Specification Level Optimization Based CAD of Waveguide Bandpass Filters

A Thesis
Submitted to the College of Graduate Studies and Research
in Partial Fulfillment of the Requirements
of the Degree of
Master of Science
in the Department of Electrical Engineering
University of Saskatchewan
Saskatoon

by

Jason Thomas Tucker

Spring 2000

© Copyright Jason Thomas Tucker, 1999. All rights reserved.
Permission To Use

In presenting this thesis in partial fulfilment of the requirements for a Postgraduate degree from the University of Saskatchewan, I agree that the Libraries of this University may make it freely available for inspection. I further agree that permission for copying of this thesis in any manner, in whole or in part, for scholarly purposes may be granted by the professor or professors who supervised my thesis work or, in their absence, by the Head of the Department or the Dean of the College in which my thesis work was done.

It is understood that any copying or publication or use of this thesis or parts thereof for financial gain shall not be allowed without my written permission. It is also understood that due recognition shall be given to me and to the University of Saskatchewan in any scholarly use which may be made of any material in my thesis.

Request for permission to copy or to make use of material in this thesis in whole or part should be addressed to:

Head of the Department of Electrical Engineering
University of Saskatchewan
57 Campus Drive
Saskatoon, Saskatchewan, Canada
S7N 5A9
Abstract

Modern satellite and terrestrial telecommunication systems usually employ a large number of waveguide filters operating at microwave frequencies. Traditionally, waveguide filters are designed empirically using approximate discontinuity models and extensive prototyping. Over the last two decades, the demand for accurate analysis and design techniques has generated a tremendous output of numerical methods for microwave applications. With the advent of powerful computers, field theory based models became desirable tools for computer aided design of waveguide filters. In this thesis, complete computer aided design methods for H-plane iris coupled waveguide bandpass filters are developed. A new technique using closed form equations to model iris discontinuities is discussed, as well as a new optimization approach which uses three optimization variables regardless of the order of the filter. Several filters designed using the new optimization approach are presented in this thesis. With the computer aided design methods discussed in this thesis, H-plane iris coupled waveguide bandpass filters can be quickly and accurately designed without post production tuning.
Acknowledgements

The author is indebted to Dr. Protap Pramanick for his supervision, financial support and constant encouragement. Thanks are due to the Department of Electrical Engineering, University of Saskatchewan, Saskatoon for providing the facilities to pursue this research.

It is a pleasure to acknowledge MCI Newhampshire and K & L Microwaves for providing the experimental results used within this work.

The author would also like to recognize his parents for the blessings that they are to him. They have been a constant source of love and encouragement throughout the author's life. Without them this research would not have been possible.

Finally, the author would like to acknowledge his Lord and Saviour, Jesus Christ, for His love and mercies. Without Jesus, life would not be as fulfilling as it is, and this research would never have been done. Jesus, you keep life interesting.
Dedication

This thesis is dedicated to Jesus Christ for His love, blessings and protection. This thesis is also dedicated to the author’s parents for their love and trust.
# Table of Contents

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Permission To Use</td>
<td>i</td>
</tr>
<tr>
<td>Abstract</td>
<td>ii</td>
</tr>
<tr>
<td>Acknowledgements</td>
<td>iii</td>
</tr>
<tr>
<td>Dedication</td>
<td>iv</td>
</tr>
<tr>
<td>Table of Contents</td>
<td>v</td>
</tr>
<tr>
<td>List of Tables</td>
<td>ix</td>
</tr>
<tr>
<td>List of Figures</td>
<td>x</td>
</tr>
<tr>
<td>List of Symbols</td>
<td>xii</td>
</tr>
<tr>
<td>1 Introduction</td>
<td>1</td>
</tr>
<tr>
<td>1.1 Filters</td>
<td>2</td>
</tr>
<tr>
<td>1.2 Rectangular Waveguide</td>
<td>5</td>
</tr>
<tr>
<td>1.2.1 Waveguide Modes</td>
<td>6</td>
</tr>
<tr>
<td>1.2.2 Waveguide Impedance</td>
<td>9</td>
</tr>
<tr>
<td>1.2.3 TE_{lo} - Mode</td>
<td>10</td>
</tr>
<tr>
<td>1.3 Microwave Filters</td>
<td>10</td>
</tr>
<tr>
<td>1.3.1 Traditional &amp; Conventional Filter Design Approaches</td>
<td>12</td>
</tr>
<tr>
<td>1.3.2 New Approach for Optimization</td>
<td>14</td>
</tr>
<tr>
<td>1.4 Objective of Research</td>
<td>16</td>
</tr>
<tr>
<td>1.5 Thesis Overview</td>
<td>17</td>
</tr>
<tr>
<td>2 Literature Review and Background</td>
<td>19</td>
</tr>
</tbody>
</table>
List of Tables

6.1 Design Parameters for Initial and Optimized Symmetric X Band Filter 80
6.2 Iris Width Dimensions for Initial and Optimized Symmetric X Band Filter 80
6.3 Resonator Lengths for Initial and Optimized Symmetric X Band Filter 80
6.4 Design Parameters for Initial and Optimized Asymmetric X Band Filter 83
6.5 Iris Width Dimensions for Initial and Optimized Asymmetric X Band Filter 83
6.6 Resonator Lengths for Initial and Optimized Asymmetric X Band Filter 83
6.7 Design Parameters for Initial and Optimized Symmetric KU Band Filter 86
6.8 Width Dimensions for Initial and Optimized Symmetric KU Band Filter 86
6.9 Resonator Lengths for Initial and Optimized Symmetric KU Band Filter 86
6.10 Design Parameters for Initial and Optimized Asymmetric KU Band Filter 89
6.11 Width Dimensions for Initial and Optimized Asymmetric KU Band Filter 89
6.12 Resonator Lengths for Initial and Optimized Asymmetric KU Band Filter 89
6.13 Design Parameters for Initial and Optimized Double Iris Filter 92
6.14 Width Dimensions for Initial and Optimized Double Iris Filter 92
6.15 Resonator Lengths for Initial and Optimized Double Iris Filter 92
6.16 Design Parameters for Initial and Optimized Double Iris Filter 95
6.17 Width Dimensions for Initial and Optimized Double Iris Filter 95
6.18 Resonator Lengths for Initial and Optimized Double Iris Filter 95
## List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>Response for filters: a) Low-Pass, b) High-Pass, c) Band-Pass, d) Band-Stop.</td>
<td>2</td>
</tr>
<tr>
<td>1.2</td>
<td>General filter representation.</td>
<td>3</td>
</tr>
<tr>
<td>1.3</td>
<td>Different types of filter response: a) Butterworth, b) Chebyshev, c) Elliptic.</td>
<td>4</td>
</tr>
<tr>
<td>1.4</td>
<td>Rectangular waveguide.</td>
<td>6</td>
</tr>
<tr>
<td>1.5</td>
<td>a) Iris coupled filter. b) Septum coupled filter.</td>
<td>11</td>
</tr>
<tr>
<td>2.1</td>
<td>Step discontinuities: a) H-plane, b) E-plane, c) EH-plane.</td>
<td>22</td>
</tr>
<tr>
<td>2.2</td>
<td>Symmetric and asymmetric junctions and their equivalent circuits</td>
<td>23</td>
</tr>
<tr>
<td>2.3</td>
<td>H-plane iris in a rectangular waveguide: a) Symmetric, b) Asymmetric, c) Double symmetric iris.</td>
<td>24</td>
</tr>
<tr>
<td>2.4</td>
<td>An N-port junction illustrating scattering waves.</td>
<td>25</td>
</tr>
<tr>
<td>2.5</td>
<td>Section of lossless uniform transmission line.</td>
<td>28</td>
</tr>
<tr>
<td>3.1</td>
<td>3.D view of an H-plane step discontinuity.</td>
<td>33</td>
</tr>
<tr>
<td>3.2</td>
<td>Front view of: a) Asymmetrical discontinuity, b) Centred discontinuity.</td>
<td>37</td>
</tr>
<tr>
<td>3.3</td>
<td>Composed scattering matrices for an iris of finite thickness.</td>
<td>43</td>
</tr>
<tr>
<td>4.1</td>
<td>Typical frequency response of a bandpass filter.</td>
<td>46</td>
</tr>
<tr>
<td>4.2</td>
<td>Waveguide bandpass filter with H-plane iris discontinuities.</td>
<td>48</td>
</tr>
<tr>
<td>4.3</td>
<td>a) Impedance inverter. b) Impedance inverter used to convert a parallel admittance into an equivalent series impedance.</td>
<td>49</td>
</tr>
<tr>
<td>4.4</td>
<td>Equivalent network of a two-step symmetrical H-plane discontinuity.</td>
<td>50</td>
</tr>
</tbody>
</table>
4.5 Equivalent models of corrugated waveguide filter including impedance inverters: a) Bandpass filter prototype, b) Bandpass filter containing impedance inverters & shunt resonators.

4.6 Frequency response of a asymmetric iris coupled filter designed by the K-inverter method.

4.7 Flowchart for rational function approximation.

4.8 Frequency response of a symmetric iris coupled filter designed by the closed form equations.

5.1 Brute force optimization flowchart.

5.2 K-inverter optimization flowchart.

5.3 Specification level optimization flowchart.

5.4 Scheme for the computer optimization.

6.1 Isolation loss for initial and optimized X-band symmetric filter design.

6.2 Return loss for initial and optimized X-band symmetric filter design.

6.3 Isolation loss for initial and optimized X-band asymmetric filter design.

6.4 Return loss for initial and optimized X-band asymmetric filter design.

6.5 Isolation loss for initial and optimized KU-band symmetric filter design

6.6 Return loss for initial and optimized KU-band symmetric filter design.

6.7 Isolation loss for initial and optimized KU-band asymmetric filter design.

6.8 Return loss for initial and optimized KU-band asymmetric filter design.

6.9 Isolation loss for unoptimized, optimized and manufactured double iris filter.

6.10 Return loss for unoptimized, optimized and manufactured double iris filter.

6.11 Isolation loss for unoptimized, optimized and manufactured double iris filter.

6.12 Return loss for unoptimized, optimized and manufactured double iris filter.
## List of Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a)</td>
<td>broad dimension of rectangular waveguide</td>
</tr>
<tr>
<td>(b)</td>
<td>narrow dimension of rectangular waveguide</td>
</tr>
<tr>
<td>(c)</td>
<td>speed of light</td>
</tr>
<tr>
<td>(f)</td>
<td>frequency</td>
</tr>
<tr>
<td>(f_L)</td>
<td>lower cutoff frequency of bandpass filter</td>
</tr>
<tr>
<td>(f_H)</td>
<td>upper cutoff frequency of bandpass filter</td>
</tr>
<tr>
<td>(f_{c.n.m})</td>
<td>waveguide cutoff frequency for mode n,m</td>
</tr>
<tr>
<td>(F_{on})</td>
<td>bandpass filter centre frequency</td>
</tr>
<tr>
<td>(R)</td>
<td>either the width of the iris or the phase associated with the obstacle</td>
</tr>
<tr>
<td>(I)</td>
<td>coupling integral</td>
</tr>
<tr>
<td>(P_f)</td>
<td>penalty function</td>
</tr>
<tr>
<td>(P_{LR})</td>
<td>power loss ratio</td>
</tr>
<tr>
<td>(\beta_{n.m})</td>
<td>propagation factor</td>
</tr>
<tr>
<td>(\varepsilon_0)</td>
<td>permittivity of free space</td>
</tr>
<tr>
<td>(\mu_0)</td>
<td>permeability of free space</td>
</tr>
<tr>
<td>(\lambda)</td>
<td>wavelength</td>
</tr>
<tr>
<td>(\lambda_g)</td>
<td>guide wavelength</td>
</tr>
<tr>
<td>(\lambda_{go})</td>
<td>midband guide wavelength</td>
</tr>
<tr>
<td>(\lambda_{gL})</td>
<td>guide wavelength in the resonator section at (f_L)</td>
</tr>
<tr>
<td>(\lambda_{gH})</td>
<td>guide wavelength in the resonator section at (f_H)</td>
</tr>
</tbody>
</table>
\( \lambda_o \quad \text{free space wavelength} \\
\lambda_{c,n,m} \quad \text{cutoff wavelength for mode n,m} \\
k_{c,n,m} \quad \text{wave number for mode n,m} \\
E \quad \text{electric field} \\
E_t \quad \text{tangential electric field} \\
H \quad \text{magnetic field; coupling matrix} \\
H_t \quad \text{tangential magnetic field} \\
\eta \quad \text{impedance of free space} \\
N_f \quad \text{filter order} \\
Z_n \quad \text{impedance of distributed elements} \\
Z_o \quad \text{characteristic impedance} \\
Z_{n,m} \quad \text{characteristic impedance for mode n,m} \\
\text{IL} \quad \text{insertion loss} \\
\text{RL} \quad \text{return loss} \\
t \quad \text{thickness of the iris} \\
[a^-] \quad \text{amplitude of reflected waves} \\
[a^+] \quad \text{amplitude of incident waves} \\
[S] \quad \text{scattering matrix} \\
[S]^t \quad \text{transpose scattering matrix} \\
[S]^* \quad \text{conjugate scattering matrix} \\
[U] \quad \text{unitary matrix} \\
[Y] \quad \text{admittance matrix} \\

xiii
A⁺ amplitudes of waves on larger side of iris discontinuity
B⁺ amplitudes of waves on smaller side of iris discontinuity
γ isolation bandwidth factor
\(P, Q, N, M\) integers, number of modes in the waveguide
\(k, p, q, n, m\) integers, waveguide mode numbers
\(j\) \(\sqrt{-1}\)
i, j integers, indices
e passband ripple
\(L_A\) isolation
\(B W\) fractional bandwidth
\(x_s\) normalized series impedance
\(X_s\) unnormalized series impedance
\(x_p\) normalized parallel susceptance
\(X_p\) unnormalized parallel susceptance
\(k'\) impedance inverter value
\(K_{n,n+1}\) normalized impedance inverter value
\(K_{v}\) impedance inverter value before scaling
\(K\) impedance inverter value after scaling
\(l_{Ri}\) length of waveguide resonator section i
\(w_i\) width of iris opening i
\(A_w, B_w, C_w\) coefficients for the width equation
\(D_w, E_w, F_w, G_w\)
$A_p$, $B_p$, $C_p$, coefficients for the phase equation

$D_p$, $E_p$, $F_p$, $G_p$

$A_F$, $B_F$, $C_F$, $D_F$

$\phi$, associated phase with impedance inverter

$\alpha$, impedance scaling parameter

$\psi$, impedance inverter scaling factor

$P_o$, total power flowing through a waveguide

TE, transverse electric

TM, transverse magnetic

TEM, transverse electric and magnetic

V, voltage

weight, relationship between the pass and stop band in the penalty function
Chapter 1

Introduction

The term microwaves is used to describe electromagnetic waves with wavelengths ranging from 1 m to 1 cm. The corresponding frequency range for the waves is 300 MHz up to 30 GHz. One of the most important aspects of microwaves and their components is that the size of the structure is comparable to that of the wavelength. Due to the small wavelengths, there are several advantages in the use of microwaves. For example, the short wavelength allows microwaves to travel through the ionosphere, enabling information to be sent or received by line-of-sight. The high frequencies of microwaves also provide a larger bandwidth (information-carrying capacity) in communication links. All of these unique properties have led to the wide use of microwaves in modern communication systems, defence, and industry. Due to the ever growing demand for faster and better methods of wireless communication, a considerable amount of research and development in various areas of microwave engineering has been done over the past fifty years.

Conventional circuit theories, such as Kirchhoff’s laws and other voltage-current
concepts, can not be used directly to solve microwave network problems accurately. This is because the size of wavelengths and the size of the structure become comparable at microwave frequencies. In order to overcome this obstacle, rigorous field analysis based on Maxwell’s equations, which is generalized circuit theory, and the boundary conditions must be used. It is necessary to adapt Maxwell’s equations to the desired network in order to solve for the different field components.

1.1 Filters

A filter is used to control the frequency response between a pair of ports in an electrical circuit by providing transmission at frequencies within the passband of the filter, and attenuation in the stop-band of the filter [1]. Filters are classified as low-pass, high-pass, band-pass and band-stop. The frequency responses of these basic types of filters are shown in Figure 1.1.

Figure 1.1: Response for filters: a) Low-Pass, b) High-Pass, c) Band-Pass, d) Band-Stop.
Figure 1.2 shows the general representation of a filter in a microwave network. At the
input plane of the filter, the power is divided into three components:

- $P_{in}$, the incident power from the generator.
- $P_R$, the power reflected towards the generator.
- $P_A$, the power absorbed by the filter.
- $P_L$, the power transmitted to the load.

![Diagram of filter representation](image)

*Figure 1.2: General filter representation.*

By conservation of energy,

$$P_{in} = P_R + P_A$$

If $P_L = P_A$, the filter is said to be lossless; in addition $P_L = P_{in}$ implies that there are no
reflections and the filter is lossless and matched. The insertion loss (IL) and the return
loss (RL) at a particular frequency are defined as,

$$IL = -10\log\frac{P_L}{P_{in}}, \quad RL = -10\log\frac{P_R}{P_{in}}.$$ (1.2)

The units for both IL and RL are decibels (dB). Ideally, the passband insertion loss
should be 0 dB and the stop band insertion loss should approach infinity. However, filters in practice have a frequency response deviating from the ideal filter response.

Filters can be classified as Butterworth, Chebyshev or Elliptic type filters based on their frequency response characteristics as shown in Figure 1.3.

Butterworth filters have a flat passband response. Chebyshev filters have an equal-ripple characteristic in the passband, while elliptic filters exhibit equal-ripple characteristics in both the passband and the stopband. Chebyshev filters have a sharp cutoff and are often preferred over Butterworth filters. Elliptic filters have an even
sharper cutoff but are very difficult to design precisely for use at microwave frequencies.

The focus of this thesis is on the design of Chebyshev bandpass filters.

### 1.2 Rectangular Waveguides

Transmission lines and waveguides are primarily used to realize passive elements at microwave frequency and to distribute microwave signals from one point to another. Waveguides come in many different shapes and sizes, but rectangular and circular waveguides are the most common. Of the two, rectangular waveguides are the most widely used, and in this thesis rectangular waveguides are adopted for the design of waveguide filters. Two different kinds of wave propagation can exist in a rectangular waveguide, these being transverse electric (TE) and transverse magnetic (TM). If the structure supports waves with a purely magnetic field (H) in the direction of propagation, then the waves are said to be TE. Similarly, if there exists a purely electric field (E) in the direction of propagation, then the waves are said to be TM. There are some situations with inhomogeneous dielectric filling or anisotropic material filling in the waveguide where the waveguide modes have both longitudinal electric and magnetic fields. These are known as hybrid modes.

A rectangular waveguide, having width $a$ and height $b$, that can propagate both TE and TM waves is illustrated in Figure 1.4. The waveguide transmits electromagnetic waves through its interior in the $z$ direction by multiple reflections from its metallic walls.
Perfectly conducting walls are located at $x = 0, a$ and $y = 0, b$. The walls are perfectly conducting, because it is assumed that it is a lossless system. The waveguide is filled with a dielectric of permittivity $\varepsilon_r$. Throughout this thesis the width and height of all waveguides are represented as $a$ and $b$ with $a > b$, and air ($\varepsilon_r = 1$) is the dielectric filling.

![Figure 1.4: Rectangular waveguide.](image_url)

1.2.1 Waveguide Modes

One of the most important properties of empty loss-free waveguides is that there exists an infinity of possible solutions for both TE and TM waves. These waves, or modes, are labelled by identifying integer subscripts $n$ and $m$. All propagating waves are labelled in the following manner, $\text{TE}_{nm}$. The integers $n$ and $m$ pertain to the number of standing-wave interference maxima occurring in the field solutions that describe the fields along the two transverse coordinates ($X$ and $Y$ directions) [2]. Each mode has its own
characteristic cutoff frequency $f_{c,nm}$. Below this cutoff frequency the mode does not propagate and above which the mode does propagate. The cutoff frequency is dependent on the geometrical shape (the width and height) of the waveguide. The cutoff frequency $f_{c,nm}$ is given by

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

(1.3)

where $c$ is the velocity of light ($c = \frac{1}{\sqrt{\mu_0\varepsilon_0}}$).

Propagation of waves in the $z$ direction is governed by the function $e^{\pm\beta_{nm}z}$. The value $e^{\pm\beta_{nm}z}$ is the position dependent component of a plane wave moving in the positive or negative $z$ direction. The propagation constant $\beta_{nm}$ is given by

$$\beta_{nm} = \sqrt{(k_o^2 - k_{c,nm}^2)},$$

(1.4)

where

$$k_o = \omega \sqrt{\mu_0\varepsilon_0}, \quad k_{c,nm} = 2\pi f_{c,nm} \sqrt{\mu_0\varepsilon_0},$$

(1.5)

and $\mu_0$ and $\varepsilon_0$ are the permeability ($4\pi \times 10^{-7}$ H/m) and permittivity ($8.845 \times 10^{-12}$ F/m) of free space respectively, and $\omega = 2\pi f$, where $f$ is the operating frequency. The guide wavelength is given by

$$\lambda_g = \frac{\lambda_o}{\sqrt{1 - \frac{\lambda_o^2}{\lambda_{c,nm}^2}}},$$

(1.6)
where $\lambda_o$ is the free-space wavelength of plane waves, and the cutoff wavelength is given by

$$ \lambda_{c, nm} = \frac{2\pi}{k_{c, nm}}. \quad (1.7) $$

Maxwell’s equations for free space are

$$ \nabla \times H = j\omega e E, \quad (1.8a) $$

$$ \nabla \times E = -j\omega \mu H, \quad (1.8b) $$

$$ \nabla \cdot E = 0, \quad (1.8c) $$

$$ \nabla \cdot H = 0. \quad (1.8d) $$

Since sources are not considered, the electric and magnetic fields are solutions of the Helmholtz (or reduced) wave equation. The Helmholtz equation is given by

$$ \nabla^2 H + k_o^2 H = 0 \quad \text{or} \quad (1.9a) $$

$$ \nabla^2 E + k_o^2 E = 0. \quad (1.9b) $$

The field components for the TE_{nm} mode are found by expanding Maxwell’s equations and solving Helmholtz equation (1.9a) for the axial magnetic field ($H_z$) at the cutoff frequency. The field components are

$$ H_z = \cos \frac{n\pi x}{a} \cos \frac{m\pi y}{b} \ e^{-j\beta_{nm} z}, \quad (1.10) $$
\[ H_x = \frac{j \beta_{nm} n \pi}{ak_{c, nm}^2} \sin \left( \frac{n \pi x}{a} \right) \cos \left( \frac{m \pi y}{b} \right) e^{-j \beta_{nm} Z}, \quad (1.11) \]

\[ H_y = \frac{j \beta_{nm} n \pi}{bk_{c, nm}^2} \cos \left( \frac{n \pi x}{a} \right) \sin \left( \frac{m \pi y}{b} \right) e^{-j \beta_{nm} Z}, \quad (1.12) \]

\[ E_z = 0, \quad (1.13) \]

\[ E_x = Z_{h, nm} H_y, \quad (1.14) \]

\[ E_y = -Z_{h, nm} H_x, \quad (1.15) \]

where the characteristic wave impedance for the \( nm \)th TE mode \( Z_{h, nm} \) is given by

\[ Z_{h, nm} = \frac{120 \pi}{\sqrt{1 - \left( \frac{\lambda_o}{\lambda_{c, nm}} \right)^2}} \text{ ohms}. \quad (1.16) \]

**1.2.2 Waveguide Impedance**

The wave impedance in a waveguide is defined as

\[ Z_{h, nm} = \frac{E_t}{H_t}, \quad (1.17) \]

where the subscript \( t \) is used to denote the components of the field perpendicular to the direction of propagation. For TE-modes

\[ Z_{h, nm} = \eta_o \frac{\lambda_g}{\lambda_o}, \quad (1.18) \]
where the impedance of free space \( \eta_0 \) is given by

\[
\eta_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} \approx 120\pi \text{ ohms.} \tag{1.19}
\]

For TE modes in a waveguide, the wave impedance is a constant for any particular mode irrespective of the position in the cross-section of the waveguide at which \( E_i \) or \( H_i \) are considered. The characteristic impedance, with respect to the voltage \( V \) at \( x = \frac{a}{2} \) and the total power \( P_o \) flowing through the waveguide is given by

\[
Z_0 = \frac{V^2}{2P_o} = \frac{2b}{a} Z_{h,nn}, \tag{1.20}
\]

which is equal to the wave impedance when \( a = 2b \).

### 1.2.3 \( \text{TE}_{10} \) - Mode

The \( \text{TE}_{10} \) mode, the mode for which the cutoff frequency is minimum when \( a > b \), is called the dominant mode of a rectangular waveguide. In this thesis, the dominant mode is considered to be the incident wave on the waveguide filter unless specified otherwise. The cutoff wavelength for the dominant mode is given by (see Equations (1.3 to 1.7))

\[
\lambda_{c,10} = 2a. \tag{1.21}
\]

### 1.3 Microwave Filters

Filters at microwave frequencies are realized using distributed elements (transmission
ines and waveguides) or a combination of lumped (capacitors and inductors) and
distributed elements. Electric and magnetic fields have to be considered in the design of
filters using distributed elements, rather than the voltage and current as in the case of
lumped elements. Designing filters using distributed elements is more involved, and the
synthesis involves complicated procedures requiring empirical adjustment to
experimental filters for obtaining a filter with satisfactory frequency response.

Waveguide bandpass filters are realized using coupled resonators. Coupling is achieved
using either irises or septums. The concept of resonators, irises and septums is
discussed in later chapters. Examples of iris coupled and septum coupled filters can be
seen in Figure 1.8. In this thesis the filters are realized using iris coupling.

![Figure 1.5: a) Iris coupled filter. b) Septum coupled filter. [3]](image)
Bandpass filters are the most typical passive component in any microwave network. Waveguides lend themselves to use in high power and low loss applications. It is because of this fact that they are used extensively in cellular communication, satellite communication and defence applications.

1.3.1 Traditional Filter Design Approaches

The traditional method of waveguide filter design leaves something to be desired. Design was based on empirical models for the discontinuities, and to achieve the proper frequency response a trial and error approach was adopted [4]. Upon completion of a design, the waveguide would have to be fine tuned by the use of tuning screws. The presence of tuning screws in the filter structures caused undesired vibration problems and loss in the signal. Moreover, the use of tuning screws makes the human element an important part of the manufacturing process, which in turn slows it down and makes the filters expensive. Present filter design allows engineers to create waveguide filters which do not need post production tuning. The use of modern network theory, field theory and brute force optimization techniques makes it possible for such filters to be realized [5,6,7]. In brute force optimization techniques, the variables are the resonator lengths $l_{ri}$ and the iris widths $w_i$ (see Figure 1.8a). This means that with brute force optimization, the iris widths and resonator lengths are directly adjusted in each iteration.

With brute force optimization the likelihood of convergence to a local minima is relatively high, and therefore the initial design of the filter is very important for
convergence to a final solution. Traditionally, the optimization is done using the quasi-
Newton method [8] with variables altered following certain rules until the desired
frequency response is obtained. Since the needed variables for optimization are the
resonator lengths \( l_r \) and the iris widths \( w_i \), in the design of an \( N_f \) pole filter of a
symmetrical structure, it would require a set of \( N_f + 1 \) variables for optimization. From
this it can be seen that the time to completion for the design is directly related to the
order of the filter. In addition, the number of independent variables contributes to the
possibility of convergence of the optimization to a local minimum.

Another approach for the design of filters, is the optimization of K-inverters [5,6]. A K-
inverter is an idealized impedance inverter. An impedance inverter has the ability to
make a series impedance appear like a shunt admittance. For example, a series
inductance with an inverter on each side looks like a shunt capacitance from its exterior
terminals. The subject of K-inverters is covered in more depth in a later chapter. In the
synthesis of a filter, \( N_f + 1 \) K-inverters are needed to realize it, which in the case of a
symmetric structure leads to \( N_f/2 + 1 \) variables needed for optimization. This method is
greater than that of optimizing the widths of the irises and lengths of the resonators
directly, but time to completion of a design and the possibility of convergence of the
optimization process to an undesired local minimum is still dependent on the order of
the filter. This leads to problems when large order filters are to be designed.
1.3.2 New Approach for Optimization

Brute force optimization is an effective tool for synthesizing waveguide filters. However, as previously mentioned, there are two problems with this method of design; those being the time to completion for one filter design and convergence of the optimization process to a local minimum. Currently, for a waveguide with given cross-sectional dimensions (See Figure 1.8a), the variables to be optimized are either the resonator lengths and widths of the K-inverter forming irises [5,6], or the K-inverter values themselves. This leads to a problem with larger order filters. The problem is the larger the order of the filter, the more variables with which the designer must work. This not only creates a long time to design completion problem, but it also increases the likelihood of convergence to a local minimum in optimization. What is needed is an optimization based design approach where the number of optimization variables are kept to a minimum. Also the number of variables should be independent of the order of the filter.

This thesis takes a different approach than those mentioned earlier. It appears that any time a filter is designed to a specification based on circuit theories, analysis will show characteristics of a filter other than that which was desired. For example, it is known that in using the K-inverter method of design there will result a filter response that has a narrower bandwidth, lower return loss and shifted centre frequency than that which was desired. This means that for any desired filter design there is generated a wrong filter response. Knowing this, there must be a “shadow” design specification out there that,
when used as the initial input of the design program, will produce the needed response. Therefore, the object of this new technique is to find the shadow filter design which will in turn produce the correct response. Any design program using this method will create a filter and subsequently analyze it and check to see if the response matches the desired one. If the response does not match the desired response, the input parameters to the filter are changed and the design process is started all over again. This is continued until the correct filter response is finally reached. The fact that variation of K-inverters indirectly varies the filter specifications was mentioned in [5], although Reference [5] optimized the K-inverters only. Reference [5] dealt with very low order filters using symmetric irises ignoring frequency dependence of the prototype inverters.

If this technique is used in optimization, the CAD program will only have to deal with three optimization variables. In other words, the number of variables is independent of the order of the filter. The variables needed for optimization are the upper and lower cutoff frequencies and the passband return loss. It should be noted that as these input parameters are changed, the order of the filter is kept the same throughout the design process. This method of optimization will greatly cut down on the time to completion for a design and the chance of convergence to a local minimum, because the number of optimization variables will remain constant at three no matter what the order of the filter is. The speed of this approach is directly effected by the method of modeling used for the filter structure. In order for this method of optimization to reach its full potential, a fast and accurate method of modeling the K-inverters is needed.
1.4 Objective of Research

Computers are powerful tools which can be used to perform tedious iterative calculations faster, cheaper and more accurately than a human can. The advances in computational technology and numerical algorithms over the last few decades has given engineers the ability to design microwave components at the very cutting edge. With the increased demand for faster and more accurate design of microwave components, full-wave analysis using the field theory approach became a necessity. The superiority of the field-theory approach over the equivalent circuit models has become increasingly attractive to design engineers.

Waveguide filters are an important part of modern communication systems. With the swift growth in personal communication systems, the demand for accurate design and analysis of waveguide filters has been on the increase. With the combination of modern network theory, field theory and CAD (Computer Aided Design) tools, the fast, accurate and compact design of waveguide filters is possible.

The main objectives of the research described in this thesis are:

1. To develop efficient computer aided design and analysis tools for H-plane iris coupled filters.

2. To develop a faster method of modeling using the K-inverter design approach.
3. To implement a new optimization technique that will utilize 3 optimization variables, irrespective of the filter order, in the design of H-plane iris coupled bandpass waveguide filters.

4. To reduce the likelihood of convergence to a local minimum in optimization.

5. To use the developed CAD software to design symmetrical, asymmetrical and double iris coupled waveguide filter structures.

The filters designed in this thesis do not require post-production tuning. They are superior in performance to their experimentally tuned counterparts, because of the absence of tuning screws which can introduce many problems, including loss and vibration. This will be advantageous when it comes to manufacturing and mass production of the filters.

1.5 Thesis Overview

The rest of this thesis is divided into the following chapters:

Chapter 2 presents the background and the literature survey of the related subject of research. Waveguide discontinuities along with the general concept of wave scattering is introduced in this chapter. The theory and the analytical aspects behind the analysis approach (the mode matching method) taken in this thesis is presented in Chapter 3.

Chapter 4 discusses general filter synthesis. This chapter also introduces two
approaches for the synthesis of H-plane band pass waveguide filters; the K-inverter
design approach with root seeking and the modified K-inverter design approach with
interpolation. Chapter 5 is dedicated to the discussion of optimization. This chapter
begins with a general discussion of existing optimization techniques, and then it presents
a new optimization technique. Validation of the design technique can be found in
Chapter 6. Chapter 7 contains the conclusions of the research.
Chapter 2

Literature Review and Background

2.1 Literature Survey

Microwave filters are a typical passive component found in microwave networks. Work on microwave filters commenced prior to the Second World War, but in particular significant developments started in 1937. Driven by the war effort, major advances were made at various laboratories in the United States and a few other countries during the years of 1941 to 1945. A good historical account of the research done in microwave filter design and development can be found in [9]. Concise summaries of the development in the theory and design of microwave filters are given in [3,8,10,11]. Much of the earlier work done in filter design was based on Marcuvitz’s [10] empirical formulae and models for waveguide discontinuities. Many engineers held Marcuvitz’s work as the foundation for waveguide component design.

With the increase in the availability and capability of computers, researchers have shifted their focus to the numerical characterization and modelling of waveguide filter components. The manipulation and solution of large sets of linear simultaneous
equations is basic to most of the techniques employed in computer solutions of field problems. An introduction to numerical analysis is given in [12]. Although complete numerical 3D analysis techniques, such as the finite element or finite difference approaches, have been commercially available, the analytical mode-matching method proves to be the fastest and the most efficient for analysing and optimizing components having well defined boundaries [3,5,13,14].

Transverse and longitudinal waveguide junctions form the building blocks of waveguide filters, directional couplers and periodic structures. It is because of the significance of waveguide junctions that considerable research in solving waveguide discontinuity problems has been undertaken. The problems of scattering from a rectangular-to-rectangular waveguide junction were exactly solved with convergent numerical results using the mode-matching (TE-mode and TM-mode expansions) method and the principle of conservation of complex power in [3,11,13-16]. This method allows the user to analyse an arbitrary waveguide cross section deformation (discontinuity) and deduce it’s reflection and transmission parameters.

As mentioned previously, most waveguide filter designs were founded on empirical relationships and approximate models for waveguide discontinuities taken directly from [10]. Designs were based on low-pass prototypes or the quarter-wave transformer prototypes, and upon fabrication the filters were tuned with tuning screws in order to get the desired bandwidth and centre frequency. Today better characterizations of
waveguide discontinuities is possible using a multimodal field approach, coupled with numerical techniques. One such approach is taken in [5,6,7,14], where double-step discontinuities are used as K-inverters. The accurate prediction of the K-inverters is not guaranteed because of the higher order mode interactions between adjacent K-inverters. Another approach is to solve for the exact scattering parameters of the junction at a few frequencies, for a few different dimensions. This information can be used to create equations which in turn can be used to model the K-inverter parameters continuously. A similar method can be found in [7], and will be referred to often throughout this thesis.

In this thesis, a successful attempt is made to develop a computer aided method for the design of H-plane coupled resonator bandpass waveguide filters. In the process, a new design and optimization technique is implemented which greatly reduces, as compared to other design methods, the time to completion for a design, and also reduces the likelihood of convergence to a local minima in frequency response optimization.

2.2 Waveguide Discontinuities

Waveguide components are composed of regions containing not only uniform or nonuniform waveguide sections, but also discontinuities. Discontinuities are regions within the waveguide where there is an abrupt change in the cross-sectional shape. These discontinuities are deliberately introduced into the waveguide to perform a certain electrical function. There is an infinite number of possibilities of shapes for discontinuities, but the most common for a rectangular waveguide are the E-plane, H-
plane, and EH-plane step discontinuities. An H-plane step is a junction of two rectangular guides of unequal widths but equal heights. An E-plane step is a junction of two rectangular guides of unequal heights but equal widths. An EH-plane step is a junction of two rectangular guides of unequal widths and heights. These structures can be seen in Figure 2.1.

![Figure 2.1: Step discontinuities: a) H-plane, b) E-plane, c) EH-plane.](image)

Discontinuities can also be classified as symmetric and asymmetric step discontinuities, which can be seen in Figure 2.2. This thesis focuses on both symmetric and asymmetric H-plane discontinuities.
Symmetrical Inductive discontinuity

Asymmetrical Inductive discontinuity

Symmetrical Capacitive discontinuity

Asymmetrical Capacitive discontinuity

Figure 2.2: Symmetric and asymmetric junctions and their equivalent circuits.

2.2.1 Iris Discontinuity

In its simplest form, a waveguide iris can be thought of as two waveguides with exacting widths and heights separated by a waveguide having smaller dimensions. The iris itself is the physically smaller waveguide that is placed between the other two waveguides. In an H-plane iris, the height of the waveguide remains constant while the width of the waveguide changes. Similarly, E-plane and EH-plane step junctions can be connected back to back with a small waveguide section to form E-plane and EH-plane irises. In this thesis only H-plane irises are considered. Figure 2.3 shows examples of small waveguide sections with an H-plane iris of thickness t and width w inserted.
Figure 2.3: H-plane iris in a rectangular waveguide: a) Symmetric, b) Asymmetric, c) Double symmetric iris.

A complete characterization of the waveguide iris involves the determination of the frequency-dependent transmission and reflection coefficients associated with the discontinuity.

2.3 Scattering Parameters

Practical problems exist when trying to measure voltages and currents at microwave frequencies. The quantities that are directly measurable are the standing wave ratio, location of a field minimum, and power. Other parameters that are directly measurable are the transmission and reflection coefficients, these being a relative measurement of the amplitude and phase of the transmitted and reflected waves as compared with those of the incident wave. The scattering parameters are defined in terms of the incident and reflected waves, and require transmission lines that are terminated in their characteristic impedances.

Consider the N-port junction in Figure 2.4. The normalized amplitudes of the incident waves are denoted by the ‘+’ superscript while the amplitudes of the reflected waves are
denoted by the ' -' superscript, and the port number is denoted as the subscript of the amplitude wave.

When $a_1^-$ is incident on port 1 and all other ports are terminated with a matched load, a reflected wave $a_1^- = S_{11} a_1^-$ is produced, where $S_{11}$ is known as the reflection or scattering coefficient at port 1. Similarly, for all the other ports, $a_i^- = S_{ii} a_i^+$ for $i = 2,3,...N$ are defined. The waves, in addition to reflection, will be transmitted to other ports and will have a proportional output of the incident wave. $a_i^- = S_{ik} a_k^+$ for $i = 2,3,...N$ and $k = 1,2,3,...N$, gives the transmission coefficient $S_{ik}$, which

\[ \text{Figure 2.4: An N-port junction illustrating scattering waves.} \]

\[ a_1^- \quad a_1^+ \quad a_2^- \quad a_2^+ \quad a_3^- \quad a_3^+ \]

Reference Plane
represents the wave transmitted from port $k$ to the $i^{th}$ port. Thus, when waves are incident on all ports simultaneously

\[
\begin{bmatrix}
a_1^-
\end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & \cdots & S_{1N} \\
S_{21} & S_{22} & \cdots & S_{2N} \\
\vdots & \vdots & \ddots & \vdots \\
S_{N1} & S_{N2} & \cdots & S_{NN}
\end{bmatrix} \begin{bmatrix}
a_1^+
\end{bmatrix},
\]

or in a simplified form

\[
[a^-] = [S][a^+],
\]

where $[S]$ is called the scattering matrix of the network.

For a passive, reciprocal, lossless network

\[
[S]' [S]^* = [U],
\]

where $[U]$ is the identity matrix. This is known as the unitary condition and the proof of this is given in References [1,2]. The superscripts ‘t’ and ‘*’ denote the transpose and the conjugate respectively of the matrix. This condition is useful for checking the conservation of power. The power into and out of a circuit at the $j^{th}$ port is defined as follows

\[
\text{Power into the circuit} = a_j^- a_j^{**} \quad (2.3)
\]

\[
\text{Power out of the circuit} = a_j^- a_j^{**}. \quad (2.4)
\]
Since this thesis considers only a physical two port network, the following is known to be true.

\[ a_1^- = S_{11} a_1^+ + S_{12} a_2^+ \]  
(2.5)

\[ a_2^- = S_{21} a_1^+ + S_{22} a_2^+ \]  
(2.6)

From the previous two equations,

\[ S_{11} = \left. \frac{a_1^-}{a_1^+} \right|_{a_2^+ = 0} \quad S_{12} = \left. \frac{a_1^-}{a_2^-} \right|_{a_1^+ = 0} \]  
(2.7)

\[ S_{21} = \left. \frac{a_2^-}{a_1^+} \right|_{a_2^+ = 0} \quad S_{22} = \left. \frac{a_2^-}{a_2^-} \right|_{a_1^+ = 0} \]  
(2.8)

where \( S_{11} \) is the reflection coefficient at port 1, with port 2 terminated in a matched load. \( S_{21} \) is the transmission coefficient into port 2 from port 1, with port 2 terminated in a matched load. Similar definitions apply to \( S_{12} \) and \( S_{22} \) with respect to port 2.

### 2.3.1 Scattering Matrix of a Matched Transmission Line Section

Consider the transmission line section of length \( L \) and characteristic impedance \( Z_0 \) as shown in Figure 2.5. The line is assumed to be terminated at both ends by impedances equal to \( Z_0 \) (matched). At a fixed frequency the change in the scattering-matrix elements arising from a shift in the terminal plane location can be found.
Consider the scattering-matrix of the section of transmission line shown in Figure 2.5.

The line is viewed as a two-port junction with incident voltage waves $a_1^+$, $a_2^+$ and reflected waves of $a_1^-$, $a_2^-$ for ports 1 and 2, respectively. If the terminal plane is shifted from 1 - 1' to 2 - 2' then there will be a corresponding electrical phase shift of $\beta L$, where $\beta$ is the propagation phase constant for the line. In a perfectly lossless line there should be no reflections, and therefore the scattering matrix of a lossless transmission line from reference planes 1 -1' to 2 - 2' is given by

$$
S = \begin{bmatrix}
0 & e^{-j\beta L} \\
e^{-j\beta L} & 0
\end{bmatrix}.
$$

\[(2.9)\]

### 2.4 Physical and Electrical Ports of a Multimode Network

The number of physical terminal-pare present in the network gives the number of physical ports present in the network. In microwave structures, consisting of waveguide...
components, the single incident mode scatters into an infinite number of modes or waves at the discontinuity. Depending on the structure, some of these modes may have little effect on the desired response of the circuit and can be ignored. However, the modes close in frequency to the fundamental mode cannot be ignored because they will have a measurable effect on the performance of a circuit. Therefore, it is possible to have several transmitted waves resulting from one incident wave. Due to these extra transmitted modes, a two-port network can in fact have several electrical ports. Each scattering parameter no longer remains a single element, but now becomes a matrix of its own. The scattering matrix $[S]$ of the network is expressed as a matrix of matrices

$$
[S] = \begin{bmatrix}
[S_{11}] & [S_{12}] \\
[S_{21}] & [S_{22}]
\end{bmatrix}.
$$

(2.10)

In this form $[S_{11}]$ contains all coefficients for the incident and resulting reflections at port 1. Furthermore $[S_{21}]$ contains all the coefficients for the incident waves at port 1 and resulting reflections at port 2. Similarly $[S_{12}]$ and $[S_{22}]$ can be defined in a like manner. In the case where two modes are taken into account the scattering matrix would be given by

$$
[S] = \begin{bmatrix}
S_{11}(0,0) & S_{11}(0,1) & S_{12}(0,0) & S_{12}(0,1) \\
S_{11}(1,0) & S_{11}(1,1) & S_{12}(1,0) & S_{12}(1,1) \\
S_{21}(0,0) & S_{21}(0,1) & S_{22}(0,0) & S_{22}(0,1) \\
S_{21}(1,0) & S_{21}(1,1) & S_{22}(1,0) & S_{22}(1,1)
\end{bmatrix}.
$$

(2.11)
Concluding the TE_{10} mode as the incident, the TE_{20} mode propagates in addition to the incident mode. In the case of this matrix, the S_{ij} values with i = 1, 2 and j = 1, 2 have the same meaning as before. The variables (X, Y), with X = 0, 1 and Y = 0, 1, represents the incident and reflected modes respectively. A value of 0 represents a TE_{10} wave, while a value of 1 represents a TE_{20} wave. For example S_{11}(0, 1) represents the reflection coefficient of incident TE_{10} mode at port one resulting in a reflected TE_{20} mode at port one.

The contribution of higher order modes to the fundamental modes can drastically affect the accurate representation of the waveguide network. Hence, for accurate representation of the network, the higher order modes and their effects must be considered.

2.5 Summary

This chapter briefly describes the different types of waveguide discontinuities (E-plane, H-plane and EH-plane) and their placement within the waveguide (symmetric or asymmetric). The type of discontinuity that the chapter focuses on is the H-plane iris discontinuity. The chapter also introduces the concept of multimode scattering, and the scattering parameters of a transmission line.
Chapter 3

The Mode Matching Method

3.1 Introduction

The analysis of electromagnetic field scattering at a waveguide discontinuity is a complex problem. The technique of modal analysis, however, is relatively straightforward and similar in principle to reflection/transmission problems. In addition, modal analysis is a rigorous and versatile technique that can be applied to many coax, waveguide and planar transmission line discontinuity problems, and it lends itself well to computer implementation. The main advantage of modal analysis is that the interaction of higher order modes can be included in the final solution. The characterization of waveguide discontinuities by modal analysis can be programmed with much less effort and with similar accuracy to other techniques like the Finite Element Method (FEM) and the Finite Difference Time Domain (FDTD) method. Good references on the topic can be found in [3,8,11,13,14, 17].

In the formulation of the final solution, the total fields in each waveguide region are
expanded in terms of a complete set of normalized modes whose amplitudes are
adjusted so as to satisfy the boundary conditions at the discontinuity. This leads into the
formation of an infinite set of linear simultaneous equations for the unknown modal
coefficients. Since it is impossible to find the exact solution to an infinite system of
equations, a large but finite number of modes is taken into account.

Due to the infinite system of equations, problems can arise in the numerical calculations
as a result of convergence problems. The numerical solutions use only a finite number
of normal modes, and as a result they may converge to an incorrect value if an improper
ratio is chosen between the number of modal terms retained in the different regions.
This event is known as relative convergence, and is discussed in [11].

The solutions of modal equations lead directly to a multimode scattering matrix that
characterizes the isolated discontinuity. Since the scattering matrix contains information
on all the modes either above or below cutoff, a useful wideband equivalent network can
be constructed for the analysis of interacting discontinuities in waveguides. Also, the
method along with network theory can be implemented for the design of passive
structures like waveguide filters, which is one of the objectives of this thesis.

### 3.2 H-Plane Discontinuity Characterization

Figure 3.1 shows the three dimensional view of an H-plane step discontinuity in a
rectangular waveguide. Such a discontinuity is the building block of an iris coupled
waveguide filter. The analysed cross-sectional plane is set at $z = 0$.

![Diagram of H-plane step discontinuity](image)

**Figure 3.1:** 3-D view of an H-plane step discontinuity.

An important thing to note is that it is assumed that only the dominant TE$_{10}$ mode is propagating in guide 1 ($z < 0$), and that such a mode is incident on the junction ($z = 0$). Because of the discontinuity there will be reflected and transmitted waves in both guides, consisting of infinite sets of TE$_{n0}$ modes, where $n = 1, 2, 3, ..., \text{ in guides 1 and 2.}$ Only the TE$_{10}$ mode will propagate in guide 1, but the higher-order modes are also important because they are created at $z = 0$ and also remain localized around $z = 0$ in order to account for the stored energy localized near $z = 0$. Because there is no field
variation in the y direction of the discontinuity, TE<sub>nm</sub> modes for m ≠ 0 and TM modes are not excited. In other words, all modes are TE<sub>e0</sub> type.

The first step to be taken is to solve for the normalized electric and magnetic fields. These can be derived from Equations [1.10 to 1.15], which already take into account the boundary conditions of the waveguide. Assuming that the incident wave is the TE<sub>10</sub> mode (Note: n = 1 & m = 0) and using Equations [1.10 to 1.15] the fields are given by

\[
H_x = \frac{A_{10}}{Z_{h,10}} \sin \left( \frac{\pi x}{a} \right) e^{-j\beta_{10}Z},
\]

\[
E_y = -A_{10} \sin \left( \frac{\pi x}{a} \right) e^{-j\beta_{10}Z},
\]

\[
H_y = E_z = E_x = 0.
\]

The coefficient A<sub>10</sub> is the coefficient related to the power from the input of the TE<sub>10</sub> mode, and Z<sub>h,10</sub> is the wave impedance of the TE<sub>10</sub> mode as defined in Equation (1.16).

When the dominant TE<sub>10</sub> mode approaches the discontinuity from the wider side of the waveguide discontinuity, it is assumed that its electric field has a strength having an amplitude of A<sup>+</sup><sub>10</sub>. The positive superscript shows that the field travels in the positive z-direction. Once the field strikes the discontinuity it creates higher order modes on both sides, with the amplitudes on side 1 being represented by A<sup>+</sup><sub>n0</sub> (n = 1,2,3 ... ∞), and B<sup>±</sup><sub>p0</sub> (p = 1,2,3 ... ∞) on side 2. The electric fields on both sides respectively now become
\[ E_y = -\sum_{n=1}^{\infty} \left( A_{n0}^+ + A_{n0}^- \right) \sin \frac{n\pi x}{a_1} e^{\pm j\beta_{n0}z} \quad \text{for } z<0, \]  
\[ E_y = -\sum_{p=1}^{\infty} \left( B_{p0}^+ + B_{p0}^- \right) \sin \frac{p\pi (x-h_x)}{a_2} e^{\pm j\beta_{p0}z} \quad \text{for } z>0. \]  

The boundary conditions are taken into account by observing the continuity of the total transverse electric and magnetic fields across the aperture and the zero tangential electric field on the surface of the iris wall at \( z = 0 \). Equating the fields on both sides gives

\[ -\sum_{n=1}^{\infty} \left( A_{n0}^+ + A_{n0}^- \right) \sin \frac{n\pi x}{a_1} = -\sum_{p=1}^{\infty} \left( B_{p0}^+ + B_{p0}^- \right) \sin \frac{p\pi (x-h_x)}{a_2}. \]  

Multiply both sides of (3.6) by \( \sin \frac{r\pi x}{a_1} \), where \( r \) is an integer. Once this is done, integrate both sides over the range \( 0 \leq x \leq a_1 \) which gives

\[ \sum_{n=1}^{\infty} \left( A_{n0}^+ + A_{n0}^- \right) \int_0^{a_1} \sin \frac{r\pi x}{a_1} \sin \frac{n\pi x}{a_1} dx = \sum_{p=1}^{\infty} \left( B_{p0}^+ + B_{p0}^- \right) \int_0^{a_1} \sin \frac{r\pi x}{a_1} \sin \frac{p\pi (x-h_x)}{a_2} dx. \]

Taking into account the orthogonality property

\[ \int_0^{a_1} \sin \frac{r\pi x}{a_1} \sin \frac{n\pi x}{a_1} dx = \begin{cases} 0 & \text{if } r \neq n, \\ \frac{a_1}{2} & \text{if } r = n, \end{cases} \]  

Equation (3.7) becomes

\[ \sum_{n=1}^{\infty} \frac{a_1}{2} \left( A_{n0}^+ + A_{n0}^- \right) = \sum_{p=1}^{\infty} I \left( B_{p0}^+ + B_{p0}^- \right). \]
For the case of a symmetric discontinuity

\[ I = \int_{h_x}^{h_x + a_2} \sin \frac{r \pi x}{a_1} \sin \frac{p \pi (x - h_x)}{a_2} \, dx. \quad (3.10) \]

Equating the integral gives

\[ I = \frac{a_2}{2} \cos \left( \frac{n \pi h_x}{a_1} \right) \quad \text{when} \quad \frac{n \pi}{a_1} = \frac{p \pi}{a_2}, \quad (3.11 \text{ a}) \]

and

\[ I = \frac{p \pi}{a_2} \left( \frac{n \pi}{a_1} \right)^2 \left( \frac{p \pi}{a_2} \right)^2 \left( (-1)^p \sin \left[ \frac{n \pi}{a_1} (h_x + a_2) \right] - \sin \left( \frac{n \pi}{a_1} h_x \right) \right) \quad (3.11 \text{ b}) \]

when \( \frac{n \pi}{a_1} \neq \frac{p \pi}{a_2} \).

For the case of asymmetrically positioned and centred discontinuities the same steps are required as the symmetrically positioned discontinuities. In these cases the integration is preformed over different regions. Figure 3.2 shows the regions of integration for both the asymmetrical and centred discontinuities.
Figure 3.2: Front view of: a) Asymmetrical discontinuity, b) Centred discontinuity.

For the case of the asymmetrically positioned discontinuity $I$ is defined as

$$I = \int_{a_1}^{a_2} \sin \frac{r \pi x}{a_1} \sin \frac{p \pi x}{a_2} dx.$$  

(3.12)

Solving for the integral gives

$$I = \frac{a_1 a_2}{2 \pi} \left( \sin \left( \frac{\pi (a_2 r - a_1 p)}{a_1} \right) a_1 \sin \left( \frac{\pi (a_2 r + a_1 p)}{a_1} \right) \right)$$

(3.13 a)

when $a_2 r - a_1 p \neq 0$, and

$$I = \frac{a_2}{2} - \frac{a_1 \sin \left( \frac{2 \pi a_2 r}{a_1} \right)}{4 \pi r}$$

(3.13 b)

when $a_2 r - a_1 p = 0$.

For the case of the centred discontinuity the coupling between the larger section of the waveguide and the two smaller sections must be taken into account. The coupling between the larger waveguide section and the smaller section on the left side of the waveguide is given by $I_1$. The coupling between the larger waveguide section and the
smaller section on the right side of the waveguide is given by $I_2$. $I_1$ and $I_2$ are defined as

\[
I_1(t) = \int_0^a \sin \frac{r \pi x}{a} \sin \frac{p \pi x}{a - t} \, dx,
\]

\[
I_2(t) = \int_{a - t}^a \sin \frac{r \pi x}{a} \sin \frac{p \pi (x - \frac{a + t}{2})}{a - t} \, dx.
\]

Solving for the integrals gives

\[
I_1 = \frac{r \pi (-1)^m}{\left( \frac{r \pi}{a} \right)^2 - \left( \frac{2p \pi}{a - t} \right)^2} \cdot \frac{\sin \left( \frac{2r \pi}{a} (a - t) \right)}{\sqrt{a(a - t) / 2}} \quad \text{when } r \neq p,
\]

\[
I_1 = \sqrt{\frac{a(a - t)}{8}} \quad \text{when } r = p,
\]

and

\[
I_2 = \frac{r \pi}{\left( \frac{r \pi}{a} \right)^2 - \left( \frac{2p \pi}{a - t} \right)^2} \cdot \frac{\sin \left( \frac{r \pi}{2a} (a + t) \right)}{\sqrt{a(a - t) / 2}} \quad \text{when } r \neq p,
\]

\[
I_2 = \sqrt{\frac{a(a - t)}{8}} \cdot \cos \left( \frac{r \pi}{2a} (a + t) \right) \quad \text{when } r = p.
\]

When taking a finite number of modes on each side of the discontinuity, the equation can be written in matrix form as

\[
\begin{bmatrix} A^+ + A^- \end{bmatrix}_{N \times 1} = \frac{2}{a_1} \left[ I \right]_{N \times P} \left[ B^+ + B^- \right]_{P \times 1}.
\]

38
The matrices \([A^+ + A^-]_{N \times 1}\) and \([B^+ + B^-]_{P \times 1}\) are the column vectors of the field amplitudes considered on sides 1 and 2 respectively. The corresponding dimensions are shown at the bottom right hand corner of the vectors. The mode coupling matrix is defined as

\[
[H]_{N \times P} = \frac{2}{a_1} [I]_{N \times P}.
\] (3.19 a)

For the centred discontinuity the coupling matrix is defined as

\[
[[H_1]_{N \times P} [H_2]_{N \times P}] = \frac{2}{a_1} [[I_1]_{N \times P} [I_2]_{N \times P}].
\] (3.19 b)

It is called the mode coupling matrix because it is the matrix that links between the TE wave modes on the two sides of the discontinuity.

The magnetic fields on the two sides of a discontinuity are given by

\[
H_x = \sum_{n = 1}^{\infty} \left( \frac{A^+_n - A^-_n}{Z_{TE_1}^{n0}} \right) \sin \frac{n \pi x}{a_1} e^{-j\beta_{n0} z} \quad \text{for } z > 0,
\]

\[
H_x = \sum_{p = 1}^{\infty} \left( \frac{B^+_p - B^-_p}{Z_{TE_2}^{p0}} \right) \sin \frac{p \pi (x - h_x)}{a_2} e^{-j\beta_{p0} z} \quad \text{for } z < 0.
\]

Equating the magnetic fields on both sides of the discontinuity and performing the same operations that were done on the electric fields the equations become
\begin{align*}
\frac{1}{2} \sum_{n=1}^{N} \frac{a_{n}}{Z_{T E_{n} n 0}} Y_{1}^{n 0} \left(A_{n 0}^{+} - A_{n 0}^{-}\right) &= \sum_{p=1}^{P} I \frac{1}{Z_{T E_{p} p 0}} Y_{2}^{p 0} \left(B_{p 0}^{+} - B_{p 0}^{-}\right), \\
\text{(3.22)}
\end{align*}

where $Y_{1}^{n 0} = \frac{I}{Z_{T E_{1} n 0}}$ in side one of the discontinuity and $Y_{2}^{p 0} = \frac{I}{Z_{T E_{2} p 0}}$ in side two.

It should be noted that in Equation (3.22), a limited number of modes ($N$ and $P$) are being considered on both sides of the discontinuity. If the proper number of modes are taken into account on both sides of the discontinuity, both speed and accuracy will be maintained.

The discontinuity Equation (3.22) can be written in matrix form as

\begin{equation}
\begin{bmatrix}
Y_{1}^{N} & A^{+} - A^{-} & N_{x 1} \\
H & Y_{2}^{P} & B^{+} - B^{-}
\end{bmatrix}, \\
\text{(3.23)}
\end{equation}

where \([Y_{1}]_{N \times N}^{N} \text{ and } [Y_{2}]_{P \times P}^{P} \) are diagonal matrices of the form

\begin{align*}
[Y_{1}] &= \begin{bmatrix}
Y_{1}^{10} & 0 & \ldots & 0 \\
0 & Y_{1}^{20} & \ldots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \ldots & Y_{1}^{N 0}
\end{bmatrix}, \\
[Y_{2}] &= \begin{bmatrix}
Y_{2}^{10} & 0 & \ldots & 0 \\
0 & Y_{2}^{20} & \ldots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \ldots & Y_{2}^{P 0}
\end{bmatrix}. \\
\text{(3.24)}
\end{align*}

The choice of the number of modes is very crucial in avoiding the relative convergence problem. The number of modes can be kept the same on both sides of the discontinuity, but it is possible to choose a different number of modes for each side of interest. In order to avoid the convergence problem the number of modes in the smaller section ($P$) and
the larger section \( (N) \) should be chosen according to
\[
P = \frac{a_j}{a_i} N.
\] (3.25)

### 3.3 Analysis Using the Generalized Scattering Matrix Method

Equations (3.18) and (3.23) are the equation set for the S-matrix of an H-plane discontinuity. Combining the two equations, the following equation is formed.

\[
\begin{align*}
[H] [Y_2] \left( [U] + [Y_2]^{-1} [H]^{-1} [Y_1] [H] \right) [B^+] &= 2[Y_1] [A^+] + [H] [Y_2] \left( [U] - [Y_2]^{-1} [H]^{-1} [Y_1] [H] \right) [B^-],
\end{align*}
\] (3.26)

where \([U]\) is the identity matrix of compatible order.

From the definition of scattering parameters in Equations (2.7) and (2.8),

\[
\begin{align*}
[A^-] &= \begin{bmatrix} S_{11} & 0 \end{bmatrix} [A^+] \left[ \begin{bmatrix} B^- \\ 0 \end{bmatrix} - [0] \right], \\
[A^-] &= \begin{bmatrix} S_{12} & 0 \end{bmatrix} [B^-] \left[ \begin{bmatrix} A^- \\ 0 \end{bmatrix} - [0] \right],
\end{align*}
\] (3.27 a,b)

\[
\begin{align*}
[B^+] &= \begin{bmatrix} S_{21} & 0 \end{bmatrix} [A^+] \left[ \begin{bmatrix} B^- \\ 0 \end{bmatrix} - [0] \right], \\
[B^-] &= \begin{bmatrix} S_{22} & 0 \end{bmatrix} [B^-] \left[ \begin{bmatrix} A^- \\ 0 \end{bmatrix} - [0] \right],
\end{align*}
\] (3.27 c,d)

where \([0]\) is a null column matrix. With some matrix manipulation the overall multimode scattering matrix of the discontinuity can be obtained (see Equation (2.10))

\[
\begin{align*}
[S_{11}] &= [H] [S_{21}] - [U], \\
[S_{12}] &= [H] \left( [S_{22}] + [U] \right),
\end{align*}
\] (3.28 a,b)

\[
\begin{align*}
[S_{21}] &= 2 \left( [U] + [Y_2]^{-1} [H]^T [Y_1] [H] \right) \left( [U] - [Y_2]^{-1} [H]^T [Y_1] [H] \right),
\end{align*}
\] (3.28 c)

\[
[S_{22}] &= \left( [U] + [Y_2]^{-1} [H]^T [Y_1] [H] \right) \left( [U] - [Y_2]^{-1} [H]^T [Y_1] [H] \right).
\] (3.28 d)
3.4 Cascading of Scattering Matrices

In practice, a waveguide filter is never formed from one discontinuity, but rather is the result of the combination of many coupled discontinuities. The desired end result is to have the overall scattering matrix of the entire filter, and this is why a technique is required to combine the scattering matrices of adjacent discontinuities. For the calculation of the overall scattering matrix, a waveguide filter is decomposed into two main building blocks: step discontinuities and transmission lines.

The problem is approached with the knowledge that the mode matching method has the ability to determine the scattering matrix of any discontinuity, and the scattering matrix for a transmission line is already known (see Equation (2.9)). Starting at the input port, analysis is done by combining the known scattering matrices of the corresponding intermediate waveguide sections. The cascaded combination of all the matrices will result in the overall scattering matrix of the corresponding coupled filter. Figure 3.3 shows the general approach for one waveguide iris.
Figure 3.3: Composed scattering matrices for an iris of finite thickness.

For any combination of scattering matrices of cascaded elements the relationship among $[S^{123}], [S^1]$ and $[S^2]$ are as follows [11,14]:

\[
[S_{11}^{123}] = [S_{11}^1] + [S_{12}^1] [S_{11}^2] ([U] - [S_{22}^1] [S_{11}^2])^{-1} [S_{21}^1],
\]

(3.29a)

\[
[S_{12}^{123}] = [S_{12}^1] ([U] + [S_{11}^2] ([U] - [S_{22}^1] [S_{11}^2])^{-1} [S_{22}^1]) [S_{12}^2],
\]

(3.29b)

\[
[S_{21}^{123}] = [S_{21}^2] ([U] - [S_{22}^1] [S_{11}^2])^{-1} [S_{21}^1],
\]

(3.29c)

\[
[S_{22}^{123}] = [S_{22}^2] + [S_{21}^2] ([U] - [S_{22}^1] [S_{11}^2])^{-1} [S_{22}^1] [S_{12}^2].
\]

(3.29d)

Derivation of the Equations (3.29a) through (3.29d) uses

\[
[S_{11}^1] = [S_{22}^2], \ [S_{12}^1] = [S_{21}^2] \text{ and } [S_{21}^1] = [S_{12}^2]
\]

43
because the two step discontinuities are the mirror images of each other.

3.5 Summary

In this chapter, the characterization of waveguide H-plane discontinuities by using the mode matching method, and the generalized multimode scattering matrix are discussed in detail. The mode matching method is a fast and effective numerical method for solving waveguide discontinuity problems, and is the method of choice in this thesis for the analysis of filter structures and characterization of resonator coupling irises.
Chapter 4

Filter Synthesis

In the previous chapter, the characterization of waveguide discontinuities using the mode matching method is explained. In this chapter, a new CAD method for the design of H-plane iris coupled waveguide bandpass filters is presented in detail. The CAD method for designing waveguide filters involves an approximate synthesis method, an analysis method and an optimization method. The synthesis routine determines the approximate physical dimensions of the filter, while the analysis method accurately simulates the frequency response of the filter for verification of the filter design. The optimization routine tunes the design so that the analysed response matches the desired response.

The synthesis method is based on distributed transmission line theory and conventional low frequency network theory. As stated earlier, it determines the geometrical dimensions of each coupling iris and the distance between any two consecutive irises in the filter structure. The design procedure greatly relies on the accurate characterization of the waveguide discontinuities that are found in this type of filter. The mode matching
method is more than adequate for the characterization of waveguide discontinuities. The mode matching method is used in the design procedures in this chapter.

4.1 Filter Design Parameters

In the design process, there are several parameters which are of importance to the design engineer. These parameters characterize the frequency response of the filter. Figure 4.1 shows the typical insertion (isolation) and return loss characteristics as a function of frequency of a bandpass filter. For this thesis, the phase response of the filter is not considered, only the frequency response is evaluated.

![Figure 4.1: Typical frequency response of a bandpass filter.](image)
The most common terms used to describe the filter frequency response are as follows:

- Lower cutoff frequency, $f_L$ (GHz)
- Upper cutoff frequency, $f_H$ (GHz)
- Fractional bandwidth, \(BW = \frac{2(f_H - f_L)}{f_H + f_L}\)
- Passband return loss, RL (dB)
- Isolation bandwidth factor, \(\gamma = \frac{\Delta f_i}{f_H - f_L}\)
- Isolation loss, $L_A$ (dB)

Further, the centre frequency is the mean of the lower and upper cutoff frequencies. The bandwidth is usually expressed as an absolute frequency or as a relative percentage of the centre frequency.

Isolation is expressed in decibels (dB) and it is the minimum stopband insertion loss. In bandpass filter design a specific isolation loss ($L_A$) is desired at specified upper and lower frequency points. The magnitude of this frequency range is called the isolation bandwidth ($\Delta f_i$). The isolation bandwidth factor ($\gamma$) is the ratio of isolation bandwidth to ripple bandwidth of the filter. In addition, the dimensions of the waveguide (width and height) and the iris thickness are parts of the design specifications. For the design of the filters in this thesis, the required input parameters are as follows:

- Lower cutoff frequency ($f_L$)
- Upper cutoff frequency ($f_H$)
- Isolation bandwidth ($\Delta f_i$)
- Isolation loss ($L_A$)
- Return loss (RL)
- Waveguide width ($a$)
- Waveguide height ($b$)
- Iris thickness ($t$)
4.2 Synthesis Method

The bandpass filters in this thesis are realized using H-plane iris coupled resonators.

The schematic diagram of an H-plane iris coupled bandpass filter can be seen in Figure 4.2. The filter shown has 5 resonator cavities (designated c₁ through c₅) and 6 irises (widths designated w₁ through w₆) of constant thickness t, therefore making it a 5th order filter. Before illustrating the design process some circuit definitions are needed.

![Figure 4.2: Waveguide bandpass filter with H-plane iris discontinuities.](image)

4.2.1 Impedance Inverters

An impedance inverter, also known as a K-inverter, is an ideal quarter-wave transmission line transformer, and can be seen in Figure 4.3a. It operates like a quarter-wavelength transmission line of characteristic impedance K ohms at a specific frequency. Therefore, if it is terminated with an impedance Zᵣ on one end, as shown in
example, for a constant iris thickness the iris width determines the K-inverter value.

Figure 4.4 shows the equivalent dominant (TE\textsubscript{10}) mode network of a symmetrical two-step H-plane discontinuity in a waveguide.

\[ K = \left| \tan\left( \frac{\Phi}{2} + \tan^{-1} x_s \right) \right|, \quad (4.2) \]

\[ \Phi = -\tan^{-1}(2x_p + x_s) - \tan^{-1}(x_s), \quad (4.3) \]

\[ jx_s = \frac{(1 - S_{12} + S_{11})}{(1 - S_{11} + S_{12})}, \quad (4.4) \]

\[ jx_p = \frac{2S_{12}}{(1 - S_{11})^2 - S_{12} S_{12}}, \quad (4.5) \]

where \( x_s = \frac{X_s}{Z_0}, x_p = \frac{X_p}{Z_0} \) and \( S_{ij} \) (\( i, j = 1, 2 \)) are the dominant mode scattering parameters of the iris.

50
4.2.3 General Synthesis Method

The synthesis method is based on an approach suggested in [4] for the design of waveguide coupled bandpass filters. A Chebyshev type of response is used in the design method. The generalized lumped element prototype for the bandpass filter is shown in Figure 4.5a, and the equivalent distributed prototype containing impedance inverters is shown in Figure 4.5b.

The distributed prototype consists of a cascade of half-wavelength transmission line resonator sections connected by K-inverters. The coupling capacitors of the bandpass filter in Figure 4.5a are replaced by K-inverters as shown in Figure 4.5b, and the parallel resonant circuits are replaced with sections of transmission lines. These resonators are realized by waveguide sections of length $\lambda_{g0}/2$, where $\lambda_{g0}$ is the guide wavelength of the line at the centre frequency of the filter.

![Figure 4.5](image-url)
In the waveguide structure shown in Figure 4.2, the iris width (w) is adjusted in the X-direction (refer to Figure 1.4) to realize the K-inverter values. Since the waveguide sections between any two irises are equal in width, and equal in height for the H-plane structure, it is said to be uniformly corrugated. It is possible to have waveguide sections of unequal width or height between any two iris discontinuities and such a structure is known as a nonuniform corrugated waveguide filter [13,14]. This thesis deals with uniform corrugated filters only.

The general idea for the synthesis of the waveguide filters is to find the widths of the irises and lengths of the resonators that will give the lumped element values for the filter prototype. Even though one may achieve the prototype values in the synthesis, an analysis of the filter will show a response that is not the same as what was desired. This is due to the fact that higher order mode interaction between any two consecutive irises is not taken into account in the overall cascading of K-inverters in the initial synthesis. Each K-inverter is synthesized very accurately, however the proximity of other K-inverters is not accounted for in the synthesis process. This modal interaction manifests itself in producing a filter which has a response that is not desired. Typically, a shrinkage in ripple bandwidth, shift in centre frequency and deterioration in return loss performance are observed.

4.2.4 Determination of K-Inverter Values From Network Theory

The following presents the step by step approach used in the synthesis module for the
filter design. Background information on the K-inverter method can be found in [5,6].

1) The order of the filter (see Figure 4.2), $N_f$, gives the number of resonators and $N_f + 1$ is the number of iris couplers needed to meet the design specifications. The filter order is given by [2]:

$$N_f \geq \frac{L_A + RL + 6}{20\log(\gamma + \sqrt{\gamma^2 - 1})}. \quad (4.6)$$

2) The midband guide wavelength $\lambda_{go}$ is determined by numerically solving the following equation [6]:

$$\lambda_{gL} \sin\left(\frac{\pi \lambda_{go}}{\lambda_{gL}}\right) + \lambda_{gH} \sin\left(\frac{\pi \lambda_{go}}{\lambda_{gH}}\right) = 0, \quad (4.7)$$

where $\lambda_{gL}$ and $\lambda_{gH}$ are the guide wavelengths in the resonator section at the lower and upper cutoff frequencies. For a narrow-band case

$$\lambda_{go} = \frac{\lambda_{gL} + \lambda_{gH}}{2}. \quad (4.8)$$

A suitable numerical method is applied for solving Equation (4.7).

3) A scaling parameter is given by

$$\alpha = \frac{\lambda_{go}}{\lambda_{gL} \sin\left(\frac{\pi \lambda_{go}}{\lambda_{gL}}\right)}. \quad (4.9)$$
4) Calculate the impedance, \( Z_n \), of the distributed elements and impedance inverter values, \( k' \), by [6]

\[
Z_n = \frac{1}{4\alpha} \sqrt{\frac{y^2 + \sin^2 \left( \frac{N_f}{N_f} \right)}{\sin \left( \frac{(2n - 1)\pi}{2N_f} \right)}}
\]

where \( n = 1, \ldots, N_f \) and

\[
k'_{n,n + 1} = \sqrt{\frac{y^2 + \sin^2 \left( \frac{N_f}{N_f} \right)}{y}},
\]

where

\[
y = \sinh \left( \frac{L}{N_f} \sinh^{-1} \frac{1}{\epsilon} \right).
\]

The variable \( \epsilon \) is the absolute value of the passband ripple.

5) The normalized \( K \)-inverter values which are to be realized by waveguide irises for the filter are defined by

\[
K_{n,n + 1} = \frac{k'_{n,n + 1}}{\sqrt{Z_n Z_{n+1}}}, \quad n = 0, \ldots, N_f
\]

and

\[
Z_0 = Z_{n+1} = 1.
\]

6) This step involves the mode matching method discussed in Chapter 3. Taking into account the fact that the iris thicknesses are the same for each iris, and the waveguide dimensions have been set, the scattering parameters for an iris junction can
be found. An iterative procedure is used to adjust each iris width in order to realize the corresponding K-inverter value. Each time the iris width is incremented, the iris junction is rigorously analysed and the dominant mode scattering parameters for the junction are determined. Using Equations 4.2 to 4.5, the accompanying K-inverter and phase $\phi_j$ (see Figure 4.4) values which correspond to the iris width can be found. The iterative process is continued until, when analysed, the iris dimensions produce the desired K-inverter value. This process is repeated for each iris junction. Using the phases ($\phi_j$ and $\phi_{j+1}$), the length of the $j^{th}$ resonator can be obtained from [5,6]:

$$l_j = \frac{\lambda_0}{2\pi} \left[ \pi + \frac{1}{2} (\phi_j + \phi_{j+1}) \right] \quad j = 1, \ldots, N_f.$$  (4.15)

As mentioned earlier, the K-inverter approach does not take into account the proximity, and thus the mode interaction, of other such junctions. Because of this fact, any design using this technique will not be exact. An analysis of a filter designed using the K-inverter method will show a frequency response that is not the same as the desired one. This is why optimization is needed for tuning of the filter design. To show the effects of neglecting the mode interaction between adjacent irises, a waveguide filter is designed without the use of optimization. Figure 4.6 shows the frequency response of a symmetric iris coupled waveguide filter that is designed using the K-inverter synthesis approach. The input design parameters were for an X-band filter, and were as follows:
Lower cutoff frequency \( (f_L) = 9 \text{ GHz} \)
Upper cutoff frequency \( (f_H) = 10 \text{ GHz} \)
Isolation bandwidth \( (\Delta f) = 2 \text{ GHz} \)
Isolation loss \( (L_A) = 30 \text{ dB} \)
Waveguide width \( (a) = 22.86 \text{ mm} \)
Iris thickness \( (t) = 1.5 \text{ mm} \)

\[
\text{Return loss (RL) = 16 dB}
\]

Waveguide height \( (b) = 10.16 \text{ mm} \)

**Frequency Response For Asymmetric Filter Design Using the K-Inverter Method**

![Graph showing frequency response of a symmetric iris coupled filter designed by the K-inverter method.](image)

**Figure 4.6:** Frequency response of a symmetric iris coupled filter designed by the K-inverter method.
As can be seen in Figure 4.6, the frequency response does not satisfy the input parameters of the filter. It appears that the ripple bandwidth is actually smaller than desired, and the return loss is lower than 16 dB in the pass band and at the band edges. These differences can be thought of as an error, and can be taken care of using optimization.

4.2.5 Rational Function Approximation Characterizing Iris Junctions

Due to the fact that each individual K-inverter value is iteratively searched for by incrementing the iris width, the previous method can be relatively time consuming. It is not the goal of this thesis to improve the accuracy of this K-inverter synthesis based design approach, any inaccuracy can be eliminated by optimization eventually. However, one of the goals of this thesis is to increase the speed of the synthesis routine. This can be done by creating equations which will characterize the iris width and phase with respect to the K-inverter value. Therefore, instead of iteratively searching for the iris widths which will satisfy the desired K-inverter value, modelled equations, along with the desired K-inverter values, can be used to find the iris width and phase directly. To do this tables of data are created by iteratively adjusting the iris width and solving for the K-inverter value and the associated phase.

Using the tables of data, and implementing a suitable curve fitting program, equations that characterize each set of information can be found. The equations for the phase and width of irises should be with respect to the K-inverter values. These equations will be
used to characterize all desired iris junctions that will be used in the filter design. Using these equations and the normalized K-inverter values found in Step 5 of Section 4.2.4, the iris widths and accompanying phases can be found for the desired filter model.

The method of using interpolation to model an iris junction is not new, but is similar to the approach taken in [7]. However, in [7] the focus was on using the approximation equations for the analysis of iris junctions, as opposed to the synthesis of iris junctions. Furthermore, the results obtained from analysis are curve-fitted to a two-dimensional polynomial function. For example, for a parameter \( y \) (either the normalized K-inverter value or the phase), the two dimensional polynomial approximation is given by

\[
y(\tilde{w}, \tilde{f}) = \sum_{q=0}^{Q} \sum_{m=0}^{M} \xi_{mq} \tilde{f}^m \tilde{w}^q,
\]

where \( \tilde{f} \) is the normalized frequency, \( \tilde{w} \) is the normalized iris width \( \left( \frac{w}{a} \right) \), and \( \xi_{mq} \) are the coefficients of the polynomial. \( Q \) and \( M \) are the maximum orders of \( \tilde{w} \) and \( \tilde{f} \). The coefficients for this polynomial are found using a suitable curve-fitting technique.

Equation 4.16 can also be written in matrix form as

\[
y(\tilde{w}, \tilde{f}) = \begin{bmatrix} 1 & \tilde{w} & \tilde{w}^2 & \cdots & \tilde{w}^Q \end{bmatrix} \begin{bmatrix} \xi_{01} & \xi_{02} & \cdots & \xi_{0M} \\
\xi_{11} & \xi_{12} & \cdots & \xi_{1M} \\
\vdots & \vdots & \ddots & \vdots \\
\xi_{Q1} & \xi_{Q2} & \cdots & \xi_{QM} \end{bmatrix} \begin{bmatrix} \tilde{f} \\
\tilde{f}^2 \\
\vdots \\
\tilde{f}^M \end{bmatrix}.
\]

From Equation 4.17 it can be seen that the normalized K-inverter value and the phase
are characterized over a frequency band for many different widths, and over the entire
dominant mode bandwidth. In other words, this method characterizes the K-inverter
value and phase by producing many width specific curves (equations) over a wide
frequency band. This strategy tends to fall to problems in accuracy, in that the function
is not well-behaved for both the K-inverter value and the phase. Consequently, this
strategy runs into problems with the precision of characterization over the entire band, in
that the accuracy is dependent on the number of data points taken before the curve
fitting. The more data points taken, the greater the accuracy of the equation
characterization. Another problem with this method is that the time to design
completion is increased with the greater number of calculations. However, this will be
absolutely necessary only when a bandpass filter is designed whose bandwidth spans the
entire waveguide dominant mode band, which is rarely the case.

For the purpose of this thesis, it would be advantageous to have the equations as a
function of the normalized K-inverter values, which are solved for in the synthesis
routine, as opposed to functions of the iris width. To improve upon the general idea
given in [7], the following rational approximation was used,

\[ R(x) = \frac{P_n(x)}{Q_n(x)} = \frac{p_0 + p_1x + \cdots + p_nx^n}{q_0 + q_1x + \cdots + q_nx^n}, \quad (4.18) \]

where \( x \) is the value of the K-inverter and \( R(x) \) is either the width of the iris forming the
K-inverter or the phase associated with the obstacle [4]. The variables \( p_o \) through \( p_n \),
are frequency dependent coefficients that are found using a suitable curve fitting routine.

The accuracy of the characterization of the rational function approximation is related to the number of data points taken. In order to improve the accuracy of the characterization two things can be done; take a large number of data points over the entire band, or decrease the band over which the data points are taken. As mentioned previously, in creating a filter only a very small fraction of the possible range of iris widths is actually needed, because the filter bandwidth covers a fraction of the dominant mode band. Therefore, this thesis uses the approach of characterizing a smaller, yet desirable, portion of the dominant mode band.

In order to characterize the smaller band, a two-step process is implemented. First a narrower band must be defined, then a rational function must be created to characterize that band. In order to define the narrower band, a small number of evenly placed reference points is created by analysing iris junctions with varying widths in the dominant mode band. The desired K-inverter values (see Step 5 in Section 4.2.4) are compared to the K-inverter values associated with the reference points. The reference points which are closest in magnitude to the maximum and minimum desired K-inverter values will represent the upper and lower width values for the desired band of characterization. With the band of interest defined, a large number of evenly placed points is created and curve fitted to the rational function seen in Equation 4.18. This improves the accuracy of the characterization of the iris junction over the desired band,
and helps to reduce the time to completion of a design. This method is more accurate and more efficient than that used in [7]. A flowchart showing the steps taken in the synthesis routine can be seen in Figure 4.7.

![Flowchart](image)

Figure 4.7: Flowchart for rational function approximation approach

### 4.2.6 The Modified K-Inverter Modelling

This method was discovered while using the approach described in Section 4.2.4. In analysing the data points found in Step 6, closed-form equations were found that
characterize both the iris width and resonator length with respect to the desired normalized K-inverter value. Separate curve fitting software was used to analyse the data in order to get the closed-form equations. The following presents the step by step approach on how the final equations were created.

1) Using Steps 1 through 6 in Section 4.2.4, tables of data containing the phases, iris widths and the normalized K-inverter values are formed. Many sets of these tables are needed to cover a specified frequency band. In this thesis, the X-band (Dominant TE\(_{10}\) mode ranging from 8.2 - 12.4 GHz with \(a = 22.86\) mm) was used for the data collection. Each group of data was formed by using the same physical input design specifications, but in each case a different centre frequency for the filter was used. In analysing each group of data separately it was discovered that the iris width and corresponding phases could be described with the following rational functions

\[
W_i = \frac{A_w + C_w(K_v) + E_w(K_v)^2 + G_w(K_v)^3}{1 + B_w(K_v) + D_w(K_v)^2 + F_w(K_v)^3}, \quad (4.19)
\]

\[
\phi_i = \frac{A_p + C_p(K_v) + E_p(K_v)^2 + G_p(K_v)^3}{1 + B_p(K_v) + D_p(K_v)^2 + F_p(K_v)^3}, \quad (4.20)
\]

where \(K_v\) is the K-inverter value for the desired iris. The variables \(A_w - G_w\) and \(A_p - G_p\)
are coefficients which are dependent on the centre frequency of the filter. To see an example of the curve fitted width data and phase data, see Appendix A and B respectively.

2) In the previous step, closed form equations are created that characterize the iris width and phase for a filter designed with a centre frequency in the X-band. However, it is more advantageous to have equations that can be used for any desired frequency range. In order to do this, the existing equations for the coefficients must be normalized with respect to frequency and the broadwall dimension of the waveguide. This is accomplished by substituting the following value for the centre frequency

\[
F_{on} = \frac{300a}{22.86\lambda_o},
\]

(4.21)

where \(F_{on}\) is the normalized centre frequency, \(a\) is the broadwall length in mm, and \(\lambda_o\) is the free space wavelength in mm. This is sufficient for normalizing the coefficients for the phase, but the coefficients for the iris widths need one extra step. The normalized coefficients for the phase can be seen in Appendix C.

3) In order to normalize the coefficients for the iris widths, Step 1 must be repeated. In this case, before finding the closed form equation for the iris widths, each iris width must be divided by the waveguide broadwall length (22.86 mm for the X-band case). In other words, the final equation will characterize the ratio of iris width to broadwall...
length with respect to the K-inverter value. Therefore, the equation that characterizes the iris width is now given by

\[
W_i = a \left( \frac{A_w + C_w(K_v) + E_w(K_v)^2 + G_w(K_v)^3}{1 + B_w(K_v) + D_w(K_v)^2 + F_w(K_v)^3} \right). \tag{4.22}
\]

The normalized coefficients for the iris widths can be found in Appendix D.

4) So far, equations characterizing the iris width and phase for any desired frequency have been formed. To this point the ratio of iris thickness to broadwall length has been kept constant. It would be advantageous to have formulas that are true for any iris thickness desired. Two approaches were taken to solve this problem.

A) The first approach is to use a scaling factor. It was noted that as the ratio of iris thickness to broadwall length was changed the K-inverter value also changed. Using curve fitting software, a relationship was discovered that allows for this correction. In order to determine the correction factor, all design parameters must be kept the same except for the iris thickness. This factor (\(\psi\)) is multiplied to the desired K-inverter value, thus making it suitable for the derived width and phase equations. The factor is found to be

\[
\psi = A_F + B_RE + C_Re^{Ro} + D_Re^{-Ro}, \tag{4.23}
\]

\(64\)
\[ \frac{t}{6.5617 \times 10^{-2}} = \frac{a}{Ro} \]  

(4.24)

The variable \( \psi \) is the K-inverter correction factor, \( t \) is the iris thickness in mm, and \( a \) is the broadwall length in mm. The constant in the denominator is the ratio of the chosen reference iris thickness (1.5 mm) to the X-band broadwall length (22.86 mm). The variables \( A_F - D_F \) are coefficients which are dependent on the original K-inverter \( (K_v \) before it has been scaled). The coefficients for the above equation can be found in Appendix E.

The problem with this method is that the accuracy is dependent on the precision of the characterization of the formulas over the range of concern.

B) The second method is to use interpolation. If Steps 1-3 are repeated for other ratios of iris thicknesses to broadwall lengths, the results can be used to interpolate other ratios that are within the range of data points. If enough points are taken, straight line approximation can also be used. The problem with this method is that the accuracy is dependent on the number of data points taken.

The major advantage of closed-form equations is the speed it brings to the filter synthesis. In this method there are very few matrices and matrix manipulations that need to be done by the computer. The more matrix manipulation can be avoided, the
faster the program will be. A drastic reduction in the initial synthesis time is achieved by avoiding the iterative search for the iris widths corresponding to the desired K-inverter values. These equations can be used to replace the iterative search used in Section 4.2.4, or the rational function approximation of Section 4.2.5.

Just like the approaches discussed in the two previous sections, the proximity of other iris junctions, and thus the mode interaction between adjacent irises, is neglected in the construction of the closed-form equations. Due to this fact, an analysis of a filter designed by this method will show a frequency response that is not the same as the desired one. It is because of this deviation in the frequency response from the desired response that optimization is needed for tuning the filter design. In order to show the effects of neglecting the mode interaction, and the effects of equation approximation error, a waveguide filter was designed without the use of optimization. Figure 4.8 shows the frequency response of a symmetric iris coupled filter that was created using this modified K-inverter approach. The input design parameters are for an X-band filter, and were as follows:

- Lower cutoff frequency \( f_L \) = 9 GHz
- Upper cutoff frequency \( f_H \) = 10 GHz
- Isolation bandwidth \( \Delta f_i \) = 2 GHz
- Isolation loss \( L_A \) = 30 dB
- Waveguide width \( a \) = 22.86 mm
- Iris thickness \( t \) = 1.5 mm
- Return loss \( RL \) = 16 dB
- Waveguide height \( b \) = 10.16 mm
As can be seen by Figure 4.8, the frequency response does not satisfy the input specifications of the filter. It appears that the centre frequency is lower than expected, and the ripple bandwidth is larger than desired. In contrast to the normal K-inverter method and the rational function approximation, it seems that the equations in the
modified K-inverter method exaggerate the coupling between adjacent resonators, causing the larger bandwidth. As with the other methods, these differences can be taken care of using optimization.

Despite the advantages, this method has the same drawbacks that [7] has, in that it covers a large range of widths and phases. This will present a problem in accurately characterizing the iris junctions. It is also limited in the fact that the formulas have built in normalized ranges in which the formulas are accurate. For example, the ratio of iris thickness to broadwall length can not be lower than 6.562×10⁻², or higher than .1312. With a broadwall length of 22.86 mm (for the X-band), these values are equivalent to an iris thickness of 1.5 mm and 3 mm respectively. Due to these limitations, the method in Section 4.2.5 is used for the synthesis of the filter designs.

4.3 Summary

This chapter briefly discusses the theory behind the design approach for iris coupled bandpass waveguide filters. Three different design approaches are examined, the K-inverter method, rational function approximation, and the modified K-inverter method. The K-inverter approach is a slow iterative search for the iris width which will produce the desired K-inverter value. The modified K-inverter approach uses closed-form equations to characterize the iris width and phase with respect to the desired K-inverter value. These equations tend to exaggerate iris coupling. Rational function approximation maintains both accuracy and speed in the synthesis of iris junctions.
Chapter 5

Optimization

When the initial approach to the design of a filter results in a frequency response that differs from that which is specified, some method, whether manual or by CAD, is required to tune the filter dimensions to achieve the desired frequency response. Practical problems such as component parasitics may prevent a pure theoretical solution for the design of waveguide filters. When only a few variables in a design require adjustment, manual tuning of the filter dimensions can be an effective tool. As the number of variables increase, visualization of the multidimensional variable space is difficult, and manual tuning becomes less effective. The difference between the desired and the achieved frequency response can be considered an error. This error can be defined in terms of an objective (error) function, and the smaller the value of the error function, the closer the achieved frequency response is to the desired response. The optimization program, through an iterative process, reduces the error to a minimum, arriving at a final filter design in terms of the optimized filter parameters.
There are two important steps in optimization: the determination of a search direction and the search for the minimum in that direction. There are two different ways of carrying out the determination of the search direction: the gradient method and the direct search method. The gradient method uses information about derivatives of the performance functions for arriving at the modified set of parameters. Direct search algorithms do not use gradient information, and parameter modification is carried out by searching for the optimum in a systematic manner. The direct search method has been used in this thesis.

5.1 Traditional Optimization Approaches

In the past, optimization was done by using a brute force approach. In brute force approaches optimization is accomplished by repeatedly analysing the filter response, and directly changing the iris widths and resonator lengths until the desired response is achieved. Since the optimization variables are both the iris widths and resonator lengths, for a symmetric \( N_f \) - pole filter, this produces \( N_f + 1 \) optimization variables. This approach to optimization is time consuming and memory intensive, especially when dealing with large order filters, i.e. \( N_f \) is large. The brute force method also has a tendency to converge to local minima, thus forcing the designer to change the input design parameters and begin the design process all over again. Figure 5.1 shows a general flowchart for this method of optimization.
Figure 5.1:  Brute force optimization flowchart.

Another approach to optimization is to optimize the K-inverters [5]. As mentioned previously, for an $N_f$-pole filter of symmetrical structure, there are $N_f + 1$ K-inverters, out of which $N_f^2 + 1$ have different values [5]. In other words, every set of iris widths and resonator lengths corresponds to a set of K-inverters. When the K-inverter values are optimized, the iris widths and resonator lengths can be found by back calculation.
Therefore, using the optimization of K-inverters rather than directly optimizing the filter dimensions reduces the number of optimization variables to $N/2 + 1$. Because of the smaller number of variables that must be handled, this method is faster than the brute force method. However, this method is still slow when dealing with large order filters, and also has the same problem with convergence to local minima that the brute force method has, although at a reduced level. Figure 5.2 shows the general flowchart for the K-inverter optimization method [5].

![Flowchart](image)

**Figure 5.2:** K-inverter optimization flowchart.
Although Reference [5] used optimization of K-inverters, it mentioned that adjusting the
K-inverters will eventually alter the three input specifications, provided the order of the
filter is kept constant. Those specifications are the bandwidth, the centre frequency and
the return loss of the filter.

5.2 Specification Level Optimization Approach

As discussed in detail in Section 1.3.2, this thesis proposes a new approach to the
optimization process. Instead of using the physical dimensions or the K-inverter values
as the optimization variables, the return loss and the upper and lower cutoff frequencies
will be used instead. This keeps the optimization variables to a total of three no matter
what the order of the filter, thus decreasing the time needed for optimization.

The idea behind this method is that every time the filter specifications are entered into
the synthesis program, the program will design a filter with a frequency response that is
different than that which was desired. Therefore, there must be filter specifications out
there which when entered into the synthesis program will produce the desired frequency
response. The object of this technique is to find the “shadow” filter specifications
which, when used by the filter synthesis algorithm, will produce the correct response.

Figure 5.3 shows the general flowchart for this new method of optimization. Any design
program using this method will create a filter, analyse the structure and check to see if
the response matches the desired one. If the response is different than that which was
desired, the initial specifications of the filter are changed and the design process is
started all over again. This will continue until the correct filter response is finally
reached. It should be noted that the order of the filter remains the same throughout the
entire design process, even though the return loss is being changed. Optimization is
carried out by a routine found in the IMSL library, which can be seen in Appendix G.

![Flowchart](image)

Figure 5.3: Specification level optimization flowchart.
5.3 Penalty Function

Each set of component values results in an error from the desired response. This error is defined in terms of an objective function, and the object of the optimization routine is to bring the value of this function to a minimum. This function is referred to as the penalty function of the system. The penalty function is not unique to the system, but rather it can be defined in many different ways. For example the penalty function can be defined in such a way as to cause the optimization routine to converge as close as possible to a specified value, or it can be defined to converge to a design in which the system tolerances have been exceeded. The penalty function used for the optimization routine in governing the optimization method developed in this thesis is given by [8]

\[ P_F = \text{weight} \sum_{q=1}^{\text{band point}} \left( \frac{RL(\text{desired pass band})}{RL(\text{Frequency}_q)} \right)^2 + \frac{1}{\text{weight}} \sum_{r=1}^{\text{band point}} \left( \frac{L_A(\text{desired lower stop band})}{L_A(Frequency_r)} \right)^2 + \frac{1}{\text{weight}} \sum_{s=1}^{\text{band point}} \left( \frac{L_A(\text{desired upper stop band})}{L_A(Frequency_s)} \right)^2. \] (5.1)

The variable weight defines the relation between the role of the pass band and stop band in the penalty function, band point is the number of uniformly placed frequency points in the band of interest. The value of weight for all designs in this thesis was 1. In order to minimize the penalty function it is desired to make the isolation loss $L_A$ (dB) in the stop band as large as possible and the return loss $RL$ (dB) over the passband as large as possible. The return and isolation loss are found by analysing the filter structure at
uniformly placed frequency points using a very accurate analysis routine. The desired values are part of the original design parameters. Figure 5.4 shows the areas of interest in the pass band and stop band for solving the penalty function.

The particular penalty function used in this thesis will not cause convergence to an exact design specification, but rather it will cause the optimization routine to try and maximize the bandwidth and isolation loss in the stop bands, while maximizing the return loss in the passband.

5.4 Benefits of the New Optimization Approach

One important aspect of this technique is that the initial accuracy of the synthesis approach is not as important to the convergence of a final solution as it is in other optimization methods. The goal of many filter design methods is to produce an initial
design which has physical dimensions approaching the desired filter dimensions. A more accurate initial design will decrease the number of iterations needed to reach a final solution, and it helps avoid convergence to a local minima. However, more accurate design techniques tend to be slow, thus forcing an engineering trade off between speed and accuracy. Since this new optimization technique is dependent on the speed of the synthesis process, a faster and maybe less accurate design approach can be chosen.

Another benefit to the specification level optimization design approach is that there is a low chance of the optimization software finding a local minima. Convergence to a local minima is a real problem that must be addressed when using optimization software. The problem comes from the fact that the more variables used in the penalty function, the more local minima will be produced. When a local minima results, the input design parameters must be adjusted, and the design must be started over again. This wastes precious time. When designing structures which contain several filter components, such as multiplexers, this problem of convergence to a local minima can become overwhelming and time consuming. Decreasing the number of optimization variables will help reduce the time wasted in converging to a local minima.

One of the arguments for using a slower and more accurate synthesis process is that it will reduce the likelihood of converging to a local minima. The hope of the design engineer is that the more accurate approach will produce an initial design that will
prevent convergence to a local minima in larger order filters. This means that the design engineer has to chose an approach that will be accurate enough for larger order filters, while maintaining speed for the lower order filters. Since the number of optimization variables in this new technique remains constant, the likelihood of converging to a local minima does not increase with larger order filters, thus defending the choice to use a faster and less accurate initial circuit theory based synthesis approach, which is subsequently used as the input to the optimization routine.

5.5 Summary

This chapter contains a general discussion on the traditional approaches, brute force and K-inverter, for filter optimization. Also presented in this chapter is a new approach which uses only three optimization variables, regardless the order of the filter. The optimization variables used are the upper and lower cutoff frequency and the return loss. Because of the fixed number of variables, the optimization routine will require smaller matrices and therefore less processing time to converge to a final solution.
Chapter 6

Results and Discussion

This chapter presents the results of six band-pass waveguide iris coupled filters, obtained from the developed CAD software based on the design methods discussed in previous chapters. It is very expensive and impractical to verify the design process experimentally, therefore the analysis method, mode matching, discussed in Chapter 3 will be used to verify the filter responses. As discussed in previous chapters, the new optimization technique does not require an exact synthesis method. However, it is critical for the optimization technique to have an exact analysis method. The validity of the analysis routine based on the mode matching method discussed in Chapter 3 can be seen in Appendix F.

Figures 6.1 through 6.12 show the frequency responses and Tables 6.1 through 6.18 show the physical designs for six different filters: two X-band filters, two KU-band filters and two double iris coupled filters. Each set of results contains the initial design specifications and shadow design specifications, which when synthesised produce the desired response. The responses for the manufactured double iris filters are also included. Following the results there is a discussion of the findings.
6.1 Symmetric X Band Filter Results

The input design specifications for the filter are as follows:

Lower cutoff frequency \( (f_L) = 9 \text{ GHz} \)
Upper cutoff frequency \( (f_H) = 10 \text{ GHz} \)
Isolation bandwidth \( \Delta f = 2 \text{ GHz} \)
Isolation loss \( L_A = 30 \text{ dB} \)
Waveguide width \( a = 0.9 \text{ inches} \)
Waveguide height \( b = 0.4 \text{ inches} \)
Iris thickness \( t = 0.0394 \text{ inches} \)

Return loss \( RL = 16 \text{ dB} \)

Table 6.1: Initial and Shadow Design Parameters for the Symmetric X-Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>Lower Cutoff Frequency (GHz)</th>
<th>Upper Cutoff Frequency (GHz)</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>9</td>
<td>10</td>
<td>16</td>
</tr>
<tr>
<td>Shadow</td>
<td>8.929</td>
<td>10.0183</td>
<td>17.8019</td>
</tr>
</tbody>
</table>

Table 6.2: Iris Width Dimensions for Initial and Optimized Symmetric X Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>( W_1 ) (inches)</th>
<th>( W_2 ) (inches)</th>
<th>( W_3 ) (inches)</th>
<th>( W_4 ) (inches)</th>
<th>( W_5 ) (inches)</th>
<th>( W_6 ) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.5263</td>
<td>0.3981</td>
<td>0.3645</td>
<td>0.3645</td>
<td>0.3981</td>
<td>0.5263</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.5387</td>
<td>0.4130</td>
<td>0.3781</td>
<td>0.3781</td>
<td>0.4130</td>
<td>0.5387</td>
</tr>
</tbody>
</table>

Table 6.3: Resonator Lengths for Initial and Optimized Symmetric X Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>( I_{R1} ) (inches)</th>
<th>( I_{R2} ) (inches)</th>
<th>( I_{R3} ) (inches)</th>
<th>( I_{R4} ) (inches)</th>
<th>( I_{R5} ) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.6411</td>
<td>0.7232</td>
<td>0.7388</td>
<td>0.7232</td>
<td>0.6411</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.6226</td>
<td>0.7043</td>
<td>0.7211</td>
<td>0.7043</td>
<td>0.6226</td>
</tr>
</tbody>
</table>
Figure 6.1: Isolation loss for initial and optimized X-band symmetric filter design.
Figure 6.2: Return loss for initial and optimized X-band symmetric filter design.
6.2 Asymmetric X Band Filter Results

The input design specifications for the filter are as follows:

- Lower cutoff frequency \((f_L) = 9\, \text{GHz}\)
- Upper cutoff frequency \((f_H) = 10\, \text{GHz}\)
- Isolation bandwidth \((\Delta f) = 2\, \text{GHz}\)
- Isolation loss \((L_A) = 40\, \text{dB}\)
- Return loss \((RL) = 16\, \text{dB}\)
- Waveguide width \((a) = 0.9\, \text{inches}\)
- Waveguide height \((b) = 0.4\, \text{inches}\)
- Iris thickness \((t) = 0.0394\, \text{inches}\)
- Filter order = 6

**Table 6.4:** Initial and Shadow Design Parameters for the Asymmetric X Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>Lower Cutoff Frequency (GHz)</th>
<th>Upper Cutoff Frequency (GHz)</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>9</td>
<td>10</td>
<td>16</td>
</tr>
<tr>
<td>Shadow</td>
<td>8.9866</td>
<td>10.0130</td>
<td>17.900</td>
</tr>
</tbody>
</table>

**Table 6.5:** Width Dimensions for Initial and Optimized Asymmetric X Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>(W_1) (inches)</th>
<th>(W_2) (inches)</th>
<th>(W_3) (inches)</th>
<th>(W_4) (inches)</th>
<th>(W_5) (inches)</th>
<th>(W_6) (inches)</th>
<th>(W_7) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.5889</td>
<td>0.4835</td>
<td>0.4547</td>
<td>0.4501</td>
<td>0.4547</td>
<td>0.4835</td>
<td>0.5889</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.5998</td>
<td>0.4917</td>
<td>0.4602</td>
<td>0.4546</td>
<td>0.4602</td>
<td>0.4917</td>
<td>0.5998</td>
</tr>
</tbody>
</table>

**Table 6.6:** Resonator Lengths for Initial and Optimized Asymmetric X Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>(l_1) (inches)</th>
<th>(l_2) (inches)</th>
<th>(l_3) (inches)</th>
<th>(l_4) (inches)</th>
<th>(l_5) (inches)</th>
<th>(l_6) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.6607</td>
<td>0.7421</td>
<td>0.7597</td>
<td>0.7597</td>
<td>0.7421</td>
<td>0.6607</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.6502</td>
<td>0.7354</td>
<td>0.7556</td>
<td>0.7556</td>
<td>0.7354</td>
<td>0.6502</td>
</tr>
</tbody>
</table>
Isolation Loss For Initial and Optimized X Band
Asymmetric Filter Design

Figure 6.3: Isolation loss for initial and optimized X-band asymmetric filter design.
Figure 6.4: Return loss for initial and optimized X-band asymmetric filter design.
6.3 Symmetric KU Band Filter Results

The input design specifications for the filter are as follows:

Lower cutoff frequency \( f_L \) = 12.3 GHz
Upper cutoff frequency \( f_H \) = 12.7 GHz
Isolation bandwidth \( \Delta f \) = 1 GHz
Isolation loss \( L_A \) = 40 dB
Return loss \( RL \) = 26 dB
Waveguide width \( a \) = 0.6220 inches
Waveguide height \( b \) = 0.3110 inches
Iris thickness \( t \) = 0.0197 inches
Filter order = 6

Table 6.7: Initial and Shadow Design Parameters for the Symmetric KU Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>Lower Cutoff Frequency (GHz)</th>
<th>Upper Cutoff Frequency (GHz)</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>12.3</td>
<td>12.7</td>
<td>26</td>
</tr>
<tr>
<td>Shadow</td>
<td>12.2877</td>
<td>12.7091</td>
<td>27.0398</td>
</tr>
</tbody>
</table>

Table 6.8: Width Dimensions for Initial and Optimized Symmetric KU Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>( W_1 ) (inches)</th>
<th>( W_2 ) (inches)</th>
<th>( W_3 ) (inches)</th>
<th>( W_4 ) (inches)</th>
<th>( W_5 ) (inches)</th>
<th>( W_6 ) (inches)</th>
<th>( W_7 ) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.3357</td>
<td>0.2150</td>
<td>0.1872</td>
<td>0.1831</td>
<td>0.1872</td>
<td>0.2150</td>
<td>0.3357</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.3415</td>
<td>0.2211</td>
<td>0.1919</td>
<td>0.1874</td>
<td>0.1919</td>
<td>0.2211</td>
<td>0.3415</td>
</tr>
</tbody>
</table>

Table 6.9: Resonator Lengths for Initial and Optimized Symmetric KU Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>( l_1 ) (inches)</th>
<th>( l_2 ) (inches)</th>
<th>( l_3 ) (inches)</th>
<th>( l_4 ) (inches)</th>
<th>( l_5 ) (inches)</th>
<th>( l_6 ) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.6027</td>
<td>0.6731</td>
<td>0.6832</td>
<td>0.6832</td>
<td>0.6731</td>
<td>0.6027</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.5973</td>
<td>0.6698</td>
<td>0.6809</td>
<td>0.6809</td>
<td>0.6698</td>
<td>0.5973</td>
</tr>
</tbody>
</table>
Figure 6.5: Isolation loss for initial and optimized KU-band symmetric filter design.
Return Loss For Initial and Optimized KU Band Symmetric Filter Design

Figure 6.6: Return loss for initial and optimized KU-band symmetric filter design.
6.4 Asymmetric KU Band Filter Results

The input design specifications for the filter are as follows:

Lower cutoff frequency \( f_L \) = 12.3 GHz  
Upper cutoff frequency \( f_H \) = 12.7 GHz  
Isolation bandwidth \( \Delta f \) = 1 GHz  
Isolation loss \( L_A \) = 40 dB  
Return loss \( RL \) = 26 dB  
Waveguide width \( a \) = 0.6220 inches  
Waveguide height \( b \) = 0.3110 inches  
Iris thickness \( t \) = 0.0197 inches  
Filter order = 6

Table 6.10: Initial and Shadow Design Parameters for the Asymmetric KU Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>Lower Cutoff Frequency (GHz)</th>
<th>Upper Cutoff Frequency (GHz)</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>12.3</td>
<td>12.7</td>
<td>26</td>
</tr>
<tr>
<td>Shadow</td>
<td>12.2816</td>
<td>12.7158</td>
<td>28.36</td>
</tr>
</tbody>
</table>

Table 6.11: Width Dimensions for Initial and Optimized Asymmetric KU Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>( W_1 ) (inches)</th>
<th>( W_2 ) (inches)</th>
<th>( W_3 ) (inches)</th>
<th>( W_4 ) (inches)</th>
<th>( W_5 ) (inches)</th>
<th>( W_6 ) (inches)</th>
<th>( W_7 ) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.3983</td>
<td>0.2997</td>
<td>0.2754</td>
<td>0.2714</td>
<td>0.2754</td>
<td>0.2997</td>
<td>0.3983</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.407</td>
<td>0.3091</td>
<td>0.2828</td>
<td>0.2785</td>
<td>0.2828</td>
<td>0.3091</td>
<td>0.407</td>
</tr>
</tbody>
</table>

Table 6.12: Resonator Lengths for Initial and Optimized Asymmetric KU Band Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>( l_1 ) (inches)</th>
<th>( l_2 ) (inches)</th>
<th>( l_3 ) (inches)</th>
<th>( l_4 ) (inches)</th>
<th>( l_5 ) (inches)</th>
<th>( l_6 ) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.605</td>
<td>0.6759</td>
<td>0.686</td>
<td>0.686</td>
<td>0.6759</td>
<td>0.605</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.5946</td>
<td>0.6698</td>
<td>0.6818</td>
<td>0.6818</td>
<td>0.6698</td>
<td>0.5946</td>
</tr>
</tbody>
</table>
Isolation Loss For Initial and Optimized KU Band
Asymmetric Filter Design

Figure 6.7: Isolation loss for initial and optimized KU-band asymmetric filter design.
Return Loss For Initial and Optimized KU Band Asymmetric Filter Design

Figure 6.8: Return loss for initial and optimized KU-band asymmetric filter design.
6.5 Double Iris Filter (Centre Frequency 8.389 GHz) Results

The input design specifications for the double iris filter are as follows:

Lower cutoff frequency \( (f_L) = 8.3120 \) GHz
Upper cutoff frequency \( (f_U) = 8.4660 \) GHz
Isolation bandwidth \( (\Delta f) = 0.5600 \) GHz
Isolation loss \( (L_A) = 68 \) dB
Waveguide width \( (a) = 1.1220 \) inches
Iris thickness \( (t) = 2.5400 \) inches

Return loss \( (RL) = 20 \) dB
Waveguide height \( (b) = 0.4970 \) inches
Filter order = 6

Table 6.13: Initial and Shadow Design Parameters for the Double Iris Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>Lower Cutoff Frequency (GHz)</th>
<th>Upper Cutoff Frequency (GHz)</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>8.312</td>
<td>8.466</td>
<td>20</td>
</tr>
<tr>
<td>Shadow</td>
<td>8.311</td>
<td>8.465</td>
<td>24.3003</td>
</tr>
</tbody>
</table>

Table 6.14: Width Dimensions for Initial and Optimized Double Iris Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>( W_1 ) (inches)</th>
<th>( W_2 ) (inches)</th>
<th>( W_3 ) (inches)</th>
<th>( W_4 ) (inches)</th>
<th>( W_5 ) (inches)</th>
<th>( W_6 ) (inches)</th>
<th>( W_7 ) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.5094</td>
<td>0.3711</td>
<td>0.3492</td>
<td>0.3458</td>
<td>0.3492</td>
<td>0.3711</td>
<td>0.5094</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.52</td>
<td>0.3809</td>
<td>0.3546</td>
<td>0.35</td>
<td>0.3546</td>
<td>0.3809</td>
<td>0.52</td>
</tr>
<tr>
<td>Manufactured</td>
<td>0.513</td>
<td>0.374</td>
<td>0.351</td>
<td>0.347</td>
<td>0.351</td>
<td>0.374</td>
<td>0.513</td>
</tr>
</tbody>
</table>

Table 6.15: Resonator Lengths for Initial and Optimized Double Iris Filter.

<table>
<thead>
<tr>
<th>Filter Design</th>
<th>( l_1 ) (inches)</th>
<th>( l_2 ) (inches)</th>
<th>( l_3 ) (inches)</th>
<th>( l_4 ) (inches)</th>
<th>( l_5 ) (inches)</th>
<th>( l_6 ) (inches)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>0.7872</td>
<td>0.862</td>
<td>0.868</td>
<td>0.868</td>
<td>0.862</td>
<td>0.7872</td>
</tr>
<tr>
<td>Optimized</td>
<td>0.775</td>
<td>0.8582</td>
<td>0.8661</td>
<td>0.8661</td>
<td>0.8582</td>
<td>0.775</td>
</tr>
<tr>
<td>Manufactured</td>
<td>0.783</td>
<td>0.861</td>
<td>0.867</td>
<td>0.867</td>
<td>0.861</td>
<td>0.783</td>
</tr>
</tbody>
</table>
Isolation Loss For Unoptimized, Optimized and Manufactured Double Iris filter

---

Isolation loss of unoptimized double iris filter
Isolation loss of optimized double iris filter
Isolation loss of manufactured double iris filter

**Figure 6.9:** Isolation loss for initial, optimized and manufactured double iris filter design.
Return Loss For Unoptimized, Optimized and Manufactured Double Iris filter

Figure 6.10: Return loss for initial, optimized and manufactured double iris filter design.
6.6 Double Iris Filter (Centre Frequency 8.039 GHz) Results

The input design specifications for the double iris filter are as follows:

Lower cutoff frequency \( (f_L) = 7.962 \text{ GHz} \)
Upper cutoff frequency \( (f_H) = 8.116 \text{ GHz} \)
Isolation bandwidth \( (\Delta f_p) = 0.5600 \text{ GHz} \)
Isolation loss \( (L_A) = 68 \text{ dB} \)
Return loss \( (RL) = 20 \text{ dB} \)
Waveguide width \( (a) = 1.1220 \text{ inches} \)
Waveguide height \( (b) = 0.4970 \text{ inches} \)
Iris thickness \( (t) = 2.5400 \text{ inches} \)
Filter order = 6

| Table 6.16: Initial and Shadow Design Parameters for the Double Iris Filter. |
|-----------------------------|-----------------|-----------------|-----------------|
| Filter Design               | Lower Cutoff Frequency (GHz) | Upper Cutoff Frequency (GHz) | Return Loss (dB) |
| Initial                     | 7.962            | 8.116           | 20              |
| Shadow                      | 7.9607           | 8.1167          | 20.191          |

| Table 6.17: Width Dimensions for Initial and Optimized Double Iris Filter. |
|-----------------------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|
| Filter Design               | \( W_1 \) (inches) | \( W_2 \) (inches) | \( W_3 \) (inches) | \( W_4 \) (inches) | \( W_5 \) (inches) | \( W_6 \) (inches) | \( W_7 \) (inches) |
| Initial                     | 0.5214           | 0.3851          | 0.3627          | 0.3593          | 0.3627          | 0.3851          | 0.5214          |
| Optimized                   | 0.5257           | 0.3899          | 0.3652          | 0.3618          | 0.3652          | 0.3899          | 0.5257          |
| Manufactured                | 0.524            | 0.389           | 0.364           | 0.361           | 0.364           | 0.389           | 0.524           |

| Table 6.18: Resonator Lengths for Initial and Optimized Double Iris Filter. |
|-----------------------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|
| Filter Design               | \( l_1 \) (inches) | \( l_2 \) (inches) | \( l_3 \) (inches) | \( l_4 \) (inches) | \( l_5 \) (inches) | \( l_6 \) (inches) |
| Initial                     | 0.8411           | 0.9234          | 0.9304          | 0.9304          | 0.9234          | 0.8411          |
| Optimized                   | 0.8351           | 0.9215          | 0.9293          | 0.9293          | 0.9215          | 0.8351          |
| Manufactured                | 0.836            | 0.922           | 0.93            | 0.93            | 0.922           | 0.836           |
Isolation Loss For Unoptimized, Optimized and Manufactured Double Iris filter

Figure 6.11: Isolation loss for initial, optimized and manufactured double iris filter design.
Figure 6.12: Return loss for initial, optimized and manufactured double iris filter design.
6.7 Discussion of Results

In the above six examples it is obvious to see that the initial design does not meet the desired design specifications. In each case the initial design has a narrower bandwidth, a shifted centre frequency and the return loss is lower than desired. It can also be seen that after optimization the design specifications have been met, excluding the upper isolation cutoff. The reason why the upper isolation specification is not met is because higher order modes are effecting the response of the filter. As the frequency increases the next higher order mode (TE_{20} for asymmetric filters and TE_{30} for symmetric filters) appears. In order to meet both the bandwidth and upper isolation specification, the width of the waveguide would have to be shortened. This will cause the physical length of the filter to increase, and it will raise the cutoff frequency of the next higher mode. The problem with doing this is that the width and heights for waveguides are standardized. In order to accommodate the standardized connections, tapering would have to be implemented. Implementing tapered waveguide filters is not one of the goals of this thesis. Taking into account that without tapering the upper isolation cutoff will not be met, the previous graphs show that the new optimization approach does in fact converge to a correct and final solution.

It can also be seen that when the upper and lower cutoff frequencies and the return loss are used as the optimization variables, the filter dimensions will in fact be optimized. When using the brute force method, each dimension must be individually dealt with and
adjusted until convergence is met. In the case of the specification level optimization approach, adjusting the three optimization variables will ultimately effect the filter dimensions. Optimizing the cutoff frequencies and the return loss will have the same effect on the filter dimensions as if each iris width and length was individually adjusted until convergence. However, in the case of the brute force method of optimization, the time to convergence is magnitudes higher than the new optimization method.

It was also observed that the specification level optimization approach decreased the likelihood of convergence to a local minima. For example all six filters dealt with in this chapter, when using the brute force method of optimization, converged to a local minimum after an average of two hours of computation time. When this happens the filter specifications must be changed, and the design process must be started all over again. When the brute force method does converge to a final solution, the program will still take about three hours to reach a final design. However, when using the specification level optimization approach, each of the six filters converged to a final solution in the first attempt. The program converged to a final solution in a time of about seventeen to thirty five minutes. This is a significant reduction in time as compared to the brute force method of optimization.

The reason for this significant reduction can be understood when it is recognized that the penalty function is a highly nonlinear and complex function. A system of nonlinear equations has the form
A general system of $n$ nonlinear equations in $n$ unknowns can alternatively be represented by defining a functional (function of several functions) $F$.

$$F(f_1(x_1, x_2, \ldots, x_n), f_2(x_1, x_2, \ldots, x_n), \ldots, f_n(x_1, x_2, \ldots, x_n)) = 0$$  \hspace{1cm} (6.2)

In a linear system, the method of finding the minimum would be to take the derivative of the system and equate it to zero. The derivative of a single variable function describes how the values of the function change relative to a change in its independent variable. The functional $F$ has $n$ different variables $x_1, x_2, \ldots, x_n$ and $n$ different component functions $f_1, f_2, \ldots, f_n$, each of which can change as any one of the variables changes. The natural way of representing the derivative of the system is by an $n \times n$ matrix.

$$J(x) = \begin{bmatrix}
\frac{\partial f_1(x)}{\partial x_1} & \frac{\partial f_1(x)}{\partial x_2} & \cdots & \frac{\partial f_1(x)}{\partial x_n} \\
\frac{\partial f_2(x)}{\partial x_1} & \frac{\partial f_2(x)}{\partial x_2} & \cdots & \frac{\partial f_2(x)}{\partial x_n} \\
\vdots & \vdots & \ddots & \vdots \\
\frac{\partial f_n(x)}{\partial x_1} & \frac{\partial f_n(x)}{\partial x_2} & \cdots & \frac{\partial f_n(x)}{\partial x_n}
\end{bmatrix} \hspace{1cm} (6.3)$$
This is known as the Jacobian matrix of the functional. Looking at the matrix it can be seen that the number of maxima and minima created depends on number of variables contained in the component functions. In the case of the specification level optimization design approach, the number of variables is always three. Other optimization approaches which utilize more than three optimization variables will produce more local maxima or minima. This problem will be even more severe when large order filters are designed. However, since when using the specification level optimization approach the number of optimization variables remains constant at three, irrespective of the order of the filter, the likelihood of convergence to a local minima will be minimized. When the order of the filter is small, third order or less, there will be no observable difference between the different optimization techniques.

The dimensions of the manufactured X-band double iris filters can be seen in Table 6.15 and Table 6.18. The physical dimensions were analyzed, and the results can be seen in Graphs 6.9 through 6.12. As can be seen the manufactured filters have a different response than the optimized filter. This difference is due to mechanical tolerances. The process used for manufacturing the filters is unknown. Appendix F shows the results of a filter designed using the EDM process. The EDM process enables a mechanical tolerance of $4000^{th}$ of an inch. The double iris filters were not manufactured using the EDM process, and therefore mechanical tolerances effect the response of the filter. Had the EDM process been used, the optimized and manufactured responses would have been comparable. Proof of this is given in Appendix F.
Chapter 7

Conclusions

The objectives of this research were:

1. To develop efficient computer aided design and analysis tools for H-plane iris coupled filters.

2. To develop a faster method of modeling using the K-inverter design approach.

3. To implement a new optimization technique that will utilize 3 optimization variables, irrespective of the filter order, in the design of H-plane iris coupled bandpass waveguide filters.

4. To reduce the likelihood of convergence to a local minimum in optimization.

5. To use the developed CAD software to design symmetrical, asymmetrical and double iris coupled waveguide filter structures.

In accordance with the above objectives, the conclusions of the thesis are as follows:
A complete computer aided design method was developed for the design of H-plane iris coupled waveguide bandpass filters. Several X and KU-band symmetric and asymmetric H-plane iris bandpass filters, and two X-band double H-plane iris filters were designed. The initial design of the filters showed that the synthesis routine was incapable of designing a filter with the correct frequency response. Fine tuning was accomplished by using a specification level optimization approach. The final designs met the design specifications, excluding the upper isolation cutoff, without the use of tuning screws.

A faster method of modeling using the K-inverter design approach was achieved by using closed form equations to model an iris junction. There was a significant increase in the speed of the synthesis program, but there was a decrease in the accuracy of the modeling. It appears that the closed form equations exaggerate the coupling between adjacent resonators. The exaggeration of the coupling was significant enough that the closed form equations were not used in the final synthesis program. Instead, equations which characterize the iris junction are created by interpolating tables of data which contain the K-inverter and phase values for iris junctions at varying widths. This technique is fast and accurate, making it suitable for use with the optimization routine.

A systems level optimization technique that utilizes three optimization variables irrespective of the filter order was implemented in the design of H-plane iris coupled waveguide bandpass filters. There was a significant reduction in the time required for
convergence to a final solution. One of the reasons for the reduction in time is that the number of optimization variables remains constant. With other methods of optimization in which the number of optimization variables is dependent on the order of the filter, the optimization routine must deal with large matrices. It is this manipulation of large matrices which tends to slow down the speed of the design program. Keeping the number of optimization variables constant and to a minimum allows for smaller matrices, thus increasing the speed of the optimization program.

Another reason for the increase in speed of the design is that it was discovered that the likelihood of convergence to a local minima was reduced. Other optimization approaches, which have optimization variables that are dependent on the order of the filter, have a tendency to converge to a local minima when larger order filters are designed. The problem can be found in the fact that it is impossible to find an analytical solution to the penalty function. If it was possible to find an analytical solution, optimization would not be necessary. The penalty function is a nonlinear function, and the number of maxima and minima created depends on the number of variables contained in the component functions. Therefore, the more variables needed to define the objective function the more maxima and minima are created. Reducing the number of variables will in turn decrease the number of maxima and minima, thus minimizing the possibility of convergence to a local minima.

The specification level optimization approach was implemented in CAD software which
designs symmetric, asymmetric and double iris waveguide bandpass filters. The specification level optimization design approach is a fast and accurate method for designing waveguide filters. For each filter designed in this thesis, optimization was carried out using the specification level and brute force method. When using the brute force method, the design process, if a local minima was not converged to, took close to two hours to complete. When using the specification level optimization approach, the design process never converged to a local minima, and the process took about twenty minutes to complete. The reduction in computation time and the unlikelihood of convergence to a local minima makes it ideal for use in design of more complicated structures which are comprised of multiple filters (e.g. multiplexers). Reducing the time to completion for individual filter designs will in turn decrease the time to completion for the structure as a whole. Application for such an optimization approach is widespread.

7.1 Future Work

In this thesis, the specification level optimization approach was applied solely to the design of H-plane iris coupled bandpass waveguide filters. Future work could involve the design of iris filter structures with E-plane and both EH-plane discontinuities. The design of septum coupled filters would also be another area in which the advantages of the new optimization approach could be seen.

Bandpass filters play an important role in the design of more complex structures like multiplexers. The number of filters in a design depends on the number of channels in
the multiplexer. For example, if a 10-channel multiplexer with 9-pole Chebyshev function filters was to be designed, it would require 121 optimization variables in total to optimize the multiplexer. One hundred of these variable are required just for the design of the filters. Using the specification level optimization approach, only 30 variables are needed in the optimization of the filters, bringing the total to 51 variables required for the multiplexer. As can be seen, the design of large channel multiplexers becomes very complicated and time consuming when using the brute force method. The advantages of decreasing the computation time using the new optimization approach would clearly be seen in the design of a multiplexer.

Another area of future study would be in the penalty function. This thesis observed the effects of the new optimization approach when using a penalty function that gives the maximum isolation loss and minimum return loss in the pass band. Other penalty functions that may cause convergence to desired values should also be studied.

In conclusion, the specification level optimization approach is not restricted to the field of microwave filter design. It can be used in many other branches of engineering where optimization and minimization are necessary.
References


Appendix A

Figure A1 shows the relationship between the K-inverter value and its corresponding normalized Iris width. This particular graph was obtained by maintaining the iris thickness at 1.875 mm and the frequency of interest at 10 GHz.

**K-Inverter Value Verses Iris Width Graph**

Width vs K-Inverter Value for Center Frequency of 10 GHz (t = 1.875 mm)

\[ \text{Width} = (A_w + C_w K_v + D_w K_v^2 + G_w K_v^3) / (1 + B_w K_v + D_w K_v^2 + F_w K_v^3) \]

\( r^2 = 0.99999942 \)
\( \text{DF Adj. } r^2 = 0.99999966 \)
\( \text{FStat} = 25340.277 \)

\( A_w = 0.031643728 \)
\( B_w = 30.425942 \)
\( C_w = 10.573237 \)
\( D_w = 22.066135 \)
\( E_w = 33.331228 \)
\( F_w = 31.930339 \)
\( G_w = -19.581803 \)

**Figure A1:** K-inverter verses iris width graph.
Appendix B

Figure B1 shows the relationship between the K-inverter value and its corresponding iris phase. This particular graph was obtained by maintaining the iris thickness at 2.25 mm and the frequency of interest at 11 GHz.

**K-Inverter Value Verses Iris Phase Graph**

Phase vs K-Inverter Value for Center Frequency of 11 GHz (t = 2.25mm)

\[
\text{Phase} = \frac{A_p + C_p K_v + E_p K_v^2 + G_p K_v^3}{1 + B_p K_v + D_p K_v^2 + F_p K_v^3}
\]

\[
r^2 = 0.999999997 \quad \text{DF Adj} \quad r^2 = 0.9973429 \quad \text{FitStdErr} = 0.001393935 \quad \text{Fstat} = 442.70214
\]

\[
A_p = 0.0300042414, B_p = 30.280765, C_p = 3.04507241, D_p = 1.30019993
\]

\[
E_p = 128.86899, F_p = 68.501675, G_p = 154.2567
\]

**Figure B1:** K-inverter verses iris phase graph.
Appendix C

Normalized Phase Coefficients

Equations (C1 to C7) are the normalized coefficients for the rational function which describes the K-inverter value and phase relationship of an iris junction. The equations are

\[ A_p = 579.68319 + 94.649118 F_{on} - 7.8088959 F_{on}^{1.5} \]

\[ - 170.92378 \ln(F_{on})^2 - 1620.0489 \frac{\ln(F_{on})}{F_{on}} , \quad (C1) \]

\[ B_p = 884903.19 + 191952.37 F_{on} - 249.46294 F_{on}^2 \ln(F_{on}) \]

\[ - 305223.6 F_{on}^{0.5} \ln(F_{on}) - 2277878.3 \frac{\ln(F_{on})}{F_{on}} , \quad (C2) \]

\[ C_p = 26174.004 - 27651.506 F_{on} + 3542.7538 F_{on} \ln(F_{on}) \]

\[ + 27021.345 F_{on}^{0.5} \ln(F_{on}) - 5279.7357 \ln(F_{on})^2 , \quad (C3) \]

\[ D_p = 417711.84 - 6485.4026 F_{on} + 3.5156384 F_{on}^3 \]

\[ - \frac{1025070.4}{\ln(F_{on})} + 3857285.8 \frac{\ln(F_{on})}{F_{on}^2} , \quad (C4) \]
\[ E_p = -3396224.8 - 555948.4F_{on} + 45894752F_{on}^{1.5} \]

\[ + 1003064.7 \ln(F_{on})^2 + 9494470.4 \frac{\ln(F_{on})}{F_{on}} , \]  
(C5)

\[ F_p = -314085.05 - 68443.672F_{on}^{1.5} + 89.371321F_{on}^{2} \ln(F_{on}) \]

\[ + 108729.09 F_{on}^{0.5} \ln(F_{on})^{0.5} + 808865.09 \frac{\ln(F_{on})}{F_{on}} , \]  
(C6)

\[ G_p = 2219643.7 + 482697.67F_{on} - 628.99578F_{on}^{2} \ln(F_{on}) \]

\[ - 767133.83 F_{on}^{0.5} \ln(F_{on}) - 5715229.3 \frac{\ln(F_{on})}{F_{on}} . \]  
(C7)
Appendix D

Normalized Width Coefficients

Equations (D1 to D7) are the normalized coefficients for the rational function which describes the K-inverter value and phase relationship of an iris junction. The equations are

\[ A_w = -61.46887 - 4.5962388F_{on} + 0.011544835F_{on}^3 \ln(F_{on}) \]
\[ + 12.994323(\ln(F_{on}))^2 + 156.20802 \frac{\ln(F_{on})}{F_{on}} \], \hspace{1cm} (D1)\]

\[ B_w = -2162.3324 + 28071.802F_{on} - 10.771302F_{on}^{2.5} \]
\[ - 69549.123F_{on}^{0.5} \ln(F_{on}) + 43623.021 \ln(F_{on})^2 \], \hspace{1cm} (D2)\]

\[ C_w = -812.32245 + 10534.997F_{on} - 4.039303F_{on}^{2.5} \]
\[ - 26104.511F_{on}^{0.5} \ln(F_{on}) + 16376.232 \ln(F_{on})^2 \], \hspace{1cm} (D3)\]

\[ D_w = 79094.599 - 451.83602F_{on} - \frac{664546.31}{F_{on}^{0.5}} + \frac{1972930.7}{F_{on}^{2}} \]
\[ - 2680854.5 \frac{\ln(F_{on})}{F_{on}^2} \], \hspace{1cm} (D4)\]
\[ E_w = 286264.56 - 189372.72F_{on} + 40057.792F_{on}\ln(F_{on}) \]
\[ - 47.245F_{on}^{2.5} + 132036.41\ln(F_{on})^2, \]  
\[ (D5) \]

\[ F_w = -422.71551 - 4.429783F_{on}^{353144.6} + \frac{353144.6}{F_{on}^{1.5}} \]
\[ - 464355.19\frac{\ln(F_{on})}{F_{on}^{2}} - 312436.37e^{-F_{on}}, \]  
\[ (D6) \]

\[ G_w = 84172.56 - 2448.3182F_{on} + 275.74285F_{on}^{1.5} \]
\[ - \frac{194685.8}{\ln(F_{on})} + 700700.84\frac{\ln(F_{on})}{F_{on}^{2}}. \]  
\[ (D7) \]
Appendix E

Coefficients For The K-Inverter Scaling Factor

Equations (E1 to E4) are the coefficients for the K-inverter scaling factor. Once the K-inverter value has been scaled, it can then be used in the closed form equations to find the corresponding iris width and phase. The equations are

\[ A_F = \frac{1.0434188 + 0.249419141\ln(K_v)}{1 + 0.269613261\ln(K_v) + 0.020911585\ln(K_v)^2} \]  
\[ (E1) \]

\[ B_F = \frac{-0.055652236 + 0.068930716\ln(K_v) + 0.019779171\ln(K_v)^2}{1 + 0.297564111\ln(K_v) + 0.025563977\ln(K_v)^2} \]
\[ (E2) \]

\[ C_F = \frac{-0.0001693316 - 0.0029341502\ln(K_v) - 0.00073478982\ln(K_v)^2}{1 + 0.280592031\ln(K_v) + 0.021383755\ln(K_v)^2} \]
\[ (E3) \]

\[ D_F = 0.13929634 - 0.28712019K_v + 0.013470536\frac{\ln(K_v)}{K_v} \]
\[ + \frac{0.09300075}{K_v} + \frac{0.00033625951}{K_v^{1.5}} \]  
\[ (E4) \]
Appendix F

Proof of Accuracy for the Mode Matching Analysis Routine

Figure F1 shows the comparison between the theoretical and experimental results for a three-pole KA-band iris coupled filter. As can be seen, the results are almost identical.

![Experimental and Analysis Results For Three-Pole Filter](image)

Figure F1: Theoretical and experimental results for a three-pole KA-band filter.
Figure F2 shows the theoretical results for an 11 pole iris coupled filter. The experimental results for the fabricated filter are shown in Figure F3. A slight variation can be seen between the two graphs. This difference is due to mechanical tolerances. The 3rd order filter shown in Figure F1 was fabricated using the EDM process, which enables a mechanical tolerance of 4000th of an inch. The 11th order filter was not fabricated using the EDM process, and therefore the mechanical tolerance had a greater effect on the frequency response. Thanks go out to MCI for their experimental results.

**Theoretical Results for 11 Pole Filter**

![Theoretical Results for 11 Pole Filter Diagram](image)

**Figure F2:** Theoretical results for the 11th order waveguide filter.
Experimental Results For The Isolation Gain
(11 Pole Filter)

Figure F3: Results of Fabricated 11th order filter.
Appendix G

Optimization Algorithm

The optimization routine BCPOL, found in the IMSL library, uses the complex method to find a minimum point of a function, \( f(x) \), of \( n \) variables subject to bounds. This method is based on function comparison, where no smoothness is assumed. It starts with \( 2n \) points \( x_1, x_2, \ldots, x_{2n} \). At each iteration, a new point is generated to replace the point among the \( 2n \) points which has the largest function value, \( x_w \). The new point, \( x_k \) is constructed by

\[
x_k = c + \alpha(c - x_w), \quad (1)
\]

where,

\[
c = \frac{1}{2n - 1} \sum_{i=1}^{2n} x_i, \quad (2)
\]

\( i \neq w \), and \( \alpha \) is the reflection coefficient and must be larger than \( 0 \).

When \( x_k \) is a best point, that is, when \( f(x_k) \leq f(x_i) \) for \( i = 1, 2, \ldots, 2n \), an expansion point is computed. The expansion point is given by

\[
x_e = c + \beta(x_k - c), \quad (3)
\]

where \( \beta \) is called the expansion coefficient and must be larger than \( 1 \). If the new point is a worse point, then the complex would be contracted to get a better new point. If this is unsuccessful, the complex is shrunk by moving the vertices halfway toward the current best point. Whenever the new point generated is beyond the bound, it will be set to the
bound. This procedure is repeated until the relative error in the function values is less
then a specified tolerance $\epsilon$, with

$$f_{\text{best}} - f_{\text{worst}} \leq \epsilon(1 - |f_{\text{best}}|),$$

(4)

or

$$\sum_{i=1}^{2n} \left( f(x_i) - \frac{\sum_{j=1}^{2n} f(x_j)}{2n} \right) \leq \epsilon,$$

(5)

where $f_i = f(x_i)$, $f_j = f(x_j)$.