization \( M \):

\[
H = \frac{B}{\mu_0} - M
\]  
(2.4)

where \( \mu_0 \) is the permeability of free space. If there exists an approximately linear relationship between \( M \) and \( H \), which is the case for a large class of magnetic materials [1], \( H \) and \( B \) are also related linearly to each other and equation (2.4) simplifies to

\[
B = \mu_0 \mu H
\]  
(2.5)

where \( \mu \) is the permeability of the material. For a more detailed discussion of Maxwell’s equations refer to [1].

**Boundary conditions**

The boundary conditions at the interface between two general medias are derived from Maxwell’s equations [2]:

\[
\mathbf{n} \cdot (D_2 - D_1) = \rho_{es}, \quad (2.6a)
\]

\[
\mathbf{n} \cdot (B_2 - B_1) = \rho_{ms}, \quad (2.6b)
\]

\[
\mathbf{n} \times (E_2 - E_1) = -M_s, \quad (2.6c)
\]

\[
\mathbf{n} \times (H_2 - H_1) = J_s, \quad (2.6d)
\]

where [3] \( \rho_{es} \) is the electric surface charge density, \( \rho_{ms} \) the magnetic surface charge density, \( M_s \) the magnetic surface current density, \( J_s \) the electric surface charge density, and \( B \) and \( H \) the magnetic field intensity and magnetic flux density, respectively.
current density and $\mathbf{n}$ is the unit normal vector.

**Figure 2.1** illustrates the decomposition of the electric field and the electric flux density into their tangential and normal components at the boundary of two dielectrics with different relative permittivity. The tangential component of the electric field is continuous across that boundary,

$$
E_{1t} = E_{2t},
$$

(2.7)

whereas for the electric flux density $\mathbf{D}$ the normal component is continuous across the boundary. With $\mathbf{D} = \varepsilon \mathbf{E} = \varepsilon_0 \varepsilon_r \mathbf{E}$ the second boundary condition for the electric field then is:

$$
\varepsilon_{r1} E_{1n} = \varepsilon_{r2} E_{2n}.
$$

(2.8)

The conditions for the magnetic field intensity $\mathbf{H}$ and the magnetic flux density $\mathbf{B}$ at the interface of two media with different permeability $\mu_1 \neq \mu_2$ can be obtained similar to the conditions for $\mathbf{E}$ and $\mathbf{D}$ [1]:

$$
H_{1t} = H_{2t},
$$

(2.9)

and with $\mathbf{B} = \mu \mu_0 \mathbf{H}$

$$
\mu_1 H_{1n} = \mu_2 H_{2n}.
$$

(2.10)
Wave equation

The propagation constant $\gamma$ of a wave traveling along the $z$-direction with the form

$$\mathcal{E}(x, y, z) = \textbf{E}(x, y) \cdot e^{-\gamma z} ,$$

$$\mathcal{H}(x, y, z) = \textbf{H}(x, y) \cdot e^{-\gamma z}$$

is defined as

$$\gamma = \alpha + j\beta ,$$

where $\alpha$ is the attenuation constant and $\beta$ the phase constant.

A field which is both a function of time and space is referred to as a wave. Being more restrictive with the definition, a wave is a solution of a wave equation [4]. Thus, the electric and magnetic field have to fulfill the HELMHOLTZ equation

$$\nabla^2 \Psi + k^2 \Psi = 0 ,$$

where $k = k_0 \sqrt{\varepsilon_r}$ is the wavenumber in the region of propagation and $k_0$ the wavenumber in free space. The wavenumber is the spatial equivalent to frequency telling the number of radians per unit length and it is real valued for lossless media and complex valued for lossy media. The HELMHOLTZ equation is a
special case of the wave equation in three dimensions and gives

\[ \left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2} \right) \cdot \mathbf{E} + k^2 \cdot \mathbf{E} = 0 \right. \quad (2.14a) \\
\left. \left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2} \right) \cdot \mathbf{H} + k^2 \cdot \mathbf{H} = 0 \right. \quad (2.14b) \\

for the electric and magnetic field respectively (the coordinate system is illustrated in Figure 2.2). If a time harmonic wave \( \mathbf{E}, \mathbf{H} \sim e^{j\omega t} \) is considered, the propagating electric and magnetic wave can be described as

\[ \mathcal{E}(x, y, z, t) = \mathbf{E} \cdot e^{j(\omega t - k_x x - k_y y - k_z z)} , \quad (2.15a) \]
\[ \mathcal{H}(x, y, z, t) = \mathbf{H} \cdot e^{j(\omega t - k_x x - k_y y - k_z z)} . \quad (2.15b) \]

To get the relation for the wavenumber components of the propagating wave, the HELMHOLTZ equation for either the electric or the magnetic field is calculated. Inserting the electric wave equation (2.15a) into equation (2.14a) leads to

\[ - \left( k_x^2 + k_y^2 + k_z^2 \right) \cdot \mathbf{E} + k^2 \cdot \mathbf{E} = 0 . \quad (2.16) \]

and subsequently to the equation for the cartesian wavenumber components:

\[ k_x^2 + k_y^2 + k_z^2 = k^2 . \quad (2.17) \]

If the \( z \)-axis is defined as the direction of propagation, the \( k_z \) wavenumber component tells us for a given frequency the speed with which the nodes of the wave are propagating along the guiding structure [5] and is equal to \( \beta \) in equation (2.12). As time harmonic waves are considered, MAXWELL’s equations (2.1c) and (2.1d) can be rewritten as (using \( \frac{\partial}{\partial t} = j\omega \))

\[ \nabla \times \mathbf{H} = j\omega \varepsilon \mathbf{E} , \quad (2.18a) \]
\[ -\nabla \times \mathbf{E} = j\omega \mu \mathbf{H} . \quad (2.18b) \]
Figure 2.3: Schematic model of a WR-10 rectangular waveguide for the W band having the inner dimensions [mm] \( w = 2.54 \) and \( u = 1.27 \) [6]

2.1.2 Standard metal waveguides

Hollow metal waveguides are not supporting transversal electromagnetic (TEM) waves, but instead the propagating waves are transversal electric (TE) and transversal magnetic (TM) waves. The most common hollow waveguide has a rectangular cross-section as sketched in Figure 2.3. It is standardized with different sizes for frequencies as high as 333 GHz [6].

For TE modes for which the electric field has no component in the direction of propagation, the wavenumber components \( k_x \) and \( k_y \) can be derived by applying boundary conditions as [6]

\[
k_x = \frac{m\pi}{w}, \quad n = 0, 1, 2, \ldots \quad (2.19)
\]
\[
k_y = \frac{m\pi}{u}, \quad m = 0, 1, 2, \ldots \quad (2.20)
\]

and now equation (2.17) can be written as

\[
k_z = \sqrt{\left(\frac{2\pi}{\lambda_0}\right)^2 - \left(\frac{n\pi}{w}\right)^2 - \left(\frac{m\pi}{u}\right)^2}. \quad (2.21)
\]

For the \( TE_{10} \) mode (the distribution of the electric and magnetic field is illustrated in Figure 2.4a) the electric field in the waveguide has the form shown in Figure 2.4b,

\[
E_y = E_0 \sin k_x x e^{i\omega t - k_z z} \quad (2.22)
\]

with \( n = 0 \), and thus \( k_y = 0 \). The guide wavelength of the waveguide is calculated
Figure 2.4: (a) Electric and magnetic field lines for the $TE_{10}$ mode in the cross-sectional plane of a rectangular waveguide [6] and (b) the distribution of the electric field for some value of $z$ [5]

by [6]

$$\lambda_g = \frac{2\pi}{\beta} = \frac{\lambda_0}{\sqrt{1 - (\lambda_0/\lambda_c)^2}} = \frac{\lambda_0}{\sqrt{1 - (f_c/f_0)^2}} ,$$

(2.23)

where $\lambda_0$ is the free-space wavelength and $\lambda_c$ and $f_c$ the cut-off wavelength and frequency, respectively. For high frequencies equation (2.23) approaches the free-space wavelength $\lambda_0$, whereas for lower frequencies the value inside the square root of equation (2.21) eventually becomes negative. For the case that $\lambda_0 > 2w$ equation (2.21) for the $TE_{10}$ mode turns imaginary, $k_z = \pm jk'$, and substituting it in equation (2.22) leads to

$$E_y = E_0 \sin k_y x e^{\pm jk'z} e^{j\omega t} .$$

(2.24)

This represents a wave oscillating with time but decaying exponentially along the waveguide away from the source and which is not able to propagate in the waveguide. The cut-off frequency for the different $TE_{nm}$ modes is given by [6]

$$f_{c,nm} = \frac{c}{2\pi} \sqrt{\left( \frac{n\pi}{w} \right)^2 + \left( \frac{m\pi}{u} \right)^2} ,$$

(2.25)

where $c$ is the velocity of light. For the $W$ band with the frequency range of 75 to 110 GHz the inner dimensions of the waveguide are set by the waveguide standard WR-10 [7] to be $[\text{mm}] w = 2.54$ and $u = 1.27$. A detailed analysis of the rectangular waveguide can be found for instance in [5] and [6]. The $TE_{10}$ mode in the rectangular waveguide can be represented as the superposition of two $TEM$ waves reflected by the walls of the waveguide as shown in Figure 2.5.
2.1.3 Dielectric rod waveguides

The dielectric rod waveguide (DRW) is a dielectric waveguide having a rectangular cross-section. Compared to low permittivity dielectric rod waveguides, DRW with high permittivity material have the advantage of smaller dimensions. Thereby, it is easier to manufacture and integrate in systems working at millimeter wave frequencies and also enables transitions to standard rectangular waveguides. Nowadays it is possible to manufacture components with small dimensions with the needed precision [8]. It is therefore desirable to chose high permittivity dielectrics with relative permittivity of 10-15 and a low loss tangent [9].

For the dielectric rod waveguide two families of guided modes exist: the $E_{pq}^y$ modes and the $E_{pq}^x$ modes, with $p$ and $q$ indicating the mode order corresponding to the number of maxima the field distribution has across the structure in the $x$ - and $y$ -direction, respectively. The upper letters $x$ and $y$ indicate the strongest component of the electric field [9] in that mode. Figure 2.7 exemplarily demonstrates the field distributions for the $E_{11}^y$ mode.

Figure 2.6: Cross-sectional view of a dielectric rod waveguide with width $a$ and height $b$ having the permittivity $\varepsilon_r$
Figure 2.7: Illustration of the distribution of the field components $E_y$ and $H_x$ for the $E_{11}^y$ dielectric rod waveguide mode [10]

Table 2.1: Dielectric properties of materials at 140GHz [9]

<table>
<thead>
<tr>
<th>Material</th>
<th>Si</th>
<th>Diamond</th>
<th>GaAs</th>
<th>Sapphire</th>
</tr>
</thead>
<tbody>
<tr>
<td>Crystal structure</td>
<td>Diamond</td>
<td>Diamond</td>
<td>Zinc blende</td>
<td>Wurtzite</td>
</tr>
<tr>
<td>Relative permittivity</td>
<td>$11.6 - 11.8$</td>
<td>$5.66$</td>
<td>$10.9 - 13.03$</td>
<td>$9.39, 11.56$</td>
</tr>
<tr>
<td>$\tan \delta [\times 10^{-4}]$</td>
<td>$0.03$</td>
<td>$0.06$</td>
<td>$3 - 14$</td>
<td>$1.1 - 1.4$</td>
</tr>
</tbody>
</table>
CHAPTER 2. NON-RADIATING DIELECTRIC WAVEGUIDES

2.1.4 Non-radiating dielectric waveguides

The non-radiating dielectric waveguide (NRD guide) was first proposed by Yoneyama and Nishida [12] and, since then, has been widely adapted in the millimeter-wave field. It is a low-loss transmission line with a topology similar to the \( H \)-guide. NRD guide components can be easily employed in monolithic millimeter integrated circuits (MMIC). The main advantage of the NRD guide is the non-radiating behaviour outside of the dielectric strip between the metal plates where the spacing is less than half the free-space wavelength and the cut-off state is reached. However, in the dielectric strip the wave can propagate because the wavelength is shortened and therefore removes the cutoff. The principle of operation is illustrated in 2.8. In most of the cases, the preferred mode of operation is the longitudinal section magnetic (\( L_{SM} \)) hybrid mode [13], where the magnetic field lines are parallel to the air/dielectric interface compared to the electric ones for \( L_{SE} \) modes as shown in 2.8c. Therefore the primary \( L_{SM_{01}} \) mode has lower conducting and dielectric losses. The wave propagation inside the NRD guide for the \( L_{SM_{01}} \) mode is similar to the propagation in the rectangular waveguide, where plane waves propagate in zigzag path through the waveguide [14].

Besides the standard NRD guide shown in Figure 2.9a, there also exist a model of an asymmetric NRD guide with a gap between the dielectric rod and
one metal plate (Figure 2.9b). While manufacturing, this kind of gap can be caused for instance due to bended metal plates. Reaching a certain value, the gap \( g \) causes a shift in the phase constant \( \beta \) resulting also in a shift of the guide wavelength \( \lambda_g \) and leakage away from the dielectric rod [15]. The NRD guide is used for various antenna concepts [16] [17].

In this thesis MARCATALI’s method is extended for the use with non-radiating waveguides. The primary \( LSM \) mode of propagation in the NRD guide is similar to the \( E_{11} \) mode in the DRW, with the assumption that \( E_y \) is the main electric field component. Connecting the NRD guide to a standard metal waveguide with a proper transition, the dominant \( LSM \) mode can be excited by the \( TE_{10} \) mode of a standard metal waveguide.

Due to the symmetric geometry of the NRD guide, symmetry planes can be introduced to reduce the complexity in electromagnetic field simulations (Figure 2.10). Thus, only one quarter of the model has to be simulated. The boundary conditions for the electric symmetry plane are

\[
E_t = 0, \quad (2.26a)
\]

\[
H_n = 0, \quad (2.26b)
\]
Figure 2.10: NRD guide symmetry planes

Figure 2.11: Symmetry planes

and for the magnetic symmetry plane

\[ E_n = 0 , \quad (2.27a) \]
\[ H_z = 0 . \quad (2.27b) \]
2.2 NRD guide model

To study the propagation characteristics of the NRD guide a waveguide model was designed and simulated with Ansoft HFSS. HFSS is a finite element method solver for electromagnetic structures. The cross-section of the two NRD guide models simulated are shown in Figure 2.12 on page 15. For the model depicted in Figure 2.12a the dielectric rod having a smaller width than the distance of the metal plates \( a < d \) was placed symmetrically between the parallel metal plates, whereas Figure 2.12b illustrates the classical NRD guide with \( a = d \). The metal plates were modeled assigning perfect electric conductors (PEC) for the side wall of the air box and the electric and magnetic symmetry planes as described in Section 2.1.4 were utilized to shorten simulation time. At the top face of the airbox radiation boundary condition (RBC) was assigned. As the dielectric rod silicon with a relative permittivity of \( \varepsilon_r = 11.9 \) (value used in HFSS) was chosen.

For the non-radiating behaviour of the waveguide the distance of the metal plates has to fulfill the condition \( d < \frac{\lambda_0}{2} \) for the upper limiting frequency \( f_u \) of the desired frequency range. The upper limiting frequency was set to \( f_u = 100 \text{ GHz} \).
and thus

\[ d < \frac{\lambda_{\mu_0}}{2} = \frac{c}{2 \cdot f_u} \approx 1.5 \text{ mm} , \]  
\[ d = 1.4 \text{ mm} , \]  

with \( c \) being the speed of light. In theory the limiting frequency for the defined distance would be \( f_u \big|_{d=1.4 \text{ mm}} \approx 107.14 \text{ GHz} \). The dimensions of the silicon rod for the NRD guide shown in Figure 2.12a with \( a < d \) were derived from [9] with \( a = 0.5 \text{ mm} \) and \( b = 1 \text{ mm} \). The height of the metal plates was set to be \( h = 3 \text{ mm} \).

With HFSS port mode simulation the guide wavelength \( \lambda_g \) and the propagation constant \( k_z \) were obtained. The dispersion diagrams \( k_z/k_0 - f \) for the classical NRD guide with \( a = d \) are presented in Figures 2.13 and A.1 for different values of the dielectric rod height \( b \). The guide wavelength is shown in Figure 2.14 on page 18 for in mm \( b = 0.5 \) and \( b = 0.75 \) and in Figure A.2 on page 93 for in mm \( b = 0.6 \) and \( b = 0.7 \). It is also possible to derive the propagation constant \( k_z \) and the guide wavelength \( \lambda_g \) analytically from the simulation results as described in Appendix D.
Figure 2.13: Simulation results of the propagation constant $k_z$ of the classical NRD guide $a = d$
Figure 2.14: Simulation results of the guide wavelength $\lambda_g$ of the classical NRD guide $a = d$
2.3 Marcatili’s method for non-radiating dielectric waveguides

In contrast to circular dielectric waveguides, of which the propagation characteristics can relatively easily be calculated using Bessel functions, the geometry of rectangular shaped dielectric waveguides makes it rather difficult. The reason is, that the field of the modes is diverging in the corners of the waveguide which complicates the design of converging numerical algorithms. An accurate analysis of this problem is discussed by Sudbø [18].

The Marcatili method is an approximation method which was first presented by Enrico A. J. Marcatili in the Bell Labs Technical Journal in 1969 [19]. It can be used to calculate the propagation constant of a rectangular waveguide. Therefore a rectangular waveguide as shown in Figure 2.15 is modeled, consisting of a core with high refractive index \( n_1 \) in the middle, denoted as area I in the model, surrounded by homogenous regions II, III, IV and V with low refractive indices \( n_{2.5} \). For the case that the propagation constant is far from the cut-off, the electromagnetic field is mainly distributed in the core and only a small part of its energy spreads to the surrounding regions. The energy distributed in the shaded areas is even less and can therefore be disregarded [20].

As outlined in more detail in Appendix C the electric and magnetic field can be described by longitudinal and transverse components [21]. Using the Maxwell equations

\[
\nabla \times \mathbf{E} = -\mu \frac{\partial \mathbf{H}}{\partial t} = -j \omega \mu \mathbf{H}
\]

\[
\nabla \times \mathbf{H} = \varepsilon \frac{\partial \mathbf{E}}{\partial t} = j \omega \varepsilon \mathbf{E}
\]

The vectors of the electric and the magnetic field can be divided into a longitudinal and a transversal component

\[
\mathbf{E} = e_z E_z + \mathbf{E}_t
\]

\[
\mathbf{H} = e_z H_z + \mathbf{H}_t
\]

Furthermore the vector differential operator \( \nabla \), represented by the greek symbol...
\[ \mathbf{\nabla} U = e_z \frac{\partial}{\partial z} + \mathbf{\nabla}_t, \]  \hspace{1cm} (2.32) \\

with 
\[ \mathbf{\nabla}_t = e_x \frac{\partial}{\partial x} + e_y \frac{\partial}{\partial y}. \]  \hspace{1cm} (2.33) 

The longitudinal field components traveling along the structure can be expressed as
\[ E_z = E_z(x, y)e^{-jk_z z}, \]  \hspace{1cm} (2.34a) \\
\[ H_z = H_z(x, y)e^{-jk_z z}, \]  \hspace{1cm} (2.34b) 

and thus the derivative in the direction of propagation is
\[ \frac{\partial}{\partial z} = -jk_z. \]  \hspace{1cm} (2.35)
Substituting equations (2.32) and (2.35) into equation (2.30a) leads to

\[
\left( \nabla_t + \epsilon_z \frac{\partial}{\partial z} \right) \times (e_z E_z + E_t) = -j\omega\mu (e_z H_z + H_t), \quad (2.36)
\]

\[
\nabla_t \times e_z E_z + \nabla_t \times E_t + e_z \frac{\partial}{\partial z} \times E_t = -j\omega\mu (e_z H_z + H_t). \quad (2.37)
\]

The z and the transversal components then are given as:

<table>
<thead>
<tr>
<th>Longitudinal</th>
<th>Transversal</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \nabla_t \times E_t = -j\omega\mu e_z H_z ), \quad (2.38a)</td>
<td>( \nabla_t \times H_t = j\omega e_z E_z ). \quad (2.38b)</td>
</tr>
<tr>
<td>( \nabla_t \times e_z E_z + e_z \frac{\partial}{\partial z} \times E_t = -j\omega\mu H_t ), \quad (2.38c)</td>
<td>( \nabla_t \times e_z H_z + e_z \frac{\partial}{\partial z} \times H_t = j\omega e_z E_t ) \quad (2.38d)</td>
</tr>
</tbody>
</table>

Finally the transversal components of the electric and magnetic field can be expressed by the longitudinal components \( E_z \) and \( H_z \) as

\[
H_t = \frac{1}{k_z^2 - k_x^2 - k_y^2} (jk_z \nabla_t H_z + jk^2 \epsilon_z \times \nabla_t E_z), \quad (2.39a)
\]

\[
E_t = \frac{1}{k_z^2 - k_x^2 - k_y^2} (jk_z \nabla_t E_z - jk^2 \mu e_z \times \nabla_t H_z), \quad (2.39b)
\]

After obtaining the transversal components \( k_x \) and \( k_y \) of the wavenumber \( k \), the propagation constant \( k_z \) can be calculated:

\[
k_z^2 = k^2 - k_x^2 - k_y^2, \quad (2.40a)
\]

\[
k_z = \sqrt{k_0^2 \epsilon_r - k_x^2 - k_y^2}. \quad (2.40b)
\]

The wavelength of a waveguide mode is given by [21]

\[
\lambda_g = \frac{2\pi}{k_z}. \quad (2.41)
\]

In MARCATILI’s method the transversal components is estimated by considering vertical and horizontal dielectric slab waveguides [22].
Vertical dielectric slab waveguide embedded between parallel metal plates

To use Marcatili’s method for the chosen NRD guide the assumptions for the vertical dielectric slab waveguide have to be extended. Therefore parallel conducting plates separated by the distance $d$ in which the slab waveguide is embedded symmetrically as shown in Figure 2.16 are modeled. For the vertical dielectric slab waveguide no dependence of the field components in the vertical $y$ -direction is assumed ($\frac{\partial}{\partial y} = 0$) and equation (2.33) therefore reduces to

$$\nabla_t = e_x \frac{\partial}{\partial x}.$$  \hfill (2.42)
Now the cross products of equations (2.38a) and (2.38c) are calculated as

\[
\nabla_t \times \mathbf{e}_t = \begin{vmatrix}
\mathbf{e}_x & \mathbf{e}_y & \mathbf{e}_z \\
\frac{\partial}{\partial x} & 0 & 0 \\
E_x & E_y & 0
\end{vmatrix} = \mathbf{e}_z \frac{\partial E_y}{\partial x}, \quad (2.43a)
\]

\[
\mathbf{e}_z \times \mathbf{E}_t = \begin{vmatrix}
\mathbf{e}_x & \mathbf{e}_y & \mathbf{e}_z \\
0 & 0 & 1 \\
E_x & E_y & 0
\end{vmatrix} = -\mathbf{e}_x E_y + \mathbf{e}_y E_x, \quad (2.43b)
\]

\[
\nabla_t \times \mathbf{e}_z = \begin{vmatrix}
\mathbf{e}_x & \mathbf{e}_y & \mathbf{e}_z \\
\frac{\partial}{\partial x} & 0 & 0 \\
0 & 0 & 1
\end{vmatrix} = -\mathbf{e}_y \frac{\partial}{\partial x}, \quad (2.43c)
\]

and the components for cartesian coordinates are:

\[
x: \quad jk_z E_y = -j\omega \mu H_x, \quad (2.44a)
\]

\[
y: \quad jk_z E_x + \frac{\partial E_z}{\partial x} = j\omega \mu H_y, \quad (2.44b)
\]

\[
z: \quad \frac{\partial E_y}{\partial x} = -j\omega \mu H_z. \quad (2.44c)
\]

Using equation (2.30b) the other three equations can be derived similarly:

\[
x: \quad jk_z H_y = j\omega \varepsilon E_x, \quad (2.45a)
\]

\[
y: \quad -jk_z H_x - \frac{\partial H_z}{\partial x} = j\omega \varepsilon E_y, \quad (2.45b)
\]

\[
z: \quad \frac{\partial H_y}{\partial x} = j\omega \mu E_z. \quad (2.45c)
\]

It has to be noted that equations (2.44a), (2.44c) and (2.45b) involve only three field components \(E_y, H_x\) and \(H_z\), whereas the other three components \(E_x, H_y\) and \(E_z\) are present in equations (2.44b), (2.45a) and (2.45c). That is, there are two independent sets of equations, corresponding to \(TE\) (set 1 with \(E_z = 0\)) and \(TM\) (set 2 with \(H_z = 0\)) wave modes. The \(TE\) modes are selected, as \(E_y\) is assumed to be the main electric field component.

From our set of equations for \(TE\) modes it is possible to express the other two field components \(H_x\) and \(H_z\) with respect to \(E_y\). Converting equation (2.44a)
delivers
\[ H_x = -\frac{k_z}{\omega \mu} E_y \]  
and rewriting equation (2.44c) gives
\[ H_z = -\frac{1}{j\omega \mu} \frac{\partial E_y}{\partial x} . \]  

It is then assumed that the dominant electric field component \( E_y \) is distributed cosinusoidally inside the dielectric slab (refer to Figure 2.7 on page 11) and decaying exponentially in the air-filled region between the dielectric and the metal plates:
\[ E_y = \begin{cases} 
A \cdot \cos (k_x x), & |x| < a' \\
B_1 \cdot e^{\gamma_x x} + B_2 \cdot e^{-\gamma_x x}, & |x| > a' 
\end{cases} \]  

where \( k_x \) and \( \alpha_x \) are the wave parameters. Inserting equation (2.48) into equation (2.47) then leads to
\[ H_z = \begin{cases} 
A \cdot k_x \cdot \sin (k_x x), & |x| < a' \\
\gamma_x \cdot (B_1 e^{-\gamma_x x} - B_2 e^{\gamma_x x}), & |x| > a' 
\end{cases} \]

Furthermore the assumption that the field distribution does not depend on the vertical \( y \) -coordinate and thus \( k_y = 0 \), reduces equation (2.40a) to
\[ k^2 = k_x^2 + k_z^2 , \]  
\[ k_x^2 = k^2 - k_z^2 . \]  

For the wave propagation in the air-filled region \( a' < x < d' \) equation (2.40a) similarly leads to
\[ k_0^2 = -\gamma_x^2 + k_z^2 , \]  
\[ \gamma_x^2 = k_x^2 - k_0^2 . \]  

Now boundary conditions at the dielectric-air interface and the metal plates can be applied to eliminate the unknowns \( A, B_1 \) and \( B_2 \).
Boundary condition: \( E_y(d') = 0 \) (metal plate)

As the electric field on the surface of a conductor is perpendicular to it, the electric field component \( E_y \) tangential to the surface layer has to fulfill the boundary condition \( E_y(d') = 0 \) at the metal plate and thus equation (2.48) has to be zero for \( x = d' > a' \):  

\[
B_1 e^{-\gamma d'} + B_2 e^{\gamma d'} = 0. 
\]  
(2.54)

(2.55)

The relation between the unknowns \( B_1 \) and \( B_2 \) is then given by:

\[
B_2 = -B_1 e^{-2\gamma d'}. 
\]  
(2.56)

Inserting equation (2.59) into equation (2.48) leads to

\[
E_y = \begin{cases} 
A \cdot \cos (k_x x) & , \ |x| < a' \\
B_1 \cdot (e^{-\gamma x} - e^{\gamma(x-2d')}) & , \ |x| > a'
\end{cases} 
\]  
(2.57)

and equation (2.49) becomes

\[
H_z = \begin{cases} 
A \cdot k_x \cdot \sin (k_x x) & , \ |x| < a' \\
\frac{B_1 \cdot \gamma_x}{j\omega \mu} \cdot (e^{-\gamma x} + e^{\gamma(x-2d')}) & , \ |x| > a'
\end{cases} 
\]  
(2.58)

Boundary condition: \( E_y(a' - 0) = E_y(a' + 0) \) (dielectric–air interface)

As described in Section 2.1.1 the tangential electric field component at an interface between two media with \( \varepsilon_1 \neq \varepsilon_2 \) are equal and thus \( E_y|_{x=d'-0} \) inside the dielectric core is equal to \( E_y|_{x=d'+0} \) in the air region at the interface: Using equation (2.57) gives

\[
A \cdot \cos (k_x a') = B_1 \left( e^{-\gamma a'} - e^{\gamma(a'-2d')} \right). 
\]  
(2.59)
Boundary condition: $H_z(a - 0) = H_z(a + 0)$ (dielectric–air interface)

The same boundary condition as for $E_y$ at the dielectric-air interface also applies for the longitudinal magnetic field component $H_z$ and with equation (2.58):

$$\frac{A \cdot k_x}{j \omega \mu} \cdot \sin(k_x a') = \frac{B_1 \cdot \gamma_x}{j \omega \mu} \cdot \left( e^{-\gamma_x a'} + e^{\gamma_x (a' - 2d')} \right), \quad (2.60)$$

$$A \cdot k_x \cdot \sin(k_x a') = B_1 \cdot \gamma_x \cdot \left( e^{-\gamma_x a'} + e^{\gamma_x (a' - 2d')} \right). \quad (2.61)$$

To eliminate the leftover unknowns $A$ and $B_1$ equation (2.59) is divided through equation (2.61):

$$\frac{eq. (2.59)}{eq. (2.61)}: \quad \frac{A \cdot \cos(k_x a')}{A \cdot k_x \cdot \sin(k_x a')} = \frac{B_1 \cdot \left( e^{-\gamma_x a'} - e^{\gamma_x (a' - 2d')} \right)}{B_1 \cdot \gamma_x \cdot \left( e^{-\gamma_x a'} - e^{\gamma_x (a' - 2d')} \right)}, \quad (2.62)$$

$$\frac{1}{\tan(k_x a')} = \frac{k_x}{\gamma_x} \cdot \frac{e^{-\gamma_x a'} - e^{\gamma_x (a' - 2d')}}{e^{-\gamma_x a'} + e^{\gamma_x (a' - 2d')} \cdot C}. \quad (2.63)$$

Let us denote

$$C = \frac{e^{-\gamma_x a'} - e^{\gamma_x (a' - 2d')}}{e^{-\gamma_x a'} + e^{\gamma_x (a' - 2d')} \cdot C}. \quad (2.64)$$

Using

$$\tan(\alpha) = \cot(\beta) = \cot \left( \frac{\pi}{2} - \alpha \right) \quad (2.65)$$

equation (2.63) can then be rewritten as:

$$\tan \left( \frac{\pi}{2} - k_x a' \right) = \frac{k_x}{\gamma_x} \cdot C, \quad (2.66)$$

$$\frac{\pi}{2} - k_x a' = \arctan \left( \frac{k_x}{\gamma_x} \cdot C \right) - n\pi, \quad (2.67)$$

$$k_x a' = \frac{\pi}{2} - n\pi - \arctan \left( \frac{k_x}{\gamma_x} \cdot C \right), \quad (2.68)$$

with $n = 0, \pm 1, \pm 2, \pm 3, \ldots$ Substituting $a' = \frac{a}{2}$ and $d' = \frac{d}{2}$ gives

$$k_x a = p\pi - 2\arctan \left( \frac{k_x}{\gamma_x} \cdot C \right), \quad (2.69)$$

26
with \( p = p_e = \pm 1, \pm 3, \pm 5, \ldots \) and

\[
C = \frac{e^{-\gamma_x \frac{d}{2}} - e^{\gamma_x (\frac{d}{2} - a)}}{e^{-\gamma_x \frac{d}{2}} + e^{\gamma_x (\frac{d}{2} - a)}},
\]

(2.70)

\[
\gamma_x = \sqrt{k_0^2 (\varepsilon_r - 1) - k_x^2}.
\]

(2.71)

The parameter \( C \) can be written in a more convenient way using the hyperbolic tangent function \( \tanh (x) = \frac{e^x - e^{-x}}{e^x + e^{-x}} \):

\[
C = \frac{e^{-\gamma_x \frac{d}{2}} - e^{\gamma_x (\frac{d}{2} - a)}}{e^{-\gamma_x \frac{d}{2}} + e^{\gamma_x (\frac{d}{2} - a)}} \cdot \frac{e^{\gamma_x \frac{d}{2}}}{e^{\gamma_x \frac{d}{2}}} \cdot \frac{e^{\gamma_x \frac{d}{2}}}{e^{\gamma_x \frac{d}{2}}} = \frac{e^{-\gamma_x (a - d)} - e^{\gamma_x (a - d)}}{e^{-\gamma_x (a - d)} + e^{\gamma_x (a - d)}},
\]

(2.72)

\[
C = \tanh \left( \frac{\gamma_x}{2} (d - a) \right).
\]

(2.73)

For now \( H_z \) was assumed to be distributed sinusoidally in the vertical dielectric slab waveguide. However, if a cosinusoidal distribution is considered, a phase difference of \( \frac{\pi}{2} \) is added to equation (2.49) due to that

\[
\cos (k_x x) = \cos (-k_x x) = \sin \left( k_x x + \frac{\pi}{2} \right).
\]

(2.74)

and after similar calculations obtain:

\[
k_x a' = -n\pi - \arctan \left( \frac{k_x}{\gamma_x} \cdot C \right),
\]

(2.75)

with \( n = 0, \pm 1, \pm 2, \pm 3, \ldots \). Substituting \( a' = \frac{x}{2} \) the parameter \( p \) of the transcendental equations in that case is \( p = p_o = 0, \pm 2, \pm 4, \ldots \). Only positive propagation parameters are considered, therefore the minus signs are omitted. Now the transcendental equations for even and odd modes are merged and the parameter \( p \) becomes

\[
p = p_e + p_o = 1, 2, \ldots
\]

(2.76)

\((p = 0 \text{ is not considered as the right part of equation (2.77) will become negative in that case). The transcendental equation representing the solution for the} \)
wavenumber component $k_x$ of the NRD guide then is:

$$ k_x a = p\pi - 2\arctan\left(\frac{k_x}{\gamma_x} \cdot C\right). \quad (2.77) $$

For the special case $a = d$ – depicting the classical NRD guide as shown in Figure 2.9a on page 13 – the coefficient $C$ becomes zero and therefore equation (2.77) simplifies to the well known equation

$$ k_x = \frac{p\pi}{a}. \quad (2.78) $$

In Figure 2.17 the graphical procedure to detect the solution of the transcendental equation (2.77) is revealed exemplarily for $f_1 = 75$ GHz and $f_2 = 95$ GHz. It is shown that for the chosen parameters only the mode for $p = 1$ is supported for these two frequencies, and for the second mode $p = 2$ there is no intersection and thus no solution.
Figure 2.18: Cross-section of the horizontal dielectric slab waveguide for Marcatili’s method with thickness \( b \)

**Horizontal dielectric slab waveguide**

To determine \( k_y \), a dielectric slab waveguide stretched infinitely in the horizontal dimension is considered (Figure 2.18). In that case the field components are not dependent of the \( x \)-direction, \( \frac{\partial}{\partial x} = 0 \), and equation (2.33) then simplifies to

\[
\nabla_t = e_y \frac{\partial}{\partial y}.
\]

(2.79)

Now the cross products of equations (2.38a) and (2.38c) will be:

\[
\nabla_t \times e_t = \begin{vmatrix}
  e_x & e_y & e_z \\
  \frac{\partial}{\partial y} & 0 & 0 \\
  E_x & E_y & 0 \\
\end{vmatrix} = -e_z \frac{\partial E_x}{\partial y},
\]

(2.80a)

\[
\mathbf{e}_z \times \mathbf{E}_t = \begin{vmatrix}
  e_x & e_y & e_z \\
  0 & 0 & 1 \\
  E_x & E_y & 0 \\
\end{vmatrix} = -e_x E_y + e_y E_x,
\]

(2.80b)

\[
\nabla_t \times \mathbf{e}_z = \begin{vmatrix}
  e_x & e_y & e_z \\
  \frac{\partial}{\partial y} & 0 & 0 \\
  0 & 0 & 1 \\
\end{vmatrix} = e_x \frac{\partial}{\partial y},
\]

(2.80c)
with which the equations for the three cartesian coordinates are given as:

\[
x: \quad jk_z E_y + \frac{\partial E_z}{\partial y} = -j\omega \mu H_x , \tag{2.81a}
\]

\[
y: \quad jk_z E_x = j\omega \mu H_y , \tag{2.81b}
\]

\[
z: \quad \frac{\partial E_x}{\partial y} = j\omega \mu H_z . \tag{2.81c}
\]

Using equation (2.30b) the other three equations are derived similarly:

\[
x: \quad jk_z H_y + \frac{\partial H_z}{\partial y} = j\omega \varepsilon E_x , \tag{2.82a}
\]

\[
y: \quad jk_z H_x = -j\omega \varepsilon E_y , \tag{2.82b}
\]

\[
z: \quad \frac{\partial H_x}{\partial y} = -j\omega \varepsilon E_z . \tag{2.82c}
\]

Again equations (2.81a), (2.82b) and (2.82c) involve only the three field components \(E_y\), \(E_z\) and \(H_x\) corresponding to the set of equations for \(TM\) wave modes with \(H_z = 0\) and equations (2.81b), (2.81c) and (2.82a) forms the set of equations for \(TE\) wave modes with \(E_z = 0\) and \(E_x\), \(H_y\) and \(H_z\) present. As \(E_y\) is assumed to be the main component \(TM\) modes are chosen and \(E_z\) and \(H_x\) are expressed through \(E_y\).

From equation (2.82b) \(H_x\) is written as

\[
H_x = -\frac{\omega \varepsilon}{jk_z} \cdot E_y , \tag{2.83}
\]

and substituting \(H_x\) in equation (2.82c) with equation (2.83) \(E_z\) is expressed through \(E_y\):

\[
E_z = \frac{1}{k_z} \cdot \frac{\partial E_y}{\partial y} . \tag{2.84}
\]

As for the vertical dielectric slab waveguide, it is again assumed that the \(E_y\) component is distributed cosinusoidally in the dielectric \(|y| < b'\), whereas it decays
exponentially outside the dielectric $|y| > b'$:

$$E_y = \begin{cases} 
A \cdot \cos(k_y y) & , |y| < b' \\
B \cdot e^{-\gamma_y y} & , |y| > b'
\end{cases}$$

(2.85)

where $k_y$ and $\alpha_y$ are the wave parameters. Inserting equation (2.85) into equations (2.83) and (2.84) gives

$$H_x = \begin{cases} 
-\frac{\omega \epsilon}{jk_z} \cdot A \cdot \cos(k_y y) & , |y| < b' \\
-\frac{\omega \epsilon}{jk_z} \cdot B \cdot e^{-\gamma_y y} & , |y| > b'
\end{cases}$$

(2.86)

and

$$E_z = \begin{cases} 
\frac{1}{k_z} \cdot A \cdot k_y \cdot \sin(k_y y) & , |y| < b' \\
\frac{1}{k_z} \cdot B \gamma_y \cdot e^{-\gamma_y y} & , |y| > b'
\end{cases}$$

(2.87)

Assuming no dependence of the field distribution on the vertical $x$-coordinate and thus $k_x = 0$ in the dielectric $|y| < b'$, equation (2.40a) simplifies to

$$k^2 = k_y^2 + k_z^2 ,$$

(2.88)

$$k_y^2 = k^2 - k_z^2 ,$$

(2.89)

whereas for the wave propagation in the air-filled region $|y| > b'$ it leads to

$$k_0^2 = -\gamma_y^2 + k_z^2 ,$$

(2.90)

$$\gamma_y^2 = k^2 - k_0^2 .$$

(2.91)

Now boundary conditions at the dielectric–air interface for the tangential field components $E_z$ and $H_x$ to eliminate the unknowns $A$ and $B$ can be applied.
Boundary condition: \( H_x(b' - 0) = H_x(b' + 0) \) \textbf{ (dielectric–air interface)}

Similarly to the dielectric-air interface for the vertical dielectric slab waveguide, again boundary conditions for the tangential magnetic field component \( H_x \) approaching from both sides of the interface are applied:

\[
\begin{align*}
  -\frac{\omega \varepsilon_r \varepsilon_0}{k_z} \cdot A \cdot \cos (k_y b') &= -\frac{\omega \varepsilon_0}{k_z} \cdot B \cdot e^{-\gamma_y b'}, \\
  \varepsilon_r \cdot A \cdot \cos (k_y b') &= B \cdot e^{-\gamma_y b'}. 
\end{align*}
\tag{2.92}
\tag{2.93}
\]

Boundary condition: \( E_z(b' - 0) = E_z(b' + 0) \) \textbf{ (dielectric–air interface)}

The same boundary condition again is applied to the tangential electric field component \( E_z \) at the dielectric-air interface:

\[
\begin{align*}
  -\frac{1}{k_z} \cdot A \cdot k_y \cdot \sin (k_y b') &= -\frac{1}{k_z} \cdot B \cdot \gamma_y e^{-\gamma_y b'}, \\
  A \cdot k_y \cdot \sin (k_y b') &= \gamma_y \cdot B \cdot e^{-\gamma_y b'}. 
\end{align*}
\tag{2.94}
\tag{2.95}
\]

To eliminate the unknowns \( A \) and \( B \) equation (2.93) is divided through equation (2.95):

\[
\frac{\text{eq. (2.93)}}{\text{eq. (2.95)}}: \quad \frac{\varepsilon_r \cdot A \cdot \cos (k_y b')}{A \cdot k_y \cdot \sin (k_y b')} = \frac{B \cdot e^{-\gamma_y b'}}{\gamma_y \cdot \varepsilon_r \cdot B \cdot e^{-\gamma_y b'}}, \\
\frac{1}{\tan (k_y b')} = \frac{k_y}{\gamma_y \varepsilon_r}, 
\tag{2.96}
\tag{2.97}
\]

and using equation (2.64) the transcendental equation for the wavenumber component \( k_y \) is affiliated:

\[
\tan \left( \frac{\pi}{2} - k_y b' \right) = \frac{k_y}{\gamma_y \varepsilon_r}, \tag{2.98}
\]

\[
\frac{\pi}{2} - k_y b' = \arctan \left( \frac{k_y}{\gamma_y \varepsilon_r} \right) + n\pi, \tag{2.99}
\]

\[
k_y b' = \frac{\pi}{2} - n\pi - \arctan \left( \frac{k_y}{\gamma_y \varepsilon_r} \right), \tag{2.100}
\]
with \( n = 0, \pm 1, \pm 2, \pm 3, \ldots \) Substituting \( b' = \frac{b}{2} \) gives

\[
k_y b = q\pi - 2 \arctan \left( \frac{k_y}{\gamma_y \varepsilon_r} \right)
\]  \hspace{1cm} (2.101)

with \( q = \pm 1, \pm 3, \pm 5, \ldots \) Similarly to the transcendental equation for the wavenumber component \( k_x \), the parameter \( q \) of equation (2.101) is assembled by adding the even and odd mode parameters:

\[
q = q_e + q_o.
\]  \hspace{1cm} (2.102)

**Propagation constant \( k_z \)**

Now having obtained both \( k_x \) and \( k_y \) for a given frequency \( f_0 \) (with the wavenumber in free space \( k_0 \)) with \( p = q = 1 \) for the \( E_{11}^1 \) mode, the propagation constant \( k_z \) is then calculated by

\[
k_z = \sqrt{k_0^2 \varepsilon_r - k_x^2 - k_y^2}.
\]  \hspace{1cm} (2.103)

In order for the wave to propagate along the structure in the \( z \)-direction, \( k_z \) has to be real, otherwise the wave will decay exponentially (called evanescent mode).
This puts limitations to the dimensions of the NRD guide which control the wavenumber components $k_x$ and $k_y$ for a given frequency and therefore dictates the cut-off. The cut-off wavenumber is derived from equation (2.103) for $k_z = 0$:

$$k_c^2 = k_{c0}^2 = k_x^2 + k_y^2. \tag{2.104}$$

The propagation constant $k_z$ can be calculated quite easily with a numerical computing environment like MATLAB or GNU Octave. The m-file calculating the propagation constant for the NRD guide taking the dimensions $a$, $b$ and $d$ as parameters is printed in Listing B.1 on page 115. Figures 2.20 and 2.21 display the dispersion diagram $k_z/k_0 - f$ and the guide wavelength $\lambda_g - f$ for a dielectric rod with dimensions in mm $a = 0.5$ and $b = 1$ for several values of the plate distance $d$. For the chosen dielectric rod it is shown that the NRD guide operates in the slow wave region with $k_z > k_0$ in the depicted frequency range from 75 to 110 GHz unless the metal plates are placed fairly close to the dielectric rod (compare the red line for $d = 0.75$ mm in Figure 2.20 with $a = 0.5$ mm) and the NRD guide enters the fast wave region for lower frequencies, near to the cut-off. In that case leakage away from the dielectric rod occurs as described in [15]. Placing the metal plates farther apart reduces the impact of the plates on the electric field and the NRD guide approaches the characteristics of the DRW, where $d \gg a$.

The dispersion characteristics for $b = 0.6$ mm and $b = 0.7$ mm are illustrated in Figures A.5 and A.6 on pages 96 and 97. It can be seen that the cut-off frequency is decreasing with increasing dielectric rod width $b (a = d = \text{const.})$. 
Figure 2.20: Dispersion diagram $k_z/k_0 - f$ for the NRD guide using the developed MARCATILI method with dimensions in mm $a = 0.5$ and $b = 1$ for different values for the distance between the metal plates $d$ and for the open dielectric rod waveguide (DRW) with $d \gg a$.

Figure 2.21: Guide wavelength of the NRD guide using the developed MARCATILI method with dimensions in mm $a = 0.5$ and $b = 1$ for different values for the distance between the metal plates $d$ and for the open dielectric rod waveguide (DRW) with $d \gg a$. 
Figure 2.22: Calculated dispersion diagram $k_z/k_0$ vs $f$ of the classical NRD guide with $a = d$
Figure 2.23: Calculated guide wavelength $\lambda_g$ of the classical NRD guide with $a = d$
Figure 2.24: Comparison of the calculated and the simulated results for the propagation constant of the NRD guide with dimensions in mm $a = 0.5$, $b = 1$ and varying plate distance $d$

Comparison of the numerical simulation results and MARCATILI’s approximation method

In Figure 2.24 the simulated and calculated results of the normalized propagation constant $k_z/k_0$ are compared for the NRD guide with a narrow dielectric rod $a = 0.5 \text{ mm} < d$. The MARCATILI results are lower than the simulated ones, which are considered to be numerically correct. This is because the field distributed in the shaded corner regions was not taken into account. For frequencies far higher than the cut-off this is not as prominent as for frequencies close to the cut-off [23]. There are approaches for dielectric waveguides to improve the accuracy of MARCATILI’s method, such as several perturbation methods [23]. Kumar et al. for instance proposed an advancement to MARCATILI’s method by also taking the fields in the shaded areas into account [24]. For our results the calculated values are higher than the simulated ones for higher frequencies. This can be due to convergence problems within the HFSS simulations.

For the classical NRD guide with $a = d$ where no neglected shaded areas as for the DRW and the NRD guide with a narrow dielectric rod exists, the compared values presented in Figures 2.25 and 2.26 are below the simulated results,
as anticipated. This effect is also more prominent for frequencies close to the cut-off, whereas it’s diminishing but not disappearing for higher frequencies. The reason for that is that it is assumed that $E_y$ is the main field component in the LSM mode of the NRD guide and the $E_x$ component was not taken into account. Even so MARCATILI’s method is a suitable method to approximate the propagation constant, especially for higher frequencies far from the cut-off and if possible inaccuracies of the simulation results are considered. The comparison of the calculated and simulated attenuation constant $\alpha$ shown in Figure 2.27 reveals that the developed extended MARCATILI method is not suitable to approximate the attenuation constant.
Figure 2.25: Comparison of the calculated and the simulated dispersion characteristic $k_z/k_0 - f$
Figure 2.26: Comparison of the calculated and the simulated dispersion characteristic $\lambda_g - f$
Figure 2.27: Comparison of the calculated and the simulated attenuation constant $\alpha$ for the classic NRD guide with $a = d$ and $b = 0.5$ mm
2.4 Standard metal waveguide to NRD guide transition

In order to connect a non-radiating dielectric waveguide to a standard metal waveguide, a suitable transition is needed. This transition has to show low reflection and good transmission characteristics and should be easy to manufacture. In [9] a standard metal waveguide to dielectric rod waveguide transition was developed. Based on that transition, a standard metal waveguide to NRD guide transition is presented in this Section.

As described in detail in Section 2.1.4, the desired mode of propagation for the NRD guide is the low-loss *LSM* mode with the magnetic field lines perpendicular to the parallel metal plates shown in Figure 2.8c on page 12. This mode can be excited by the *TE*$_{10}$ mode of a rectangular waveguide. There are different approaches towards transitions from rectangular waveguide to the NRD guide, such as the rectangular waveguide to image NRD guide presented by Hattori et al. (2001) [25]. Malherbe et al. (2005) developed a transition from a rectangular to nonradiating dielectric waveguide [26]. In this Section an easy transition for the W band is studied.

2.4.1 Design

Four different kinds of transitions were studied (illustrated in Figure 2.28):

(I) the pyramidal transition from the standard metal waveguide sidewalls to the NRD guide sidewalls,

(II) the opening of the sidewalls of the standard metal waveguide towards the NRD guide parallel plates,

(III) the direct connection of the two waveguides with shield plates closing the parts of the aperture of the standard metal waveguide outside of the NRD guide,

(IV) the direct connection as in (III) with additional shield plates at the NRD guide aperture not attached to the standard metal waveguide.

The dimensions of the NRD guide, the rectangular waveguide (WR-10 standard for 75 to 110 GHz) and the dielectric rod taper section are given in Figure 2.28. The dimensions of the dielectric rod and its taper sections are taken
Figure 2.28: Metal waveguide to NRD guide transition types: (I) pyramidal transition, (II) side wall opening, (III) direct connection and (IV) direct connection with additional shield plates

Whereas the NRD guide is attached directly to the rectangular waveguide for types (III) and (IV), a transition section with variable length is placed between the two waveguides for types (I) and (II) and the effect of the length on the reflection and transmission characteristics was studied. The NRD guide was modeled as described in Section 2.2 with a length of $l = 20$ mm. The leakage out of the NRD guide is affected by the height of the sidewalls, therefore its impact on the transmission characteristics was studied.

### 2.4.2 Simulation results

The cross-section of the NRD guide in the HFSS simulation models for the transition is shown in Figure 2.12a on page 15. To minimize simulation time and
computing ressources the symmetry of the structure is utilised by introducing
electric and magnetic symmetry planes as described in Section 2.2.

The reflection and transmission characteristics of types (I) and (II) are shown
in Figures A.7 to A.10 on pages 99 to 102. Although both transitions show com-
parable low reflection and good transmission behaviour for some of the studied
configurations, their comparable complex build-up do not predestine them for in-
tegrated solutions.

The desired type of transition is type (III) which is easy to manufacture. The
s -parameters obtained for transition (III) are illustrated in Figures 2.30 and 2.31.
As for the other types of transition, the reflection coefficient $s_{11}^{B}$ stays well
below $-20 \text{ dB}$. For the reflection coefficient $s_{21}^{B}$ it is seen from Figure 2.31 that for
$h = 3 \text{ mm}$ leakage out of the NRD guide occurs for lower frequencies, whereas
for frequencies higher than $f \approx 100 \text{ GHz } s_{21}$ decreases again. The reason for
that is the distance of the metal plates $d$ defined in Section 2.2, which is set to
d $= 1.4 \text{ mm}$ with an upper frequency of $f_u \approx 100 \text{ GHz}$. This effect is not that
prominent for bigger values of $h$. The spikes appearing in the transmission dia-
gram for frequencies higher than $f_u$ are possibly due to resonances in the NRD
guide. However, this has not been investigated explicitly in this thesis. Transition
type (III) with additional shield plates at the NRD guide sidewalls as seen in Fig-
ure 2.28IV does not show better reflection and transmission characteristics than
transition type (III), and therefore it is not studied any further.
The transition of choice is transition type (III), where the two waveguides are attached to each other directly with a tapered dielectric rod (Figure 2.32 shows a schematic model of the NRD guide with two transitions). Looking at the reflection coefficient $s_{11}^{\text{dB}}$ for the classical NRD guide with $a = d = 1.4$ mm in Figure 2.33 spikes of high reflection at $f_1 \approx 95$ GHz and $f_2 \approx 104$ GHz can be seen, caused at the transition and decreasing the transmission characteristic for that frequencies.

**Figure 2.30:** Reflection coefficient $s_{11}^{\text{dB}}$ of transition type (III) with dimensions in mm $a = 0.5$, $b = 1$ and $d = 1.4$
Figure 2.31: Transmission coefficient $s_{21}^{dB}$ of transition type (III) with dimensions in mm $a = 0.5, b = 1$ and $d = 1.4$

Figure 2.32: HFSS simulation model of the chosen NRD guide with transitions of type (III) with $l = 20$ mm
Figure 2.33: Reflection coefficient $s_{11}^{dB}$ of transition type (III) for $a = d$ and $b = 0.75 \text{ mm}$

Figure 2.34: Transmission coefficient $s_{21}^{dB}$ of transition type (III) for $a = d$ and $b = 0.75 \text{ mm}$
2.4.3 Measurement results

The measurement setup is illustrated schematically in Figure 2.35. For the experimental verification of the simulation results of the transition, two aluminium side-walls with appropriate connectors for the standard metal waveguide segments were fabricated. The dielectric rod was placed symmetrically between the side-walls, adjusted and held by styrofoam spacers and the sidewalls were screwed to metal waveguide flanges. The cross-section and the top-view schematics of the experimental NRD guide are shown in Figures 2.36 and 2.37 on page 50. This setup formed the device under test (DUT) attached to the test port connectors of the HP8530A vector network analyzer (VNA). Figure 2.38 shows a picture of the DUT in the test measurement setup.

The NVA delivers the reflection and the transmission coefficient as a complex number \( s_{x1} = \Re(s_{x1}) + j \cdot \Im(s_{x1}) \) which can be transformed into the polar form \( s_{x1} = \text{mag}(s_{x1}) \cdot e^{j\varphi} \) to get the magnitude and the phase of the complex reflection or transmission coefficient in the post-processing. Often the reflection coefficient is expressed in dB and refered to as return loss [27]

\[
\frac{dB}{11} = L_{\text{return}} = -20 \cdot \log |s_{11}| \text{ dB ,} \tag{2.105}
\]

and for the transmission coefficient \( s_{21} \) the gain:

\[
\frac{dB}{21} = G = 20 \cdot \log |s_{21}| \text{ dB .} \tag{2.106}
\]
Figure 2.36: Cross-section of the transition measurement setup with dimensions in mm

Figure 2.37: Top view of the transition measurement setup with dimensions in mm
As the setup consisted of a very experimental NRD guide with a narrow dielectric rod, it can be seen from the cross-section and the top view that the correct symmetric placement of the NRD guide sidewalls to the RWG flanges and the dielectric rod between the sidewalls and the metal waveguide by hand was challenging. Also the correct position of the dielectric rod could not be ultimately verified. Furthermore the used RWG flanges placed between the NRD guide and the VNA port connectors introduced further sources of error.

The measured reflection coefficient $s_{11}$ in dB is shown in Figure 2.39. The aforementioned difficulties with the experimental setup caused higher reflections in certain frequency areas compared to the simulation results. The simulated
reflection coefficient stayed well below $-20 \text{ dB}$ for the whole frequency range from 75 to 110 GHz, whereas the measured results only shows good reflection characteristics with $s_{11} < -15 \text{ dB}$ for the frequency range between 80 and 95 GHz (with spikes occurring around 90 GHz). Analyzing the transmission coefficient $s_{21}$ in Figure 2.40 shows the poor transmission characteristics of the assembled NRD guide for frequencies below $\sim 80$ GHz. However, above that frequency the investigated NRD guide with the transitions demonstrates comparable good transmission characteristics having the experimental nature of the studied NRD guide in mind, besides the spikes caused by unwanted reflections. The design of the NRD guide was set to have a distance of the metal plates $d = 1.4 \text{ mm} < \lambda_{0,100 \text{ GHz}}$ and thus its non-radiating attitude was limited to 100 GHz, which explains the decrease of the transmission properties for frequencies above 100 GHz (refer to the simulation model of the NRD guide in Section 2.2). For a better evaluation a different setup would be needed, where the placement of the dielect rod at the transition can be fine-tuned. With a more precisely manufactured waveguide (especially for the classical NRD guide with $a = d$) the metal waveguide to NRD guide transition is supposed to demonstrate a significantly improved reflection and transmission characteristic compared to the experimental one presented in this Section.
3 NON-RADIATING DIELECTRIC WAVEGUIDE ANTENNAS

The non-radiating properties of the NRD guide offers interesting applications of this waveguide for antenna purposes. In this chapter two different NRD guide antennas are presented. Based on the NRD guide developed in Chapter 2 and on the concept of a DRW antenna proposed in [9], an NRD guide end-fire antenna was designed, simulated and verified experimentally. Introducing discontinuities along the guiding structure of the NRD guide and modifying the metal plates, it is possible to alter the propagation properties to make the antenna structure leaky. The metal plates can either be foreshortened, enabling radiation out of the waveguide permanently, or the properties of the metal plates are changed by applying a special texture on its surface and thus making them high impedance surfaces which allows to controll the radiation. A concept of periodic leaky-wave antennas using NRD guides with HIS is presented in this Chapter.
3.1 Basic principles

3.1.1 Antenna parameters

For antenna characteristics a spherical coordinate system as illustrated in Figure 3.1 is used. Per definition the principal direction of propagation of the simulated antenna structures is set to be the y-direction.

Principal radiation patterns

Very often the performance characteristics of linearly polarized antennas are described in terms of their E- and H-plane far-field radiation patterns, refering to the plane containing the E or H field in the direction of maximum gain [28] as shown in Figure 3.2 for a vertical polarized antenna: the electric field lies in the elevation plane with $\phi = \frac{\pi}{2}$ and the magnetic field in the azimuthal plane with $\theta = \frac{\pi}{2}$. 

![Figure 3.2: Definition of the E- and H-plane of a vertical polarized antenna](image)
Figure 3.3: Near-field and far-field region surrounding an antenna with diameter \( D \) [30]

Field Regions

Usually the space an antenna is enclosed to is classified into two regions: the near-field (NF) and the far-field (FF) region – also referred to as FRAUNHOFER region – shown in Figure 3.3, whereat the near-field region is subdivided into the reactive and the radiating near-field (FRESNEL) region [28]. The boundary between the NF and the FF range is defined by the FRAUNHOFER distance [29]

\[
R = \frac{2A^2}{\lambda}
\]  

(3.1)

where \( A \) is the antenna diameter and \( \lambda \) the free space wavelength of the radiated electromagnetic wave [28]. The far-field describes that region compared to an antenna where the angular field distribution is independent from the distance to the antenna [28].
CHAPTER 3. NON-RADIATING DIELECTRIC WAVEGUIDE ANTENNAS

Directivity, gain and radiation efficiency

Directivity

\[ D (\phi, \theta) = \frac{P (\phi, \theta)}{P_{\text{isotropy}}} \]  \hspace{1cm} (3.2)

\[ D_0 = D (\phi, \theta) \big|_{\text{max}} \]  \hspace{1cm} (3.3)

antenna radiation efficiency

\[ e = e_{cd} = e_c e_d \]  \hspace{1cm} (3.4)

with \( e_c \) being the conductive and \( e_d \) the dielectric efficiency. Because \( e_c \) and \( e_d \) are very difficult to compute, they are usually determined experimentally \cite{28}.

From the definition of the radiated power

\[ P_{\text{rad}} = e P_{\text{in}} \]  \hspace{1cm} (3.5)

it is possible to describe the antenna radiation efficiency with

\[ e = e_{cd} = \frac{P_{\text{rad}}}{P_{\text{in}}} . \]  \hspace{1cm} (3.6)

Compared to the directivity, the gain of an antenna also takes losses into account which occur in the antenna structure, namely conductive and dielectric losses, but not losses due to mismatching and is related to the directivity by the antenna radiation efficiency:

\[ G (\phi, \theta) = e D (\phi, \theta) \]  \hspace{1cm} (3.7)

and in a similar way the maximum gain is interrelated with the maximum directivity:

\[ G_0 = G (\phi, \theta) \big|_{\text{max}} = e_{cd} D (\phi, \theta) \big|_{\text{max}} \]  \hspace{1cm} (3.8)

\[ G_0 = e D_0 . \]  \hspace{1cm} (3.9)
3.1.2 Leaky-wave antennas

Uniform leaky-wave antennas

The physical structure of a uniform leaky-wave antenna consists of a waveguide of a certain length $L$ at which the leakage occurs through a proper leakage mechanism. The oldest concept of a leaky-wave antenna is a rectangular waveguide with a continuous slit along its sidewall [31]. In order to be suitable as a uniform leaky-wave antenna, the dominant mode has to operate in the fast-wave region ($v_{ph} > c$) and therefore its phase constant $\beta$ has to fulfill the condition

$$\frac{\beta}{k_0} < 1$$

(3.10)

where $k_0$ is the wavenumber in free space. After deriving the phase constant $\beta$ and the attenuation constant $\alpha$ of the guiding structure, the beam angle $\theta_m$ and the beam width $\Delta \theta$ can then be calculated by the simple equations

$$\sin (\theta_m) \approx \frac{\beta}{k_0}$$

(3.11)

$$\Delta \theta \approx \frac{1}{(L/\lambda_0) \cos (\theta_m)}$$

(3.12)

where $\theta$ is measured perpendicular to the direction of propagation inside the guiding structure (Figure 2.2 on page 6 illustrates the used coordinate system with the wave traveling along the z-axis in the waveguide).

The conventional NRD guide can be modified to make it leaky. One way is to foreshorten the parallel plates at the top, which prevent the waveguide to radiate. However, this only works if the phase constant of the dominant propagation mode operates in the fast-wave region and fulfills equation (3.10). The concept is illustrated in Figure 3.4. The parameter $d$ indicating the distance the parallel plates are continued above the dielectric then controls the leakage of the antenna. An overview of uniform NRD guide leaky-wave antennas is given in [32].
Figure 3.4: Foreshortened-top leaky-wave NRD guide antenna developed by Sanchez and Oliner [33]. The leakage of the antenna is controlled by the stub length $s$.

Figure 3.5: Principle of operation of a periodic leaky-wave antenna with discontinuities separated by periodic distance $d$. The beam angle $\theta_m$ is counted from the perpendicular direction of the guiding structure.

Periodic leaky-wave antennas

In the case that the dominant mode of propagation operates in the slow-wave region its phase constant $\beta_0$ is greater than the wavenumber in free space $k_0$:

$$\frac{\beta_0}{k_0} > 1. \quad (3.13)$$

Therefore the dominant mode is purely bound to the guiding structure. Adding discontinuities introduces an infinity number of spatial harmonics [34] that can modulate the phase constant in a way that it fulfills equation (3.10). The resulting phase constant $\beta_n$ is related to the phase constant $\beta_0$ of the dominant mode with

$$\beta_n d = \beta_0 d + 2n\pi, \quad -\infty \leq n \leq \infty \quad (3.14)$$
where $d$ is the periodic distance between two subsequent discontinuities (compare Figure 3.5) and $n$ is the order of a given space harmonic. Thereby $n = 0$ stands for the basic wave. Equation (3.14) can be rewritten in the following form

$$\frac{\beta_n}{k_0} = \frac{\beta_0}{k_0} + \frac{2\pi n}{k_0 d} = \frac{\beta_0}{k_0} + \frac{n\lambda_0}{d}, \quad (3.15)$$

$$\frac{\beta_n}{k_0} = \frac{\lambda_0}{\lambda_g} + \frac{n\lambda_0}{d}. \quad (3.16)$$

Introducing discontinuities along the guiding structure means introducing small radiation sources, which together form an antenna array. Explained otherwise, in a periodic structure formed by discontinuities, Floquet spatial harmonics appear, and some of them can be in the fast-wave region.

As the waveguide mode is operating in the slow wave region fulfilling equation (3.13), it can easily been seen from equation (3.16) that $\frac{\beta_n}{k_0}$ can be less than unity if $n < 0$ is considered. Therefore only negative spatial harmonics were observed. Substituting $\frac{\beta}{k_0}$ in equation (3.10) with $\frac{\beta_n}{k_0}$ from equation (3.15) leads to

$$\left| \frac{\beta_n}{k_0} \right| = \left| \frac{\beta_0}{k_0} + \frac{n\lambda_0}{d} \right| < 1. \quad (3.17)$$

Compared to uniform leaky-wave antennas it is possible to have backward radiation at periodic leaky-wave antennas if

$$-1 < \frac{\beta_0}{k_0} + \frac{n\lambda_0}{d} < 0. \quad (3.18)$$

In order to cause leakage radiation for a given free space wavelength $\lambda_0$, period $d$ is limited to the range (derived from equation (3.17)):

$$\frac{-n\lambda_0}{\beta_0/k_0 + 1} < d < \frac{-n\lambda_0}{\beta_0/k_0 - 1} \quad (3.19)$$

$$n = -1:\quad \frac{\lambda_0}{\beta_0/k_0 + 1} < d < \frac{\lambda_0}{\beta_0/k_0 - 1}. \quad (3.20)$$

The distance $d$ of two subsequent discontinuities is also limited by the fact that it
is not desirable to have more than one lobe, which means the phase constant $\beta_{-2}$ of the second spatial harmonic should in the ideal case not fulfill equation (3.10) and therefore:

$$\frac{\beta_{-2}}{k_0} < -1$$  \hspace{1cm} (3.21)

$$\frac{\beta_0}{k_0} - \frac{2\lambda_0}{d} < -1$$  \hspace{1cm} (3.22)

$$d < \frac{2\lambda_0}{\beta_0 / k_0 + 1}.$$  \hspace{1cm} (3.23)

The spatial harmonics are caused by reflection of the incident wave at the discontinuities. The phase difference between two consecutive strips is [35]

$$\Delta \phi = \beta_n d = \beta_0 d = \left( \beta_{-1} + \frac{2\pi}{d} \right) d = \beta_{-1} d + 2\pi.$$  \hspace{1cm} (3.24)

The phase difference is $\Delta \phi = 2\pi$ for broadband radiation ($\theta_m = 0$) with $\beta_{-1} = 0$, which means that the reflected waves are in phase. This causes strong reflection for the given broadband radiation frequency and a strong attenuation of the propagating wave in that stop band. This behaviour can be reduced by using double-strips instead of single-strips for strip discontinuities [35].

### 3.1.3 High impedance surfaces

A perfect magnetic conductor (PMC) reflects the electromagnetic wave without changing the phase of the electric field (reflection coefficient $R = 1$), compared to a phase change of $180^\circ$ at the reflection at perfect electric conductors (corresponding to a reflection coefficient of $R = -1$) [36]. In real life no perfect magnetic conductor exits. However, some artificial surfaces can achieve nearly the same behaviour for a specific frequency.

It is possible to alter the radiofrequency surface properties of a conductor
by adding a special texture onto it. If the period of the surface texture defining the unit cell is much smaller than the wavelength of the desired frequency, the structure can be described by its surface impedance [38]. Sievenpiper presented in [37] an easy to manufacture mushroom-like structure with its cross-section illustrated in Figure 3.6 and its top-view in Figure 3.7. For the design frequency HIS behave like a PMC. In electromagnetic simulation tools HIS can be modeled as perfect magnetic conductors and thus there is no necessity to simulate the whole surface structure.

The original Sievenpiper HIS is set as a frequency selective surface (FSS) for a special frequency by design. The size of the unit cell defines the frequency for which it acts as a high impedance surface. Chicherin et al. (2008) introduced a voltage controlled high impedance surface for the W frequency band [39]. The proposed HIS consists of an array of electrically small MEMS capacitors placed atop of a ground plane and coupled to each other. The resonant frequency of the FSS is

\[ f_{\text{res}} = \frac{1}{2\pi \sqrt{L_{\text{eff}} C_{\text{eff}}}}. \]  (3.25)

Applying a control voltage to the MEMS capacitors changes their capacitance and therefore the resonant frequency for which the texture acts as a HIS and the phase shift it creates. In [39] different applications for this tuneable HIS are discussed. Furthermore, with this approach the HIS can be seen as switchable.
3.2 NRD guide end-fire antenna

If a DRW with both ends attached to a metal waveguide is considered, an end-fire antenna can be easily implemented by letting one end open instead of matching it with a transition. The DRW with the taper section left open then works as an end-fire antenna. HFSS simulation results for this antenna are presented in simulation results in Appendix E. Based on that antenna introduced in [9], an NRD guide end-fire antenna with the NRD guide designed in Section 2.2 was developed (the schematic of this antenna is shown in Figure 3.8).

3.2.1 Design

The design of the antenna is identical to the NRD guide studied in Chapter 2. One end is attached to a metal waveguide, whereas the other is left open and the taper section acts as the radiating end. The antenna structure is sketched in Figure 3.8. The antenna consists of a silicon dielectric rod with two taper section at each end with the dimensions given in Figure 2.29 and with length \( l \), which is placed symmetrically between two metal plates. For the simulation model \( l \) was set to 20 mm, whereas for the measurement setup the same dielectric rod as in Section 2.4.3 with \( l = 32 \) mm was used.

![Figure 3.8: NRD guide end-fire antenna model with prepended metal waveguide to NRD guide transition](image)
Figure 3.9: Cross-sectional view of the NRD guide HFSS simulation model for the end-fire antenna with \( E \) and \( H \) symmetry planes with dimensions in mm: \( a = 2a' = 0.5 \), \( b = 2b' = 1 \), \( d = 2d' = 1.4 \), \( t = 0.1 \) and \( h = 2h' \)

### 3.2.2 Simulation results

Similarly to the simulated NRD guide model in Section 2.2, it is possible to reduce the simulation time of the simulation model by taking advantage of the electric and magnetic symmetry of the NRD guide. The cross-section of the HFSS simulation model with dimensions is shown in Figure 3.9. As the height of the metal plates has an influence on the radiation patterns, the variation of the height from 3 to 9 mm with a step size of 2 mm was investigated. As in this research only a conceptual model of the NRD guide end-fire antenna is presented, the radiation efficiency of the antenna was not studied and the dielectric rod was considered to be loss-less. The radiation characteristics of the NRD guide end-fire antenna were studied for \( f_1 = 75 \) GHz, \( f_2 = 85 \) GHz and \( f_3 = 95 \) GHz and the results for the \( E \) - and \( H \) -plane are presented in Figure 3.11. For illustration purposes the gain range was set from \(-10\) to \(10\) dB. Figure 3.12 displays the tridimensional radiation patterns of the NRD guide end-fire antenna with \( s = 0 \) (Figures 3.12a to 3.12c) and \( s = 3 \) mm (Figures 3.12d to 3.12f) with the same gain range.

The maximum gain \( G_0 \) listed in Table 3.1 varies with the height of the metal plates and with frequency. For \( h_3 = 7 \) mm the effect of extending the metal plates half way over the radiating taper end, \( s = 3 \) mm, was studied. Interestingly thereby the maximum gain for higher frequencies \( f_2 \) and \( f_3 \) was increased,
Table 3.1: Maximum gain $G_0^\text{dB}$ of the NRD guide and the DRW end-fire antenna

<table>
<thead>
<tr>
<th>$f$ /GHz</th>
<th>$h$ /mm</th>
<th>3</th>
<th>5</th>
<th>7</th>
<th>7.3</th>
<th>9</th>
<th>DRWA</th>
</tr>
</thead>
<tbody>
<tr>
<td>75</td>
<td>7.92</td>
<td>8.32</td>
<td>8.70</td>
<td>6.79</td>
<td>9.42</td>
<td>8.65</td>
<td></td>
</tr>
<tr>
<td>95</td>
<td>7.64</td>
<td>7.39</td>
<td>8.11</td>
<td>9.68</td>
<td>8.31</td>
<td>8.54</td>
<td></td>
</tr>
</tbody>
</table>

whereas decreased for $f_1$ compared to $s = 0$.

Looking at the plot of the maximum gain $G_0$ in respect to frequency in Figure 3.10 the effect of the height of the metal plates can be seen. For $h = 9$ mm the antenna offers stable maximum gain $G_0 \approx 9.5$ dB for all three studied frequencies. For $h < 9$ mm and $s = 0$ the peak of the maximum gain appears for $f_2 = 85$ GHz.

![Figure 3.10: Maximum gain $G_0^\text{dB}$ (refer to Table 3.1)](image-url)
Figure 3.11: Radiation patterns of the NRD guide end-fire antenna
Figure 3.12: HFSS 3D polar radiation plot of gain $G_{\text{dB}}$ for the NRD guide end-fire antenna (a) – (c) without ($s = 0$) and (d) – (f) with extended metal plates ($s = 1.5 \text{ mm}$), $h = 7 \text{ mm}$
CHAPTER 3. NON-RADIATING DIELECTRIC WAVEGUIDE ANTENNAS

Comparison to DRW end-fire antenna

The non-radiating behaviour of the NRD guide and the shielding effect of the metal plates improves the radiation characteristics compared to that of the DRW end-fire antenna presented in [9]. Figure 3.13 shows the comparison between the NRD guide end-fire antenna with metal plate height $h = 3$ mm and $h = 7$ mm and the DRW end-fire antenna discussed in Appendix E for $f_1 = 75$ GHz. The comparison of the radiation patterns for $f_2 = 85$ GHz and $f_3 = 95$ GHz are illustrated in Figures A.18 and Figure A.19 on page 109 and 110 respectively. As can be seen from the figures, the radiation patterns are smoothed by the introduction of the parallel metal plates canceling out the ripples of the DRW antenna in the range from $-90^\circ$ to $90^\circ$. Therefore, the proposed NRD guide end-fire antenna improves the radiation characteristics compared to the DRW end-fire antenna.
Figure 3.13: Comparison of the radiation patterns of the NRD guide and the DRW end-fire antenna for different values of the plate height $h$ in mm for $f_1 = 75$ GHz
3.2.3 Measurement results

To measure the far-field radiation patterns of the NRD guide end-fire antenna a planar scanner from Nearfield Systems Inc.\(^1\) (NSI) was used. The front view of the scanner is shown schematically in Figure 3.14. The probe consisting of an open rectangular waveguide is placed on the scanning unit and can be moved in the horizontal \(x\)- and vertical \(y\)-direction to scan the radiation patterns in two dimensions. The equipment of the planar scanner is shielded with absorbers to prevent undesired reflections. The measurement setup is sketched in Figures 3.15 and 3.16 on page 70 and was set up in a temperature and humidity controlled anechoic chamber. Both the antenna under test (AUT) and the probe were connected to a millimeter vector network analyzer (MVNA) from AB Millimetre\(^2\). The MVNA-8-350 system\(^3\) is capable of covering a frequency range from 8 to 350 GHz. The measurement procedure and the record of the radiation patterns were controlled by a computer with the NSI2000\(^4\) antenna measurement software. With that software it is possible to program the scanner to scan a predefined area with a possible step size of \(\Delta s = 1\) mm. The AUT was placed at a distance of \(d_m = 15\) cm away from the probe. Due to setup limitations the maximum scanning angles in the vertical and horizontal plane were limited from the movement of the scanning unit.

\(^1\)http://www.nearfield.com/
\(^2\)http://www.abmillimetre.com/
\(^3\)http://www.abmillimetre.com/Products.htm
\(^4\)http://www.nearfield.com/sales/software.htm
Figure 3.15: Side view of the end-fire antenna measurement setup with $\theta_0 = \theta_{\text{max}}$ indicating the maximum vertical scanning angle in the $E$-plane due to setup limitations.

Figure 3.16: Top view of the end-fire antenna measurement setup with $\varphi_0 = \varphi_{\text{max}}$ indicating the maximum horizontal scanning angle in the $H$-plane due to setup limitations.
Figure 3.17: Scanning area with dimensions in m

The limitations for the horizontal and vertical movement of the scanning unit were \( x_{\text{max}} = 1.00 \, \text{m} \) and \( y_{\text{max}} = 0.62 \, \text{m} \). Therefore the maximum scanning angles are

\[
\theta_0 = \theta_{\text{max}} = 2 \cdot \arctan \left( \frac{0.50 \, \text{m}}{0.15 \, \text{m}} \right) = 2 \cdot 73.3^\circ = 146.6^\circ .
\]

(3.26)

and

\[
\varphi_0 = \varphi_{\text{max}} = 2 \cdot \arctan \left( \frac{0.31 \, \text{m}}{0.15 \, \text{m}} \right) = 2 \cdot 64.2^\circ = 128.4^\circ .
\]

(3.27)

The end-fire antenna assembled and presented in this section was the same experimental NRD guide with a silicon dielectric rod with tapered ends studied in Section 2.4.3, besides that only one end was attached to a metal waveguide forming the transition with the other end left open acting as the radiating end of the antenna. The radiation patterns for four different frequencies were recorded: \( f_1 = 75 \, \text{GHz} \), \( f_2 = 85 \, \text{GHz} \), \( f_3 = 95 \, \text{GHz} \) and \( f_4 = 105 \, \text{GHz} \). The scanning area is shown in Figure 3.17. The correct placement of the dielectric rod inside the waveguide could not be verified due to the experimental assembly
of the NRD guide. Therefore the center position was estimated and the scanning unit was programmed to scan the area $\pm 15\text{mm}$ from the center position ($\Delta x = \Delta y = 30\text{mm}$) with a step size of $\Delta s = 3\text{mm}$ for both the $E$- and the $H$-plane measurement. For $f_3 = 85\text{GHz}$ a complete scan of the whole scanning area $x_{\text{max}} \times y_{\text{max}} = 1.00\text{m} \times 0.62\text{m}$ with the step size $\Delta s = \Delta s_x = \Delta s_y = 3\text{mm}$ was produced. The NSI2000 software delivers the data in cartesian coordinates which were later transformed into polar data in the post-processing. To get the correct radiation patterns of the NRD guide end-fire antenna, the characteristics of the open-ended rectangular waveguide acting as the probe also needed to be taken into account. Therefore the radiation patterns of a similar probe than the one used on the scanning unit were recorded.

To correct the uncompensated far-field patterns the *gain transfer method* was used. Thereby the unknown gain of the AUT is compared to a reference probe [40]. Two open-ended rectangular waveguides are considered to have the same radiation characteristic and therefore the radiation pattern of the probe was derived by measuring the radiation pattern of one using the second as the probe. The measured gain is the multiplied gain of both the receiving and the transceiving antenna, corresponding to the sum of the gains in dB [41]:

$$ G_{m}^{\text{dB}} = G_r^{\text{dB}} + G_t^{\text{dB}} , $$

with $G_m$ being the measured gain, $G_r$ the gain of the receiving antenna and $G_t$ the gain of the transmitting antenna in dB. If both probes are considered to have the same radiation pattern with $G_{\text{probe}} = G_r = G_t$, the probe gain can be calculated by

$$ G_{\text{probe}} = 0.5 \cdot G_{m|\text{probe}} . $$

Now having obtained the gain of the probe, the AUT was calculated similarly:

$$ G_{\text{AUT}} = G_{m|\text{AUT}} - 0.5 \cdot G_{m|\text{probe}} . $$

Due to inaccuracies introduced by the measurement setup and the probes itself, the reference radiation pattern of the probe had to be post-processed in order to get feasible results. Therefore the mean value of $n = \pm 5$ consecutive values were taken to smoothen the radiation pattern. The effect can exemplarily
be seen in Figure A.21 on page 112.

The comparison of the simulated and measured radiation patterns for $f_1 = 75$ GHz and $f_3 = 75$ GHz are illustrated in Figures 3.18 and 3.19 (the radiation patterns for $f_4 = 105$ GHz are shown in Figure A.20 on page 111). Before plotting the comparison figures, the raw measured data was processed in order to smooth the radiation pattern. Thereby the mean value of $n = \pm 5$ consecutive values was taken. The effect of smoothing can be seen in Figure A.21 on page 112 for the probe radiation pattern.

Evaluating the $H$ -plane radiation pattern in Figure 3.18, corresponding to the horizontal movement of the scanning unit in the $x$ -direction, reveals the asymmetric placement of the dielectric rod in the NRD guide. In spite of this bad excitation the measured antenna radiation patterns did not show ripples even with the discussed measurement inaccuracies. The comparison of the $E$ -plane radiation pattern in Figure 3.19 also shows the asymmetric position of the dielectric rod and the NRD guide as a whole. However the measured data for the $E$ -plane shown in the figure is more deformed than for the $H$ -plane. This could be caused by the assembly of the experimental NRD guide and inaccuracies of the measurement setup. The contour plot of the whole scanning area for $f_2 = 85$ GHz is plotted in Figure 3.20. From this plot the asymmetric placement with bad excitation can be seen as well, compared to the tridimensional plot in Figure 3.12b.

Other causes of error than the ones discussed in Section 2.4.3 for the experimental NRD guide are possible reflections due to non-optimal arrangement of the absorbers around the AUT and the measurement probe and the error introduced by the gain transfer method (the two reference probes were not exactly the same).
Figure 3.18: Comparison of the simulated and measured far-field H-plane radiation patterns of the AUT
Figure 3.19: Comparison of the simulated and measured far-field E-plane radiation patterns of the AUT
Figure 3.20: 2D contour plot of the measured radiation pattern for $f_2 = 85$ GHz
3.3 NRD guide leaky-wave antennas

There have been different approaches on uniform leaky-wave antennas based on NRD guides. However, this thesis is focused on NRD guides working in the slow-wave region and therefore only periodic leaky-wave antennas were studied.

One method is to introduce periodic strips on the dielectric rod [42]. As described in Section 3.1.2, these discontinuities change the propagation characteristic to make the NRD guide leaky. Due to the non-radiating property of the area between the metal plates above the dielectric rod, the height of these metal plates need to be reduced in order to let the leaked wave propagate. Another method is to alter the properties of the sidewalls by adding a special texture onto it to turn it into a high impedance surface (see Section 3.1.3). Thus electromagnetic waves are able to propagate away from the dielectric rod. In this section a concept for NRD guide period leaky-wave antennas using high impedance surfaces is presented. HIS developed in [39] can be modulated by an external voltage and therefore the antenna can be switched on and off for different frequencies.

3.3.1 Design

To make the NRD guide end-fire antenna discussed in the previous section leaky, discontinuities were placed along the waveguide. For this concept of a leaky-wave antenna two kinds of discontinuities shown in Figure 3.21 were chosen: (a) PEC strips and (a) PEC stubs placed periodically on top of the dielectric rod. The discontinuities were modeled as flat PEC planes placed on top of the dielectric rod. In order to radiate out of the waveguide along the structure, the parallel metal plates needed to be modified to let the electromagnetic wave propagate in that region. The HIS structure is modeled in the HFSS simulation as PMC. The design of the original concept is shown in Figure 3.22. The schematic simulation model of the conceptual leaky-wave antenna is illustrated in Figure 3.23. Thereby PEC elements are positioned periodically above the dielectric rod. However, in this concept the PEC elements are hanging in the air above the dielectric, which is difficult to manufacture. It would be easier to realize this concept by putting the PEC elements directly on top of the dielectric rod, with the offset parameter $\alpha$ equal to zero.
Figure 3.21: Top view of the NRD guide with discontinuities separated by period $p$ on top of the dielectric: (a) printed strips with $dw = a$ and (b) printed stubs with $dw < a$.

Figure 3.22: Concept of NRD guide leaky-wave antenna: (a) side view and (b) cross-section with dimensions in mm $a = 1.4$, $b = 0.5$, $s = 1.6$, $t = 0.25$, $p = 2$, $r = 0.01$ and $o = 0.3$. 
To study the concept for a leaky-wave antenna based on the studied NRD guide with HIS surfaces on the inside of the metal plates, the period \( p = 2 \text{ mm} \) was chosen as the length of the unit cell of the leaky-wave antenna. The height of the parallel plates was set to be \( h = 10 \text{ mm} \), whereby PMC boundary conditions were applied to the plates above the dielectric rod on the side were the discontinuities were placed. The length of the NRD guide appointed to be \( l = 20 \text{ mm} \), corresponding to \( n = l/p - 1 = 9 \) placed discontinuities. The dispersion characteristic \( |\beta_{-1}/k_0| - f \) is shown in Figure 3.24. The center frequency at which broadside radiation occurs \( (\beta_{-1}c = 0) \) for \( p = 2 \text{ mm} \) then is \( f_c \approx 96.7 \text{ GHz} \). The antenna structure is leaky in the frequency range from \( \sim 85 \) to \( 110 \text{ GHz} \) with \( |\beta_{-1}| < k_0 \). For \( f < f_c \) the antenna has backfire radiation \( \theta_m < 0^\circ \), whereas for \( f > f_c \) the beam angle is \( \theta_m > 0^\circ \).

[43]
Figure 3.24: Calculated dispersion characteristics $\beta_{-1}/k_0 - f$ of the classical NRD guide with $a = d$ and dimensions in mm $a = 1.4$ and $b = 0.5$, $f_c \approx 96.7$ GHz
3.3.2 Simulation results

The $E$-plane radiation pattern for the original leaky-wave antenna concept is illustrated in Figure A.22 (with dimensions of the discontinuities given in Figure 3.22). This concept shows good radiation characteristics for the frequency range from 85 to 110 GHz with the gain of the side lobes being at least 10 dB less than the gain of the main lobe. However, as mentioned before, this concept is not easy to assemble with discontinuities hanging in the air. Placing the PEC elements directly on the dielectric rod, $o = 0$, while leaving the other dimensions untouched, did not deliver feasible radiation characteristics. Shortening the length of the stubs to $s = 0.5$ mm leads to the $E$-plane radiation pattern given in Figure A.23. The beam angle $\theta_m$ of the main lobe is shifted compared to the aforementioned setup and the beam form is distorted. Overall, the concept of placing these discontinuities directly on the dielectric rod did not prove to be a a good method for leaky-wave antennas based on the NRD guide.

Modeling the discontinuities not as PEC elements as described in Figure 3.22, but as flat PEC surfaces on top of the dielectric (considered as printed on the dielectric material, with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$), delivers the $E$-plane radiation pattern illustrated in Figure 3.25 with good radiation characteristics in the frequency range from 95 to 105 GHz. This
method proved to be more suitable for a leaky-wave antenna than the aforementioned method with PEC elements. In Figure 3.26 the beam angle $\theta_m$ of the simulation results is compared to the calculated beam angle. The calculated beam angle was based on MARCATILI’s method extended in Section 2.3, which was not completely matching the simulation results. This, and the change of the propagation characteristics of the NRD guide by the discontinuities, explains the shift of the beam angle. The maximum gain $G_0^{dB}$ of the main lobe is shown in Figure 3.27 However, within this concept the size and shape of the discontinuities is crucial. For dimensions in mm $dw = 0.1$ and 0.5 the developed leaky-wave antenna offers interesting radiation characteristics. Changing the size of the stubs as in Figure A.25, the resulting radiation pattern for some values makes the concept unfeasible. The effect of the stubs (or strips) shape and size on the radiating behaviour still needs to be investigated. The tridimensional radiation pattern of the stub loaded NRD guide leaky-wave antenna in the frequency range from 92 to 109 GHz are shown in Figures 3.28 to 3.30.

There are approaches to avoid the high reflections (the reflection coefficient $s_{11}^{dB}$ for the discussed leaky-wave antenna is shown in Figure A.24) in the frequency range close to broadside radiation by introducing double strips [35]. It
Figure 3.27: Maximum gain $G_0 \text{dB}$ of the main lobe of an NRD guide leaky-wave antenna with discontinuities of type (b) shown in Figure 3.21b on page 78 with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$

has to be studied if this is also applicable for this kind of leaky-wave antenna.
Figure 3.28: Tridimensional radiation pattern of the NRD guide leaky-wave antenna with discontinuities shown in Figure 3.21b for the frequency range from (a) 92 to (f) 97 GHz with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$
Figure 3.29: Tridimensional radiation pattern of the NRD guide leaky-wave antenna with discontinuities shown in Figure 3.21b for the frequency range from (a) 98 to (f) 103 GHz with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$.
Figure 3.30: Tridimensional radiation pattern of the NRD guide leaky-wave antenna with discontinuities shown in Figure 3.21b for the frequency range from (a) 104 to (f) 109 GHz with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$
Figure 3.31: E-plane radiation pattern of an NRD guide leaky-wave antenna with strip discontinuities of type (a) as shown in Figure 3.21a having the dimensions in mm $a = d = 1.4$, $b = 0.5$, $dl = 0.1$ and $p = 2$

The E-plane radiation pattern for the NRD guide leaky-wave antenna with strips as discontinuities as shown in Figure 3.21a is illustrated in Figure 3.31 for in GHz $f_1 = 90$, $f_2 = 95$ and $f_3 = 100$. It can be seen that for the chosen parameters this type of discontinuities is not delivering appropriate radiation characteristics. However, the effect of the dimension of the strip width $dl$ needs to be studied more thoroughly in order to judge about the feasibility of this kind of discontinuity for the proposed leaky-wave antenna concept.
4 CONCLUSIONS

The extension of Marcatili’s method to be used to analyze the propagation constant of the primary \( \text{LSM} \) mode of the NRD guide developed in this thesis offers an easy way to study the propagation characteristics of the NRD guide. Its suitability for the design of periodic leaky-wave antennas based on NRD guides was proven. However, considering the simulation results to be numerically correct, unlike for dielectric waveguides the calculated propagation constant does not converge totally to the one obtained from simulating the structure but stays below that, depending on the geometry of the dielectric rod. The reason for that, except of a possible error introduced for instance by converging problems occurring within the simulation, is the assumption that the main electric field component is the \( E_y \) component and the \( E_x \) field component is neglected.

The presented transition from standard metal waveguide to NRD guide marks a suitable and easy to manufacture way to connect both waveguide types. Aside from the difficulties at the measurement setup to verify the suitability of the transition, the measurement of the experimental NRD guide showed its potential to be used within integrated circuits. Though, in order to achieve both better reflection and transmission characteristics the NRD guide needs to be assembled more thoroughly. Especially the position of the dielectric rod between the metal plates and the placement of the taper section in the standard metal waveguide has to be optimized in order to get more reliable measurement results. As the same prototype as for the transition was used for the NRD guide end-fire antenna, the same problems due to the experimental setup occurred for the measurement of the radiation pattern. However, the developed end-fire antenna offers interesting applications within the millimeter wave area.

As the studied NRD guide is operating in the slow-wave region, introducing discontinuities along the dielectric rod causes the antenna to become leaky. The
presented concept of an NRD guide, where PMC boundary conditions were assigned to the inner surface of the parallel plates and PEC discontinuities were placed above the dielectric rod, describes a promising design for NRD guide periodic leaky-wave antennas. As the simulation results obtained, certainly not every type of discontinuity is suitable for this kind of leaky-wave antenna.
5 OUTLOOK

In order to verify the suitability of the presented transition more accurately, a more precise prototype needs to be manufactured and measured. A way has to be found to place the rod exactly in the right position, which is difficult to achieve by assembling the prototype by hand. A possible way is to prefabricate proper spacers so that the dielectric rod is both placed symmetrical between the metal plates and exactly in the right position inside the metal waveguide. For the classical NRD guide, where the dielectric rod is placed and attached directly between the metal plates, a prototype with two tapered sections of the dielectric rod can be microfabricated in an easy way to prove the capabilities of the studied transition. As the developed NRD guide end-fire antenna is based on the developed transition, the same arrangements apply for the more accurate verification process of its radiating characteristics. Furthermore the measurement setup has to be enhanced and probes with known radiation characteristics have to be used.

The developed extended MARCATILI's method proved to be a suitable method to estimate the propagation constant of the NRD guide mode of interest in order to realize the proposed NRD guide leaky-wave antennas. The HIS used to model PMC in real applications can be easily placed on the inner surface of the metal plates. To verify the presented concept the prototype has to be microfabricated as whole, with the discontinuities placed on top of the dielectric rod. Considering the effect the discontinuities have on the propagation characteristics and therefore the radiation characteristics, further research on feasible discontinuities design needs to be done.
A FIGURES

A.1 NON-RADIATING DIELECTRIC WAVEGUIDES

A.1.1 NRD guide model
**Figure A.1:** Simulation results of $k_z/k_0$ of the classical NRD guide with $a = d$
Figure A.2: Simulation results of the guide wavelength $\lambda_g$ of the classical NRD guide
$a = d$
A.1.2 MARCATILI’s method for non-radiating dielectric waveguides

Figure A.3: Calculated attenuation constant $\alpha$ of the classical NRD guide with $a = d$ for $b = 0.5$ mm
Figure A.4: Calculated dispersion diagram $k_z/k_0 - f$ of the classical NRD guide with $a = d = 1.4$ mm for different values of the height of the dielectric rod $b$. 
Figure A.5: Calculated dispersion diagram $k_z/k_0 - f$ of the classical NRD guide with $a = d$
Figure A.6: Calculated guide wavelength $\lambda_g$ of the classical NRD guide with $a = d$
A.1.3 Standard metal waveguide to NRD guide transition
Figure A.7: Reflection coefficient $s_{11}^{dB}$ of transition type (l) for varied transition length $v$
Figure A.8: Transmission coefficient $s_{21}^{\text{BB}}$ of transition type (l) for varied transition length $\nu$.
Figure A.9: Reflection coefficient $s_{11}^{dB}$ of transition type (II) for varied transition length $v$
Figure A.10: Transmission coefficient $s_{21}^{dB}$ of transition type (ll) for varied transition length $\nu$. 
Figure A.11: Reflection coefficient $s_{11}^{dB}$ of transition type (IV)

Figure A.12: Transmission coefficient $s_{21}^{dB}$ of transition type (IV)
Figure A.13: Reflection coefficient $s_{11}^{dB}$ of transition type (III) for in mm $a = d = 1.4$ and $b = 0.5$

Figure A.14: Transmission coefficient $s_{21}^{dB}$ of transition type (III) for in mm $a = d = 1.4$ and $b = 0.5$
Figure A.15: Transmission coefficient $s_{21}^{dB}$ of transition type (III) for in mm $a = d = 1.4$ and $b = 0.5$ in the frequency range 85 to 110 GHz
Figure A.16: Reflection coefficient $s_{21}^{dB}$ of transition type (III) for $a = d = 1.4$ mm
Figure A.17: Transmission coefficient $s_{21}^{dB}$ of transition type (III) for $a = d = 1.4$ mm
A.2 NON-RADIATING DIELECTRIC WAVEGUIDE ANTENNAS

A.2.1 NRD guide end-fire antenna
Figure A.18: Comparison of the radiation patterns of the NRD guide and the DRW end-fire antenna for different values of the plate height $h$ in mm at $f_2 = 85$ GHz.
Figure A.19: Comparison of the radiation patterns of the NRD guide and the DRW end-fire antenna for different values of the plate height $h$ in [mm] at $f_3 = 95$ GHz
Figure A.20: Measured E- and H-plane radiation pattern for $f_4 = 105$ GHz
Figure A.21: Probe compensation E-plane radiation pattern showing the smoothing effect
A.2.2 NRD guide leaky-wave antennas

Figure A.22: E-plane radiation pattern of the NRD guide leaky-wave antenna concept shown in Figure 3.22

Figure A.23: E-plane radiation pattern of the NRD guide leaky-wave antenna concept shown in Figure 3.22, with in mm \( o = 0 \) and \( s = 0.5 \)
**Figure A.24:** Reflection coefficient $s_{11}^{dB}$ of an NRD guide leaky-wave antenna with discontinuities of type (b) shown in Figure 3.21b on page 78 with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$

**Figure A.25:** E-plane radiation pattern for different dimensions of the discontinuities with $f = 95$ GHz and $p = 2$ mm
B  LISTINGS

Listing B.1: NRD.m

function [y] = NRD_param (a, b, d)

global eps

E1 = 3.41^2;
%a = 0.5; b = 1.0; d = 1.2;
y = []; fmin = 75; fmax = 110; step = 0.1;
kx = []; ky = [];
freq = [fmin:step:fmax];
k = [2*pi*freq/300];

for f = [fmin:step:fmax]
    k0 = 2*pi*f/300; eps = 11.9;
    x = find_kx (k0);
    kx = [kx, x];
    %eps = (k0^2*E1-x^2)/k0^2; %effective epsilon
    ky = [ky, find_ky (k0)];
end

y = (sqrt((k.x^2*E1-kx.x^2-ky.x^2)./k))';
plot(freq, sqrt((k.x^2*E1-kx.x^2-ky.x^2)./k), 'b')

function [kx] = find_kx (k0)

    kx = fzero (@(x) Fx(x, k0), 3);
function [result]=Fx(kx,k0)
    %global a d k0 eps;
    gammax=sqrt(k0^2*(eps-1)-kx^2);
    C=tanh(gammax/2*(d-a));

    result=kx*a-pi+2*atan(kx/gammax*C);
end

function [ky]=find_ky(k0)
    ky=fzero(@(x) Fy(x,k0),k0*2);
end

function [result]=Fy(ky,k0)
    %global a d k0 eps;
    gammay=sqrt(k0^2*(eps-1)-ky^2);

    result=ky*b-pi+2*atan(ky/gammay/eps);
end
C LONGITUDINAL AND TRANSVERSE FIELD COMPONENTS

\[ \nabla \times \mathbf{H} = j \omega \varepsilon \mathbf{E} \], \hspace{1cm} (C.1a) \hspace{1cm} -\nabla \times \mathbf{E} = j \omega \mu \mathbf{H} \]. \hspace{1cm} (C.1b)

As described in [21] the electric and magnetic field can be expressed in longitudinal and transverse field components. Using the definition

\[ \nabla = \nabla_t + e_z \frac{\partial}{\partial z} \] \hspace{1cm} (C.2)

the electric and magnetic field can be described as

\[ \mathbf{E} = e_z \mathbf{E}_z + \mathbf{E}_t \], \hspace{1cm} \mathbf{H} = e_z \mathbf{H}_z + \mathbf{H}_t \]. \hspace{1cm} (C.3)

Inserting equation (C.2) and equation (C.3) into equations (C.1) leads to

\[ \left( \nabla_t + e_z \frac{\partial}{\partial z} \right) \times \left( \mathbf{H}_t + e_z \mathbf{H}_z \right) = j \omega \varepsilon \left( e_z \mathbf{E}_z + \mathbf{E}_t \right), \]

\[ \nabla_t \times \mathbf{H}_t + e_z \frac{\partial}{\partial z} \times \mathbf{H}_t + \nabla_t \times e_z \mathbf{H}_z = j \omega \varepsilon \left( e_z \mathbf{E}_z + \mathbf{E}_t \right) \] \hspace{1cm} (C.4)

and

\[ \left( \nabla_t + e_z \frac{\partial}{\partial z} \right) \times \left( \mathbf{E}_t + e_z \mathbf{E}_z \right) = -j \omega \mu \left( e_z \mathbf{H}_z + \mathbf{H}_t \right), \]

\[ \nabla_t \times \mathbf{E}_t + e_z \frac{\partial}{\partial z} \times \mathbf{E}_t + \nabla_t \times e_z \mathbf{E}_z = -j \omega \mu \left( e_z \mathbf{H}_z + \mathbf{H}_t \right). \] \hspace{1cm} (C.5)

These equations can be divided into a longitudinal \((z)\) and a transverse \((t)\) compo-
APPENDIX C. LONGITUDINAL AND TRANSVERSE FIELD COMPONENTS

\[ z: \nabla_t \times E_t = -j \omega \mu e_z H_z , \]  
\[ \nabla_t \times H_t = j \omega \varepsilon e_z E_z , \]  
\[ t: e_z \frac{\partial}{\partial z} \times E_t + \nabla_t \times e_z E_z = -j \omega \mu H_t , \]  
\[ e_z \frac{\partial}{\partial z} \times H_t + \nabla_t \times e_z H_z = j \omega \varepsilon E_t . \]

To eliminate \( E_t \) (C.9) is written in the form

\[ E_t = \frac{1}{j \omega \varepsilon} \left( e_z \frac{\partial}{\partial z} \times H_t + \nabla_t \times e_z H_z \right) \]  
and insert it in equation (C.8) with \( \frac{\partial}{\partial z} = -jk_z \) (according to ??) and \( k^2 = \omega^2 \mu \varepsilon \) being the wave number in the filling material:

\[ -j \omega \mu H_t = e_z \frac{\partial}{\partial z} \times \left( \frac{1}{j \omega \varepsilon} \left( e_z \frac{\partial}{\partial z} \times H_t + \nabla_t \times e_z H_z \right) \right) + \nabla_t \times e_z E_z , \]
\[ = -\frac{k_z}{\omega \varepsilon} (\nabla_t H_z + jk_z H_t) + \nabla \times e_z E_z , \]
\[ H_t = -j \frac{k_z}{k^2} \nabla_t H_z + \frac{k^2}{k^2} - j \frac{\omega \varepsilon}{k^2} e_z \times \nabla_t E_z , \]

using the identities

\[ e_z \times (\nabla_t \times e_z H_z) = \nabla_t H_z , \]  
\[ e_z \times (e_z \times H_t) = -H_t , \]  
\[ \nabla_t \times e_z E_z = -e_z \times \nabla_t E_z . \]

Now it is possible to write \( H_t \) depending on the longitudinal components \( E_z \) and \( H_z \) only:

\[ H_t = \frac{1}{k_z^2 - k^2} (jk_z \nabla_t H_z + j \omega \varepsilon e_z \times \nabla_t E_z) . \]

The equation for \( E_t \) can be derived analogously:

\[ E_t = \frac{1}{k_z^2 - k^2} (jk_z \nabla_t E_z - j \omega \mu e_z \times \nabla_t H_z) . \]
D PROPAGATION CONSTANT

![Diagram of an NRD model showing simulation boundaries and dimensions.](image)

It is possible to obtain the dispersion curve of the NRD guide from simulation results in an analytical way. Therefore a structure as shown in Figure D.1 has been simulated in HFSS. The parallel metal plates were presented as perfect electric conductor (PEC) boundaries and radiation boundaries (RBC) were assigned to the top and bottom layer of the model.

After simulation the phase shift $\varphi|_{z=1}$ of the wave at the output port compared to the incident wave at the input port was studied (Figure D.2. The propagation constant $k_z$ can be calculated by

$$k_z = \frac{\varphi}{L}$$  \hspace{1cm} (D.1)

where $L$ is the length of the guiding structure and $\varphi$ the phase shift the field experiences along the length of the waveguide. As the phase shift at the output port...
is only relative to the phase at the input port the distribution of the electric field along the waveguide had to be examined for the lowest frequency of the studied frequency range $f_l = 75 \text{ GHz}$. The field distribution is presented in Figure D.3. The propagation constant $k_z$ is the spatial equivalent to frequency representing the number of nodes of the field per unit length. The overall phase shift of the wave along the waveguide is then determined by

$$\phi = \phi_{s21} - 2\pi \cdot N$$

(D.2)

where $N$ is the number of rotations the incident wave experiences along the guiding structure obtained from Figure D.3. With $N_{75\text{ GHz}} = 8$ the phase shift $\phi$ is shown in Figure D.4 and using equation (D.1) the dispersion diagram $k_z/k_0 - f$, with $k_0 = \frac{2\pi}{\lambda_0}$ being the wavenumber in free space, shown in Figure D.5 was plotted. With the relation of the guide wavelength $\lambda_g$ to the propagation constant $k_z$

$$\lambda_g = \frac{2\pi}{k_z}$$

(D.3)

the guide wavelength was calculated (Figure D.6) which is crucial for the design of leaky-wave antennas based on NRD guides.
Figure D.3: Magnitude of the electric field in the vertical plane of the NRD guide model at $f_l = 75$ GHz showing the number of full rotations the electromagnetic wave experience along the guide structure ($N = 8$)

Figure D.4: Cumulative phase shift $\phi_{911}$ along the NRD guide structure at the output port
**Figure D.5:** Dispersion curve $k_z/k_0 - f$ of the NRD guide model

**Figure D.6:** Guide wavelength $\lambda_g - f$ of the NRD guide model
E DIELECTRIC ROD WAVEGUIDE ANTENNA

Figure E.1: Schematic model of the dielectric rod waveguide antenna with RWG to DRW transition

In [9] a dielectric rod waveguide antenna excited by a rectangular waveguide for the W frequency band from 75 to 110 GHz was introduced (its schematic is shown in Figure E.1). The dimensions of the dielectric rod are in mm $a = 0.5$, $b = 1$ and $m = 6$ for the length of the taper section (compare to Figure 2.29 on page 45) and its simulation model with E and H symmetry planes is sketched in Figure E.2. The polar radiation patterns for $f_1 = 75$ GHz, $f_2 = 85$ GHz and $f_3 = 95$ GHz are shown in Figure E.3 and the normalized gain $G^{\text{dB}} - G_0^{\text{dB}}$ in Figures E.4 to E.6.
Figure E.2: (a) Cross-section of the metal waveguide to DRW transition and (b) DRW end-fire antenna simulation cross-section with dimensions in mm $a = 2a' = 0.5$ and $b = 2b' = 1$
Figure E.3: Radiation patterns of the DRW end-fire antenna
Figure E.4: Normalized gain $G_{\text{dB}} - G_0^{\text{dB}}$ of the DRW antenna for $f_1 = 75$ GHz

Figure E.5: Normalized gain $G_{\text{dB}} - G_0^{\text{dB}}$ of the DRW antenna for $f_2 = 85$ GHz
Figure E.6: Normalized gain $G \text{dB} - G_0 \text{dB}$ of the DRW antenna for $f_3 = 95$ GHz
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BIBLIOGRAPHY


131


List of Figures

2.1 Decomposition of the (a) electric field and (b) electric flux density at the interface of two dielectrics with different relative permittivity into tangential and nominal components .................................. 5
2.2 Cartesian and polar coordinate system .............................................. 6
2.3 Schematic model of a WR-10 rectangular waveguide for the W band having the inner dimensions [mm] w = 2.54 and u = 1.27 [6] ................................................................. 8
2.4 Electric and magnetic field lines for the $TE_{10}$ mode in the cross-sectional plane of a RWG ................................................................. 9
2.5 Decomposition of the $TE_{10}$ mode into two plane waves inside a RWG ................................................................. 10
2.6 Cross-sectional view of a dielectric rod waveguide with width $a$ and height $b$ having the permittivity $\varepsilon_r$ .............................................. 10
2.7 Illustration of the distribution of the field components $E_y$ and $H_x$ for the $E_{11}'$ dielectric rod waveguide mode [10] ...................... 11
2.8 Principle of operation of the NRD guide: (a) Rectangular waveguide working below cutoff, (b) parallel plates waveguide still below cutoff and (c) low-loss LSM$_{01}$ mode electric and magnetic field lines for non-radiating dielectric waveguides [11] .............................................. 12
2.9 Cross-sectional view of the NRD guide .............................................. 13
2.10 NRD guide symmetry planes ......................................................... 14
2.11 Electric and magnetic symmetry plane ........................................... 14
2.12 HFSS simulation model cross-sections of the NRD guide using electric and magnetic symmetry planes with $a' = a/2$, $b' = b/2$, $d' = d/2$ and $h' = h/2$ ...................... 15
2.13 Simulation results of the propagation constant $k_z$ of the classical NRD guide $a = d$ ...................................... 17

133
2.14 Simulation results of the guide wavelength $\lambda_g$ of the classical NRD guide $a = d$ .......................................................... 18
2.15 The cross-sectional view of a dielectric rod waveguide in MARCATILI’s method .......................................................... 20
2.16 Cross-sectional view of the vertical dielectric slab waveguide for MARCATILI’s method embedded between parallel metal plates (refer to Figure 2.9c) with width $a = 2a'$ and distance of the metal plates $d = 2d'$ .......................................................... 22
2.17 Graphical method to obtain the solutions for the transcendental equation (2.68) for $k_x$ with dimensions in mm $a = 0.5$, $b = 1$ and $d = 1.4$; limited by $k_{xd} = k_0 \sqrt{\varepsilon_r - 1}$, with $k_{xd/75GHz} = 5.19$ mm$^{-1}$ and $k_{xd/95GHz} = 6.57$ mm$^{-1}$ .......................................................... 28
2.18 Cross-section of the horizontal dielectric slab waveguide for MARCATILI’s method with thickness $b$ .......................................................... 29
2.19 Graphical method to obtain the solutions for the transcendental equation (2.101) for $k_y$ with dimensions in mm $a = 0.5$, $b = 1$ and $d = 1.4$; limited by $k_{yd} = k_0 \sqrt{\varepsilon_r - 1}$, with $k_{yd/75GHz} = 5.19$ mm$^{-1}$ and $k_{yd/95GHz} = 6.57$ mm$^{-1}$ .......................................................... 33
2.20 Dispersion diagram $k_z/k_0 - f$ for the NRD guide using the developed MARCATILI method with dimensions in mm $a = 0.5$ and $b = 1$ for different values for the distance between the metal plates $d$ and for the open dielectric rod waveguide (DRW) with $d \gg a$ .......................................................... 35
2.21 Guide wavelength of the NRD guide using the developed MARCATILI method with dimensions in mm $a = 0.5$ and $b = 1$ for different values for the distance between the metal plates $d$ and for the open dielectric rod waveguide (DRW) with $d \gg a$ .......................................................... 35
2.22 Calculated dispersion diagram $k_z/k_0 - f$ of the classical NRD guide with $a = d$ .......................................................... 36
2.23 Calculated guide wavelength $\lambda_g$ of the classical NRD guide with $a = d$ .......................................................... 37
2.24 Comparison of the calculated and the simulated results for the propagation constant of the NRD guide with dimensions in mm $a = 0.5$, $b = 1$ and varying plate distance $d$ .......................................................... 38
2.25 Comparison of the calculated and the simulated dispersion characteristic $k_z/k_0 - f$ .......................................................... 40
2.26 Comparison of the calculated and the simulated dispersion characteristic $\lambda_g - f$ .......................................................... 41
Comparison of the calculated and the simulated attenuation constant $\alpha$ for the classic NRD guide with $a = d$ and $b = 0.5$ mm.

Metal waveguide to NRD guide transition types: (I) pyramidal transition, (II) side wall opening, (III) direct connection and (IV) direct connection with additional shield plates.

NRD guide, RWG cross-section and taper side view dimensions.

Reflection coefficient $s_{11}^{dB}$ of transition type (III) with dimensions in mm $a = 0.5$, $b = 1$ and $d = 1.4$.

Transmission coefficient $s_{21}^{dB}$ of transition type (III) with dimensions in mm $a = 0.5$, $b = 1$ and $d = 1.4$.

HFSS simulation model of the chosen NRD guide with transitions of type (III) with $l = 20$ mm.

Reflection coefficient $s_{11}^{dB}$ of transition type (III) for $a = d$ and $b = 0.75$ mm.

Transmission coefficient $s_{21}^{dB}$ of transition type (III) for $a = d$ and $b = 0.75$ mm.

Metal waveguide to NRD guide transition measurement setup with $l = 32$ mm.

Cross-section of the transition measurement setup with dimensions in mm.

Top view of the transition measurement setup with dimensions in mm.

Picture of the measurement setup showing the experimental NRD guide attached to two metal waveguide sections connected to the VNA.

Measured reflection coefficient $s_{11}^{dB}$ of the experimental transition.

Measured transmission coefficient $s_{21}^{dB}$ of the experimental transition.

Spherical coordinate system used for antenna parameter computations.

Definition of the $E$ - and $H$ -plane of a vertical polarized antenna.

Near-field and far-field region surrounding an antenna with diameter $D$.

Foreshortened-top leaky-wave NRD guide antenna developed by Sanchez and Oliner. The leakage of the antenna is controlled by the stub length $s$.

Principle of operation of a periodic leaky-wave antenna.
3.6 Cross-section of a Sievenpiper high impedance surface structure \[37\] 60
3.7 Top view of a Sievenpiper high impedance surface structure \[37\] 61
3.8 NRD guide end-fire antenna model ................................. 62
3.9 Cross-sectional view of the NRD guide HFSS simulation model for the end-fire antenna with \(\mathbf{E}\) and \(\mathbf{H}\) symmetry planes with dimensions in mm \(a = 2a' = 0.5, b = 2b' = 1, d = 2d' = 1.4, t = 0.1\) and \(h = 2h'\) ................................................................. 63
3.10 Maximum gain \(G_0^{dB}\) (refer to Table 3.1) .................... 64
3.11 Radiation pattern of the NRD guide end-fire antenna ....... 65
3.12 HFSS 3D polar radiation plot of gain \(G^{dB}\) for the NRD guide end-fire antenna (a) – (c) without \((s = 0)\) and (d) – (f) with extended metal plates \((s = 1.5\ mm), h = 7\ mm\) ..................................................... 66
3.13 Comparison of the radiation patterns of the NRD guide and the DRW end-fire antenna for different values of the plate height \(h\) in mm for \(f_1 = 75\ GHz\) ................................................. 68
3.14 Front view of the NSI planar scanner .............................. 69
3.15 Side view of the end-fire antenna measurement setup ....... 70
3.16 Top view of the end-fire antenna measurement setup ....... 70
3.17 Scanning area with dimensions in m .............................. 71
3.18 Comparison of the simulated and measured far-field radiation patterns ................................................................. 74
3.19 Comparison of the simulated and measured far-field radiation patterns ................................................................. 75
3.20 2D contour plot of the measured radiation pattern for \(f_2 = 85\ GHz\) ................................................................. 76
3.21 Top view of the NRD guide with discontinuities separated by period \(p\) on top of the dielectric: (a) printed strips with \(dw = a\) and (b) printed stubs with \(dw < a\) ................................................. 78
3.22 Concept of NRD guide leaky-wave antenna: (a) side view and (b) cross-section with dimensions in mm \(a = 1.4, b = 0.5, s = 1.6, t = 0.25, p = 2, r = 0.01\) and \(o = 0.3\) .................................................. 78
3.23 Concept of periodic leaky-wave NRD guide antennas with PEC discontinuities loaded on the dielectric rod separated by distance \(p\) ................................................................. 79
3.24 Calculated dispersion characteristics \(\beta_{-1}/k_0 = f\) of the classical NRD guide with \(a = d\) and dimensions in mm \(a = 1.4\) and \(b = 0.5, f_c \approx 96.7\ GHz\) ................................................................. 80
3.25 E-plane radiation pattern of an NRD guide leaky-wave antenna with discontinuities of type (b) shown in Figure 3.21b with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$ ................................. 81

3.26 Beam angle $\theta_m$ of an NRD guide leaky-wave antenna with discontinuities of type (b) shown in Figure 3.21b on page 78 with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$ ................................. 82

3.27 Maximum gain $G_{dB}$ of the main lobe of an NRD guide leaky-wave antenna with discontinuities of type (b) shown in Figure 3.21b on page 78 with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$ ................................. 83

3.28 Tridimensional radiation pattern of the NRD guide leaky-wave antenna with discontinuities shown in Figure 3.21b for the frequency range from (a) 92 to (f) 97 GHz with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$ ................................. 84

3.29 Tridimensional radiation pattern of the NRD guide leaky-wave antenna with discontinuities shown in Figure 3.21b for the frequency range from (a) 98 to (f) 103 GHz with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$ ................................. 85

3.30 Tridimensional radiation pattern of the NRD guide leaky-wave antenna with discontinuities shown in Figure 3.21b for the frequency range from (a) 104 to (f) 109 GHz with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$ ................................. 86

3.31 E-plane radiation pattern of an NRD guide leaky-wave antenna with strip discontinuities of type (a) as shown in Figure 3.21a having the dimensions in mm $a = d = 1.4$, $b = 0.5$, $dl = 0.1$ and $p = 2$ ................................. 87

A.1 Simulation results of $k_z/k_0$ of the classical NRD guide with $a = d$ ................................. 92

A.2 Simulation results of the guide wavelength $\lambda_g$ of the classical NRD guide $a = d$ ................................. 93

A.3 Calculated attenuation constant $\alpha$ of the classical NRD guide with $a = d$ for $b = 0.5$ mm ................................. 94

A.4 Calculated dispersion diagram $k_z/k_0 - f$ of the classical NRD guide with $a = d = 1.4$ mm for different values of the height of the dielectric rod $b$ ................................. 95

A.5 Calculated dispersion diagram $k_z/k_0 - f$ of the classical NRD guide with $a = d$ ................................. 96
A.6 Calculated guide wavelength $\lambda_g$ of the classical NRD guide with $a = d$ ........................................ 97
A.7 Reflection coefficient $s_{11}^{dB}$ of transition type (I) for varied transition length $\nu$ ........................................ 99
A.8 Transmission coefficient $s_{21}^{dB}$ of transition type (I) for varied transition length $\nu$ ........................................ 100
A.9 Reflection coefficient $s_{11}$ of transition type (II) .......................... 101
A.10 Transmission coefficient $s_{21}$ of transition type (II) for varied transition length $\nu$ ........................................ 102
A.11 Reflection coefficient $s_{11}^{dB}$ of transition type (IV) ..................... 103
A.12 Transmission coefficient $s_{21}^{dB}$ of transition type (IV) ..................... 103
A.13 Reflection coefficient $s_{11}^{dB}$ of transition type (III) for in mm $a = d = 1.4$ and $b = 0.5$ ........................................ 104
A.14 Transmission coefficient $s_{21}$ of transition type (III) for in mm $a = d = 1.4$ and $b = 0.5$ ........................................ 104
A.15 Transmission coefficient $s_{21}^{dB}$ of transition type (III) for in mm $a = d = 1.4$ and $b = 0.5$ in the frequency range 85 to 110 GHz ............ 105
A.16 Reflection coefficient $s_{21}^{dB}$ of transition type (III) for $a = d = 1.4$ mm 106
A.17 Transmission coefficient $s_{21}^{dB}$ of transition type (III) for $a = d = 1.4$ mm 107
A.18 Comparison of the radiation patterns of the NRD guide and the DRW end-fire antenna for different values of the plate height $h$ in mm at $f_2 = 85$ GHz ........................................ 109
A.19 Comparison of the radiation patterns of the NRD guide and the DRW end-fire antenna for different values of the plate height $h$ in [mm] at $f_3 = 95$ GHz ........................................ 110
A.20 Measured $E$ - and $H$ -plane radiation pattern for $f_4 = 105$ GHz .... 111
A.21 Probe compensation $E$ -plane radiation pattern showing the smoothing effect ........................................ 112
A.22 $E$ -plane radiation pattern of the NRD guide leaky-wave antenna concept shown in Figure 3.22 ........................................ 113
A.23 $E$ -plane radiation pattern of the NRD guide leaky-wave antenna concept shown in Figure 3.22, with in mm $o = 0$ and $s = 0.5$ .... 113
A.24 Reflection coefficient $s_{11}^{dB}$ of an NRD guide leaky-wave antenna with discontinuities of type (b) shown in Figure 3.21b on page 78 with dimensions in mm $a = 1.4$, $b = 0.5$, $p = 2$, $dl = 0.5$ and $dw = 0.1$ 114
A.25 $E$ -plane radiation pattern for different dimensions of the discontinuities with $f = 95$ GHz and $p = 2$ mm .......................... 114
### LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>D.1</td>
<td>HFSS NRD model showing the simulation boundaries and the dimensions of the NRD guide and the dielectric rod in mm</td>
<td>119</td>
</tr>
<tr>
<td>D.2</td>
<td>Phase shift $\phi_{s21}$ at the output port compared to the incident phase at the input port</td>
<td>120</td>
</tr>
<tr>
<td>D.3</td>
<td>Magnitude of the electric field in the vertical plane of the NRD guide model at $f = 75 \text{ GHz}$ showing the number of full rotations the electromagnetic wave experience along the guide structure ($N = 8$)</td>
<td>121</td>
</tr>
<tr>
<td>D.4</td>
<td>Cumulative phase shift $\phi_{s21}$ along the NRD guide structure at the output port</td>
<td>121</td>
</tr>
<tr>
<td>D.5</td>
<td>Dispersion curve $k_z/k_0 - f$ of the NRD guide model</td>
<td>122</td>
</tr>
<tr>
<td>D.6</td>
<td>Guide wavelength $\lambda_g - f$ of the NRD guide model</td>
<td>122</td>
</tr>
<tr>
<td>E.1</td>
<td>Schematic model of the dielectric rod waveguide antenna with RWG to DRW transition</td>
<td>123</td>
</tr>
<tr>
<td>E.2</td>
<td>(a) Cross-section of the metal waveguide to DRW transition and (b) DRW end-fire antenna simulation cross-section with dimensions in mm $a = 2a' = 0.5$ and $b = 2b' = 1$</td>
<td>124</td>
</tr>
<tr>
<td>E.3</td>
<td>Radiation patterns of the DRW end-fire antenna</td>
<td>125</td>
</tr>
<tr>
<td>E.4</td>
<td>Normalized gain $G_{dB} - G_{0dB}^*$ of the DRW antenna for $f_1 = 75 \text{ GHz}$</td>
<td>126</td>
</tr>
<tr>
<td>E.5</td>
<td>Normalized gain $G_{dB} - G_{0dB}^*$ of the DRW antenna for $f_2 = 85 \text{ GHz}$</td>
<td>126</td>
</tr>
<tr>
<td>E.6</td>
<td>Normalized gain $G_{dB} - G_{0dB}^*$ of the DRW antenna for $f_3 = 95 \text{ GHz}$</td>
<td>127</td>
</tr>
</tbody>
</table>