REFLECTED GROUP DELAY METHOD WITH SPACE MAPPING TECHNIQUES
FOR COUPLED-RESONATOR FILTER DESIGN

A Thesis
Submitted to the Faculty of Graduate Studies and Research
In Partial Fulfillment of the Requirements
For the Degree of

Master of Applied Science
in
Electronic Systems Engineering
University of Regina

By
Xiaolin Fan
Regina, Saskatchewan
April, 2015

Copyright 2015: Xiaolin Fan
Xiaolin Fan, candidate for the degree of Master of Applied Science in Electronic Systems Engineering, has presented a thesis titled, *Reflected Group Delay method with Space Mapping Techniques for Coupled-Resonator Filter Design*, in an oral examination held on April 27, 2015. The following committee members have found the thesis acceptable in form and content, and that the candidate demonstrated satisfactory knowledge of the subject material.

External Examiner: Dr. Andrei Volodin, Department of Mathematics & Statistics

Supervisor: Dr. Paul Laforge, Electronic Systems Engineering

Committee Member: Dr. Raman Paranjape, Electronic Systems Engineering

Committee Member: Dr. Christine Chan, Software Systems Engineering

Chair of Defense: Dr. Farshid Torabi, Petroleum Systems Engineering
Abstract

The RF (radio frequency) and microwave filter is widely used in various applications in the fields of radio broadcasting, radar, telecommunication and satellite technologies. The design methods of RF and microwave filters are important topics, especially for electromagnetic (EM) based design. In this thesis, a novel design method for sequential resonator-coupled bandpass filter is proposed by implementing the reflected group delay design approach with space mapping techniques.

The theory of the reflected group delay method is discussed in detail. An improvement to the traditional reflected group delay method is proposed in which reflected group delay values at selected sweep frequency points are exploited as the target goals for each design stage instead of the whole curve used in the traditional method. Several filter design examples are given to verify the efficiency and accuracy of the improved reflected group delay method.

An EM based design method is first proposed by implementing the aggressive space mapping technique as the optimization algorithm for the improved reflected group delay method in designing a 5-pole microstrip hairpin filter. The optimization routine using the reflected group delay method is represented mathematically and a design procedure is proposed for the integration of the aggressive space mapping technique and the reflected group delay method. The design steps are summarized and the filter is fabricated and tested.

Another method to integrate the reflected group delay design approach with the implicit space mapping technique is also proposed. This method is applied to the design
of a 6-pole microstrip hairpin filter. Detailed design theory and procedures are given. The 6-pole microstrip hairpin filter is designed using the Sonnet EM simulator. By using the proposed methods, the computation time and space mapping iterations are significantly reduced.

The proposed methods are proven to be very efficient and accurate for EM-based sequential coupled resonator filter design compared to traditional EM based filter design methods.
Acknowledgements

The completion of the work presented in this thesis is not possible without the help of a lot of people. First and foremost, I would like to express my great appreciation to my supervisor Dr. Paul Laforge for his great guidance, ideals, advice, and full support to my research work throughout my graduate studies. I also want to thank to all the members in our research group.

Additionally, I want to acknowledge to the scholarship I received from the Faculty of Graduate Studies and Research.

Finally, I would like to express my deepest appreciation and gratitude to my parents for their infinite, unconditional love and inspiration, and their continuous support and encouragement to help me overcome all the obstacles during these years. I would like to dedicate this work and all my appreciation to my parentis for their love.
Table of Contents

Abstract ............................................................................................................................................... I
Acknowledgements .......................................................................................................................... III
Tables of Contents ........................................................................................................................... IV
List of Figures ...................................................................................................................................... VI
List of Tables ....................................................................................................................................... VIII

CHAPTER 1 Introduction ...................................................................................................................... 1

1.1 Outline ........................................................................................................................................ 1
1.2 Motivation .................................................................................................................................... 2
1.3 Thesis Organization ...................................................................................................................... 3

CHAPTER 2 Literature Review .............................................................................................................. 5

2.1 Basic Concepts and Theory of Filter ............................................................................................... 5
  2.1.1 General Definitions .................................................................................................................. 5
  2.1.2 Characteristic Polynomial and Transfer Function ................................................................. 7
  2.1.3 Lowpass Prototype Filters and Elements ............................................................................... 8
  2.1.4 Coupling Matrix Synthesis ................................................................................................... 11
2.2 EM Based Optimization using Space Mapping Techniques .......................................................... 14
  2.2.1 Space Mapping Basic Concepts ............................................................................................ 15
2.3 Filter Design using Reflected Group Delay Method ....................................................................... 18

CHAPTER 3 Reflected Group Delay Method for Sequential Coupled Resonator Filter ................................................................. 23

  3.1 Improvement to the Reflected Group Delay Method ................................................................. 23
3.2 Reflected Group Delay in Band Pass Prototype ................................................................. 25
3.3 Application of Improved Reflected Group Delay to Filter Designs ................................. 29
  3.3.1 Example of Calculated Goals for Design of a 12-pole Chebyshev Band-pass Filter ................................................................. 29
  3.3.2 Application of the Goals to a 12-pole End-coupled Microstrip Band-pass Filter ....... 31
  3.3.3 Application of the Goals to a 12-pole Parallel-coupled Band-pass Filter............... 34
  3.3.4 Application of the Goals to a 12-pole Hairpin Band-pass Filter ............................... 37
  3.3.5 Application of the Goals to a 12-pole Hairpin Band-pass Filter with Different Dielectric Material and Substrate Configurations ................................................................. 40

CHAPTER 4  Reflected group delay method with aggressive space mapping ................. 44
  4.1 Introduction .................................................................................................................. 44
  4.2 Reflected Group Delay Method with Aggressive Space Mapping Design Theory ........ 46
  4.3 Reflected Group Delay Method with Aggressive Space Mapping Design Procedure .... 47
  4.4 Application to a 5-pole Microstrip Filter .................................................................... 50

CHAPTER 5  A Sequentially Coupled Filter EM based Design Approach Using the
Reflected Group Delay Method and the Implicit Space Mapping Technique ............. 60
  5.1 Introduction .................................................................................................................. 61
  5.2 Proposed Reflected Group Delay Method and the Implicit Space Mapping Design Approach ....................................................................................................................... 61
  5.3 Application .................................................................................................................. 62

CHAPTER 6  Conclusion ....................................................................................................... 72

APPENDIX A ...................................................................................................................... 71

Reference .......................................................................................................................... 72
LIST OF FIGURES

Fig. 2.1 Doubly terminated lossless transmission network ................................................. 5

Fig. 2.2 General form of lowpass prototype (a) shunt-series configuration, (b) series-shunt configuration [8]........................................................................................................... 9

Fig. 2.3 Equivalent circuit of n-coupled resonators for loop-equation formulation [3] ........................................................................................................................................ 12

Fig. 2.4 Space mapping implementing concept [18].......................................................... 17

Fig. 2.5 General Lowpass Prototype ................................................................................... 19

Fig. 3.1 Equivalent-circuit of the sequentially coupled band pass filter [15] .................... 25

Fig. 3.2 Equivalent circuit of the one-port network for each reflected group delay design stage. ........................................................................................................................................... 26

Fig. 3.3 Circuit schematic for the 12-pole hairpin end-coupled filter ......................... 31

Fig. 3.4 ADS simulation result of the 12-pole end-coupled bandpass filter .................. 33

Fig. 3.5 Circuit schematic for the 12-pole parallel-coupled bandpass filter .................... 34

Fig. 3.6 ADS simulation result of the 12-pole parallel-coupled band-pass filter ............ 36

Fig. 3.7 Circuit schematic for the 12-pole hairpin bandpass filter ............................ 37

Fig. 3.8 ADS simulation result of the 12-pole hairpin band-pass filter....................... 39

Fig. 3.9 Circuit schematic for the 12-pole hairpin bandpass filter using alumina ....... 41

Fig. 3.10 ADS simulation result of the 12-pole hairpin band-pass filter using Alumina ................................................................. 43

Fig. 4.1 Coarse Model Schematic ($L_o = L - L_m$): (a) first resonator (steps 1 and 3) (b) first two resonators (step 6) (c) entire filter (step 8). ................................................................. 51
Fig. 4.2 Sonnet Layout: (a) first resonator (step 2 and 4) (b) first two resonators (step 7) (c) entire filter (step 8). .................................................................................. 53

Fig. 4.3 Fine Model Reflected Group Delay response for each space mapping iteration. (a) First resonator. (b) Frist and Second Resonators. The cross lines are the results of first space mapping iteration, the circle lines are the second iteration, the solid lines are the optimum response................................................................. 56

Fig. 4.4 EM simulation return loss of the first iteration (dotted line), second iteration (solid line) and final filter design (bold line). ................................................................. 57

Fig. 4.5 The fabricated microstrip 5-pole hairpin bandpass filter ........................................ 57

Fig. 4.6 Measured results of fabricated filter................................................................. 57

Fig. 5.1 Geometry in Sonnet for each reflected group delay stage (a) stage 1 (b) stage 2 (c) stage 3 (d) entire filter................................................................. 65

Fig. 5.2 Calibrated Coarse Model for reflected group delay design stages (a) stage 1, (b) stage 2, (c) stage 3, (d) entire filter ................................................................. 68

Fig. 5.3 Final EM simulation result of the 6 pole microstrip hairpin bandpass filter. Dotted line is the initial response, solid line is the response after one iteration. ................. 70
LIST OF TABLES

Table 2.1 Group-delay values at center frequency in term of low-pass prototype [1]
.................................................................................................................................................. 21

Table 3.1 Input impedance in terms of g values and frequency variables for each
group delay stage based on the low-pass prototype circuit start with shunt capacitor...... 24

Table 3.2 Formulas to calculate reflected group delay in terms of adjacent
couplings and resonator resonance frequencies ........................................................................... 28

Table 3.3 Target reflected group delay goals of each design stage for design of a
12-pole Chebyshev bandpass filter ............................................................................................... 30

Table 3.4 Design Parameters of a 12-pole end-coupled bandpass filter in ADS.............. 32

Table 3.5 Design Parameters of a 12-pole parallel-coupled Bandpass filter in ADS........ 35

Table 3.6 Design Parameters of a 12-pole hairpin Bandpass filter in ADS ..................... 38

Table 3.7 Design Parameters of a 12-pole hairpin Bandpass filter using alumina in
ADS ............................................................................................................................................... 42

Table 4.1 The target reflected group delay goals of each stage for design of a filter
with center frequency of 1GHz, fractional bandwidth of 15% .............................................. 52

Table 4.2 Coarse model optimizing results of each reflected group delay design
stage ............................................................................................................................................... 55

Table 4.3 Space Mapping Iterations in each RGD stage & Final Optimum results ......... 55

Table 5.1 Selected frequency points and goals for each reflected group delay stage ....... 63

Table 5.2 Coarse model design parameters and optimizing result for each reflected
group delay stage ......................................................................................................................... 64
Table 5.3 fine model design parameters and space mapping result for each reflected group delay stage

................................................................. 69
Chapter 1
Introduction

1.1 Outline
Microwave filters play very important roles in RF applications used for most broadcast radio, television, wireless communication, space applications. They are used to select and confine RF signals within assigned frequency spectrum in the range from MegaHertz to GigaHertz. They are widely included in various RF and microwave devices as important filtering blocks for the transmitting and receiving of RF signals. The design of a filter is an important topic due to the computational expensive nature of electromagnetic field analysis and the requirements of stringent filter specifications to achieve higher performance, smaller size, lighter weight and lower cost. Over the years, research has been performed in the areas of simplifying the design process, using circuit simulations appropriately, and implementing electromagnetic (EM) simulations efficiently.

The reflected group delay [1] method provides a filter tuning and design procedure in which resonators are successively added stage by stage. In each step, design parameters are tuned and optimized to match the group delay of the reflected signal to a set of objective curves calculated from the low-pass prototype model. This method has been proven to be a great method for the design of sequentially coupled resonator filter.

The space mapping (SM) technique [2] introduced a very efficient approach to carry out the computational expensive EM based optimizing problem. The idea of space mapping is to establish a mapping between the spaces of the EM based “fine model” and an inaccurate but faster “coarse model”. In this way, the optimizing process is directed to the
faster coarse model while the simulation accuracy is ensured by the EM based “fine model”.

In this thesis, an improvement is proposed to the reflected group delay method by introducing specific sweep frequencies and group delay goals in each design stage. The mathematical theory required to obtain the required sweep frequencies and group delay goals based on low pass prototype circuit model is given in detail. A set of new equations to characterize the reflected group delay of a bandpass prototype circuit model is first proposed. A 12-pole hairpin bandpass filter is designed in Keysight ADS to verify the accuracy and efficiency of the proposed improvement.

An EM based filter design method combining the original reflected group delay approach and the space mapping technique is proposed.

A mathematical description of the optimization processes required for implementing aggressive space mapping with reflected group delay method [26] is presented. The application of the proposed design procedure to a 5-pole hairpin microstrip band pass filter is given.

Another design approach exploiting implicit space mapping with reflected group delay method is proposed. An application of a 6-pole hairpin microstrip band pass filter is given to verify this proposed method.

1.2 Motivation

The reflected group delay method divides a filter design process into smaller steps and reduced the number of required optimizing parameters. However, due to limitations of hardware and optimizing algorithm, the optimization becomes a problem in the filter design especially for high-pole structures which often lead to long optimizing time and non-convergence. The improvement to the reflected group delay method proposed in Chapter 3
can significantly speed up the optimization process and lead to accurate design solutions.

The EM based filter design is always a problem due to the computational expensive simulate environment and the complexity of filter structure. The combined EM based design method using reflected group delay approach and space mapping techniques proposed in Chapter 4 and Chapter 5 can result in a large savings in computation time as the space mapping techniques significantly reduced the optimizing iterations and the reflected group delay approach ensure that each design stage contains less parameters than that required in space mapping, which reduces the time for parameter extraction and reduces the time to determine the optimum design.

1.3 Thesis Organization

Following the introduction given in Chapter 1, Chapter 2 presents an overview of the synthesis, techniques and methods used in this thesis, including the reflected group delay method, the space mapping techniques and the general filter synthesis and coupling matrix synthesis.

Chapter 3 presents detailed mathematical synthesis of improvements to the reflected group delay method by introducing selected points for the target reflected group delay goals. Examples are given to validate that these improvements are efficient and accurate for the design of filters with different structures, materials and desired specifications. A set of equations for characterizing the reflected group delay of the band pass prototype circuit is first reported which can be directly exploited for filter design.

Chapter 4 proposes a novel design procedure combining the reflected group delay design approach and the aggressive space mapping technique [26]. Detailed mathematical concepts and representations are given. A five-pole microstrip hairpin band pass filter is
designed, fabricated and tested with the proposed design method.

Chapter 5 proposes the design procedure combining the reflected group delay design approach and the implicit space mapping technique. Detailed mathematical concepts and representations are given. The selections of pre-assigned parameters are discussed. A 6-pole microstrip band-pass filter is designed with the Sonnet EM simulator to validate the proposed method. Future work and conclusions are presented in Chapter 6.
Chapter 2
Literature Review

2.1 Basic Concepts and Theory of Filter

2.1.1 General Definitions

From filter synthesis, a specified filter design can be achieved by carrying out appropriate frequency and impedance level scaling to the corresponding lumped element lossless lowpass filter. The normalized lumped element lossless lowpass filter network is named the lowpass prototype circuit which is normalized to a cutoff frequency of 1 rad/s and terminated in resistor of 1 Ω. This section deals with the circuit theory approximation for the design of lossless lowpass prototype filter transfer functions. Fig. 2.1 illustrates a lossless two-port network terminated in resistors. This is a general representation of a doubly terminated filter network capable of maximum power transfer.

Fig.2.1 Doubly terminated lossless transmission network
The maximum available power $P_{\text{max}}$ is generated from the ideal voltage source. $P_{\text{through}}$ is the power delivered to the load $R_2$. In a passive lossless two-port networks, $P_{\text{through}}$ is equal to or less than $P_{\text{max}}$. Thus, a characteristic function $k(s)$ is defined to express the power transmission in the network as:

$$\frac{P_{\text{max}}}{P_{\text{through}}} = 1 + k(s)^2$$

(2.1)

Where $s = j\omega$

According to transmission-line theory,

$$|\rho|^2 = \frac{P_{\text{reflect}}}{P_{\text{max}}}$$

$$|t|^2 = \frac{P_{\text{through}}}{P_{\text{max}}}$$

(2.2)

Here $\rho$ is reflection coefficient and $t$ is the transmission coefficient which is synonymous with $S_{11}$ and $S_{21}$, respectively. The physical meaning of $S_{11}$ is the input reflection with the output of the network terminated by a matched load; $S_{21}$ is the forward transmission from source to load.

By conventional definition, the insertion loss $L_A$ can be computed as:

$$L_A = 10 \log_{10} \frac{1}{|S_{21}|^2} \text{ (dB)}$$

(2.3)

transmission group delay ($\Gamma d$) of the filter can be found as:

$$\Gamma d = -\frac{\partial \theta_{21}}{\partial \omega} \text{ (sec)}$$

(2.4)

Where $\theta_{21}$ is the transmission phase.
2.1.2 Characteristic Polynomial and Transfer Function

The reflection coefficient for a two-port network is given by

\[
\rho = \frac{Z_{\text{in}} - R}{Z_{\text{in}} + R} = \frac{z - 1}{z + 1}
\]  

(2.5)

Here \( z = \frac{Z_{\text{in}}}{R} \) is the normalized impedance, and \( Z_{\text{in}} \) is input impedance.

The normalized impedance \( z \) can be also expressed as the ratio of the numerator and denominator polynomials:

\[
z(s) = \frac{n(s)}{d(s)}
\]

(2.6)

Hence

\[
\rho = \frac{z - 1}{z + 1} = \frac{n - 1}{d - 1} = \frac{n - d}{n + d} = \frac{F(s)}{E(s)}
\]

(2.7)

\[
t = \frac{P(s)}{E(s)}
\]

(2.8)

Where the three polynomials \( P(s), F(s) \) and \( E(s) \) are defined as the characteristic polynomials.

\[
k = \frac{P(s)}{F(s)}
\]

is defined as the transfer function of the filter network.

The roots of polynomial \( F(s) \) are the zeros of reflection which stand for the frequencies at which all the power is transmitted and none is reflected. The roots of
polynomial \( P(s) \) are commonly called attenuation poles or transmission zeros at which no power is transmitted through the network filter function and response shapes are characterized by characteristic polynomials. For all-pole filter prototype, the polynomial \( p = 1 \) for all frequencies variables, the response shape is determined by \( F(s) \).

\[
\tau = \frac{P(s)}{E(s)} = \frac{1}{E(s)}
\]

\[
\rho = \frac{F(s)}{E(s)}
\] (2.9)

Here, for a Butterworth response:

\[
F(s) = s^n
\]

For Chebyshev response:

\[
F(s) = s^n (s^2 + a_1^2)(s^2 + a_2^2)\ldots(s^2 + a_n^2)
\]

For prototype filter functions with finite transmission zeros and equal ripple passband response:

\[
P(s) = (s^2 + b_1^2)(s^2 + b_2^2)\ldots(s^2 + b_n^2)
\]

\[
F(s) = s^n (s^2 + a_1^2)(s^2 + a_2^2)\ldots(s^2 + a_n^2)
\] (2.10)

### 2.1.3 Lowpass Prototype Filters and Elements

The realization of filter transfer function is usually carried out with the lowpass prototype filter configuration. As describe in the previous section, the lowpass prototype filter is a normalized lumped-element lossless filter network containing a number of
reactive elements as shown in Fig. 2.2. The ‘g’ values are used to denote the inductance of a series inductor or the capacitance of a shunt capacitor. In the design process, it is of importance to determine the g values from the transfer function and characteristic polynomial of a desired lowpass prototype filter.

\[
|S_{21}|^2 \overset{\text{even}}{=} \frac{1}{1 + k^2} \quad (2.11)
\]

For Butterworth filter

\[
k^2 = \xi \omega^2 \quad (2.12)
\]

For Chebyshev filter
\[ k^2 = e^{2T_n^2 \left( \frac{\omega}{\omega_c} \right)} \]  

(2.13)

Here \( T_n^2(x) = \cos \left( n \cos^{-1} x \right) \) is the mathematical expression for Chebyshev type response.

As long as the transfer function \( k \) and scattering parameters are obtained, the input impedance and \( g \) values of the lowpass prototype filter can be determined.

For Butterworth filters, \( g \) values can be derived from the following formulas:

\[ g_0 = 1 \]

\[ g_k = 2 \sin \left( \frac{(2k-1)\pi}{2n} \right) \]  

(2.14)

\[ g_{n+1} = 1 \]

For Chebyshev filters, \( g \) values can be derived from the following formulas:

\[ g_0 = 1 \]

\[ g_1 = 2 \sin \left( \frac{\pi}{2n} \right) \]

\[ g_k = \frac{1}{g_{k-1}} \frac{4 \sin \left( \frac{(2k-1)\pi}{2n} \right) \sin \left( \frac{(2k-3)\pi}{2n} \right)}{\gamma^2 + \sin^2 \left( \frac{(k-1)\pi}{n} \right)} \]  

for \( k = 2, 3, \ldots, n \)  

(2.15)

\[ g_k = \begin{cases} 1 & \text{for } n \text{ odd} \\ \coth^2 \left( \frac{\beta}{4} \right) & \text{for } n \text{ even} \end{cases} \]

Where \( \gamma = \sinh \left( \frac{\beta}{2n} \right) \) and \( \beta = \ln \left( \coth \left( \frac{L_{AB}}{17.37} \right) \right) \). \( L_{AB} \) is the pass band ripple in dB.
It is important to note that the lowpass prototype ladder network of Fig. 2.2 locates all its transmission zeros at infinity. As a result, only all-pole filters like the Butterworth and the Chebyshev filter can be synthesized by such a configuration. For filters with finite transmission zeros, the coupling matrix synthesis is considered to be more preferable.

2.1.4 Coupling Matrix Synthesis

In recent years cross coupled microwave resonator have been widely applied in microwave filter design. By using cross-coupled resonators to generate finite transmission zeros, the filter synthesis technique with a coupling matrix was developed. The coupling matrix concept was developed in the 1970s by Atia and Williams [10]. Carmon [11]-[12] gives a scheme to determine the filtering function with arbitrary placed transmission zeros. Once the system function is obtained, the coupling matrix is found by extracting element values. In this section, the synthesis method of designing a microwave filter with a coupling matrix is reviewed.

An equivalent prototype circuit of coupled-resonator filter [3] is given in Fig. 2.3 where \( e \) represent the voltage source, \( i \) represent the loop current, \( R, L, C \) represent ideal resistance, inductance and capacitance.
Fig. 2.3 Equivalent circuit of n-coupled resonators for loop-equation formulation [3]

The loop equation of this network is

\[
\begin{bmatrix}
R_s + j\omega L_1 + \frac{1}{j\omega C_1} & -j\omega L_{12} & \cdots & -j\omega L_{1n} \\
-j\omega L_{21} & j\omega L_2 + \frac{1}{j\omega C_2} & \cdots & -j\omega L_{2n} \\
\vdots & \vdots & \ddots & \vdots \\
-j\omega L_{n1} & -j\omega L_{n2} & \cdots & R + j\omega L_n + \frac{1}{j\omega C_n}
\end{bmatrix}
\begin{bmatrix}
i_1 \\
i_2 \\
\vdots \\
i_n
\end{bmatrix}
= \begin{bmatrix}
e_s \\
0 \\
\vdots \\
0
\end{bmatrix}
\tag{2.16}
\]

Suppose that the resonators of the filter are synchronously tuned (resonate at the same frequency), \( \omega_0 = \frac{1}{\sqrt{LC}} \) is obtained where \( L = L_1 = \cdots = L_N \), and \( C = C_1 = \cdots = C_N \) such that (2.14) is rewritten as

\[
\begin{bmatrix}
\frac{R_s}{\omega_0 L \cdot FBW} + p & -j\frac{\omega L_{12}}{\omega_0 L \cdot FBW} & \cdots & -j\frac{\omega L_{1n}}{\omega_0 L \cdot FBW} \\
-j\frac{\omega L_{21}}{\omega_0 L \cdot FBW} & \frac{\omega L_2}{\omega_0 L \cdot FBW} & \cdots & -j\frac{\omega L_{2n}}{\omega_0 L \cdot FBW} \\
\vdots & \vdots & \ddots & \vdots \\
-j\frac{\omega L_{n1}}{\omega_0 L \cdot FBW} & -j\frac{\omega L_{n2}}{\omega_0 L \cdot FBW} & \cdots & \frac{R}{\omega_0 L \cdot FBW} + p
\end{bmatrix}
\begin{bmatrix}
i_1 \\
i_2 \\
\vdots \\
i_n
\end{bmatrix}
= \begin{bmatrix}
e_s \\
0 \\
\vdots \\
0
\end{bmatrix}
\tag{2.17}
\]

where \( FBW = \Delta \omega / \omega_0 \) and \( p = j \frac{1}{FBW} \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) \)

The normalized coupling coefficient, \( m_{ij} = \frac{L_{ij}}{L \cdot FBW} \) in which \( L_{ij} \) represents the mutual inductance between resonators \( i \) and \( j \) is defined.

Hence, the current-voltage relationship of the network is:
Where \( [M] = \begin{bmatrix} m_{11} & m_{12} & \cdots & m_{1n} \\ m_{21} & m_{22} & \cdots & m_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ m_{n1} & m_{n2} & \cdots & m_{nn} \end{bmatrix} \) is defined as the coupling matrix, \([I]\) is a unity diagonal matrix, \([R]\) is termination impedance matrix. The scattering parameters of the two-port network are expressed as:

\[
S_{21} = 2 \sqrt{R_s R_L} \cdot i_s \\
S_{11} = 1 - 2 R_i \cdot i_i 
\]

By combining those equations, the S parameters can be calculated in terms of the coupling matrix elements as:

\[
S_{21} = -2 j \sqrt{R_s R_L} \cdot [A]_{s1}^{-1} \\
S_{11} = 1 + 2 j R_s \cdot [A]_{11}^{-1}
\]

From Cameron’s method \([11]-[12]\), coupling matrix can be directly constructed from a filter’s transfer and reflection polynomials. A recursive procedure is used to obtain the filter polynomial \(F(s)\), \(P(s)\) and \(E(s)\) for filters with transmission zeros. Then, the scattering parameters can be calculated. To obtain a coupling matrix for a desired scattering parameter specification, it is normally required to firstly convert the scattering parameters
into admittance parameters, then using Eigendecomposition and Orthonormalization [5] to construct the eigenvector matrix $T$ in which the first and last row are calculated from the admittance parameter and the remaining orthogonal rows are constructed by the Gram-Schmitt Orthonormalization process [6]. Finally, the coupling matrix $M$ is synthesized by using the following equation

$$-M = T \cdot \Lambda \cdot T'$$  \hspace{1cm} (2.21)

Where $\Lambda$ is

$$\begin{bmatrix}
\lambda_1 & 0 & \cdots & 0 \\
0 & \lambda_2 & \cdots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & \cdots & 0 & \lambda_n
\end{bmatrix}$$

and $\lambda_i$ is the eigenvalue of $-M$. $T'$ is the transpose of $T$.

2.2 EM Based Optimization using Space Mapping Techniques

A number of computational electromagnetic methods have been proposed in the past decades for modeling real world high frequency electromagnetic problems in RF microwave industry. Computational numerical methods are developed to derive valid solutions to Maxwell equations for arbitrary shaped designs under EM environment. EM modeling methods such as finite element method, finite-difference time-domain method and the method of Moment are widely used for modeling and analyzing involutes relations of media, electromagnetic fields, boundary conditions and many other microwave concepts. Due to the complexity of EM modeling concepts and limitations of computer hardware performance, most of these techniques are prohibitively computational-expensive. It is very hard to directly apply these modeling techniques into design and optimizing processes. In the past 20 years, many EM-based design and optimizing techniques have been comprehensively researched. In this section one of the most famous EM based optimizing methods, space mapping technique is reviewed.
In the RF and microwave area, circuit-theory based modeling techniques and computer aided design (CAD) tools are developed and widely used to execute a rough microwave design approximation. Different numerical optimizing algorithms are comprehensively applied to the designing process benefiting from its fast computation speed with moderate sacrifice of approximating accuracy.

The space mapping method[2] introduced by Bandler, Biernacki, Chen, Grobelny and Hemmers in 1994 proposed a superior way to carry out EM based computational-expensive optimization efficiently with the assistance of a circuit based fast but low accuracy surrogate modeling. The computational-expensive and accurate EM based model is defined as the “fine model”, the fast and low accuracy circuit based surrogate model is defined to be the “coarse model”. The goal of space mapping is to establish an appropriate mapping relationship between the two models and achieve specification requirements within a minimum number of the computational-expensive EM based fine model evaluations.

2.2.1 Space Mapping Basic Concepts

Optimizing technique has been used in microwave design for decades. The target of optimization is to determine the values of some circuit or physical parameters in order to make the output response satisfy required specifications. Traditional optimization techniques directly optimize parameters and available derivatives to force the output response converge to required specifications. However, most of these optimization techniques are costly for nonlinear microwave optimizing problems. They often require many model simulations which could be very computer-intensive and time consuming for an accurate EM simulator.
In 1994, John Bandler successfully applied the space mapping technique to microwave design. In the recent years, more and more researches [9], [16]-[20] have taken to improve and extend the space mapping approach.

The space mapping technique introduced a simple mathematical synthesis to link or map design parameters in two different models such that the two models give the same output response. One model is an inaccurate but much faster coarse model. Another is a more accurate but time-intensive fine model. By applying space mapping technique, the original optimization process in the time-intensive fine model was carried out by a much faster coarse model while the accuracy of the result was preserved by the support of more accurate fine model analysis.

In the microwave area, circuit based CAD tools such as Agilent ADS [27] are faster but inaccurate simulators and EM simulators such as ADS Momentum, Sonnet EM [28], or Ansoft HFSS are more accurate but time-intensive. Most researchers prefer to use circuit based CAD tools as the coarse model while selected EM simulators are used for the fine model. A block diagram showing the approach of space mapping is given in Fig. 2.4.
Parameter Extraction (PE) is a sub problem in the space mapping methodology. In the PE step, coarse model parameters are optimized such that the output response is matched to a fine model response. This step is an optimizing process, which may lead to non-uniqueness solutions. The non-uniqueness of PE solution can lead to intensive PE time or a divergence or oscillation of the space mapping progress.
From the research performed and presented in this thesis, one of the best ways to overcome this problem is to reduce the number of optimizing parameters. A new method combining the reflected group delay method and space mapping is presented in chapters 4 and 5 to give a significant improvement in the filter design efficiency.

2.3 Filter Design using Reflected Group Delay Method

The reflected group delay method was first introduced by Ness in 1998[1]. Ness proposed the theory that group delay of the input reflected signal of sequentially tuned resonators contains all the information required for the design, measurement and tuning of sequentially coupled resonator filters.

In Ness’s method, the equations are derived to show that the reflected group delay value at the center frequency have direct mathematical relationship with the coupling between the connecting lines and the input and output resonators $Q$, and internal couplings between resonators $k_{ij}$. Thus, the group delay of the reflected signal can determine coupling values in a filter and determine the filter response.

Ness proposed a design approach in which each resonator was added and successively tuned until the input reflected group delay at the center frequency satisfies specific values and keeps the group delay response symmetric about the center frequency. The desired reflected group delay values are calculated based on low-pass prototype lumped element circuit in Fig. 2.5.
Fig. 2.5 General Lowpass Prototype

The group delay of \( S_{11} \) in the low-pass prototype model is defined as the derivative of reflected signal phase \( \phi \) with respect to frequency \( \omega \).

\[
\Gamma_{\phi}(\omega) = -\frac{\delta \phi}{\delta \omega} 
\]  

(2.22)

In a bandpass filter, the group delay of \( S_{11} \) can be obtained directly from the low pass prototype by

\[
\Gamma_{\phi}(\omega) = -\frac{\delta \phi \cdot \delta \omega^1}{\delta \omega} 
\]  

(2.23)

The frequency transformation from low pass to bandpass using

\[
\omega^1 \rightarrow \frac{\omega}{\omega_2 - \omega_1} \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) 
\]  

(2.24)

Where \( \omega^1 \) the frequency of low pass prototype, \( \omega_0 \) is center frequency of bandpass filter, \( \omega_1 \) is lower frequency of bandpass filter and \( \omega_2 \) is the upper frequency of the bandpass filter. \( \omega_2 - \omega_1 \) is the absolute bandwidth.

Thus, the group delay of the reflected signal \( S_{11} \) in this case can be derived as:

\[
\Gamma_{\phi}(\omega) = -\frac{\delta \phi}{\delta \omega} = -\frac{\delta \phi}{\delta \omega^1} \frac{\delta \omega^1}{\delta \omega} = -\frac{\delta \phi \cdot \delta \omega^1}{\delta \omega^1 \cdot \omega^2} \frac{\omega^2 + \omega_0^2}{\omega_2 - \omega_1} 
\]  

(2.25)
From transmission line theory, the reflected signal $S_{11}$ is defined as

$$S_{11} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} = \frac{jX_{in} - Z_0}{jX_{in} + Z_0}$$  \hspace{1cm} (2.26)$$

Where $Z_{in} = jX_{in}$ for the lossless case.

The phase of $S_{11}$ is expressed as:

$$\phi = \tan^{-1} \left( \frac{X_{in}(\omega)}{-Z_0} \right) - \tan^{-1} \left( \frac{X_{in}(\omega)}{Z_0} \right) = -2 \tan^{-1} \left( \frac{X_{in}(\omega)}{Z_0} \right)$$ \hspace{1cm} (2.27)$$

Therefore, a general formula can be derived to calculate reflected group delay in the bandpass circuit case:

$$\Gamma_g(\omega) = 2 \frac{\omega^2 + \omega_0^2}{\omega^2 (\omega_2 - \omega_1)} \left( \frac{\omega^2}{\delta \omega^4} \right) \frac{X_{in}(\omega^4)}{Z_0}$$ \hspace{1cm} (2.28)$$

For first resonator (with all other resonators detuned):

$$Z_0 = g_0$$

$$X_{in}^1 = -\frac{1}{\omega g_1}$$ \hspace{1cm} (2.29)$$

$$\therefore \Gamma_{g1}(\omega) = 2 \frac{\omega^2 + \omega_0^2}{\omega^2 (\omega_2 - \omega_1)} \frac{g_0 g_1}{1 + \left( g_0 g_1 \omega_1 \right)^2} = \frac{2 \left( \omega^2 + \omega_0^2 \right) g_0 g_1}{\omega^2 (\omega_2 - \omega_1) \left( 1 + \left( \frac{\omega}{\omega_2 - \omega_1} \frac{\omega_0}{\omega - \omega_0} \right)^2 \right)}$$

Ness gave the formulas to calculate the reflected group delay for first $n$ resonators (with all other resonators disconnected) at center frequency in terms of the low pass
prototype \( g \) values. Additionally, he proposed the mathematical relationship between the reflected group delay and coupling coefficients at center frequency in inverter coupled filter. The equations are shown in Table2.1

Table2.1 Group-delay values at center frequency in term of low-pass prototype \([1]\)

<table>
<thead>
<tr>
<th>First n resonator</th>
<th>( \Gamma_{d1} = \frac{4g_0g_1}{\Delta \omega} = \frac{4Q_E}{\omega_0} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( n=1 )</td>
<td></td>
</tr>
<tr>
<td>( \Gamma_{d2} = \frac{4g_2}{g_0\Delta \omega} = \frac{4}{\omega_0Q_Ek_{12}^2} )</td>
<td></td>
</tr>
<tr>
<td>( n=2 )</td>
<td></td>
</tr>
<tr>
<td>( \Gamma_{d3} = \frac{4g_0(g_1 + g_3)}{\Delta \omega} = \Gamma_{d1} + \frac{4Q_Ek_{12}^2}{\omega_0k_{23}^2} )</td>
<td></td>
</tr>
<tr>
<td>( n=3 )</td>
<td></td>
</tr>
<tr>
<td>( \Gamma_{d4} = \frac{4(g_2 + g_4)}{g_0\Delta \omega} = \Gamma_{d2} + \frac{4k_{23}^2}{\omega_0Q_Ek_{12}^2k_{34}^2} )</td>
<td></td>
</tr>
<tr>
<td>( n=4 )</td>
<td></td>
</tr>
<tr>
<td>( \Gamma_{d5} = \frac{4g_0(g_1 + g_3 + g_5)}{\Delta \omega} = \Gamma_{d3} + \frac{4Q_Ek_{12}^2k_{34}^2}{\omega_0k_{23}^2k_{45}^2} )</td>
<td></td>
</tr>
<tr>
<td>( n=5 )</td>
<td></td>
</tr>
<tr>
<td>( \Gamma_{d6} = \frac{4(g_2 + g_4 + g_6)}{g_0\Delta \omega} = \Gamma_{d4} + \frac{4k_{23}^2k_{45}^2}{\omega_0Q_Ek_{12}^2k_{34}^2k_{56}^2} )</td>
<td></td>
</tr>
</tbody>
</table>

Based on equations from Table 2.1, a conclusion is reached that the reflected group delay value at the resonant frequency in each stage is only affected by the low-pass prototype \( g \) values and the absolute filter bandwidth. In other words, the reflected group delay values in each stage are independent of filter designed center frequency. Ness emphasizes the importance to keep the reflected group delay response symmetric about the
center frequency. However, it is not easy to keep the group delay symmetric, especially for wider bandwidth. Later research [7] shows that this method only works for narrow band filter. As the filter bandwidth goes up, the reflected group delay response is no longer symmetric about the center frequency. In order to overcome this limitation, a correction factor is introduced to the original reflected group delay formula [7]

$$\Gamma_{d, symmetric}(f) = A(f)\Gamma_d(f)$$

(2.30)

Here: $A(f) = \frac{2f^2}{f^2 + f_0^2}$

By applying this correction factor, it is able to keep the reflected group-delay response logarithmic symmetric about the center frequency at any condition.

Additionally, Laforge presented that the reflected group delay method can be used to directly design the EM based filter [7]. The filter design process is divided into smaller design stages. Adding only one more resonator at each stage, consequently a few selected parameters are optimized in each stage to match the reflected group delay response from EM simulator to the reflected group delay of an ideal lumped circuit. Therefore, the computation and optimization time is less than the time required to design the entire filter in one optimization step. When all the design stages are complete, the EM based filter design is accomplished. Different from Ness’s method, he proposed that a matching of the group delay curve instead of single center frequency point is recommended during design process [7].
Chapter 3

Reflected Group Delay Method for Sequential Coupled Resonator Filter

3.1 Improvement to the Reflected Group Delay Method

In Chapter 2, the reflected group delay method proposed by Ness in 1998[1] was reviewed. This method introduces a routine of tuning and designing filter by successively adding resonator stage by stage. In each stage, design parameters are tuned to match its reflected group delay value at center frequency to a calculated value meanwhile the group delay response should be kept symmetric to the center frequency.

Recent research indicates the group delay curve is approximately symmetric only for filter fractional bandwidth less than 1% [7]. When tuning and designing a higher bandwidth EM based filter, [7][13] – [14] match the reflected group delay of EM based resonators to a pre-designed lumped element equivalent circuit in a circuit simulator. In this case, the entire reflected group delay is matched to the ideal response, which is asymmetric.

In this chapter, general mathematical formulas are proposed to calculate reflected group delay value at any frequency in each design stage based on the low-pass prototype. It will also be shown that instead of matching the entire reflected group curve, only a few frequency points are required to be matched in each design stage. By applying these formulas and the simplification of the matching configuration, circuit based filter design becomes much faster, an efficient computer-aided EM based filter design algorithm is also implemented.

In order to ascertain the ideal group delay curve at each stage, general formulas are required to be generated for the calculation of the group delay value for any frequency.
Based on formulas (2.22)-(2.27) mentioned in Chapter 2, the key is to generate the formulas to characterize the input impedance \( X_{in} \) in each design stage for all frequencies. The formulas to calculate \( X_{in} \) in terms of low-pass prototype \( g \) values are listed in Table 3.1

Table 3.1 Input impedance in terms of \( g \) values and frequency variables for each group delay stage based on the low-pass prototype circuit start with shunt capacitor

<table>
<thead>
<tr>
<th>Number of Resonators</th>
<th>( X_{in} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>( X_{in1} = -\frac{1}{\omega g_1} )</td>
</tr>
<tr>
<td>2</td>
<td>( X_{in2} = -\frac{g_2 \omega}{(\omega^2 - 1)} )</td>
</tr>
<tr>
<td>3</td>
<td>( X_{in3} = \frac{g_2 g_3 (\omega^1)^2 - 1}{\omega^1 - g_1 g_2 g_3 (\omega^1) + g_3 + g_1} )</td>
</tr>
<tr>
<td>4</td>
<td>( X_{in4} = \frac{g_2 g_3 (\omega^1)^2 + (g_2 + g_4)\omega}{g_3 g_4 (\omega^1) + g_3 g_2 (\omega^1)^2} - 1 )</td>
</tr>
<tr>
<td>5</td>
<td>( X_{in5} = \frac{(g_2 g_3 + g_2 g_4 + g_4 g_3) (\omega^1)^2 - g_1 g_2 g_3 (\omega^1)^4 - 1}{\omega (g_1 + g_3 + g_5 - g_1 g_2 g_3 (\omega^1) - g_2 g_3 g_5 (\omega^1)^2 - g_1 g_4 g_5 (\omega^1) - g_3 g_4 g_5 (\omega^1) - g_1 g_2 g_3 (\omega^1) + g_1 g_2 g_3 g_5 (\omega^1)^4)} )</td>
</tr>
</tbody>
</table>

By applying the input impedance into equation (2.28) where \( Z_0 = g_0 \), it is possible to obtain the general formulas to calculate the reflected group delay value in each stage for any given frequency. (These formulas can be found in Appendix A).
\[ \Gamma'_{\phi}(\omega) = 2 \frac{\omega^2 + \omega_0^2}{\omega^2 (\omega_2 - \omega_1)} \frac{\delta \tan^{-1} X_{\text{int}}(\omega')}{\delta \omega'} \frac{R_0}{\delta \omega'} \quad \text{where } i = 1, 2, \ldots, N \] (3.1)

In this case, instead of introducing an extra circuit model simulator, it is able to mathematically obtain the required reflected group values to match the entire response. When designing a filter, it is important to keep the number of frequency points to a minimum as the simulation time increases with an increase in the number of simulation points.

### 3.2 Reflected Group Delay in Band Pass Prototype

The reflected group delay method proposed by Ness is based on low-pass prototype model. In this section, a unique way to apply the reflected group delay method directly in sequential band-pass filter model is proposed.

A sequentially coupled resonator band pass filter [15] can be synthesized by the well-known ideal model given in Fig. 3.1.

![Fig. 3.1 Equivalent-circuit of the sequentially coupled band pass filter [15]](image)

Here \( M_{i,j} \) is the frequency independent reactance representing the couplings
between adjacent resonators. $R_1$ and $R_n$ are the equivalent resistances representing the input
and output couplings. The individual resonance frequency of the $i^{th}$ resonator is represented by:

$$\omega_{ui} = \frac{1}{\sqrt{L_i C_i}}$$

(3.2)

In a lossless one port band pass network given in Fig. 3.2, the reflected group delay is expressed as

$$\Gamma_d = -\frac{\partial \phi}{\partial \omega} - \frac{\partial \phi}{\partial \omega} = 2 \frac{\partial \tan^{-1} \left( \frac{X_\omega}{R_1} \right)}{\partial \omega}$$

(3.3)

Here $X_\omega$ is the band pass network input impedance which is purely imaginary for lossless case.

Fig.3.2 Equivalent circuit of the one-port network for each reflected group delay design stage.

Thus the reflected group delay response for a series of one port sub-networks in
terms of frequency independent reactance $M_{ij}$, resonance frequencies $\omega_0$ of each individual resonator and frequency variable $\omega$ for sequential coupled band pass filter can be generated from (3.3) and the formulas are listed in Table 3.2.

The proposed equations can be directly exploited to generate target reflected delay goals for a sequential band-pass filter network in terms of the adjacent couplings and resonance frequencies of individual resonators.
Table 3.2 Formulas to calculate Reflected group delay in terms of adjacent couplings and resonator resonance frequencies

<table>
<thead>
<tr>
<th>Resonator</th>
<th>$X_{in}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$X_{i1} = \frac{\omega}{\omega_{01}} - \frac{\omega_{01}}{\omega}$</td>
</tr>
<tr>
<td>1, 2</td>
<td>$X_{i2} = \frac{\omega}{\omega_{01}} - \frac{\omega_{01}}{\omega} - \frac{M_{12}^2}{\omega_{02}} - \frac{M_{12}^2}{\omega}$</td>
</tr>
<tr>
<td>1, 2, 3</td>
<td>$X_{i3} = \frac{\omega}{\omega_{01}} - \frac{\omega_{01}}{\omega} - \frac{M_{12}^2}{\omega_{02}} - \frac{M_{12}^2}{\omega_{03}} - \frac{M_{23}^2}{\omega}$</td>
</tr>
<tr>
<td>1, 2, 3, 4</td>
<td>$X_{i4} = \frac{\omega}{\omega_{01}} - \frac{\omega_{01}}{\omega} - \frac{M_{12}^2}{\omega_{02}} - \frac{M_{12}^2}{\omega_{03}} - \frac{M_{12}^2}{\omega_{04}} - \frac{M_{23}^2}{\omega_{03}} - \frac{M_{23}^2}{\omega_{04}} - \frac{M_{34}^2}{\omega}$</td>
</tr>
<tr>
<td>1, 2, 3, 4, 5</td>
<td>$X_{i5} = \frac{\omega}{\omega_{01}} - \frac{\omega_{01}}{\omega} - \frac{M_{12}^2}{\omega_{02}} - \frac{M_{12}^2}{\omega_{03}} - \frac{M_{12}^2}{\omega_{04}} - \frac{M_{12}^2}{\omega_{05}} - \frac{M_{23}^2}{\omega_{03}} - \frac{M_{23}^2}{\omega_{04}} - \frac{M_{23}^2}{\omega_{05}} - \frac{M_{34}^2}{\omega_{04}} - \frac{M_{34}^2}{\omega_{05}} - \frac{M_{45}^2}{\omega}$</td>
</tr>
</tbody>
</table>
3.3 Application of Improved Reflected Group Delay to Filter Designs

In this part the required poles and zeros frequency locations and group delay values can easily be obtained by solving the derivative of the reflected group delay function from Table 3.1 and equation (3.1). An example of target goals established for designing a 12-pole Chebyshev band pass filter is calculated and applied to different types of filter structures in the next section.

3.3.1 Example of Calculated Goals for Design of a 12-pole Chebyshev Band-pass Filter

A set of target reflected group delay goals for design of a 12-pole Chebyshev band-pass filter with center frequency of 1 GHz, fractional bandwidth of 5% and ripple of 0.01dB is calculated. The required simulation frequency points and corresponding objective group delay values for each design stage are listed in Table 3.3.

Please note the goals are calculated based on purely mathematical low-pass prototype model, thus they can be applied to any filter topology and bandwidth specifications. In this section the same goals are applied to three different types of filter structures. They are end-coupled microstrip filter structure, parallel-coupled microstrip filter structure and hairpin microstrip filter structure. Two designs using different dielectric materials and substrate configurations are carried out for the hairpin examples exploiting the same goals. All the designs achieve the desired simulation results.
Table 3.3 Target reflected group delay goals of each design stage for design of a 12-pole Chebyshev band-pass filter

<table>
<thead>
<tr>
<th>Resonators</th>
<th>1</th>
<th>1,2</th>
<th>1,2,3</th>
<th>1,2,3,4</th>
<th>1,2,3,4,5</th>
<th>1,2,3,4,5,6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Response</td>
<td><img src="image1.png" alt="Graph" /></td>
<td><img src="image2.png" alt="Graph" /></td>
<td><img src="image3.png" alt="Graph" /></td>
<td><img src="image4.png" alt="Graph" /></td>
<td><img src="image5.png" alt="Graph" /></td>
<td><img src="image6.png" alt="Graph" /></td>
</tr>
<tr>
<td>((f(GHz), \Gamma d^f_{\text{goal}}(\text{ns})))</td>
<td>(0.999,10.526)</td>
<td>(1.016,24.179)</td>
<td>(1.021,49.996)</td>
<td>(1.023,72.969)</td>
<td>(1.024,108.291)</td>
<td>(1.019,89.811)</td>
</tr>
<tr>
<td></td>
<td>(0.984,24.971)</td>
<td>(9.912,31.247)</td>
<td>(1.40772)</td>
<td>(1.017,66.345)</td>
<td>(1.018,90.18)</td>
<td></td>
</tr>
<tr>
<td></td>
<td>(1.18,457)</td>
<td>(9.793,46.919)</td>
<td>(0.99,49.283)</td>
<td>(1.006,53.34)</td>
<td>(0.981,93.326)</td>
<td></td>
</tr>
<tr>
<td></td>
<td>(0.998,33.925)</td>
<td>(0.999,58.988)</td>
<td>(0.993,73.988)</td>
<td>(0.981,93.326)</td>
<td>(0.981,93.326)</td>
<td></td>
</tr>
<tr>
<td></td>
<td>(1.009,30.693)</td>
<td>(1.01,4.834)</td>
<td>(0.994,54.022)</td>
<td>(1.011,68.69)</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(1.143,4.663)</td>
<td>(0.977,113.561)</td>
<td>(0.976,159.077)</td>
<td>(0.976,159.077)</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(0.986,47.959)</td>
<td>(0.985,69.269)</td>
<td>(0.982,93.51)</td>
<td>(0.982,93.51)</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(0.983,68.665)</td>
<td>(1.025,151.528)</td>
<td>(1.006,73.029)</td>
<td>(1.006,73.029)</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(1.015,67.225)</td>
<td>(0.989,70.19)</td>
<td>(1.63,719)</td>
<td>(1.63,719)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(\Gamma d^f_{\text{goal}}(f = 0.975GHz))</td>
<td>6.3482</td>
<td>19.115</td>
<td>38.476</td>
<td>64.791</td>
<td>98.759</td>
<td>141.718</td>
</tr>
<tr>
<td>(\Gamma d^f_{\text{goal}}(f = 1.025GHz))</td>
<td>6.162</td>
<td>18.791</td>
<td>38.49</td>
<td>66.12</td>
<td>102.67</td>
<td>148.852</td>
</tr>
</tbody>
</table>
3.3.2 Application of the Goals to a 12-pole End-coupled Microstrip Band-pass Filter

The calculated goals listed in Table 3.3 are applied to the design of an end-coupled microstrip band-pass filter in Keysight ADS. Since the filter structure is designed to be physically symmetric, this design consists of 6 reflected group delay design stages and an additional stage to fix the magnitude response of the entire filter. Detailed design parameters for each reflected group delay stage are listed in Table 3.4.

By successively adding resonators and optimizing corresponding parameters to match a few points in the group delay curve, the input/output couplings and internal coupling between resonators are well characterized. The optimizing time is significantly reduced to less than 20 seconds for each design stage because of the reduction of simulation points. The entire filter schematic is shown in Fig. 3.3. The final simulation result is shown in Fig. 3.4. This design example demonstrates the accuracy of this improvement.

Fig. 3.3 Circuit schematic for the 12-pole hairpin end-coupled filter
Table 3.4 Design Parameters of a 12-pole end-coupled bandpass filter in Keysight ADS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>N=1</th>
<th>N=2</th>
<th>N=3</th>
<th>N=4</th>
<th>N=5</th>
<th>N=6</th>
<th>N=12</th>
<th>Entry filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>C01 (pf)</td>
<td>1.07664</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>1.07664</td>
</tr>
<tr>
<td>L1 (mil)</td>
<td>2825.23</td>
<td>2757.68</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>2757.68</td>
</tr>
<tr>
<td>C12 (pf)</td>
<td>0.229024</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>0.229024</td>
</tr>
<tr>
<td>L2 (mil)</td>
<td>3051.38</td>
<td>3005.5</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>3005.5</td>
</tr>
<tr>
<td>C23 (pf)</td>
<td>0.154855</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>0.154855</td>
</tr>
<tr>
<td>L3 (mil)</td>
<td>3074.54</td>
<td>3033.25</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>3033.25</td>
</tr>
<tr>
<td>C34 (pf)</td>
<td>0.14051</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>0.14051</td>
</tr>
<tr>
<td>L4 (mil)</td>
<td>3078.6</td>
<td>3039.28</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>3039.28</td>
</tr>
<tr>
<td>C45 (pf)</td>
<td>0.135845</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>0.135845</td>
</tr>
<tr>
<td>L5 (mil)</td>
<td>3079.4</td>
<td>3041.29</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>3041.29</td>
</tr>
<tr>
<td>C56 (pf)</td>
<td>0.13386</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>0.13386</td>
</tr>
<tr>
<td>L6 (mil)</td>
<td>3079.59</td>
<td>3042.13</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>3042.13</td>
</tr>
<tr>
<td>C67 (pf)</td>
<td>0.133345</td>
<td>0.133345</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>0.133345</td>
</tr>
</tbody>
</table>
Fig. 3.4 ADS simulation result of the 12-pole end-coupled bandpass filter
3.3.3 Application of the Goals to a 12-pole Parallel-coupled Band-pass Filter

To validate that the calculated goals can be applied to different filter structures, a parallel-coupled band pass filter is designed in Keysight ADS using the exact same goals used in the end-coupled filter design. The entire design consists of 6 reflected group delay stages and one final stage to adjust the final magnitude response for the entire filter. Detailed design parameters for each design stage are listed in Table 3.5. The entire filter schematic is shown in Fig. 3.5 and the final simulation response is given in Fig. 3.6.

![Fig. 3.5 Circuit schematic for the 12-pole parallel-coupled bandpass filter](image)
Table 3.5 Design Parameters of a 12-pole parallel-coupled Bandpass filter in Keysight ADS

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>8.83594</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>8.83594</td>
</tr>
<tr>
<td>L1</td>
<td>283.159</td>
<td>257.849</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>257.849</td>
</tr>
<tr>
<td>S12</td>
<td>102.225</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>102.225</td>
</tr>
<tr>
<td>L2</td>
<td>295.051</td>
<td>270.101</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>270.101</td>
</tr>
<tr>
<td>S23</td>
<td></td>
<td>137.494</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>137.494</td>
</tr>
<tr>
<td>L3</td>
<td>295.214</td>
<td>271.184</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>271.184</td>
</tr>
<tr>
<td>S34</td>
<td></td>
<td>148.253</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>148.253</td>
</tr>
<tr>
<td>L4</td>
<td></td>
<td></td>
<td>271.914</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>271.914</td>
</tr>
<tr>
<td>S45</td>
<td></td>
<td></td>
<td>151.833</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>151.833</td>
</tr>
<tr>
<td>L5</td>
<td></td>
<td></td>
<td></td>
<td>271.653</td>
<td></td>
<td></td>
<td></td>
<td>271.653</td>
</tr>
<tr>
<td>S56</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>153.604</td>
<td></td>
<td></td>
<td>153.604</td>
</tr>
<tr>
<td>L6</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>271.909</td>
<td></td>
<td>271.909</td>
</tr>
<tr>
<td>S67</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>154.055</td>
<td>154.055</td>
</tr>
</tbody>
</table>
Fig. 3.6 ADS simulation result of the 12-pole parallel-coupled band-pass filter
3.3.4 Application of the Goals to a 12-pole Hairpin Band-pass Filter

A hairpin band pass filter is also designed in Keysight ADS using the exact same goals from the previous two designs. The entire design consists of 6 reflected group delay stages and one final stage to adjust the final magnitude response for the entire filter. Detailed design parameters for each design stage are listed in Table 3.6. The entire filter schematic is shown in Fig. 3.7 and the final simulation response is given in Fig. 3.8.

![Fig. 3.7 Circuit schematic for the 12-pole hairpin bandpass filter](image)
Table 3.6 Design Parameters of a 12-pole hairpin Bandpass filter in Keysight ADS

<table>
<thead>
<tr>
<th>ER=4.8</th>
<th>H=62 mil</th>
<th>W=109 mil</th>
<th>coupled line 1350 mil</th>
</tr>
</thead>
<tbody>
<tr>
<td>Parameters</td>
<td>N=1</td>
<td>N=2</td>
<td>N=3</td>
</tr>
<tr>
<td>units (mil)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Tin</td>
<td>1247.89</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>L1</td>
<td>180.105</td>
<td>139.863</td>
<td>-</td>
</tr>
<tr>
<td>S12</td>
<td>100.282</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>L2</td>
<td>168.782</td>
<td>136.684</td>
<td>-</td>
</tr>
<tr>
<td>S23</td>
<td>134.648</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>L3</td>
<td>161.624</td>
<td>131.518</td>
<td>-</td>
</tr>
<tr>
<td>S34</td>
<td>144.293</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>L4</td>
<td>160.055</td>
<td>131.128</td>
<td>-</td>
</tr>
<tr>
<td>S45</td>
<td>147.945</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>L5</td>
<td>159.023</td>
<td>131.147</td>
<td>-</td>
</tr>
<tr>
<td>S56</td>
<td>149.392</td>
<td>149.392</td>
<td>-</td>
</tr>
<tr>
<td>L6</td>
<td>157.936</td>
<td>159.599</td>
<td>159.599</td>
</tr>
<tr>
<td>S56</td>
<td>149.823</td>
<td>149.823</td>
<td>-</td>
</tr>
</tbody>
</table>
Fig. 3.8 ADS simulation result of the 12-pole hairpin band-pass filter
### 3.3.5 Application of the Goals to a 12-pole Hairpin Band-pass Filter with Different Dielectric Material and Substrate Configurations

In this part, a hairpin band pass filter with a different dielectric material and substrate configurations is designed in Keysight ADS using the exact same goals from the previous design. In the previous hairpin design, the dielectric material is chosen to be FR4 with an expected dielectric constant of 4.8 and a substrate height of 62 mil. In this design example, the dielectric material is changed to alumina with expected dielectric constant of 10.2 and substrate height of 25 mil.

For microstrip filter, the selection of dielectric material and substrate height are of the most importance to the design process. The dielectric constant of the material and height of substrate layer can significantly affect the characteristic impedance and electrical length of the resonator at a specified resonance frequency. For 25 mil aluminum, the selected physical width of the resonator is 23 mil which leads to a \(50\,\Omega\) characteristic impedance. The physical lengths of the coupled lines are set to 950 mil which is close to 90 degree of electrical length. For 64 mil FR-4, the physical width of each resonator is set to 109 mil to give a characteristic impedance of \(50\,\Omega\). The physical lengths of coupled lines are set to 1300 mil which is close to 90 degree electrical length.

The calculation of target reflected group delay goals are directly from filter specifications and does not related to material and substrate settings. Thus in this example the goals for each design stage is exactly the same from Table 3.3.

The entire design consists of 6 reflected group delay stages and one final stage to adjust the final magnitude response for the entire filter. Detailed design parameters for each design stage are listed in Table 3.7. The entire filter schematic is shown in Fig. 3.9 and the
final simulation response is given in Fig. 3.10.

Fig. 3.9 Circuit schematic for the 12-pole hairpin bandpass filter using alumina
Table 3.7 Design Parameters of a 12-pole hairpin Bandpass filter using alumina in Keysight ADS

<table>
<thead>
<tr>
<th>Parameters</th>
<th>N=1</th>
<th>N=2</th>
<th>N=3</th>
<th>N=4</th>
<th>N=5</th>
<th>N=6</th>
<th>N=12</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>units (mil)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Tin</td>
<td>906.6</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>906.6</td>
</tr>
<tr>
<td>L1</td>
<td>297.767</td>
<td>279.409</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>279.409</td>
</tr>
<tr>
<td>S12</td>
<td>39.0219</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>39.0219</td>
</tr>
<tr>
<td>L2</td>
<td>299.9</td>
<td>286.915</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>286.915</td>
</tr>
<tr>
<td>S23</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>50.2902</td>
</tr>
<tr>
<td>L3</td>
<td>294.08</td>
<td>282.441</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>282.441</td>
</tr>
<tr>
<td>S34</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>53.3978</td>
</tr>
<tr>
<td>L4</td>
<td>292.618</td>
<td>281.863</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>281.863</td>
</tr>
<tr>
<td>S45</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>54.5828</td>
</tr>
<tr>
<td>L5</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>281.784</td>
</tr>
<tr>
<td>S56</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>55.0442</td>
</tr>
<tr>
<td>L6</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>281.7</td>
<td>281.7</td>
<td>281.7</td>
<td>281.7</td>
</tr>
<tr>
<td>S67</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>55.2083</td>
<td>55.2083</td>
<td>55.2083</td>
<td>55.2083</td>
</tr>
</tbody>
</table>
The theory in Section 3.1 is for design of sequentially coupled resonator filter without cross couplings. The four applications are ideal filter models without inherent cross couplings. It is obvious that the calculated goals are only related to filter specifications and can be used for the design of different filter structures with different types of materials. The proposed improvement to the original reflected group delay method is proven to be efficient and accurate enough to carry out Chebyshev filter design problems.
Chapter 4

Reflected Group Delay Method with Aggressive Space Mapping

4.1 Introduction

The improvement to the reflected group delay method described in Section 3.1 introduced a way to minimize the required simulation points such that each single simulation becomes faster. Another big issue in the EM based filter design is the choice of optimizing techniques and the corresponding convergent iterations.

The target of optimization in a filter design is to obtain the optimum set of values for the designed filter parameters such that the desired filter specifications can be satisfied. Traditional optimizing techniques apply error functions and numerical algorithms directly on the analysis of the filter response and design parameters. For many complex filter structures, even if a single simulation time is well reduced, the number of optimizing iterations for a basic optimizing technique can still be very high while convergence is not guaranteed. Higher level optimizing techniques often require the derivative of the response parameters in terms of physical design parameters to be available. For many EM based simulators, the system matrix generated during the simulation is not available or hard to analyze. Practically, the choices of EM based optimizing techniques are very limited. It becomes a sub-problem in filter design and tuning.

As reviewed in Chapter 2, the space mapping technique [2] introduced by Bandler in 1994 has been developed and proven to be very efficient in microwave engineering optimizing problems. In the recent years, more and more research [16]-[20] has been carried out on various space mapping and related studies.
Space mapping techniques aim at establishing a mapping between the spaces of a computational-expensive but accurate EM based fine model and a faster but low accuracy circuit base coarse model. In this way, the expensive optimizing processes are directed to the faster coarse model while the accuracy is confirmed by the accurate fine model. By applying the space mapping technique, a complex EM based optimizing problem is able to be solved within a few EM based simulations.

A significant stage in the space mapping technique is parameter extraction (PE), in which the design parameters of coarse model are optimized to match the simulated response from the fine model. The non-uniqueness of the PE process has become a sub-problem in the space mapping research area. This non-uniqueness can lead to divergence or oscillations of the space mapping iterations. Research has been carried out to solve the non-uniqueness of parameter extraction [21] - [25].

Instead of modifying the algorithm or introducing limitations to the PE process, an easier way to solve the non-uniqueness problem is to reduce the number of optimizing parameters. The reflected group delay method provides a design procedure in which the entire filter design is divided into smaller stages. In each stage, the number of parameters is limited to 2 to 4. Thus, by implementing with the space mapping technique, the number of EM simulations can be significantly reduced compared to direct EM based optimizing. Meanwhile, the non-uniqueness problem of parameter extraction is removed because of the reduced number of parameters in the reflected group delay method.

In this chapter, an EM-based design approach using the reflected group delay method with aggressive space mapping technique is proposed. The design procedure is presented. An example of a five pole hairpin band pass filter is given as an example.
4.2 Reflected Group Delay Method with Aggressive Space Mapping Design Theory

The reflected group delay (RGD) method has been well proven in [1],[7],[13]-[14] to be an efficient filter design method by dividing an entire filter design into smaller stages to simplify the design process. In each design stage, a resonator is added successively to match a calculated objective reflected group delay response. For sequentially coupled filters with negligible inherent cross coupling, most parameters determined in one stage are keep constant and do not take part in later optimizing stages. Thus, the optimization problem in any stage $i$ using the reflected group delay method can be expressed as

$$\{ x'_i \} = \arg \min_{x'_i} \{ \Gamma d (x'_i) - \Gamma d'_{\text{goal}} \}$$

(4.1)

Here $x'_i$ is the set of design parameters and $\Gamma d (x'_i)$ denotes the simulated reflected group delay response. $\Gamma d'_{\text{goal}}$ is the objective reflected group delay values of a corresponding low-pass prototype circuit model calculated from the theory proposed in Chapter 3 or directly simulated from an ideal lumped-element circuit. The target design parameter values $\{ x'_i \}$ are obtained through direct tuning and optimizing techniques until the simulated group delay response matches the calculated objective curve.

It is important to note that the calculated reflected group delay from low pass prototype is purely mathematical in terms of basic filter specifications (filter order, center frequency, return loss, bandwidth, and characteristic impedance). Thus, it does not have filter structure limitations or design bandwidth limitations. As long as the calculated group delay objectives can be achieved, the designed filter specifications can be certainly achieved in any sequentially coupled filter structure.
When designing an actual filter, the EM based optimizing is still computationally expensive. It takes many iterations to achieve a match in a single design stage. By implementing the space mapping technique, the number of optimizing iterations is significantly reduced such that a fast convergence to the objective group delay is able to be achieved within 2 or 3 EM simulations.

For sequentially coupled filters, the use of space mapping and the reflected group delay method can show a great improvement in computation time over the traditional space mapping technique, as the individual design stages require a smaller number of parameters to be optimized. This can reduce the computation time for parameter extraction and reduce the time to achieve the desired filter response.

4.3 Reflected Group Delay Method with Aggressive Space Mapping Design Procedure

The design of a symmetric filter of order $N$ is first divided into $N/2$ (even order) or $(N-1)/2$ (odd order) design stages where the filter is optimized such that the reflected group delay meets the ideal response. An additional design stage using traditional ASM where a few required parameters are optimized on the entire filter to meet the desired filter response is required. In each design stage, aggressive space mapping (ASM) iterations are applied to achieve faster convergence.

For the proposed design procedure, the design parameter sets of the coarse and fine models in the $i^{th}$ RGD stage are defined as $x_{i}^c, x_{i}^f$. The optimum solutions to the design parameter sets in the coarse and fine models for the $i^{th}$ RGD stage are defined as $\{x_{i}^{c'}\}, \{x_{i}^{f'}\}$. The parameters in the previous $k$ RGD stages that keep optimum constant
values in the $i^{th}$ stages are defined as $\{x_c^{k \text{ const}}\}_i, \{x_f^{k \text{ const}}\}_i$ where $k = [1, 2, ..., i - 1]$. The values of design parameters obtained in the $i^{th}$ design stage and $j^{th}$ space mapping iteration are defined as $\{x_c^{i,j}\}_i, \{x_f^{i,j}\}_i$. The reflected group delay responses in coarse and fine models are defined as $\Gamma d_c, \Gamma d_f$. The magnitude response denoted by $R$ is used in the final design stage.

By integrating the reflected group delay method with the space mapping technique, the entire design problem can be achieved by completing the following problem, stage by stage:

$$\{x_f^{(*)}\} = \arg \min_{x_f} \left\{ \Gamma d_f \left( x_c^{k \text{ const}}, x_f^{i}\right) - \Gamma d_c \left( x_c^{k \text{ const}}, x_f^{(*)}\right) \right\} \quad (4.2)$$

After all group delay objectives are achieved, the final design stage uses the traditional aggressive space mapping technique to achieve the desired magnitude response.

The proposed design procedure is:

Step 1: Add the first resonator in the coarse model and optimize $x_c^1$ to match $\Gamma d_c^1$ using

$$\{x_c^{1,*}\} = \arg \min_{x_c} \left\{ \Gamma d_c \left( x_c^{1}\right) - \Gamma d_c^1 \right\} \quad (4.3)$$

Step 2: Set $x_f^1 = x_c^{1,*}$ and record the simulated response $\Gamma d_f \left( x_f^{1}\right)$. 

Step 3: Use parameter extraction technique to get $\{x_c^{i,j}\}$ using

$$\{x_c^{i,j}\} = \arg \min_{x_c} \left\{ \Gamma d_c \left( x_c^{i}\right) - \Gamma d_f \left( x_f^{i,j}\right) \right\} \quad (4.4)$$

Step 4: Approximate the next fine model parameter $\{x_f^{i+1}\}$ by ASM. If $\Gamma d_f \left( x_f^{i+1}\right)$ converges to the objective reflected group delay response, it is assigned to be the optimum solution $\{x_f^{(*)}\} = \{x_f^{i+1}\}$, otherwise, repeat steps 3 and 4.
Step 5: Denote the parameters required to be optimized in the \( i^{th} \) stage to be \( x^{i}_{(c,f)} \). If there are shared parameters with the previous \( k \) stages, the values of shared parameters from previous stages are ignored. The rest of the parameters are kept constant and defined as \( x^{k,\text{const}}_{(c,f)} \).

Step 6: Optimize \( x^{i} \) until the optimum solution to reflected group delay objective is obtained using

\[
\{x^{i*}_{c}\} = \text{arg min}_{x^{i}_{c}} \left\{ \Gamma d \left( x^{k,\text{const}}_{c}, x^{i}_{c}\right) - \Gamma d^{i}_{\text{goal}} \right\}
\] (4.5)

Step 7: Set \( x^{i,1}_{f} = x^{i*}_{c} \) and follow step 3 and step 4 until the reflected group delay response in the fine model achieves the objective curve using

\[
\{x^{f*}_{c}\} = \text{arg min}_{x^{i}_{c}} \left\{ \Gamma d \left( x^{k,\text{const}}_{c}, x^{i}_{c}\right) - \Gamma d^{i}_{\text{goal}} \left( x^{k,\text{const}}_{c}, x^{f*}_{c}\right) \right\}
\] (4.6)

Repeat steps 5 to 7 until all required reflected group delay design stages are completed.

Step 8: Repeat step 5 to obtain \( x^{k,\text{const}}_{c} \). A final optimization to adjust \(|S_{11}|\) and \(|S_{21}|\) is carried out in coarse model using

\[
\{x^{\text{final,}*}_{c}\} = \text{arg min}_{x^{\text{final}}_{c}} \left\{ R \left( x^{k,\text{const}}_{c}, x^{\text{final}}_{c}\right) - R_{\text{specification}} \right\}
\] (4.7)

The aggressive space mapping technique is applied to obtain the fine model optimum solution using

\[
\{x^{\text{final,}*}_{f}\} = \text{arg min}_{x^{\text{final}}_{f}} \left\{ R \left( x^{k,\text{const}}_{f}, x^{\text{final}}_{f}\right) - R \left( x^{k,\text{const}}_{f}, x^{\text{final}}_{f}\right) \right\}
\] (4.8)
4.4 Application to a 5-pole Microstrip Filter

To validate the procedure proposed in Section 4.3, a 5-pole Chebyshev microstrip hairpin bandpass filter is designed, fabricated, and tested. The filter is specified with a center frequency of 1GHz and a fractional bandwidth of 15%. A dielectric material with an expected relative permittivity of 4.8 and a substrate thickness of 62 mil is used in this design.
Fig. 4.1. Coarse Model Schematic ($L_n = L - L_{in}$): (a) first resonator (steps 1 and 3) (b) first two resonators (step 6) (c) entire filter (step 8).

The coarse model in Fig. 4.1 is implemented with the Keysight ADS circuit simulator. The fine model is simulated using the EM simulator, Sonnet. The filter topology is symmetric such that the entire design is divided into 2 stages for adjusting group delay and
1 final stage to adjust the magnitude response. According to the theory proposed in Chapter 3, the required target reflected group delay goals are calculated and given in Table 4.1.

Table 4.1 The target reflected group delay goals of each stage for design of a filter with center frequency of 1GHz, fractional bandwidth of 15%

<table>
<thead>
<tr>
<th>Stage</th>
<th>( \Gamma d^1_{\text{goal}} )</th>
<th>( \Gamma d^2_{\text{goal}} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Selected points</td>
<td>0.99505, 3.2179</td>
</tr>
<tr>
<td></td>
<td>BW edge points</td>
<td>(0.925, 2.1495)</td>
</tr>
<tr>
<td>2</td>
<td>Selected points</td>
<td>0.94608, 7.8976</td>
</tr>
<tr>
<td></td>
<td>BW edge points</td>
<td>(0.925, 6.721)</td>
</tr>
</tbody>
</table>

The input tapped positions \( L_m \), resonator lengths \( \{ L_1, L_2, L_3 \} \) and gaps between resonators \( \{ G_{12}, G_{23} \} \) are the design parameters of the filter. The geometry layouts for each design stage can be found in Fig. 4.2.

In the first design stage, the input line and first resonator are implemented in ADS, step 1 described in the Section 4.3 is completed and \( \{ x^1 \} \) is obtained by optimizing the coarse model design parameters \( x^1 = \{ L, L_m, L_1 \} \) until the response \( \Gamma d_{x^1} \) matches the objective group delay curve, \( \Gamma d_{\text{goal}} \). The coarse model optimizing results are given in Table 4.2.

Step 2 is completed where the values \( \{ x^{1,1} \} \) are set to be equal to \( \{ x^1 \} \). Aggressive space mapping is applied to approximate the fine model parameters by repeating step 3 and
step 4 until convergence is achieved. The acceptable fine model parameters \( \{ x_r^* \} \) are reached within 2 iterations.

Fig. 4.2. Sonnet Layout: (a) first resonator (step 2 and 4) (b) first two resonators (step 7) (c) entire filter (step 8).
In the second design stage, the second resonator is added. The design parameters \( x^2 = \{ L_1, L_2, G_{12} \} \) are optimized until the reflected group delay response \( \Gamma d_e (x^{2*}) \) matches the calculated objective curve. The coarse model optimizing results are given in Table 4.2. In this step, \( L_i \) is required to be re-optimized. Thus the parameters that remain constant are \( x^{1, const} = \{ L, L_{iw} \} \). After 2 iterations the optimum set of the fine model parameters \( \{ x^{2*}_f \} \) is obtained. The detailed space mapping iterations and results are given in Table 4.3 and Fig. 4.2.
Table 4.2 Coarse model optimizing results of each reflected group delay design stage

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Ltotal (mil)</th>
<th>Tin (mil)</th>
<th>L1 (mil)</th>
<th>S12 (mil)</th>
<th>L2 (mil)</th>
<th>S23 (mil)</th>
<th>L3 (mil)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coarse Model</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{1,*}_{1,1}$</td>
<td>1300</td>
<td>948.322</td>
<td>235.349</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{2,*}_{2,1}$</td>
<td>--</td>
<td>--</td>
<td>141.885</td>
<td>20.6475</td>
<td>302.916</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{3,*}_{3,1}$</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>247.291</td>
<td>36.1975</td>
<td>231.068</td>
</tr>
<tr>
<td>ADS Optimum Dimensions</td>
<td>1300</td>
<td>948.322</td>
<td>141.885</td>
<td>20.6475</td>
<td>247.291</td>
<td>36.1975</td>
<td>231.068</td>
</tr>
</tbody>
</table>

Table 4.3 Space Mapping Iterations in each RGD stage & Final Optimum results

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Ltotal (mil)</th>
<th>Tin (mil)</th>
<th>L1 (mil)</th>
<th>S12 (mil)</th>
<th>L2 (mil)</th>
<th>S23 (mil)</th>
<th>L3 (mil)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1st Resonator</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{1,1}_{1,1}$</td>
<td>1300</td>
<td>948.322</td>
<td>235.349</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{1,2}_{1,2}$</td>
<td>1464.262</td>
<td>889.2362</td>
<td>243.74954</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{1,*}_{2,1}$</td>
<td>1381.6427</td>
<td>890.4308</td>
<td>238.9702</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1st &amp; 2nd Resonators</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{2,1}_{2,2}$</td>
<td>-</td>
<td>-</td>
<td>141.885</td>
<td>20.6475</td>
<td>302.916</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{2,2}_{2,2}$</td>
<td>-</td>
<td>-</td>
<td>207.4565</td>
<td>15.1808</td>
<td>317.7748</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{2,*}_{2,2}$</td>
<td>-</td>
<td>-</td>
<td>189.6142</td>
<td>16.83911</td>
<td>313.7765</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Full Filter</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$x^{3,1}_{3,1}$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>247.291</td>
<td>36.1975</td>
<td>231.068</td>
</tr>
<tr>
<td>$x^{3,2}_{3,2}$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>289.0164</td>
<td>33.2824</td>
<td>279.09</td>
</tr>
<tr>
<td>$x^{3,*}_{3,2}$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>286.13</td>
<td>32.9806</td>
<td>272.1347</td>
</tr>
<tr>
<td>Sonnet Optimum Dimensions</td>
<td>1381.6427</td>
<td>890.4308</td>
<td>189.6142</td>
<td>16.83911</td>
<td>286.13</td>
<td>32.9806</td>
<td>272.1347</td>
</tr>
</tbody>
</table>
Fig. 4.3: Fine Model Reflected Group Delay response for each space mapping iteration. (a) First resonator. (b) First and Second Resonators. The cross lines are the results of first space mapping iteration, the circle lines are the second iteration, the solid lines are the optimum response.
At this point, all reflected group delay goals have been achieved. The entire filter is implemented for optimizing to reach the required magnitude response. The optimized parameters are $x^{\text{final}}_{(r,f)} = \{L_2, L_3, G_{23}\}$. In this stage, the constant parameter set is $x^{2, \text{const}}_{(r,f)} = \{L, L_{in}, L_{1}, G_{12}\}$. The fine model simulation results achieved return loss of 20 dB after 1 iteration and a convergence to the design specifications within 2 parameter extraction processes and 3 EM simulations. Detailed iterations and results are given in Table 4.1 and Fig. 4.4. The image of fabricated filter is given in Fig 4.5. The measured results for the filter are also shown in Fig. 4.6.

![Graph](image)

**Fig.4.4.** EM simulation return loss of the first iteration (dotted line), second iteration (solid line) and final filter design (bold line)
Fig. 4.5. The fabricated microstrip 5-pole hairpin bandpass filter

Fig. 4.6 Measured results of fabricated filter.
For this filter design, the number of optimized parameters is reduced to 3 for each design stage due to the introduction of the RGD method, which reduced the PE time during any iteration to be less than one minute. Thus, by integrating the RGD method with the SM technique, the PE process for each design stage can be achieved within a very short time with a high likelihood of convergence.

For this design, the PE computation times for all stages and the EM simulation times for the group delay stages were small compared to the EM simulation times of the final design stage, which makes the total final stage EM simulation times the approximate optimization time for the whole design procedure. The proposed design procedure takes much less computation time than the traditional space mapping approach as PE times are reduced significantly and as the two approaches share similar EM simulation times.
Chapter 5 A Sequentially Coupled Filter EM based Design Approach
Using the Reflected Group Delay Method and the Implicit Space Mapping Technique

5.1 Introduction

In this chapter, a computer-aided technique is presented for the reflected group delay stages where very few selected frequency points are required instead of the whole curve for each design stage. Frequency sweep plans and optimization goals can be easily determined. In this way, optimizing time is significantly reduced as a result of the reduced number of simulation points.

This chapter proposes a filter design approach that integrates the reflected group delay method with the implicit space mapping technique where implicit space mapping is applied to each reflected group delay stage as an optimizing method. In each implicit space mapping iteration, selected pre-assigned parameters are extracted to match the coarse model to the fine model. Then, the calibrated coarse model is re-optimized to satisfy the group delay goals at selected frequency points. The optimized parameters are then assigned to the fine model. The use of the sweep plan and goals with reduced coarse model simulation points makes the optimization much faster and improves the chance for convergence.

The theory of the proposed method and design procedure are given in Section 5.2. Section 5.3 presents the application of the proposed design approach to a 6-pole Chebyshev hairpin microstrip filter.
5.2 Proposed Reflected Group Delay Method and the Implicit Space Mapping Design Approach

The proposed method presents an efficient technique to match the objective group delay curve by matching a limited number of selected frequency points. For a sequentially coupled filter with a Chebyshev response, the number of peaks and valleys of the target reflected group delay curves are related to the number of resonators included in the circuit. $2n + 1$ reflected group delay points are required for the $n^{th}$ design stage, a point at each band edge and a point at each peak and valley. Thus, in a filter design, a frequency sweep plan and corresponding group delay goals are built up based on calculations Table 3.1 and equation (3.1). Few frequency points are required to be simulated using the coarse model in each design stage. The space mapping optimizing time can be significantly reduced when implementing the proposed design procedure.

In each reflected group delay design stage, the implicit space mapping technique is applied to achieve a fast match of the coarse model response to the reflected group delay goals at selected frequency points to determine the optimum EM based design parameters. The set of design parameters in the coarse and fine models in the $i^{th}$ reflected group delay stage are denoted by $x'_i$ and $x'_f$. The pre-assigned parameter set for the $i^{th}$ stage is denoted by $p'_i$. The set of optimum solutions of the two models are denoted by $\{x'^*_i\}$ and $\{x'^*_f\}$.

The proposed design procedure is as follows:

Step 1: Calculate selected frequencies and group delay goals from the design specifications.
Build up the frequency sweep plan and optimizing goals for each design stage.

Step 2: Obtain all coarse model optimum parameters $\{x'^*_i\}$ in $i^{th}$ reflected group delay stage using the sweep plan and equation (3.1)
\[
\{ x^i_c \} = \arg \min_{x^i_c} \left\{ \Gamma d_c \left( x^i_c \right) - \Gamma d^i_{goal} \right\}
\]  
(5.1)

Step 3: Obtain optimum solutions for fine model parameters in the \( i^{th} \) reflected group delay stage that give the same reflected group delay response as the corresponding coarse model and the implicit space mapping algorithm given in equations (5.2) to (5.4)

\[
\{ p^{i,j+1} \} = \arg \min_{p^{i,j+1}} \left\{ \Gamma d_c \left( x^{i,j}_c, p^{i,j} \right) - \Gamma d_f \left( x^{i,j}_f \right) \right\}
\]  
(5.2)

\[
\{ x^{i,j+1}_c \} = \arg \min_{x^{i,j+1}_c} \left\{ \Gamma d_c \left( x^{i,j}_c, p^{i,j+1} \right) - \Gamma d^i_{goal} \right\}
\]  
(5.3)

\[
\{ x^{i,j+1}_f \} = \{ x^{i,j+1}_c \}
\]  
(5.4)

Step 4: After performing steps 1-3 for half of the filter structure, implementing the entire filter and adjusting the remaining few parameters in the coarse model to achieve the desired magnitude response. Then run implicit space mapping on these parameters to obtain the fine model optimum solution.

**5.3 Application**

To validate the method presented in Section 5.2, a 6-pole microstrip hairpin Chebyshev band-pass filter is designed in Sonnet full-wave EM simulator. The filter is specified with a center frequency of 1.5 GHz, a fractional bandwidth of 15% and a return loss better than -20dB. The dielectric material is chosen to be alumina with relative permittivity of 10.2 and a substrate thickness of 25 mil. The filter is designed to be physically symmetric such that the design procedure consists of three reflected group delay stages (Steps 1-3) and one final stage to adjust magnitude response (Step 4).
The input and output couplings are provided by 50Ω tapped transmission lines. The width of each resonator is set to be 23mil which is close to 50Ω. The design parameters are tapped in position, \( L_{in} \), resonator lengths \( \{L_1, L_2, L_3\} \) and gaps between resonators \( \{g_{12}, g_{23}, g_{34}\} \). The coarse model for space mapping is implemented using the Keysight ADS circuit simulator. The fine model is the Sonnet full-wave EM simulator with an accurate 1mil mesh size.

Table 5.1 Selected frequency points and goals for each reflected group delay stage

<table>
<thead>
<tr>
<th>Stage</th>
<th>(Frequency GHz, Reflected Group delay ns)</th>
<th>( \Gamma d^1_{goal} )</th>
<th>( \Gamma d^2_{goal} )</th>
<th>( \Gamma d^3_{goal} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Selected points</td>
<td>(1.493, 2.159)</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>BW edge points</td>
<td>(1.392, 1.433)</td>
<td>(1.617, 1.314)</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>Selected points</td>
<td>(1.577, 4.903)</td>
<td>(1.422, 5.437)</td>
<td>(1.503, 3.844)</td>
</tr>
<tr>
<td></td>
<td>BW edge points</td>
<td>(1.392, 4.431)</td>
<td>(1.617, 4.145)</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>Selected points</td>
<td>(1.604, 9.091)</td>
<td>(1.459, 6.563)</td>
<td>(1.401, 10.404)</td>
</tr>
<tr>
<td></td>
<td>BW edge points</td>
<td>(1.497, 6.999)</td>
<td>(1.546, 6.197)</td>
<td></td>
</tr>
</tbody>
</table>

The design procedure starts from calculating selected frequency points and corresponding reflected group delay goals for each design stage. By solving equations in
Table 3.1, three goal points are obtained for stage 1, five points are obtained for stage 2 and 7 points are obtained for stage 3. Details are given in Table 5.1.

The sweep plans and group delay goals are set up in the Keysight ADS circuit simulator. Coarse model optimum results $\{x^1_c\}, \{x^2_c\}$ and $\{x^3_c\}$ are obtained by optimizing the design parameters in each design stage to match the optimizing goals in Table 5.1. Detailed optimizing parameters and results are shown in Table 5.2. The benefit of using the proposed procedure is that only three parameters are optimized in each design stage, meanwhile, the number of simulation points is reduced by using the sweep plan. The optimization process in each space mapping iteration converges within a very short time.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>465.806</td>
<td>189.468</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>--</td>
<td>152.416</td>
<td>12.4936</td>
<td>222.739</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td></td>
<td>201.97</td>
<td>19.0814</td>
<td>209.657</td>
</tr>
<tr>
<td>Entire Filter</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>193.775</td>
<td>20.5557</td>
</tr>
<tr>
<td>Final Dimension</td>
<td>465.806</td>
<td>152.416</td>
<td>12.4936</td>
<td>201.97</td>
<td>19.0814</td>
<td>193.775</td>
<td>20.5557</td>
</tr>
</tbody>
</table>

Then the implicit space mapping technique is used to determine fine model parameters for each design stage. The adaptive band synthesis sweep available in Sonnet is
the sweep plan used for all fine model EM simulations. The geometries for each reflected group delay stage are shown in Fig 5.1.

Fig.5.1 Geometry in Sonnet for each reflected group delay stage (a) stage 1 (b) stage 2 (c) stage 3 (d) entire filter
In design stage 1, the design parameters are \( \{L_n, L_1\} \). The pre-assigned parameters are chosen to be dielectric constant and substrate thickness \( \{\varepsilon_{r_2}, h_2\} \) of “MSub2”. The pre-assigned parameters \( \{\varepsilon_{r_2}, h_2\} \) are first extracted to align the coarse model to the fine model reflected group delay response. Then the calibrated coarse model in Fig 5.2a is re-optimized using the sweep plan and goals set up at the beginning of the design. With the reduced number of simulation points, the optimizing convergence in space mapping iterations can be achieved within a very short time. The optimized results are then assigned to the fine model directly as the next fine model parameter prediction. The benefit of using the reflected group delay method is that there are only two design parameters in this stage. The fine model reaches the target reflected group delay curve with only one space mapping iteration. The detailed results are given in Table 5.3.
In design stage 2, the design parameters are \( \{G_2, L_1, L_2\} \). A constant substrate “MSub1” is assigned to the fixed components from stage 1. The dielectric constant and substrate thickness for “MSub1” are kept constant to the original \( \{10.2, 25\} \). The pre-assigned parameters are the dielectric constant and substrate thickness as shown in the calibrated coarse model in Fig 5.2b. The sweep plan and goals calculated for stage 2 are used in implicit space mapping iterations to predict the fine model parameters. In this way, the simulation points are reduced to a minimum number and the optimization can be completed within a very short time. Additionally, since only three design parameters are optimized in the space mapping, a convergence is achieved with only one iteration requiring two EM simulations. The results can be found in Table 5.3.

In design stage 3, the design parameters are \( \{G_3, L_2, L_3\} \). The pre-assigned parameters are only from newly added components in Fig 5.2c. The implicit space mapping
converges with only one iteration requiring two EM simulations in a very short time due to using the reflected group delay method and the sweep plan goals calculated for stage 3. The results can be found in Table 5.3.

Table 5.3 fine model design parameters and space mapping result for each reflected group delay stage

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>449.88</td>
<td>183.435</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>--</td>
<td>153.9</td>
<td>10.55</td>
<td>217.1</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>201.97</td>
<td>17.1384</td>
<td>204.815</td>
<td></td>
</tr>
<tr>
<td>Entire Filter</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>197.58</td>
<td>18.25</td>
</tr>
<tr>
<td>Final dimensions</td>
<td>449.88</td>
<td>153.9</td>
<td>10.55</td>
<td>201.97</td>
<td>17.1384</td>
<td>197.58</td>
<td>18.25</td>
</tr>
</tbody>
</table>

At this point, all reflected group delay goals have been reached in both the coarse and fine models. Since the filter layout is physically symmetric, the first three resonators are mirrored to implement the entire filter for final magnitude response adjustment (Step 4). Only two design parameters \( \{ L_3, G_{3,4} \} \) are required to be optimized in this stage. Since all the other parameters have been determined in stage 1 to 3, it is easy to get a very good initial response in both coarse and fine models. The two parameters are first optimized in the coarse model to satisfy the design specifications, and then implicit space mapping is used to predict the two parameters in the fine model. The number of coarse model
simulation points used in this optimizing stage is chosen to be 201 points, as the magnitude response is being considered. The good initial fine model response and a small number of design parameters leads to a final convergence with only one iteration and two EM simulations. The coarse model optimization and fine model space mapping results are given in Table 5.2 and Table 5.3. The final EM response is given in Fig 5.3.

Fig.5.3 Final EM simulation result of the 6 pole microstrip hairpin bandpass filter. Dotted line is the initial response, solid line is the response after one iteration.
In the proposed design method, the reflected group delay method reduces design parameters. The use of space mapping technique results in a fast EM based matching. The selected frequency points and goals decrease the SM optimization time due to the decrease in coarse model simulation time.

The implicit space mapping technique is implemented as the optimization algorithm for the reflected group delay method in designing a 6-pole microstrip hairpin filter. A technique is proposed to reduce simulation points by matching a few selected group delay points. A computer-aided EM based design approach is proposed for the integration of the implicit space mapping technique and the reflected group delay method. The design steps are summarized and the filter is designed in Sonnet. By using the proposed method, the computation time and space mapping iterations are significantly reduced.

In this chapter an EM based sequentially coupled filter design method is proposed by integrating implicit space mapping technique with the reflected group delay design approach. A practical technique is proposed to complete reflected group delay stages more efficiently by matching the group delay values of selected frequency points instead of the entire curve. An example of a 6-pole microstrip hairpin band pass filter is designed using this method. All parameter extraction and fine model predictions are performed using Keysight ADS with the sweep plan and optimizing goals which can dramatically reduce the optimization time. A Sonnet EM simulation result is given to validate the design approach using the reflected group delay method and implicit space mapping technique.
Chapter 6
Conclusion

This thesis focuses on discussing the methodology for the design and implementation of filter structures and coupled resonator based microwave system. The review of original reflected group delay method, space mapping techniques are presented.

An improvement to the original reflected group delay is proposed by introducing selected frequency points and corresponding group delay values in each design step. This improvement can significantly improve the optimizing time for each reflected group delay design step. Examples of filter designs with different filter structures, substrate configurations and filter specifications are given to validate the accuracy and efficiency of the proposed improvement.

A set of equations to characterize the reflected group delay of the band-pass prototype circuits are first reported. These equations can be directly used for filter design.

EM based design method for sequentially coupled resonator filter is proposed by combining the space mapping techniques with the reflected group delay design approach. Mathematical synthesis and design procedures are given in detail. Different design procedures for the combined method using aggressive space mapping and implicit space mapping are presented. A 5-pole hairpin band pass filter is designed, fabricated and tested to verify the proposed method with aggressive space mapping technique. A 6-pole hairpin band pass filter is designed using Sonnet EM simulator to verify the proposed method with implicit space mapping technique. All the proposed methods are proved to be accurate and efficient for sequentially coupled resonator filter designs.
APPENDIX A

\[ \Gamma d_1 = - \frac{g_0}{w^2} \left[ g_0 \frac{1}{2} \left( 2w^2 + 2w_0^2 \right) \right] \]

\[ + \frac{g_0^2 g_1}{w_0^2} \left[ \frac{w - w_0}{w_0 - w} \right] \frac{1}{(w_1 - w_2)^2} \left( w_1 - w_2 \right) \]

\[ \Gamma d_2 = - \frac{g_0}{w^2} \left[ g_0 \frac{1}{2} \left( 2w^2 + 2w_0^2 \right) \right] \]

\[ + \frac{g_0^2 g_1}{w_0^2} \left[ \frac{w}{w_0} \frac{w - w_0}{w_0 - w} \right] \frac{1}{(w_1 - w_2)^2} \left( w_1 - w_2 \right) \]

\[ \Gamma d_3 = - \frac{g_0}{w^2} \left[ g_0 \frac{1}{2} \left( 2w^2 + 2w_0^2 \right) \right] \]

\[ + \frac{g_0^2 g_1}{w_0^2} \left[ \frac{w}{w_0} \frac{w - w_0}{w_0 - w} \right] \frac{1}{(w_1 - w_2)^2} \left( w_1 - w_2 \right) \]

\[ + \frac{g_0^2 g_1^2}{w_0^2} \left[ \frac{w}{w_0} \frac{w - w_0}{w_0 - w} \right] \frac{1}{(w_1 - w_2)^2} \left( w_1 - w_2 \right) \]
Reference

[1] Ness, J.B. "A unified approach to the design, measurement, and tuning of coupled-
resonator filters," IEEE Transactions on Microwave Theory and Techniques, vol.46,
no.4, pp.343-351, Apr 1998.

"Space mapping technique for electromagnetic optimization," Microwave Theory and

[3] Hong, J.S., and M. J. Lancaster, Microstrip Filters for RF/Microwave Applications,


[6] J. S. Frame, Matrix functions and applications, Part IV: Matrix functions and

filters with the reflected group delay method,” IEEE Trans. Applied Superconductivity,


[10] A. Atia and Williams, “New type of waveguide bandpass filters for satellite


[12] Cameron, R. J., “Advanced coupling matrix synthesis techniques for microwave
filters,” IEEE Transactions on Microwave Theory and Techniques, vol. 51, pp. 1–10,
2003.

wideband bandpass filter at the UHF band,” IEEE Trans. Appl. Supercond., vol. 21,
No.3, pp. 538-541, June 2011.


