Dedication

To my parents who taught me all the great virtues of life
Abstract

This thesis describes the synthesis and design of two novel dual-mode microwave waveguide filter structures. The mode degeneracy is fundamental to this design technique. The first design introduces a structurally symmetric dual-mode cavity pseudo-elliptic band-pass filter, which has been implemented by only asymmetric structures so far. It is also used to shed light on the concepts of positive and negative coupling in dual-mode filters. A novel iris coupling structure is investigated allowing negative coupling coefficients required for certain polarizations of modes in dual-mode rectangular cavities. A 4th order dual-mode band-pass filter is designed at 11 GHz with a bandwidth of 200 MHz, an in-band return loss of 20 dB and two transmission zeros at normalized frequencies of ±1.6. The filter makes use of the negative coupling property of the iris for inter-resonance coupling of the modes of horizontal orientation inside the cavity. The desired performance of the filter was achieved by the initial design without any optimization.

The second design is especially convenient when temperature compensation is of prime concern. A direct design technique is developed and demonstrated with computer simulation and measurement in circular waveguide technology on a 4th order filter with a bandwidth of 50 MHz at 11 GHz, an in-band return loss of 20 dB and two normalized transmission zeros at ±1.5.
Acknowledgments

I would like to express my deepest gratitude for my supervisor Dr. Smain Amari for all the help, time and support he provided during my masters degree.

I would like to express gratitude to my friend and colleague Maged Bekheit for all the technical tips he offered as well as the moral support.

I would like to acknowledge Gary Contant from the department of physics at Queen’s University for his help with the filter fabrication as well as David Lee at the CRC and my colleague Ahmed El-Gabaly at Queen’s university for their help with the measurements of the filters.

I am also grateful to my family in Egypt who supported me during my two years of study at Queen’s University and to my friends in Kingston Abdel-Hamid Taha, Ayman Radwan and Kate Sutton for all the support and the fun times.
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List of Symbols

Γ: Reflection Coefficient.
B: Propagation Constant.
ε: Material Permittivity.
λ: Wavelength.
λ₀: Free Space Wavelength.
λ₉₉: Guided Wavelength.
μ: Material Permeability.
φ: Phase Difference.
ω: Angular frequency.
C: Capacitance.
B: Bandwidth.
f₀: Resonant Frequency.
f₉₉: Cutoff Frequency.
g: Chebyshev Polynomial Coefficient.
K: Impedance Inverter Factor.
L: Inductance.
M: Coupling Coefficient.
N: Filter Order.
R: Load/Source Resistance.
X: Reactance.
Z₉: Characteristic Impedance.
Z_L: Load Impedance.
Chapter 1

Introduction

Microwave components play a very important role in the field of wireless communication. The strict frequency requirements in different applications such as satellite, navigation, radar, biomedical, cellular systems, etc. created the need for more efficient, smaller and low-cost microwave devices used mainly in the front-end of these systems. In systems operating at microwave frequencies, which range roughly from 100 MHz to 100 GHz (or a wavelengths ranging from 3mm to 3m), passive resonant components such as filters and antennas play a vital role. The design of microwave devices varies according to the nature of the device and the frequency band of operation.

The fundamentals of microwave filter theory and design were established in the 1950’s and the 1960’s of the last century. Since then, significant progress has been made in order to meet the increasingly stringent requirements for a variety of applications. Modern microwave filters are indeed expected to exhibit low in-band insertion loss, linear phase, sharp cutoff rate, small size and light weight. In satellite communications, dual-mode microwave band-pass filters, which were introduced in the 1970’s, remain the unchallenged technology as they come very close to satisfying all these often contradictory requirements. Dual-mode filters exploit the two degenerate modes of a waveguide cavity in order to reduce the size and weight of the structure while offering the possibility of implementing sophisticated pseudo-elliptic filtering functions. They are
commonly designed through the well established theory of coupled single-resonator filters. The presence of mode degeneracy is not included in this theory.

The remainder of this chapter discusses different types of microwave filtering structures, with special emphasis on structures using dual-mode resonators which are the building blocks of the novel filters in this work.

1.1 Lumped Resonator Filters

A lumped resonator is an LC tank circuit consisting of parallel or series inductors and capacitors. Loss in the resonator is modeled by a resistor, \( R \). A lumped filtering structure is an assemblage of lumped resonators in series or in parallel. These structures have the advantage of being very compact, but the low quality factor of the resonators, especially at higher frequencies, leads to relatively poor performance.

The resonant frequency of a parallel RLC resonator circuit is the frequency at which the imaginary part of its admittance vanishes. Shown in Figure 1-1 is the special case where the circuit is lossless and the resistance, \( R \), is infinity (an open circuit). With an inductance of 1nH and a capacitor of 1pF, the resonant frequency is 5.03 GHz. At resonance, the circuit behaves like a perfect open circuit. For a characteristic impedance \( Z_c \) equal to the port or the system impedance \( Z_0 \), in a lossless system \( R=\infty \), at the resonant frequency the reflection coefficient would be exactly at the center of the Smith chart shown in the figure below. This is means the input signal is perfectly matched to the load at the output. While at DC as well as infinite frequency, the ideal parallel LC presents a short circuit and exhibits total reflection of the input signal.
For a series RLC circuit, the resonance frequency is the one at which the imaginary part of its impedance vanishes. Shown on the smith chart of Figure 1-2 is the special case where the resistance of the circuit is zero, meaning the circuit is lossless. With values of 1nH and 1pF, the resonant frequency is around 5.03 GHz. Here, the circuit behaves like a perfect short circuit. Similarly, for \( Z_c \) equal to \( Z_0 \), and \( R=\infty \), at the resonant frequency the reflection coefficient would be at the center of the Smith chart. At zero frequency (DC) as well as infinite frequency, the ideal parallel LC presents an open circuit.

Figure 1-1. Shunt RLC resonator. The chart assumes \( R=\infty \).

Figure 1-2. Series LC resonator. The chart assumes \( R=\infty \).
At low frequencies a required filter response (band-pass, band-stop, low-pass or high-pass) meeting the network matching specifications can be synthesized and the desired performance can be achieved by a combination of resonant circuits connected in parallel and in series. Other design techniques are employed when the frequency is too high for the lumped circuit assumption to hold. At frequencies higher than a few gigahertz, lumped element circuit models comprising resistors, inductors and capacitors fail to describe the propagation and phase effects.

The circuit model of lumped element filters is a combination of series and parallel resonators. The most common configurations for these networks are represented by the Pi-model or the T-model shown in Figure 1-3. The filter depicted is a 5\textsuperscript{th} order symmetric lumped element band-pass filter. The synthesis of the elements values of a similar structure according to the desired frequency response specifications is explained in details in Chapter 2.

![Figure 1-3. 5\textsuperscript{th} order symmetric, lossless, lumped resonator filter in two configurations; (a) T-network, (b) Pi-network](image)
1.2 Ultra-Wideband Filters

Ultra-Wideband (UWB) is a technology for transmitting information spread over a large bandwidth that should, in theory and under the right circumstances, be able to share spectrum with other users. Regulatory settings of the FCC (Federal Communications Commission) are intended to provide an efficient use of scarce radio bandwidth while enabling high data rate wireless connectivity and longer-range, low data rate applications as well as radar and imaging systems.

Ultra Wideband is defined in terms of a transmission from an antenna for which the emitted signal bandwidth exceeds the lesser of 500 MHz or 20% of the center frequency. UWB communications transmit in a way that doesn't interfere largely with other more traditional 'narrow band' and continuous carrier wave uses in the same frequency band. However the rise of noise level by a number of UWB transmitters puts a burden on existing communications services.

In an UWB wireless communication system, the UWB filter plays an important role in ensuring the radiation of the system meets the system specifications. A lot of engineers have been engaged in developing the components in UWB wireless system to use the unlicensed band of 3.1~10.6 GHz. Different fashions of UWB filters have been designed and published in the literature [1-7] such as microstrip open-stub-loaded ring filters, the combination of microstrip and coplanar waveguides structures, multi-stage UWB filters using the short- /open-circuited stubs to create the UWB filter and, novel UWB filters with multi-mode resonators (MMR) of one wavelength based on EM simulation were successfully developed [1-7].
Shown below are the circuit representation and the response of an UWB band-pass 6\textsuperscript{th} order filter from 3.2 to 9.8 GHz, with a 20dB in-band return loss. A possible way of synthesizing this filter is to use the parameter extraction method of ladder network using synthesized S-parameters polynomials of the network [8]. However, numerical errors are a problem for filters of high orders. An optimization technique was used to extract the exact values of the filter elements.

![Ladder Network of the UWB filter.](image)

Figure 1-4. Ladder Network of the UWB filter.

![Frequency response of the UWB filter.](image)

Figure 1-5. Frequency response of the UWB filter.
This filter can be realized in a compact microstrip, co-planar waveguide or slotline fashion on a single layer or multi-layer board.

1.3 Microwave Waveguide Filters

Closed waveguide filters are of more importance in higher frequency applications, typically higher than 10 GHz, where radiation and losses significantly affect the performance of the filter. Microwave waveguide filters come in many design approaches as discussed below.

Waveguide Stub Filters

Simply composed of open-circuited (or short-circuited) half-wavelength stub sections, $\lambda_0/2$, separated by quarter-wavelength waveguide sections, $\lambda_0/4$, at the center frequency of the pass-band. Each of the quarter wavelength transformers represents an impedance inverter while all the stubs are resonators of the same length and the same resonant frequency, $f_0$. An impedance inverter is a network component that inverts the input impedance seen of the network multiplied by a scaling factor which is the square of the characteristic impedance of the inverter itself. Impedance inverters are essential elements in all filtering and matching networks. The realization of a waveguide stub filter is one of the simplest for band-pass filters. A microstrip realization of this filter is shown in Figure 1-6.
Corrugated Waveguide Filter

Corrugated Waveguide structures typically come in rectangular waveguide configuration. The corrugations are of electrical lengths that correspond to the defined poles of the filter and the pass-band edges. A physical structure of it is depicted in Figure 1-7. They are used as low-pass filters in numerous antenna feed systems. They are most often used for the rejection of spurious harmonics from transmission. They can also be designed as band-pass filters with wide or narrow pass-band response [9-10]. Several different design approaches have been described in the literature. Historically, the first design of a corrugated waveguide filter is based on the image parameter method which proved some drawbacks in terms of design efficiency and is almost no longer used in filter synthesis and design [11-12], thus it yields a less accurate response, while an advanced synthesis method is due to Levy [13-14].
Waveguide Iris-Coupled Filters

Waveguide iris coupled filters are the most commonly used types of waveguide filters. The fact that they have a straightforward design procedure makes them preferred by filter designers. In terms of performance attributes, they have a high power handling, and a close to ideal frequency response as well as an adequate quality factor. Their fabrication is rather simple and cost efficient. Recently there has been a lot of progress in the design techniques of iris-coupled filters that no longer come only in the typical rectangular waveguide configuration but also apply to cylindrical and elliptical waveguides configurations [15-16].

The simplest forms of filters using waveguide resonators and iris opening as inter-resonator coupling structures are the E-plane and H-plane filters. The simplest form of E-plane and H-plane filters comes when the synthesized frequency response can be achieved with an in-line resonator setting. Another form of coupled resonator filters that provides size reduction but has a somewhat more complex design is when the waveguide cavities representing the resonators support more than one mode, known as dual-mode
cavities. These two modes are degenerate and they are only coupled together through the inter-resonator coupling structures.

An H-plane filter is a combination of multiple H-plane discontinuities in rectangular waveguide setting. The inter-cavity coupling is achieved using H-plane discontinuities as well as the input/output coupling, they affect the magnetic field and result an infinite number of evanescent modes around the discontinuity. An accurate field solution of these discontinuities can be found using the Mode Matching Technique (MMT) based on the approximation of unknown electric and magnetic field components by Fourier series. These discontinuities play the role of impedance or admittance inverters. The ABCD matrix of such an inverter is

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = \begin{bmatrix}
0 & K \\
1/K & 0
\end{bmatrix}
\]

Where K in this equation is the value with which the input impedance of the system before the inverter is scaled. The loading effect of these inverters, in the shape of an iris opening, is compensated for by reducing the length of the cavities. Figure 1-8 shows the structure of a second order H-plane filter whose equivalent circuit is shown in Figure 1-9.

Figure 1-8. Rectangular waveguide H-plane filter.
Figure 1-9. Equivalent circuit model of a second order H-plane filter.

E-plane filters are based on the same concept as H-plane filters with the exception that the field solution adds more field components at the edge of the step discontinuities. The solution provides the coefficients of an infinite number of evanescent modes at both edges of the step as well as one fundamental propagating mode, $TE_{10}$ with field components $E_y$, $H_x$ and $H_z$ assuming the $z$-direction is the direction of propagation.

**Dual-Mode Cavity Filters**

When the filter prototype network to be implemented has a simple filtering function such as Chebyshev or Butterworth functions with all transmission zeros at infinity and without any cross-resonator couplings, the low-pass prototype network is a simple ladder structure where H-plane or E-plane implementation is suitable. The realization of more complicated filtering functions with cross-coupled resonators and transmission zeros at finite frequencies is achieved through more complicated filter structures. An example is the use of dual-mode resonant cavities which are widely used in implementing the filtering function as well as significantly reducing the size of the structure. The design of this class of filters involves the concept of a coupling matrix as...
an equivalent circuit. It contains normalized values of the coupling coefficients between
the resonant cavities and the input/output ports as will be further explained in Chapter 2.

Dual-mode waveguide filters are widely used in high power applications because
of their low-loss performance and reduced size. In the dual-mode cavities two degenerate
modes exist at the same time, thus allowing the creation of two electrical cavities inside
only one physical cavity. The filter size and weight are consequently reduced. Often,
circular waveguides of high quality-factor cavities are used to realize waveguide dual-
mode filters. The same concept can be also implemented using different type of
rectangular cavities. The excitation of two modes in a single cavity is achieved by
introducing a perturbation in the cavity, inward screws in most of the cases, which
disturbs the electromagnetic field and creates two resonant modes of different resonant
frequencies. The larger in size the perturbation is, the higher the frequency shift between
the two modes is. Another way to generate two modes in one cavity is by using oversized
resonant cavities, for example to involve the $\text{TE}_{102}$ and $\text{TE}_{201}$ in the case of rectangular
waveguides.

For a very long time, the design techniques of dual-mode filters was based on trial
and error, relying on tuning screws as part of the design itself. Recently, straight forward
design techniques of dual-mode filters, in which tuning elements are not used as part of
the design, were introduced [39]. A simple example of a fourth order rectangular
waveguide dual-mode filter is shown in Figure 1-10. It is a cascade of two resonant dual-
mode cavities coupled through iris openings to each other and to the input ports. The
corresponding response of the ideal coupling matrix of this filter is shown in Figure 1-11.
The filter is a 4\textsuperscript{th} order dual-mode filter operating at 11 GHz with two transmission zeros. The filter has a bandwidth of 40 MHz and an in-band return loss of 30 dB.

Figure 1-10. Structure of a dual-mode 4\textsuperscript{th} order filter.

![Figure 1-10](image)

Figure 1-11. Ideal reflection and transmission of the coupling matrix of the dual-mode 4\textsuperscript{th} order filter.

![Figure 1-11](image)

In this work, the design of novel dual-mode filters is accomplished based on direct techniques using the actual physical resonant modes existing in the cavities. An
ideal response is synthesized along with a prototype circuit representation in the form of a coupling matrix, the physical structure of the filters is clearly matched to the synthesized circuit and no tuning is used in the initial design. In Chapter 2, a literature review of microwave filter synthesis and design is presented. The design of a novel structurally symmetric 4th order dual-mode filter is discussed in Chapter 3. The discussion includes the manufacturing process as well as practical measurements. A cylindrical waveguide filter design procedure with novel coupling from the sidewalls is presented in Chapter 4. This filter structure is suitable for compensation of frequency shifts due to temperature variation. Chapter 5 of this thesis presents general conclusions and recommendations for relevant future work.
Chapter 2

Review of microwave filter synthesis and design

2.1 Introduction

The design procedure of a microwave filter commences with determining the required filter response specifications such as frequency band, number and locations of transmission zeros, return loss and phase delay as well as a filtering function that fits those specifications. Then, a fundamental step in the design of a microwave filter is the synthesis of the prototype network and an appropriate electrical model represented in most of the cases by a coupling matrix. The synthesis of an electrical filter is accomplished by the determination of the lumped elements of a network which can produce the required frequency response. However, for microwave filters, which are distributed networks, the ultimate goal of the design is the determination of the physical dimensions of the device. Various synthesis techniques have been developed for direct-coupled resonators as well as the more general case of filters with couplings between non-adjacent resonators. The last step is the design procedure of the physical filter which is the implementation in one of the common waveguide structures such as striplines, coplanar waveguides and closed rectangular waveguides. Generally, a field-theory-based analysis can be an integral part of the filter design. Such analysis allows filter response to be predicted very accurately, both in the pass-band and stop-band regions [8]. Optimization is usually necessary if a filter is synthesized by trial and error procedure rather than by direct synthesis techniques. Provided that the analysis program can be
executed fast enough at multiple frequency points, optimization algorithms can be used to
drive a field solver to finalize an initial design.

2.2 Response specification

The first step of designing a microwave filter is the definition of the frequency
response specification and a filtering function that realizes this response. The
transmission coefficient, \( S_{21} \), and the reflection coefficient, \( S_{11} \), of this network identify
the filter response. Assuming they are rational functions where the numerators and the
denominators are polynomials whose orders are determined by the order of the filter and
positions of the transmission zeros as in (2.1) and (2.2).

\[
S_{21} = \frac{P_N(\omega)}{\varepsilon E_N(\omega)} \quad (2.1)
\]

\[
S_{11} = \frac{F_N(\omega)}{\varepsilon R E_N(\omega)} \quad (2.2)
\]

Here, \( \varepsilon \) and \( \varepsilon_R \) are normalization factors and \( P_N, F_N \) and \( E_N \) are the polynomials forming
the rational functions of \( S_{11} \) and \( S_{21} \). These polynomials are related to each other since
they must satisfy a number of constraints such as reciprocity and conservation of energy.
For a reciprocal, lossless, passive two-port network the following amplitude and phase
conditions must be valid for S-parameters.

\[
S_{12} = S_{21} \quad (2.3)
\]

\[
|S_{11}|^2 + |S_{21}|^2 = 1 \quad (2.4)
\]

\[
\phi_{21} = \frac{\phi_{11} + \phi_{22}}{2} \pm \frac{n\pi}{2}, \quad n = 1, 2, 3, ... \quad (2.5)
\]

The ratio between the transmission and reflection coefficients is called the filtering
function, \( C_N \), which is also another rational function of the same order as the filter. The
response usually corresponds to a low-pass prototype of normalized cutoff frequency that can be transformed to band-pass response using the well-known transformation of (2.8).

\[ C_N = \frac{F_N(\omega)}{P_N(\omega)} \quad (2.6) \]

\[ |S_{21}(\omega)|^2 = \frac{1}{1+C_N(\omega)C_N(-\omega)\varepsilon^2/\varepsilon_R^2} \quad (2.7) \]

\[ \omega_L = \frac{1}{B} \left( \frac{\omega_B}{\omega_0} - \frac{\omega_0}{\omega_B} \right) \quad (2.8) \]

In equation (2.8), \( \omega_L \) is the low-pass frequency variable, \( \omega_B \) is the bandpass frequency variable, \( \omega_0 \) is the center frequency of the band-pass filter and \( B \) is the fractional bandwidth of the filter.

The most commonly used response types are the Butterworth, the Chebyshev and the Generalized Chebyshev functions. The formulation of these filtering functions has been rigorously explained in the literature [17-19].

### 2.3 Synthesis

After determining a filtering function meeting the specifications of the filter, a prototype network that is able to realize this response is to be synthesized. This prototype is usually a loss-less two-port low-pass network that is composed of passive inductors and capacitors in different configurations. The network itself can be a simple ladder network of successive series and shunt resonators, or can have cross couplings between non-adjacent resonances. These cross-couplings can be symmetric or even asymmetric [20]. A simple Chebyshev or Butterworth filter can be realized by a network of shunt capacitors and series inductors. In the initial design the filter is primarily matched to the
input and output ports of unity port impedances, $R = 1\Omega$, and the values of the filter elements can be extracted from the well known tables found in [17 and 18].

For most of the networks operating at microwave frequencies the physical realization of parallel and series resonators simultaneously is very difficult. In this case the prototype circuit using impedance ($K$) or admittance ($J$) inverters, shown in Figure 2-1 is a more convenient representation. An inverter placed at the input of a network inverts the input impedance of that network to the value of its admittance and scales it by a scaling factor. Exploiting this transformation, a filter network of mixed parallel and series resonances can be transformed to a network of only either series or parallel resonance and inverters.

![Figure 2-1. Transformed network with impedance inverters.](image)

The values of the impedance inverters can be easily obtained from the synthesized prototype coefficients using the following equations

$$K_{01} = \sqrt{\frac{B}{2g_{j-1}g_j}} \quad (2.9)$$

$$K_{j-1,j} = \frac{B}{2\sqrt{g_1g_2}} \quad (2.10)$$
where $B$ is the fractional bandwidth of the propagation constant over the specified band of the filter, $g_i$ is the $i^{\text{th}}$ coefficient of the filtering function.

A practical problem with filter design using this technique is the approximation that the values of the inverters are constant over the whole frequency band of operation of the filter. This assumption is not valid with most waveguide structures and causes a slight difference between the optimum response and the electromagnetic simulation or the measurements of the fabricated filters. This difference is typically ignored in narrow-band filters while it cannot be neglected for wide-band and ultra-wideband filters [21].

### 2.3.1 The Coupling Matrix

For a more complicated filter response, such as a pseudo-elliptic filtering function with multiple transmission zeros, cross-coupled resonators (shown in Figure 2-2) are commonly used. The topology of the network must be able to generate the required number of transmission zeros. A coupling matrix of that topology needs to be synthesized.

The coupling matrix facilitates the design of the filter because it contains information that can be directly related to the actual physical characteristics of the filter structure. A mapping from the entries of the coupling matrix can be done to indicate the relation between attributes like the local resonant frequencies, coupling coefficients and matching to the input ports and the effect of a direct variation in the physical dimensions of the filters. This can be seen in a one-to-one way in the numbers of the coupling matrix [8]. The numerical values in a coupling matrix are normalized to a low-pass prototype. The de-normalized coupling coefficients, $K$-coefficients, are affected by the bandwidth
and the impedance level (slope parameters) and used in the design step as will be shown
further on.

Figure 2-2. General prototype circuit of a filter with cross-couplings.

The derivation of a coupling matrix of a two port network similar to the one
shown in Figure 2-2 is fairly simple. Starting from the current loop equations where $e$ is
the excitation, $R_s$ is the source impedance and $R_{\text{load}}$ is the termination impedance, the
impedance matrix $[z]$ can be expressed as

$$
[e] = [z][i] \quad (2.11)
$$

$$
[1,0,0,\ldots,0]^t = [jM + sI + R].[i_1, i_2, \ldots, i_N]^t = j[A][i_1, i_2, \ldots, i_N]^t \quad (2.12)
$$

In the previous equation, $M$ is the initial coupling matrix containing all the
coupling coefficients between the internal nodes of the network, $s$ is the frequency
variable, $I$ is the identity matrix so that the frequency effect is only seen in the diagonal of
the impedance matrix and $R$ is an all zeros matrix except for the first and the last items of
the diagonal which contain the values of the source and the load impedance. All of the 3
matrices are of the size $(N \times N)$ [8].
Figure 2-3. Two port network with an impedance matrix of equation (2.12).

In order to generate an \((N+2) \times (N+2)\) coupling matrix, we introduce separate nodes at the source and load and normalize the resistors to unity. This is done by including two inverters at the input and the output sides of the network. The result of this normalization step is an \((N+2) \times (N+2)\) matrix where the additional two rows and columns contain the values \(\sqrt{R_s}\) and \(\sqrt{R_{\text{load}}}\) as the values of the items \(M_{s1}\) and \(M_{NL}\), respectively.

The expression of the matrix before the normalization is in the following form:

\[
\begin{align*}
&|S| & |1| & |2| & \ldots & |N| & |L| \\
&S & G_s + jB_s & jM_{s1} & jM_{s2} & \ldots & jM_{sN} & jM_{sL} \\
&1 & jM_{1S} & sC_1 + jB_1 & jM_{12} & \ldots & jM_{1N} & jM_{1L} \\
&2 & jM_{2S} & jM_{12} & \ldots & \ldots & jM_{N2} & jM_{L2} \\
&\ldots & \ldots & \ldots & \ldots & \ldots & \ldots & \ldots \\
&N & jM_{SN} & jM_{1N} & jM_{2N} & \ldots & sC_4 + jB_4 & jM_{LN} \\
&L & jM_{SL} & jM_{1L} & jM_{2L} & \ldots & jM_{NL} & G_L + jB_L
\end{align*}
\]
The multiplication of each row and line with the square root of the corresponding frequency dependent component yields the required normalization and the following scaling of the non-diagonal elements

\[
M_{ij} \rightarrow \frac{M_{ij}}{\sqrt{C_i C_j}} \quad (2.13)
\]

\[
M_{S1} \rightarrow \frac{M_{S1}}{\sqrt{G_S C_1}} \quad (2.14)
\]

\[
M_{NL} \rightarrow \frac{M_{NL}}{\sqrt{G_L C_N}} \quad (2.15)
\]

A coupling matrix can be synthesized directly as explained in [8], where the complete folded N+2 matrix is synthesized starting from the given polynomials of the S-parameters. Another way of synthesizing the coupling matrix is by numerical optimization of an error function as explained in [22]. Throughout this work, all coupling matrices are synthesized by optimization.

2.3.2 Synthesis of the Coupling Matrix by Optimization

A practical and simple approach of synthesizing a coupling matrix uses the elements of the matrix as independent variables as was first introduced in [23]. Using this approach, an error function is defined to optimize these elements. In [22], a gradient-based optimization was suggested as an addition to this technique and excellent results were obtained. The first step of this technique is the calculation of the low-pass prototype transmission function of the filter using a recursion relation and then an optimization technique is used over the elements of a coupling matrix that enforces the required filter
topology. Filters with more stringent requirements and arbitrarily placed transmission zeros can be synthesized using this method without the need for long analytical approaches.

The first step to obtaining the transmission function is to define the filtering function as a rational function given by

\[
F_N(\omega) = \frac{p_N(\omega)}{d_N(\omega)} = \frac{p_N(\omega)}{\prod_{n=1}^{N} (1 - \frac{\omega}{\omega_n})} = \cosh\left(\sum_{n=1}^{N} \cosh^{-1}\left(\frac{\omega - 1/\omega_n}{1 - \omega/\omega_n}\right)\right) \quad (2.16)
\]

While the transmission function is given by

\[
|S_{21}(\omega)|^2 = \frac{1}{1 + \epsilon^2 F_N^2(\omega)} \quad (2.17)
\]

Where \(\omega_n\) is the location of the transmission \(n^{th}\) transmission zero and \(\epsilon\) is a normalization factor. Note that the band edges of the response are at unity angular frequency. In order to obtain the numerator of the filtering function \(P_N\) a simple recursion relation is derived using the identities of the hyperbolic functions and they result in the following relation [22]

\[
P_{N+1}(\omega) = -P_{N-1}(\omega) \left(1 - \frac{\omega}{\omega_N}\right)^2 \left(1 - \frac{1}{\omega_{N+1}^2}\right)^{1/2} \left(1 - \frac{1}{\omega_N^2}\right)^{1/2} \]

\[+ P_N(\omega) \left[\omega - \frac{1}{\omega_{N+1}} + \left(\omega - \frac{1}{\omega_N}\right) \cdot \left(1 - \frac{1}{\omega_{N+1}^2}\right)^{1/2} \left(1 - \frac{1}{\omega_N^2}\right)^{1/2}\right] \quad (2.18)
\]

The initial point of calculation for the polynomials \(P_0\) and \(P_1\) are calculated by

\[
P_0(\omega) = 1 \quad (2.19)
\]
Knowing the response of the filter and the corresponding polynomials of the transmission and reflection functions, the optimization technique can be applied to find the coupling matrix that enforces the filter topology. The transmission and reflection function can be expressed in terms of coupling matrix elements as

$$S_{21}(\omega) = 2\sqrt{R_{11}R_{NN}}i_N = -2j\sqrt{R_{11}R_{NN}[A^{-1}]}_{N1} \quad (2.21)$$

$$S_{11}(\omega) = 1 - 2R_{11}i_1 = 1 + 2jR_{11}[A^{-1}]_{11} \quad (2.22)$$

Where

$$-j[A] = [M + \omega I - R] \quad (2.23)$$

The optimization is done over a cost function that contains special emphasis on the location of poles and zeros of the transmission function. The function in [23] only considered the location of poles and zeros of the filtering function while the adopted function introduced in [22] takes the band edges at $\omega=\pm 1$ into consideration. The fact that the gradient of the cost function is used in the optimization, gives better results and yields less numerical error. This cost function is expressed as

$$K = \sum_{i=1}^{N} |S_{21}(\omega_{zi})|^2 + \sum_{i=1}^{P} |S_{11}(\omega_{pi})|^2 + (|S_{11}(\omega = -1)| - \frac{\epsilon}{\sqrt{1+\epsilon^2}})^2 + (|S_{11}(\omega = +1)| - \frac{\epsilon}{\sqrt{1+\epsilon^2}})^2 \quad (2.24)$$

Here, $\omega_{zi}$ is the location of the $i^{th}$ zero of the filtering function and $\omega_{pi}$ is the location of the $i^{th}$ pole of the filtering function.
The evaluation of a general term gradient of the transmission and reflection functions is expressed in details in [22]. While a simplification of the general term expression to find the gradient with respect to the elements of the coupling matrix using the symmetry of the matrix $A$ is given by

$$\frac{\partial s_{11}}{\partial M_{pq}} = -4jR_1 P_{pq}[A^{-1}]_{p1}[A^{-1}]_{q1}$$  \hspace{1cm} (2.25)

$$\frac{\partial s_{21}}{\partial M_{pq}} = 2j\sqrt{R_1 R_2}P_{pp}([A^{-1}]_{Np}[A^{-1}]_{q1} + [A^{-1}]_{Nq}[A^{-1}]_{p1})$$  \hspace{1cm} (2.26)

$$\frac{\partial s_{11}}{\partial M_{pp}} = -2jR_1 P_{pp}[A^{-1}]_{p1}[A^{-1}]_{p1}$$  \hspace{1cm} (2.27)

$$\frac{\partial s_{21}}{\partial M_{pp}} = 2j\sqrt{R_1 R_2}P_{pp}[A^{-1}]_{Np}[A^{-1}]_{p1}$$  \hspace{1cm} (2.28)

The benefit of using this synthesis method is that it applies directly to the topology used in the design procedure without the need for similarity transformation [24]. This applies for symmetric and asymmetric topologies as well as even and odd order filters. For example, for a 4\textsuperscript{th} order filter of the topology given in Figure 2-4, the starting values of the optimization process can be the values of the coupling coefficients of an inline 4\textsuperscript{th} order Chebyshev filter synthesized analytically and the initial values of the cross-couplings can be set to zero. The processing time of the optimization is minimal and similar filters of higher orders can be synthesized the same way.
2.4 Filter Design and Fabrication

The design step is the translation of the synthesized coupling matrix into actual component parameters such as the length of a resonator, the width of an iris opening or the depth of a perturbation to a resonant structure to provide a dual-mode behavior. For lower frequencies, where radiation loss is relatively low, open waveguide structures are more commonly used such as micro-strip lines, coplanar waveguides and slot lines. Some of the most common microstrip line filter structures are shown below in Figure 2-5 [32].
Figure 2-5. Different Microstrip filter configurations.

Most of the open waveguide structures operating at relatively low frequencies exhibit only single mode resonances. Rectangular and circular closed waveguides are widely used in dual-mode filter designs [25-31]. Dual-mode structures are of smaller size and facilitate cross-coupling between resonators essential for the generation of transmission zeros. Figure 2-6 depicts various cross-section shapes of possible realizations of dual-mode cavities. Some of the structures providing two degenerate
modes such as the elliptical cavity are widely used in dual-mode filter design. The size and shape of these perturbations determine the resonant frequencies of the two modes.

Figure 2-6. Dual-mode cavities; (a) rectangular waveguide with inward perturbations, (b) Rectangular waveguide with outward perturbations, (c) Circular waveguide with outward perturbations, (d) Circular waveguide with inward perturbations.

It should be mentioned that the design and fabrication steps of microwave filters comprises the use of EM simulators. A good understanding of the adopted solving method such as the Method of Moments (MoM) technique, the Mode Matching Technique (MMT), the Finite Element Method (FEM) and the Finite Difference Time Domain (FDTD) method is essential for filter designers [33-37].
2.5 Conclusion

In this chapter the synthesis process of transmission and reflection coefficients for the most common filtering functions such as the Chebyshev, the Butterworth and the Elliptical functions has been reviewed. The synthesis of a circuit representation capable of reproducing the desired frequency response in the form of a coupling matrix has been reviewed as well. Methods for coupling matrix synthesis either analytically or by optimization have been discussed. At the end of the chapter, different physical realizations of microwave filters in both closed and open waveguide configurations have been discussed, including some dual-mode structures. This straightforward design sequence is the foundation of the design of the novel dual-mode filters introduced in Chapters 3 and 4 of this work.
Chapter 3
A Novel Structurally Symmetric Dual-Mode Waveguide Filter

3.1 Introduction

A direct synthesis and design technique of microwave filters in general, and of dual-mode filters in particular, is much preferred to an optimization-based approach. An initial design that exhibits a close to ideal response along with the avoidance of mechanical tuning using inward screws saves a lot of time and effort.

In satellite transponders, aspects like compact size, light weight and the simplicity of the manufacturing process of dual-mode waveguide filters are very important features. For metallic waveguide cavities, issues like the ease of drilling and the number of pieces that constitute the whole filter influence significantly the resulting measured response. The fabrication process is further simplified by using symmetric structures. Unfortunately, even dual-mode filters implementing pseudo-elliptic responses with symmetrically placed transmission zeros are implemented by waveguide structures that are not structurally symmetric.

In this chapter, a novel fully symmetric rectangular waveguide filter is introduced. The symmetry is achieved due to the utilization of novel mixed coupling irises. A direct design technique for this filter is explained along with EM simulation and measurement. The same approach can be used for different shapes of dual-mode cavities.

The theory of design of dual-mode resonators has been widely investigated in the literature [25, 26, 38]. Typically, the two modes are coupled by the insertion of
perturbations that disturb the electromagnetic field inside a resonant cavity that supports two degenerate modes. The dual-mode behavior can be seen as a split of the initial resonance frequency of the fundamental mode in the cavity, for example \( \text{TE}_{101} \) in rectangular waveguide, into a higher-frequency resonant mode and another lower frequency mode. The existence of two degenerate modes in one cavity allows the design of filters with different cross-coupling topologies efficiently from a size, weight and cost points of view. When the filter is fully symmetric, the manufacturing process reduces to fabricating two duplicates of the same half thereby reducing milling time, especially when the irises are milled with parts of the resonators attached to them.

The adopted design theory uses the physical modes inside the resonant cavities as basis functions. The coupling matrix based on the vertical and horizontal modes of the cavities is used to determine the dimensions of the coupling irises [39]. A matrix rotation is performed to obtain the transversal coupling matrix from which the resonant frequencies are extracted and used for the design of the cavities [39]. The introduced filter can be implemented by using other types of dual-mode cavities, although the design steps remain the same.

### 3.2 Theory and Design

The first design theory of dual-mode filters was introduced by Atia and Williams [25, 26]. Another different approach was developed by Cameron [40-42]. The synthesis of the coupling matrix of a band-pass filter of order \( N \) starts by defining the response specifications such as the in-band return loss, the filter bandwidth, the center frequency and the number and locations of transmission zeros. The obtained \( N+2 \) by \( N+2 \) coupling matrix is a circuit representation based on a set of basis function which are the resonances
inside the physical structure realizing this filter. The synthesis of the coupling matrix enforcing a certain topology can be attained numerically by optimization as explained in the previous chapter. The conversion from one basis function or representation can be achieved by applying similarity transformations [24].

The most suitable matrix to use in the design of dual-mode rectangular waveguide resonator filters is the sparse matrix based on the horizontal and vertical resonances of the cavity. These resonances are considered the initial degenerate modes existing in the cavity before the insertion of the perturbation. The reliance on this matrix in the design is due to the fact that this matrix has a small number of non-zero elements. When one or more transmission zeros is required at a specific location of the stop-band, the inter-cavity coupling coefficients will have opposite signs, this implies the inversion of the polarity of the resonant modes in the cavities in tandem. For example, in the coupling matrix (3.1), the matrix represents a fourth order filter and as will be further demonstrated the coupling coefficient $M_{14}$ has an opposite sign to $M_{23}$ and $M_{12}$.

$$M_{(N+2,N+2)} = \begin{bmatrix}
0 & M_{01} & 0 & 0 & 0 & 0 \\
M_{01} & 0 & M_{12} & 0 & M_{14} & 0 \\
0 & M_{12} & 0 & M_{23} & 0 & 0 \\
0 & 0 & M_{23} & 0 & M_{12} & 0 \\
0 & M_{14} & 0 & M_{12} & 0 & M_{01} \\
0 & 0 & 0 & 0 & M_{01} & 0
\end{bmatrix}$$

(3.1)

The need for the inversion of the modes in the subsequent cavities makes the symmetry unfeasible. A possible solution is to use a coupling iris that has mixed capacitive and inductive couplings thereby providing positive and negative couplings at the same time. The iris is capacitive for one polarization and negative for the other. The use of the
conventional cross iris, Figure 3-1, is not helpful anymore. It provides inductive coupling for both polarizations, i.e., there is no negative coupling through this iris.

![Figure 3-1. Conventional cross coupling iris for dual-mode cavities.](image)

3.2.1 Resonant Cavities

A coupling matrix similar to the one in (3.1) can be realized by \( N/2 \) dual-mode inline cavities, coupled together using irises. A direct design technique of such a filter is explained in detail in [39]. The design steps start by the design of the unloaded dual-mode cavities at the frequencies corresponding to the eigen values obtained in the transversal coupling matrix. An eigen-resonance simulation is accomplished with perturbations of random dimensions as a start point. The depth of the perturbation is varied once and the length of the cavity is varied another time and two more EM simulations are performed to calculate the variation in the local resonances with each of the dimensions. The required cavity size of perturbation and length is calculated using a simple two-by-two Jacobian [43].

The following step is the design of the irises and their loading of the resonators. Finally, the input and output irises are designed and the loading effect of the resonances of the input and output cavities taken into account.
In most of the cases, the realization of a negative coupling coefficient is achieved by rotating the coupling elements in one of the cavities by 90°. This has been viewed as a virtual negative coupling effect of the coupling iris between the cavities [8]. However, the iris provides inductive coupling for both polarizations. The negative coupling is due to the rotation in the placement of the coupling elements inside the cavities as mentioned earlier.

To show the validity of this explanation, a 4th order dual-mode rectangular waveguide filter with different perturbations’ orientation in different cavities (shown in Figure 3-2a) is designed. The first cavity of the filter has inward perturbations while the other cavity has outward perturbations and both cavities have the same alignment with no rotation needed. The coupling between the two cavities is a typical iris with cross opening.

The design steps of this asymmetric filter remain the same as mentioned before, and the initial design response shows very good agreement with the response of the ideal response of the synthesized coupling matrix as shown in Figure 3-3. The simulation was done using CST software package. These results show that when perturbations of different natures are used as “coupling” elements, we no longer need to rotate them. The inward and outward perturbations provide electric and magnetic coupling, respectively. However, the resulting filter is still not structurally symmetric.
Figure 3-2. 4th order filter structure with cavities of different perturbations, a) Complete structure, b) Input and inter-cavity irises.
3.2.2 A Novel Mixed Coupling Iris

With reference to the previous discussion, the design of an iris which provide both negative and positive coupling can lead to a fully symmetric filter without the need of rotating the “coupling” elements in the cavities to provide the negative coupling. The iris remains nothing more than an impedance inverter but with the added difference that it provides a $90^\circ$ phase shift to one polarization and a $-90^\circ$ phase shift to the other. Regarding the amplitude of the coupling coefficients provided by this iris, they remain
exactly the same and can be calculated using (3.2)

\[ K_{ij} = \frac{1-|S_{11}|}{\sqrt{1+|S_{11}|}} \quad (3.2) \]

where the de-normalized coupling coefficient \( K_{ij} \) is

\[ K_{ij} = \frac{\pi}{2} B_\lambda M_{ij} \quad (3.3) \]

Here, \( B_\lambda \) is the fractional bandwidth in terms of wavelength. The loading of the cavities due to this iris calculated using (3.4) for the negatively loaded mode and (3.5) for the positively loaded mode.

\[ \Phi_{ij} = \Phi_{21} - \frac{\pi}{2} \quad (3.4) \]

\[ \Phi_{ij} = \Phi_{21} + \frac{\pi}{2} \quad (3.5) \]

where \( \Phi_{21} \) is the phase of the \( S_{21} \) of the mode across the iris. An independent EM simulation of the iris only is performed to be able to calculate the transmission and reflection coefficient and find the correct opening dimension that leads to the desired coupling values.
The iris provides coupling for both modes inside the cavity. Given that the vertical and horizontal polarizations are used in the design of the iris, the coupling of each mode is implemented predominantly through different openings. The negative and positive coupling coefficients implemented by the iris can be controlled practically independently.

An example structure of this iris in rectangular waveguide configuration is shown in Figure 3-4. The dimensions $h$, $h_1$ and $v$ provide almost independent coupling control over two modes; one is horizontally polarized and the other is vertically polarized which allows a straight-forward design of the synthesized values in the coupling matrix of (3.1). It also provides a 90° phase shift for the horizontal mode and a -90° for the vertical mode.
as shown in Figure 3-5.

![Figure 3-5. Phase responses of the iris at 11 GHz for horizontal and vertical excitation.](image)

3.2.3 Input and Output Irises

The input and output irises play the role of impedance inverters that match the load and the source to the input impedance of the filter. The difference with the input and output irises from the inter-cavity irises is that they represent an asymmetric discontinuity, the modes on one side of the iris can be taken as the horizontally and vertically polarized modes or the diagonally polarized modes, whereas the input rectangular waveguide on the left supports only a TE_{10} propagating mode within the frequency band of interest. The modeling of such an iris was shown in [39]. The input
excitation is coupled to the filter through the input iris and the amount of this coupling is controlled only by the horizontal opening $a_1$ of the iris. The value of this coupling is obtained from the coupling matrix normalized values as

$$K_{01} = \frac{\pi B}{2M_{01}}$$ \hspace{1cm} (3.6)

While the vertical dimension is used as a variable loading reactance to compensate for loading difference on both modes inside the cavity and achieve phase balance for both modes.

An independent EM simulation of the iris only is used to evaluate the amount of coupling any opening with a given dimension provides, the resulting S-parameters of this simulation help determining the required dimension that provides the correct value of coupling. The equation that relates the coupling coefficient of the input iris to the S-parameters are presented in details in Chapter 4 of this work.

### 3.3 Fourth-Order Filter Example

In this section, a fourth-order dual-mode symmetric filter is designed. The filter consists of two symmetric dual-mode rectangular waveguide cavities coupled with an inverting iris as in Figure 3-4. Figure 3-6 shows the geometry of the filter. The filter has a center frequency of 11 GHz and a bandwidth of 200 MHz, the in-band return loss of the filter is 20dB and the filter has two transmission zeros on both sides of the pass-band at normalized frequencies of $\pm 1.6$. The corresponding coupling matrix of this response is
A similarity transformation can be used to obtain the transversal coupling matrix that is based on the physical diagonally polarized modes as

\[
M' = \begin{bmatrix}
0 & 0.719 & 0.719 & 0 & 0 & 0 & 0 \\
0.719 & -0.8306 & 0 & 0.259 & -0.555 & 0 & 0 \\
0.719 & 0 & 0.8306 & -0.555 & 0.259 & 0 & 0 \\
0 & 0.259 & -0.555 & 0.8306 & 0 & 0.719 & 0 \\
0 & -0.555 & 0.259 & 0 & -0.8306 & 0.719 & 0 \\
0 & 0 & 0 & 0.719 & 0.719 & 0 & 0
\end{bmatrix}
\] (3.8)

The design steps are similar to those of a filter with an inductive cross iris. The first step is to design the unloaded cavity with the resonances corresponding to the diagonal values of the transversal matrix; this is accomplished using an eigenmode simulation in CST. This step results in a square cavity whose width \(a_0\) is 17.220mm and a perturbation of depth \(d=2.0143\)mm. The length of the perturbation is \(t_{\text{step}}=3\) mm.
Figure 3-6. Geometry of the 4\textsuperscript{th} order rectangular waveguide filter. All the following dimensions are in mm; $L_x=8.398$, $L_y=8.733$, $a_1=9.904$, $b_1=13.455$, $t=5.000$, $t_{in}=1.100$, $t_{step}=3.000$, $h=7.698$, $h_1=0.5281$, $v=4.200$.

The second step is the iris design. The coupling values it should provide are $K_{14}$ of the vertical mode and $K_{23}$ of the horizontal mode corresponding to $M_{14}$ and $M_{23}$ in (3.7) and related to each other by (3.3). Subsequently, the loading of the iris on each of the modes is calculated according to (3.4 and 3.5). The loading of the horizontal mode is $+4.45^\circ$ while that of the vertical mode is $-10.28^\circ$. The negative loading is compensated for by shortening the length of the cavity, while the horizontal loading of the iris is accounted for by increasing the height of the input iris, $b_1$, to maintain the phase balance of both modes inside the cavity.
The final step is the design of the input iris, it starts with a square iris of width, $a_I$, that gives the required input coupling $K_{01}$, from the input WR-90 rectangular waveguide. The height, $b_I$, affects only the horizontal mode and it should be increased to add a negative phase to it and satisfy the phase balance equation of both modes inside the cavity

$$\Phi_{iv} + \Phi_{14v} = \Phi_{ih} + \Phi_{23h} \quad (3.9)$$

where $\Phi_{iv}$ is the loading of the input iris on the vertical mode caused by the opening $a_I$, $\Phi_{14v}$ is the loading of the inverting iris on the vertical mode, $\Phi_{ih}$ is the loading of the input iris on the horizontal mode caused by the opening $b_I$ which is required to be calculated and $\Phi_{23h}$ is the loading of the inverting iris on the horizontal mode.

The filter was simulated using CST design package as well as HFSS, the finite element based simulator. Both results show a very good agreement. The initial design response is very close to the ideal one as shown in Figure 3-7. The ideal response S-parameters of the coupling matrix in (3.7 and 3.8) are obtained using the transformation equations in section 2.2 that give the S-parameters vectors of frequency starting from the synthesized coupling matrix.
Figure 3-7. Amplitude response of the 4\textsuperscript{th} order filter. Dotted: ideal coupling matrix, solid: CST simulation, dashed: HFSS simulation

The design example in this chapter was fabricated using CNC milling machines at the machining laboratory of the physics department at Queen’s university. With the available fabrication process, it is extremely difficult to fabricate sharp edges at 90° angles. The milling radii of the cutters were included in the EM simulation. In addition, the dimension of the opening providing the capacitive coupling in the middle iris is too small to mill, increasing the width of this opening increases the capacitive coupling of the horizontal resonance. In order to compensate for this increase, a second inductive opening for the same mode is added in the middle of the iris noting that this addition does
not affect the coupling of the other mode. Figure 3-8 shows the new structure of the fabricated filter. Figure 3-9 shows the fabricated filter.

Figure 3-8. 3-dimensional view of the fabricated structure complying with the possible manufacturing process taken from CST simulator.

Figure 3-9. Fabricated symmetric 4\textsuperscript{th} order filter.
The measurement results show a close to perfect response compared to that of the simulated response in CST as shown in Figure 3-10. The noticeable asymmetry in the response is due higher order modes emission and the variable value of the impedance inverters provided by the irises versus frequency. However, the level of asymmetry is still acceptable.

Figure 3-10. Measured S-parameters of the fabricated 4\textsuperscript{th} order dual-mode symmetric filter. Dashed: EM simulation, solid: measurement.
3.4 Conclusion

A novel iris coupling structure is investigated allowing negative coupling coefficients required for certain polarizations of modes in dual-mode rectangular cavities. A 4\textsuperscript{th} order dual-mode band-pass filter is designed at 11 GHz with a bandwidth of 200 MHz, the in-band return loss is 20dB and the filter has two transmission zeros at normalized frequencies of ±1.6. The filter makes use of the negative coupling property of the iris for the horizontal resonance. The well-known design procedure of inline dual-mode filters based on the coupling matrix representation and its transformed version has been followed. The desired performance of the filter was achieved by the initial design without any EM optimization iteration in the design step by the used simulators. The presence of the negative coupling iris allows perfect symmetry of the structure with two completely identical dual-mode cavities. To the author’s knowledge, this is the first structurally symmetric design of this type of filters.
Chapter 4

A Direct Design Technique for Side-Coupled Dual-Mode Circular Waveguide Filters

4.1 Introduction

Circular waveguide dual-mode filters exhibit a similar response to single-mode filters but can be much smaller and lighter. Their accurate design continues to attract considerable attention [30, 44]. This chapter presents a step-by-step design procedure for a novel structure of cross-coupled dual-mode circular waveguide filters. The filters consist of side-coupled circular dual-mode cavities which is a convenient realization when temperature compensation is of concern [45, 49].

Temperature stability is an important feature in channel combiners and dividers since the adjacent bands are very close to each other. Thus, the design of filters suitable for temperature compensation is greatly advantageous. In [46] the temperature characteristics of resonant cavities have been studied and a temperature compensation method was proposed, in which a dielectric sphere is used for field perturbation along with the analytical results to determine the maximum and quasi-linear regions of operation in the cavity. In [47], the characteristics of the frequency drift caused by the temperature in the coaxial and waveguide resonators, which were used to construct filters, were investigated. In [48], it is proposed that a variable length bimetal tuning screw can be used to compensate for temperature drift in waveguide resonators and filters constructed from high Coefficient of Thermal Expansion (CTE) materials such as
aluminum. Methods for designing and tuning a bimetal tuning screw and for applying this approach to filter compensation are proposed.

As mentioned previously in this work, any filter response of order $N$ is defined by a filtering function that is translated into a circuit representation in the form of a $(N+2)$ by $(N+2)$ coupling matrix. Any filter can be represented by an infinite number of coupling matrices; however, the transversal coupling matrix based on the actual resonances inside the filter cavities, as basis functions, remains the only truly physical representation of the filter. The circuit representation of the transversal coupling matrix based on the physical modes along with the commonly used representation based on the vertical and horizontal resonances inside the cavities are shown in Figure 4-1. The advantage of the vertical-horizontal representation is that it yields a sparse matrix which facilitates the design of some internal parts of the filter such as the inter-cavity coupling irises and the input irises as will be shown later on.
Although the filter considered in this work is of order 4, the design technique used for it is generic and applicable for filters with higher orders and higher number of transmission zeros.

4.2 Dual-Mode Cavities

The considered mode operation for both cavities is that of the lowest order fundamental resonance in a circular cavity which is the TE_{101} resonant mode. Therefore, the radius of the cavity is fixed to a value such that this mode is propagating and all of the higher order modes are at cut-off. Throughout the rest of the design this value of the radius will not be varied. A perturbation is introduced to have two degenerate modes inside the cavity; this perturbation can be of the form of inward screws or circular
outward “ears” in the middle of the cavity or along of its length, however, the perturbations along the cavity makes it less sensitive to changes in dimensions. Figure 4-2 shows a cross-section of the cavity and the electric field distributions of the two resonant modes inside of it. It is to note that these modes do not couple to each other inside a closed cavity, yet each of them contributes to the coupling of both modes in the next cavity through the inter-cavity coupling iris. The shape of the inter-cavity iris forces them to couple to each other with a very small coupling value.

Figure 4-2. Dual-mode cavity cross-section with the electric field distributions for the two resonant modes. a) First resonant mode, b) Second resonant mode.

The values of the required resonances inside each cavity are obtained from the diagonal elements of the transversal coupling matrix of (4.1) synthesized according to the desired frequency response. These values have to be de-normalized using the well known low-pass to band-pass transformation of (4.2)[17].
The size of the perturbation and the length of the cavity are then varied and an Eigen-mode simulation is performed using EM solver to obtain the desired resonance values. The length of the cavity shifts both resonances while the size of the perturbation has much stronger effect on the first resonant mode over the second mode shown in Figure 4-2.

4.3 Inter-cavity Iris

The magnetic coupling between the two cavities is achieved using a single rectangular opening in the middle of the side-wall through the field component $H_\Phi$. A more complicated coupling configuration for similar structures has been proposed in [49] that is also suitable for higher order resonances with improved quality factor, however, a straightforward method of extracting the iris dimensions matching the required inverters’ values of the synthesized coupling matrix is missing. The mixed boundary shape between straight and curved surfaces makes a direct design of the iris more challenging. The numerical values of the coupling coefficients, $K_{14}$ and $K_{23}$, of the coupling matrix corresponding to the circuit in Figure 4-1.a are very small compared to the input/output
couplings; therefore, strong enough magnetic coupling can be achieved by an opening in the middle of the cavity.

Figure 4-3 illustrates the magnetic field distribution inside the resonant dual-mode cavity propagating along the length of the cavity used in this work. It is obvious that side coupling of the magnetic field is easily achievable through an iris opening in the middle of the cavity, while the magnetic field is very weak at the edges of the cavity. In general, the coupling should provide the required coupling coefficient and avoid the excitation of spurious cavity modes.

![Figure 4-3. Magnetic field distribution of both modes inside the circular cavity.](image)

The dimensions of the rectangular opening is verified by [10] to provide the coupling value of $K$ in (4.3)

$$K = \frac{4M\lambda^2 q^2 J_1(1.84)^2}{\pi L^3 D 1.84^2} \quad (4.3)$$
Where $L$ is the length of the cavity, $D$ is its diameter, $\lambda$ is the guided wavelength of the initial resonance, $J_1$ is the coefficient of the first Bessel function and $M$ is the corrected magnetic polarizability defined in [10]. The polarizability depends on the aperture dimensions and is multiplied by correction factors depending the shape and length of the iris opening.

In order to be able to compensate for the loading effect of the inter-cavity iris, an indirect technique is used. Two identical cavities coupled from the top side are designed with a conventional cross iris coupling between them, the loading of the cross iris exhibiting the same required coupling values is calculated and used as a starting point of optimization for the length of the cavity to compensate for the loading. This is the only part of the design that is not directly mapped to analytical rationalization.

### 4.4 Input Iris

The input iris is an asymmetric T-network with a single mode port of the rectangular waveguide feed from one side and a dual-mode port of the circular cavity from the other side. In order to determine the correct values of the input inverter value and the loading effect on both modes a top fed circular cavity with a rectangular iris designed.

An S-parameter simulation of a side fed cavity is performed and the dimensions of the input iris that yield the same amplitude and phase response as the top are the ones used in the design. Figure 4-4 shows the configuration of both input irises.
An important aspect of the design of the side input iris is that its width should not exceed a certain limit in order not to overlap with the perturbation in the cavity that takes the form of outward “ears”. Otherwise, the resonance frequencies of the cavity designed in the step explained previously will suffer a high shift which is difficult to compensate for. An alternative for this is to design the cavity with perturbations only in the middle in the form of inward screws but a distributed perturbation over the length of the cavity is preferred because it has less sensitivity to dimensions’ variations.

Figure 4-4. Input iris configurations for side-fed and top-fed cavities.
The analytical derivation was introduced in [43] and it starts with the Z matrix of the T network (without the phase shifts), given by

\[
Z = \begin{bmatrix} jX_{s1} + jX_p & jX_p \\ jX_p & jX_{s2} + jX_p \end{bmatrix}
\]  
(4.4)

The Z matrix components in equation (4.4) can be expressed in terms of the scattering parameters of the discontinuity using an S-to-Z transformation. Derived from the scattering parameters, the circuit parameters are given by

\[
jX_p = \frac{2S_{21}}{(1-S_{11})(1-S_{22})-S_{21}^2} \quad (4.5a)
\]

\[
jX_{s1} = \frac{(1+S_{11})(1-S_{22})+S_{21}^2-2S_{21}}{(1-S_{11})(1-S_{22})-S_{21}^2} \quad (4.5b)
\]

\[
jX_{s2} = \frac{(1-S_{11})(1+S_{22})+S_{21}^2-2S_{21}}{(1-S_{11})(1-S_{22})-S_{21}^2} \quad (4.5c)
\]

where \(S_{ij}\) are the scattering parameters of the discontinuity. The purpose of this analysis is to be able to find the values of the phase shifts \(\Phi_1\) and \(\Phi_2\) as well as the inverter value \(K_0\). The next step then is to express the equivalent scattering matrix \(S_x\) in terms of the the impedance matrix in equation (4.4). This conversion is given by

\[
S_{11x} = \frac{(jX_{s1}+jX_p-1)(jX_{s2}+jX_p+1)+X_p^2}{(jX_{s1}+jX_p+1)(jX_{s2}+jX_p+1)+X_p^2} \quad (4.6a)
\]
\[ S_{12x} = S_{21x} = 2jX_p \quad (4.6b) \]

\[ S_{22x} = \frac{(jX_{s1}+jX_p+1)(jX_{s2}+jX_p-1)+X_p^2}{(jX_{s1}+jX_p+1)(jX_{s2}+jX_p+1)+X_p^2} \quad (4.6c) \]

Shifting the reference plane by \( \Phi_1/2 \) and \( \Phi_2/2 \) to the left and right of the \( T \) network will add phase terms to the \( Sx \). It is required to find the values of \( \Phi_1 \) and \( \Phi_2 \) such that the overall phase shift of the inverter is \( \pi/2 \). The scattering matrix \( S_x' \) with shifted reference planes is given by

\[
S_x' = \begin{bmatrix}
S_{11x}e^{-j\phi_1} & S_{12x}e^{-j(\phi_1+\phi_2)} \\
S_{21x}e^{-j(\phi_1+\phi_2)} & S_{22x}e^{-j\phi_2}
\end{bmatrix} \quad (4.7)
\]

Equating the phase of the transmission coefficient to \( \pi/2 \), the following equation can be obtained

\[
\phi_1 + \phi_2 = -2\tan^{-1}\left(\frac{X_{s1}+X_{s2}+X_p}{1-X_{s1}X_{s2}-X_p(X_{s1}+X_{s2})}\right) \quad (4.8)
\]

Assuming that the port parameters are normalized to the mode impedances on both side of the discontinuity, the inverter \( K_{01} \) is given by

\[
K_{01}^2 = \frac{1+\Gamma_{in}}{1-\Gamma_{in}} = \frac{1+\Gamma_{out}}{1-\Gamma_{out}} \quad (4.9)
\]
Where $\Gamma_i$ and $\Gamma_o$ are the input and output reflection coefficients respectively. Substituting with the S parameters in equation (4.9) it can be shown that

\[
\frac{S_{11x}}{S_{22x}} = e^{j(\phi_1 - \phi_2)} = \frac{(1 + X_s x_s + X_p(x_s + x_s^2)) - j(x_s - x_s^2)}{(1 + X_s x_s + X_p(x_s + x_s^2)) + j(x_s - x_s^2)}
\]  

(4.10)

Therefore

\[
\phi_1 - \phi_2 = -2\tan^{-1}\left(\frac{x_s - x_s^2}{1 + X_s x_s + X_p(x_s + x_s^2)}\right)
\]  

(4.11)

From equations (4.8) and (4.11), the $\phi_1$ and $\phi_2$ are given by

\[
\phi_1 = -\tan^{-1}\left(\frac{x_s + x_s + 2x_p}{1 - x_s x_s - X_p(x_s + x_s^2)}\right) - \tan^{-1}\left(\frac{x_s - x_s}{1 + x_s x_s + X_p(x_s + x_s^2)}\right)
\]  

(4.12)

\[
\phi_2 = -\tan^{-1}\left(\frac{x_s + x_s + 2x_p}{1 - x_s x_s - X_p(x_s + x_s^2)}\right) + \tan^{-1}\left(\frac{x_s - x_s}{1 + x_s x_s + X_p(x_s + x_s^2)}\right)
\]  

(4.13)

Also $K_{0_1}$ is given by

\[
K_{0_1} = \frac{\sqrt{1 + S_{11x} e^{-j\phi_1}}}{\sqrt{1 - S_{11x} e^{-j\phi_1}}} = \frac{\sqrt{1 + S_{22x} e^{-j\phi_2}}}{\sqrt{1 - S_{22x} e^{-j\phi_2}}}
\]  

(4.14)

This derivation has been formulated in [43] and is used for the design of the input iris to many waveguide cavities of different shapes. Although it might seem a lengthy procedure to compare and calculate the input inverter parameter based on this analysis, a simple MATLAB code is used to formulate the previous equations and to calculate the
inverter value in one step knowing the values of the scattering parameters of the simulated EM structure.

4.5 Fourth-Order Filter Example

In this section, a design example of the circular waveguide cavity filter is demonstrated. The synthesized coupling matrix of this filter is given by

\[
M = \begin{bmatrix}
0 & 1.017 & 0 & 0 & 0 & 0 \\
1.017 & 0 & 0.8306 & 0 & -0.2963 & 0 \\
0 & 0.8306 & 0 & 0.8145 & 0 & 0 \\
0 & 0 & 0.8145 & 0 & 0.8306 & 0 \\
0 & -0.2963 & 0 & 0.8306 & 0 & 1.017 \\
0 & 0 & 0 & 0 & 1.017 & 0
\end{bmatrix} \tag{4.15}
\]

The coupling values in equation (4.15) are used in the design of the inter-cavity iris as well as the input iris. A similarity transformation can be used to obtain the transversal coupling matrix that is based on the physical diagonally polarized modes in (4.16).

\[
M' = \begin{bmatrix}
0 & 0.719 & 0.719 & 0 & 0 & 0 \\
0.719 & -0.8306 & 0 & 0.259 & -0.555 & 0 \\
0.719 & 0 & 0.8306 & -0.555 & 0.259 & 0 \\
0 & 0.259 & -0.555 & 0.8306 & 0 & 0.719 \\
0 & -0.555 & 0.259 & 0 & -0.8306 & 0.719 \\
0 & 0 & 0 & 0.719 & 0.719 & 0
\end{bmatrix} \tag{4.16}
\]

The transversal coupling matrix contains values of the local resonances inside every cavity. The coupling matrices of (4.15) and (4.16) exhibit a 4th order filter response.
with an in-band return loss of 20 dB and two transmission zeros at normalized frequencies of ±1.6. The de-normalization of the coupling coefficients will be in a way that attains a center frequency of operation of 11 GHz and a bandwidth of 50 MHz.

This chosen band of operation is arbitrary to demonstrate that the same design procedure can be used to design filters at higher frequencies and with different bandwidth from the common channel bandwidth of 39 MHz used in satellite applications. The proposed filter structure is shown Figure 4-5 with rectangular waveguide input of dimensions 22.86mm by 10.16mm (WR-90 standard).

![Figure 4-5. Proposed circular waveguide dual-mode 4th order filter structure.](image)

The first step of the design is the design of the dual-mode cavity supporting two diagonal modes of Eigen the required Eigen frequencies. The dimensions of this cavity
must also assure that only the fundamental mode is resonating and all the higher order modes are at cutoff. The length and radius of the cavity that satisfy these conditions are 24.42mm and 9.5mm, respectively. The outward perturbations split the fundamental mode into two modes; they are minimal in size and the accuracy of the fabrication process has to be taken into consideration in these design steps. It has a radius of 1mm along the length of the cavity and is shifted inward by an offset of 0.592mm. Figure 4-6 shows the top cross section of the cavity to demonstrate the dimensions findings above.

![Diagram of cavity cross section](image)

**Figure 4-6. Cross section of the cavity with the relevant dimensions.**

The next step is the design of the inter-cavity iris. This iris is a rectangular opening in the side wall of the cavities. It is centered right in the middle of the length of the cavity as shown in Figure 4-5. The opening dimensions are proportional to the required bandwidth of 50 MHz. The aperture has a height of 3.76mm and a width of 7.67mm.

The input iris is designed as explained before, but with the difference that the opening in the rectangular input feed comes in the bottom of it so that all of the structure
has a flat surface. The dimensions of the input iris are 10.68mm and 5.23mm. Figure 4-7 shows the side view of the input to the filter. The loading of the irises on the cavity is compensated for by reducing the length of it until the frequency shifts up to the required values so that the cavity represents a phase shift of \( \pi \) for both resonant modes.

The complete filter structure is simulated using CST software package and gives excellent response. The S-parameters from the EM simulation almost perfectly match the ideal response of the synthesized coupling matrix as shown in Figure 4-8.

The small discrepancies between the EM simulation and the ideal response are due to the fact that although the filter is considered narrow-band, the inverters’ values are still a function in frequency. Also the synthesis and design process require a high numerical accuracy and at that high frequency the response is very sensitive to very little dimension changes, even of a few micro-meters.
The fabrication of this filter was accomplished at the machining labs at the physics department at Queen’s university using computerized CNC milling machines. Despite the fact that the accuracy of the fabrication process is not high enough for such a filter of such sensitivity to dimensions, the measurements show a good enough agreement with the desired response. The process allows the fabrication of this filter on 4 different pieces. The pieces are compiled together using screws which also affects the response of the filter. The advantage of this filter is that the cost of labor for its fabrication is very low and the ease of design saves a lot of time and effort. The fabricated filter is shown in Figure 4-9 and the measurements are depicted in Figure 4-10 along with the ideal response.
Figure 4-9. Fabricated 4\textsuperscript{th} order filter.

Figure 4-10. Measurements of the 4\textsuperscript{th} order circular waveguide filter.
The measured response shows a high level of losses in the passband as well as a non-optimum return loss, this due to the fact that the filter fabrication is done for 8 separate pieces tied together using fixing screws. The figure also shows an additional transmission zero around 11.3 GHz which can be an effect of the higher order modes and the connectors to the Vector Network Analyzer (VNA).

4.6 Conclusion

In this chapter, the design technique of a novel category of circular waveguide filters is introduced. The design theory is demonstrated and conveyed with a practical example and measurements. The designed dual-mode filter structure is suitable for temperature compensation since the alignment of the filter leaves the top and bottom surfaces of the cavities unattached to any coupling structures. The advantage of this class of filters is that all the resonant cavities and the coupling structures are simple structures that require no complicated manufacturing process. Another advantage is that the direct design requires no tuning after fabrication; the synthesized filtering topology is directly mapped to the physical structure implemented in circular waveguide technologies and the measured response shows good agreement with the analytical data.
Chapter 5

Conclusions and Future Work

This chapter includes a brief discussion of the work presented in this thesis document as well as suggestions for future research and studies in the same topics to be added to and further complete the accomplished work.

5.1 Discussion and Conclusions

The work demonstrated in this thesis comprises a broad introduction to a range of microwave filters types and the important role they play as a front-end component in various communication systems. Different design examples of filters in open and closed waveguide technologies have been presented.

The core of this work is the completion of the design procedures of two novel microwave filter structures in both rectangular and circular waveguide configurations. These filtering structures comply with the recent advances and requirements in the field of waveguide filters of specific importance in satellite communication systems.

The first filter introduced is a 4\textsuperscript{th} order rectangular waveguide filter, the design is accomplished using cross-coupled dual-mode cavities, the advantage of this filter design is the symmetry feature; since it is end-to-end symmetric, it simplifies the manufacturing process and reduces its cost. The coupling iris used in this design is a novel structure that provides independent inductive and capacitive couplings which translates to independent positive or negative phase shifts for different polarizations. This iris design gives better understanding of the concept of virtual negative coupling brought up in the literature. The
theory is supported with practical design examples, EM simulation and practical measurements.

Another design theory for dual-mode circular waveguide filters is introduced. An innovative design for the feeding of the circular waveguide cavities from the side walls leads to an original design suitable for compensation of frequency shifts in the resonances due to temperature change. The temperature compensation is allowed due the uncoupled flat surfaces of the top and bottom faces of the circular cavities, however actual temperature compensation either with conductive or dielectric materials is out of the scope of this thesis. The synthesis of a coupling matrix is performed; this coupling matrix is directly mapped to the physical design structure using the physical resonances representation inside the cavities. The performance of this filter is judged through multiple EM simulations and practical measurements and both are in excellent agreement with the analytical ideal response of the filtering function.

5.2 Future Work Recommendations

The following suggestions are for future work relevant to the topics considered in this work that can be made:

- The synthesis and design of dual-mode filters of higher orders for the fully symmetric filter with mixed coupling irises.
- The realization of actual temperature compensation for the filter designed using circular waveguide cavities.
- A development of an accurate analytical method for the loading compensation of the coupling irises on the cavities designed for use in the temperature compensated filter.
References


