A Post-Fabrication Tuning Method for a Varactor-Tuned Microstrip Filter using the Space Mapping Technique

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Song Li, candidate for the degree of Master of Applied Science in Electronic Systems Engineering, has presented a thesis titled, *A Post-Fabrication Tuning Method for a Varactor-Tuned Microstrip Filter Using the Space Mapping Technique*, in an oral examination held on April 29, 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.

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Abstract

The RF (radio frequency) and microwave filter is of great importance in the most of the microwave applications which are widely used in broadcasting radios, televisions, radar techniques, telecommunications and satellite applications. Most of the microwave devices contain microwave filter blocks for transmitting and receiving Megahertz to Terahertz frequency band signals. The technologies in the fields of materials, fabrication, design method and electromagnetic analysis are developing quickly in recent years for RF applications. This thesis focuses on an important topic in microwave filter applications, post-fabrication filter tuning.

Post-fabrication tuning processes become more and more important with the development of microwave applications and the requirement of stringent filter performance. A fabricated filter often gives very different performance compared to software designed and simulated models due to the fabrication and material tolerances. The post-fabrication tuning process aims to adjust tuning components implemented with the filter to make the filter give the expected performance. The tuning process is traditionally performed by expert technologists with strong filter knowledge and tuning experience, and it is time consuming and expensive in labor costs. Much easier, automated and accurate post-fabrication filter tuning approaches are necessary.

In this thesis, the basics of microwave filters, tuning techniques and space mapping techniques are discussed in detail. A novel post-fabrication tuning method that exploits the space mapping technique to directly make tuning decisions is first proposed. The tuning
theory and procedures are given in detail.

An application of the proposed method to directly determine the tuning voltages of a fabricated 4-pole varactor diode tuned microstrip combline filter with a center frequency of 1 GHz and an absolute bandwidth of 200 MHz is presented. In the proposed application, implicit space mapping is exploited to map the circuit based coarse model and the capacitance values of the varactor diodes of the fabricated filter. The method is shown to be accurate and efficient in which as long as an accurate mapping is established, all the tuning decisions can be directly made exploiting the fast coarse model without further testing and tuning of the fabricated filter.
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Chapter 1  Introduction

1.1 Outline

Microwave and RF (radio frequency) filters play very important roles in microwave communication systems to operating in the MegaHertz (MHz) to TeraHertz (THz) frequency band. Most microwave devices include some kind of microwave and RF filtering blocks for signal transmitting and receiving. Filters are also the basic building blocks for duplexers, multiplexers and switched filter banks which are widely used in the field of broadcasting radios, radar, television, wireless communication and satellite applications. With the development of these applications, microwave filters with stringent specifications are required. More and more advances in novel materials, fabrication techniques, filter structures, tuning techniques; full wave electromagnetic analysis methods and computer-aided (CAD) design tools are proposed and exploited in the past 50 years. Microstrip filters are one of the most popular newer types of RF and microwave filters. In recent years, more and more novel microstrip filter structures are designed, fabricated and demonstrated to give advanced filtering characteristics. Compared to traditional waveguide filters, microstrip filters are much lighter, compact and cheaper. The disadvantages are that microstrip filters usually have lower power handling capacities and higher losses, and can be more susceptible to manufacturing and material tolerance. Thus post-fabrication tuning becomes an important process for microstrip microwave filters. This thesis focuses on this important post-fabrication tuning process.

A fabricated microwave filter generally requires adjustment and post-fabrication
tuning due to manufacturing and material tolerances. The goal of post-fabrication tuning is to find the optimum solution to some tuning elements such that the measurement results can be adjusted to achieve a best fit to the desired filter specifications. The tuning elements can be tuning screws, tuning varactors, microelectromechanical systems (MEMS) components or other tunable devices. Traditionally, this post-fabrication tuning process is carried out in the form of a human performed task by an expert in filter tuning. This method can be time consuming and expensive.

Many recent research efforts [1]-[6] focus on establishing computer-aided tuning approaches for filters based on either analytical methods, such as coupling extractions and analysis, or real-time optimizing methods. The post-fabrication tuning problem can be considered a general optimizing problem in which real-world tuning parameters are optimized to achieve a certain set of tuning specifications.

The space mapping technique has been proven to be very efficient and successful for microwave optimizing problems. Many types of space mapping algorithms have been developed in recent years. The aim of space mapping is to establish a mapping between a computational expensive but accurate fine model and a fast but inaccurate coarse model. In this way, the expensive fine model optimizing process is directed to a faster coarse model. By updating the mapping iteratively, an accurate matching between the two models can be achieved within a few iterations.

In this thesis, a post-fabrication tuning method for a varactor tuned microstrip filter is proposed where the Implicit Space Mapping (ISM) technique is used to establish a
mapping between the fabricated filter and a coarse model implemented in a circuit simulator. An interpolation technique and the aggressive space sapping technique are used to model the nonlinear and unknown voltage/capacitance relationship of the varactor diodes.

A demonstration of tuning on a post-fabricated varactor-tuned combline filter is given to verify the proposed method.

1.2 Motivation

A post-fabricated filter can give very different performance from original design in software due to manufacturing and material tolerance. Post-fabrication tuning process becomes very important to adjust the fabricated filter to give desired performance. The post-fabrication tuning process is traditionally carried out manually by tuning technologist with strong microwave background and tuning experience. For large-scale microwave filter applications, post-fabricated tuning process becomes very time consuming and expensive on labor cost. Thus a computer-aided post-fabricated tuning approach to make the tuning work much easier and automated becomes necessary. Most of the proposed tuning approaches in the recent years are based on analytical methods or direct optimizing using numerical methods. The analytical methods often require strong microwave filter background and clear realization of filter characteristics. The direct optimization often takes many iterations and the convergence are not guaranteed to be achieved. The implementation of space mapping technique becomes a good choice for
developing such a tuning approach to fast and accurately tune a post-fabricated filter to give desired performance.

Hence, a novel post-fabricated tuning method is proposed in this thesis, the proposed method is proved to be efficient and accurate for post-fabricated filter tuning. With the proposed method, as long as the mapping is well established, tuning decisions can be directly made in the coarse model (a fast circuit simulator) without further implementation and analysis of the fabricated filter or the need of a human expert to tune the filter.

1.3 Thesis Organization

Followed by the introduction in Chapter 1, Chapter 2 presents a detailed review of some popular post-fabrication tuning methods. Chapter 3 gives a brief review of the space mapping techniques and some basic concepts and theory required in this thesis. Applications of basic space mapping techniques are also given.

In Chapter 4, the tuning theory and procedure is presented in detail. In Chapter 5, the method is demonstrated to tune a well-known tunable four-pole microstrip combline filter structure fabricated using printed circuit board technology. Conclusions are presented in Chapter 6.
Chapter 2 Literature Review

An important part in filter design is the post-fabrication tuning. Due to manufacturing and material tolerances, designers may get a good result in a full-wave microwave simulator like ADS Momemtun [49] and Sonnet [50]. But the fabricated product can have a very different measured performance. Traditionally, the post-fabrication tuning process requires expert skills and tuning experience. It can be very time-consuming and expensive. In the past decades, research has been performed to simplify the complexity of this tuning process by introducing different types of tuning components, tunable filter structures and tuning methods. In this Chapter, a review of recent popular filter tuning components, structures and tuning methods is presented.

2.1 Microwave Tunable Filters and Tuning Components

2.1.1 3D Tunable Filters and Tuning Screws

Most of the three dimensional (3D) microwave filters, such as waveguide filters and cavity filters, are popular tunable filter structures [7]-[11]. They are widely used in the area of radar system, telephone networks, television broadcasting and satellite communications. As shown in Fig. 2.1, the most popular tuning component for a 3D filter is the tuning screw. Tuning screws are screws that are inserted into the resonant cavity to adjust the coupling and the center frequency of the resonator by changing the tuning position of the screws
inside the resonant cavity. The 3D filter structure and the tuning screws can be well modeled by many popular 3D microwave design software. Research has been taken on the tuning technique of 3D waveguide and cavity filters.

Fig.2.1 3D waveguide tunable filter and tuning screws [12]-[13]

### 2.1.2 Microstrip Tunable Filter and Tuning Components

With the development of microstrip applications and printed circuit board (PCB) technology, more and more research has been carried out on the design of microstrip tunable filters. Compared to traditional waveguide technology, microstrip is much cheaper, lighter and compact though microstrip circuits show higher losses and lower power handling capabilities.

Hence, various types of tuning components are proposed for the purpose of post-fabrication tuning on two dimensional (2D) microstrip tunable filters. Many different microstrip tunable filter structures are proposed in the past ten years with the use of different tuning components. One of the most popular basic microstrip tunable filter
structure is the combline filter given in Fig.2.2.

As shown in Fig. 2.2, the combline filter consists of a number of adjacent coupled resonators which are short circuited at one end and have a tunable component loaded between the other end and the ground. Each resonator is designed with an electrical length of less than a quarter wavelength. In this filter structure, the tunable components are used
to adjust the resonant frequency of each resonator. Many special microstrip structures introduced in [14]-[18] are modified based on this basic combline structure. Different types of tuning components are used by many designers for different tuning purpose. The choice of tuning components can lead to very different filter performances.

In the following part, some popular types of tuning components for microstrip tunable filter are introduced.

2.1.3 Varactor Diodes (Varicap)

A varactor diode is generally a type of diode in which the capacitance of the diode varies as a function of the biasing DC controlling voltage.

Varactors work as a reverse-biased p-n junction as shown in Fig. 2.3. There is no current flow within the component.

Fig.2.3 Basic structure of Varactor diode [19]
As the reverse voltage is applied to the p-n junction, the holes in the p-type material move toward the node, and the electrons in the n-type material move towards the cathode of the diode. This leaves a region with no carriers, and acts as the depletion region. The thickness of the depletion region varies with the applied bias voltage. When increasing the bias voltage, the thickness of the depletion region and the capacitance will decrease. Since the varactor diode is an active device, there will be non-linearities associated with signals that pass through it.

Varactor diodes are widely used as tuning elements in microstrip tunable filters structures where capacitance can be tuned to change the resonant frequencies of resonators or the couplings between resonators [20]. In Chapter 4, a varactor tuned microstrip 4-pole combline filter is fabricated and tuned with varactor diode. This is a hyper-abrupt silicon varactor with a high quality factor of 2000 at 5 Volts at 50MHz. The provided capacitance can range from 0.6pF to 7pF. The varactor diode and its circuit model are shown in Fig. 2.4

Fig.2.4 Varactor diode MA46H202 [21] (a) varactor package (b) circuit schematic

Since the tuning speed of the varactor diodes is dictated by electrical means, its tuning
speed is one of its main advantages.

### 2.1.4 RF Microelectromechanical Systems (MEMs) Components

In recent years, more and more RF Microelectromechanical systems (MEMs) components are applied to microstrip tunable filter structures. The benefit of MEMs technology is that large number of elements can be fabricated on a single chip. There are many different RF MEMs structures and applications reported in [22]-[26]. The most popular RF MEMs components used for microwave tunable filter design are the RF MEMs switch and MEMs varactor.
Fig. 2.5: MEMs tunable components [27]: (a) A RF MEMs switch (b) RF MEMS RF MEMs varactors

The RF MEMs switch is a component that switches between an open or short circuit on a transmission line through mechanical movement such that the impedance of the transmission line can be controlled.

The RF MEMs varactor is a component that operates as a varactor where the separation between two plates is controlled through mechanical movement such that the capacitance of the mechanical structure varies.

The advantage of RF MEMs components is that the fabricated RF MEMs components are highly integrated and show very good linearity performance. Meanwhile, the power-requirement is much lower compared to normal varactor diodes.

The disadvantage for the RF MEMs components is that the tuning range of MEMS
tunable filters is typically low due to limitations from the range of the mechanical movement. Additionally, the unloaded quality factors of MEMs components are very low and the losses are relatively higher, though recent advancements in MEMS devices using superconducting materials have demonstrated low loss performance [52].

2.1.5 Ferroelectric Materials

Ferroelectric materials can be used as tuning components in tunable filters [28]-[29] since the permittivity of the material changes with an applied DC electric field where the external DC voltage is delivered with interdigital electrodes. The ferroelectric material and the DC electrodes can be placed wherever tuning is needed. The ferroelectric material and the substrate contribute to the effective permittivity of the microstrip structure. Changing the permittivity of the ferroelectric material results in tuning the filter as desired.

There are two main ferroelectric materials used in tunable filters. Barium strontium titanate (BST) can be used for room temperature applications, and strontium titanate (ST) can be used at cryogenic temperatures along with high-temperature superconductors (HTS).

2.1.5.1 Barium Strontium Titanate (BST) Capacitor

Barium strontium titanate (BST) is used as a ferroelectric material for making tunable capacitors at room temperature [30]-[31]. The BST capacitors are made by depositing a thin film with metalorganic chemical vapour deposition (MOCVD). The thin film is then
covered with a deposited layer of metal making a parallel plate capacitor.

2.1.5.2 Strontium Titanate (STO) Capacitor

Unlike the BST material, the strontium titanate (STO) ferroelectric material is often used at low temperatures [32], such as temperatures less than 77 Kelvin, and can be used in conjunction with high-temperature superconducting (HTS) material \(\text{YBa}_2\text{Cu}_3\text{O}_{7-\delta}\) (YBCO) on a LaAlO\(_3\) substrate.

The thin film materials have higher loss tangents and lower dielectric constants, but are easier to work with and are tunable at temperatures less than 77 Kelvin [32].

2.2 Filter Tuning Method

2.2.1 Sequential Tuning of Coupled Resonator Chebyshev Filter

J. B. Ness proposed a sequential tuning method for coupled resonator Chebyshev filter in 1998 [33] by successively adding resonators step by step. In each step the group delay response of reflected signal \(S_{11}\) is aligned to an ideal low-pass prototype model. He indicate that the group delay response of input reflected coefficient \(S_{11}\) of sequential coupled resonators provide all necessary information to characterize the couplings between resonators and resonant frequencies of individual resonators.

This tuning method is best fit into 3D post-fabricated tuning problems where resonators for post-fabricated cavity filter can be easily detuned. Meanwhile, for a practical post-fabricated filter, the filter is often attached to input/output transmission lines
which affect the locations of the 180/0 degree phase, the reference plane of the post-fabricated filter is required to be analyzed to get correct phase adjustment before taking the post-fabricated tuning process. Thus, it is not easy to be applied to microstrip tunable filter structures.

2.2.2 Coupling Matrix Extraction Method

The general idea of coupling matrix extraction method is to use an ideal coupling matrix model [34] to synthesize a coupled resonator filter structure. By optimizing the coupling matrix parameters to fit the measured post-fabricated filter response, it is able to do analysis and comparison between the extracted filter coupling matrix and a desired coupling matrix. Tuning decision are made based on the comparison analysis. This method requires direct relations between the coupling matrix elements and particular physical tuning elements.

The coupling matrix synthesis is first proposed by Atia and Williams in [34]. Several modifications of this technique have been reported [35]-[37]. They proposed a generalized coupled resonator filter circuit model as shown in Fig. 2.6
This circuit model contains a number of inter-coupled resonators made up of ideal capacitors and inductors. Each capacitor has a constant capacitance of 1F and the inductor has a constant inductance of 1H such that all resonators share the same resonant frequency of 1 rad/second. As shown in the Fig. 2.6, $M_{ij}$ are defined coupling elements to represent the internal couplings between individual resonators. $R_1$ and $R_N$ are the source and load impedance. Thus, a filter response can be directly synthesized by $R_1$, $R_N$ and coupling matrix elements $M_{ij}$, a matrix form to represent all couplings of the filter is defined as the coupling matrix and given in Fig. 2.9.
Hence in coupling matrix synthesis, the scattering parameters of a filter can be directly expressed in the coupling matrix form as:

\[
S_{21} = -2j\sqrt{R_i R_e} \left[ A^{-1} \right]_{n1},
\]
\[
S_{11} = 1 + 2jR_i \left[ A^{-1} \right]_{j1},
\]
\[
A = \lambda I - jR + M
\]
\[
\lambda = \frac{f_0}{BW} \left( \frac{f}{f_0} - \frac{f_0}{f} \right)
\]  

(2.1)

For a practical post-fabricated tuning problem, each single element in the coupling matrix is required to be related to particular tuning element. The diagonal element \(m_{ii}\) is related to the resonator resonant frequencies while \(m_{ij}\) is related to corresponding adjacent resonator coupling or cross coupling. \(R_i\) and \(R_e\) are related to input and output couplings. By optimizing the coupling matrix elements to match the measured post fabricated filter response, it is possible to extract the coupling matrix of the measured filter. Then a comparison between extracted coupling matrix and ideal matrix can be carried out to determine which tuning element is required to be tuned.

This method provides a way to extract the coupling information of a measured post-fabricated filter response for technologists to make tuning decision. Traditionally the
tuning decisions are made by skilled technologists based on their tuning experience. Some later research is then carried out to implement filter structure theoretical analysis or direct numerical optimizations. Meanwhile, since the derivative of post-fabricated filter response in terms of coupling matrix elements are unavailable, high-level gradient-based optimizing algorithms cannot be directly applied, the convergence of the optimizing is not guaranteed. In order to achieve the convergence, a good initial measurement response is very important for the optimization. Sensitivity analysis is required to be carried out before the optimizing process to achieve a reduction of tuning iterations.

2.2.3 Computer-aided Tuning Based on Poles and Zeros of The Input Reflection Coefficients

From basic filter synthesis, the filter response is directly characterized by transfer and reflection polynomials which are determined by zeros and poles [38]. The core of this tuning method is that the phase of reflected coefficients of a filter contains the information of poles and zeros of a filter which can be used to characterize all resonant frequencies of individual resonators and couplings between resonators.
Fig. 2.8 an N-resonator coupled two-port network with output port terminated in a short circuit [48]

This method is based on the fact that the zeros and poles of input reflection coefficient in a N-resonator coupled two-port network shown in Fig. 2.8 with output port terminated in a short circuit are related to individual resonator resonant frequencies and coupling coefficients. According to transmission line theory and filter synthesis, the input impedance at loop \(i\) is given as:

\[
Z_{in}^{(i)} = j \frac{Z_{0i}}{\omega_0} \frac{P_i(\omega^2)}{Q_i(\omega^2)}, i = 1, 2, \ldots n
\]  

(2.2)

Where

\[
\begin{align*}
P_i(\omega^2) &= \prod_{i=1}^{n-i+1} \left( \omega^2 - \omega_{zi}^2 \right), i = 1, 2, \ldots n \\
Q_i(\omega^2) &= \prod_{q=1}^{n-i+1} \left( \omega^2 - \omega_{pq}^2 \right), i = 1, 2, \ldots n
\end{align*}
\]  

(2.3)

\(P_i(\omega^2)\) and \(Q_i(\omega^2)\) are the polynomials made up of order \((n - i + 1)\) and \((n - 1)\).

\(Z_{0i}\) and \(\omega_0\) are the characteristic impedance and resonance frequency of resonator \(i\).

\(\omega_z\) and \(\omega_{pq}\) are the zeros and poles of the two polynomials as well as the poles and zeros of the input impedance of the load shorted one port network.

An equation relates to poles and zeros of the one port network to the resonator
resonant frequencies and couplings between resonators are given by [38]:

\[
\omega_{0i}^2 = \frac{\prod_{j=1}^{n-i+1} \omega_{zi}^{(j)^2}}{\prod_{q=1}^{n-i} \omega_{pq}^{(j)^2}}, i = 1, 2, \ldots n
\]

\[
m_{i,j+1}^2 = \sum_{i=1}^{n-i+1} \omega_{zi}^{(j)^2} - \sum_{q=1}^{n-i} \omega_{pq}^{(j)^2} - \omega_{0i}^2, i = 1, 2, \ldots n - 1
\]

\[
R_{i,n} = \frac{\prod_{j=1}^{n} \left( \omega_{R}^2 - \omega_{zi}^{(j)^2} \right)}{\omega_{R} \prod_{j=1}^{n} \left( \omega_{R}^2 - \omega_{pi}^{(j)^2} \right)}
\]

Where \( \omega_{R} \) is the frequency where a ±90° phase of the input reflected coefficient takes place.

Thus, the resonant frequencies of resonators and couplings in the filter can be obtained by measurement of the poles and zeros of the shorted one-port network. As the reference plane has been adjusted, the poles and zeros can be directly read at the frequencies where 180° and 0° phase take place. These frequencies are the required poles and zeros of the input reflection coefficients as well as the input impedance. It is important to note that the reference plane is required to be firstly adjusted before the calculation of poles and zeros.

In this way, the poles and zeros can be extracted from a post-fabricated filter measurement of the shorted N resonator network. After the comparison and theoretical analysis according to an ideal model, tuning decision can be made.

It is important to note that for a practical post-fabricated filter, the last resonator is often loaded with a transmission line and a RF connector. The loading effect is required to be considered and analyzed during the extraction of poles and zeros.
2.2.4 Time Domain Tuning Technique

Keysight technologies proposed the time domain tuning technique [39]-[40]. They proposed that the time-domain response of the input reflected coefficients $S_{11}$ characterized the resonant frequencies of resonators and all couplings between resonators of a filter. A 5 pole Chebyshev filter example was given by them. The filter response and corresponding time domain response is shown in Fig. 2.9.

![Graphs showing frequency response and time-domain response](image)

Fig. 2.9 (a) Frequency response and $S_{11}$ time-domain response when resonator 2 is detuned. (b) Frequency response and $S_{11}$ time-domain response when resonator 3 is detuned. [48]

In the example, they showed that in the time domain of $S_{11}$, there are five dips related to individual resonators resonance frequency while each peak between two dips indicates
one inter-resonator coupling. They found that tuning of one resonator or inter-resonator coupling only affect one dip or peak. In their tuning process, resonators and adjacent couplings are tuned successively to match the dip and peak in the time domain response of input reflected coefficients until the whole response are well matched. This method provides a way to divide the whole filter tuning problem into smaller sub-problems. This tuning procedure is very similar to the reflected group delay method. The limitation is that it requires experienced technologist to map the relationship of dips and the tuning components to make tuning decisions based on their microwave and filter background.

2.2.5 Tuning Method Based on Fuzzy Logic Techniques

Fuzzy logic technique was firstly introduced in filter tuning by Miraftab and Mansour [41]-[42]. The idea comes from the fact that experienced technologists often use the concept of sets during their manual tuning process. These sets are ranked with different level like very small, small, large and very large. Experienced tuning technologists can directly make the tuning decisions to particular tuning component according to the displayed filter measurement response. The fuzzy logic technology is designed to perform the human like thoughts to make required approximations and decisions.

Similar to basic Boolean logic, an element in fuzzy logic can either belong to a set or does not belong to the set. A binary value 0 or 1 called membership value is assigned to each element in a set. 0 means the element is not in this set and 1 means the element is in
this set. Fuzzy logic interprets the numerical data as linguistic rules. Then the extracted rules will be used to generate the output value of the fuzzy logic system.

Generally, a fuzzy logic system can be considered as a smart function estimator. It maps the input information into number of input fuzzy sets, and generates output fuzzy sets by applying pre-established fuzzy logic rules. The output fuzzy sets are then translated into output tuning information. These rules are normally some IF-THEN statements created based on expertise experience, numerical data and mathematical analysis. Thus, the fuzzy logic technology is able to combine all the filter and coupling matrix synthesis with the expert tuning experience from tuning technologists since the fuzzy logic processes all of these sets of information in the same way.

As shown in Fig. 2.10 a fuzzy logic system for post-fabricated tuning problem contains four parts, they are fuzzifier, fuzzy inference system, rules and defuzzifier.

![Fig. 2.10 A block diagram of the fuzzy logic system [48]](image)

The fuzzifier transfers the input information into input fuzzy sets. Fuzzy inference
procedure is the engine to generate output fuzzy sets from input fuzzy sets based on pre-created rules. There are many different types of fuzzy logic inferential procedures, normally only a few of them are used in engineering field and particular post-fabricated filter tuning problem. Just like there are a lot of optimizing algorithms or human methods for making decisions. The choice of fuzzy logic inferential procedure is dependent on the requirements of the particular goals. The rules and inference procedures are the most important in the system because they are the key to affect the accuracy and efficiency of the function approximations. The defuzzifier transfers the output fuzzy sets into the required output information.

There are many types of membership functions, the most popular types are triangular, trapezoidal piecewise linear, and Gaussian. The membership function are usually designed according to a user’s experience and numerical data provided by the system designer for a particular problem. More membership functions will lead to a better approximation but with higher computation costs.

The general steps to build up a fuzzy logic system for post-fabrication tuning problem is as follows:

Step 1: Assigning memberships to all tuning variables

Step 2: Creating IF-THEN rules

Step 3: Apply fuzzy inference and defuzzification process according to the IF-THEN rules obtained from step 2 to calculate the required output tuning variables.

Please note that the fuzzy logic method is focused on combining information and
carrying out approximations. It is very different from the other analytical method described in previous sections. The fuzzy logic method is an information analyzer and a function estimator. It is often built up with the completed collections of all the data and information obtained from the analytical methods plus human experience information for further approximation. It is able to integrate theoretical models like the coupling matrix. It is also able to include data information like poles and zeros. Thus, this method is compatible to all the other tuning method discussed in the previous sections.

An automated 3D filter tuning system was given in [48]. The block diagram is shown in Fig. 2.11:

![Block diagram of automated tuning station](image)

Fig. 2.11 a proposed automated tuning station in [48]

The tuning components are the tuning screws physically controlled by the motor arms. The VNA (vector network analyzer) is used to read measurement data which is the input data. The computer contains the fuzzy logic system collecting the input data and sending
out approximated output commands to the motor arms.
Chapter 3 Microstrip Filter and Space Mapping Techniques

3.1 Introduction to microstrip and basic concepts of filter network

Microstrip is a type of electrical transmission line which can be fabricated using the printed circuit board technology. The microstrip consists of a ground layer at bottom, a dielectric layer in the middle and a conducting layer at the top. It is very popular in the recent years for design of microwave applications like RF filters.

Compared to traditional waveguide technology, microstrip is much cheaper, lighter and more compact; the drawbacks are the low power handling and high losses.

Most microwave filters can be represented by a two port network between a source and a load. Since the voltages, currents and impedances cannot be directly measured using the voltmeters and ammeters under microwave frequencies, the scattering matrix are usually used to characterize the reflected and incident voltage waves at each port of the network. A scattering matrix for a two port filter network can be represented by

\[
\begin{bmatrix}
    b_1 \\
    b_2
\end{bmatrix} =
\begin{bmatrix}
    S_{11} & S_{12} \\
    S_{21} & S_{22}
\end{bmatrix}
\begin{bmatrix}
    a_1 \\
    a_2
\end{bmatrix}.
\]

Where \( b_i \) denotes the incident voltage wave at port \( i \) and \( a_i \) denotes the reflected voltage wave at port \( i \). The \( S_{11} \) and \( S_{22} \) are called the reflection coefficients and the \( S_{21} \) and \( S_{12} \) are called the transmission coefficients. The S-parameters are complex parameters which can be directly measured by a vector network analyzer (VNA) for characterizing a filter network.
3.2 Introduction of the Space Mapping Technique

The space mapping technique was first proposed by John W. Bandler in 1994[44]. The core of the space mapping technique is to establish a mapping between a computational expensive fine model and a fast but inaccurate coarse model. In this way, optimization of the expensive fine model can be carried out by the faster coarse model while the accuracy is ensured by taking fine model evaluations. The space mapping technique is considered to be a great contribution to engineering design especially for microwave design.

In the past 20 years, various space mapping techniques have been proposed and proven to be efficient in microwave design problems [44]-[47]. The most popular two space mapping techniques are the aggressive space mapping [46] and the implicit space mapping [47].

In this Chapter, a brief review of these two types of space mapping techniques is given. Two examples are given to show how they work.

3.3 Basic Concepts of Space Mapping:

A general microwave circuit design optimizing problem can be considered as to solve:

\[ x^* = \arg \min_x U(R(x)) \]  

(3.1)

Here \( X \) denotes the set of design parameters, \( U \) denotes the optimizing objective function. \( R \) denotes the set of resulted responses. \( x^* \) is defined as the optimal solution of
design parameters. Normally, in a microwave circuit design the optimizing process is carried out directly in a full-wave EM based simulator. The optimizing is often very expensive and time-intensive.

In the space mapping technique, two models are defined, the coarse model and the fine model. The coarse model and fine model design parameters are denoted by \( x_c \) and \( x_f \). The corresponding coarse and fine model response are denoted by \( R_c \) and \( R_f \). The mapping \( P \) established between coarse and fine models need to satisfy:

\[
\begin{align*}
x_c &= P(x_f) \quad \text{such that } \quad R_c(P(x_f)) \approx R_f(x_f)
\end{align*}
\]

By iteratively updating the established mapping information, it is possible to find the fine model optimum solution within a few fine model simulations.

### 3.4 Aggressive Space Mapping:

The aggressive space mapping algorithm exploits a quasi-newton iteration and standard Broyden updates. In each iteration, parameter extraction is taken place using the coarse model and the results are applied to Broyden updates to update the established mapping. The main process in each iteration can expressed as:

\[
x_{f}^{i+1} = x_{f}^{i} + h_{i} \quad \text{and } \quad B_{i}h_{i} = -f_{i}
\]

Here, \( B_{i} \) is the approximation of the mapping Jacobian \( J_{p} \). It is updated iteratively by

\[
B_{i+1} = B_{i} + \frac{f_{i+1}}{h_{i}^{T}h_{i}}h_{i}^{T}
\]
3.4.1 Application of Aggressive Space Mapping to An 8-pole End-coupled Band Pass Filter

An 8-pole end-coupled microstrip band pass filter is designed as an example to show the process of applying aggressive space mapping technique to microwave filter design. The filter is specified to have a center frequency of 2GHz with a fractional bandwidth of 2% and return loss better than 20dB. The dielectric material is chosen to be alumina with expected dielectric constant of 10.2 and substrate height of 25 mil. The Keysight ADS circuit simulator is exploited as the coarse model. The Sonnet EM simulator is used as the fine model.

This filter is firstly designed in the coarse model according to general filter design method. The circuit schematic and circuit simulation results are given in Fig. 3.1 and Fig. 3.2. The fine model Sonnet geometry is given in Fig. 3.3. Then by following the aggressive space mapping process, the EM simulation results converge to 20dB return loss after 6 iterations. The dimension parameters of the coarse and fine models for each iteration are given in Table 3.1. The corresponding fine model response after each iteration is given in Fig. 3.4.
Fig. 3.1 Circuit schematic of the designed 8-pole end-coupled filter in ADS simulator.

Fig. 3.2 Circuit simulation results of the coarse model; Red curve: $S_{11}$; Blue curve: $S_{21}$

Fig. 3.3 Sonnet geometry of the designed 8-pole end-coupled filter in ADS simulator.
Table 3.1 The dimension parameters of coarse and fine models in each aggressive space mapping iteration

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<th>Iterations</th>
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<th>d2</th>
<th>d3</th>
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<th>L2</th>
<th>L3</th>
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Fig. 3.4 The results of EM simulation after each aggressive space mapping iteration. The iterations stopped when the return loss within pass band is better than 20dB

3.5 Implicit Space Mapping

Different from basic explicit space mapping like the aggressive space mapping discussed in section 3.4, implicit space mapping technique introduces an algorithm where the mapping process is embedded in the coarse model. In other words, the mapping updates are carried out during the parameter extraction process. By introducing the
auxiliary parameters (pre-assigned parameters), the implicit space mapping provides an indirect way to do fine model prediction.

In the implicit space mapping technique, an implicit mapping is created between the fine model and the coarse model where \( x_f \) is used to denote the set of fine model design parameters, \( x_c \) is used to denote the set of coarse model design parameters and \( x_{aux} \) is used to denote the set of auxiliary parameters (pre-assign parameters). Different from the explicit space mapping, during the parameter extraction procedure, the implicit mapping is modeled by optimizing auxiliary parameters \( x_{aux} \) while the designable parameter set \( x_c \) is kept constant. At this point it is assumed that the coarse model is mapped to the fine model under the established implicit mapping. The next fine model prediction is then obtained by optimizing the coarse model design parameters \( x_c \) until the coarse model response matches the target specifications. By taking these steps iteratively it is able to match the fine model response to certain specifications with a few fine model simulations.

3.5.1 Application of Implicit Space Mapping to An 5-pole Parallel-coupled Band Pass Filter

A 5-pole parallel microstrip band pass filter is designed as an example to show the process of applying implicit space mapping technique to microwave filter design. The filter is specified to have a center frequency of 1GHz with a fractional bandwidth of 5%
and return loss better than 20dB. The dielectric material is chosen to be alumina with expected dielectric constant of 10.2 and substrate height of 25 mil. The Keysight ADS circuit simulator is exploited as the coarse model. The Sonnet EM simulator is used as the fine model. The design parameters are the length of each resonator \( \{ L_1, L_2, L_3 \} \) and the gaps between resonators \( \{ S_{01}, S_{12}, S_{23} \} \). As shown in the circuit schematic. There are four different substrate configurations “MSub1”, “MSub2”, “MSub3” and “MSub4” assigned to different individual components. The pre-assigned parameters are the dielectric constants and substrate heights for the four substrates and the pre-assigned width to each resonator. Thus the auxiliary parameters are \( \{ \varepsilon_r, \varepsilon_r, \varepsilon_r, \varepsilon_r, h_1, h_2, h_3, W_1, W_2, W_3 \} \)

This filter is firstly designed in the coarse model according to a general filter design method. The circuit schematic and circuit simulation results are given in Fig. 3.5 and Fig. 3.6. The fine model Sonnet geometry is given in Fig. 3.7. Then by following the implicit space mapping process, the EM simulation results converge to 20dB return loss after 5 iterations. The dimension parameters of coarse and fine model in each iteration are given in Table 3.2. The corresponding fine model response after each iteration is given in Fig. 3.8.
Fig. 3.5 Circuit schematic of the designed 5-pole parallel-coupled band pass filter in ADS simulator.

Fig. 3.6 Circuit simulation results of the coarse model.

Fig. 3.7 Sonnet geometry of the designed 5-pole parallel-coupled band pass filter in ADS simulator.
Fig. 3.8 The results of EM simulation after each implicit space mapping iteration. The iterations stopped when the return loss within pass band is better than 20dB
Table 3.2 The dimension parameters of coarse and fine models in each implicit space mapping iteration

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Chapter 4  Post-Fabrication Tuning for Microstrip Filters

4.1 Tuning Theory for Varactor-Tuned Microstrip Filter

Most of the popular post-fabricated filter tuning methods reviewed in Chapter 2 are proposed for 3D cavity and waveguide filters. Practically, for a 3D cavity or waveguide filter, tuning screws are often used as the tuning components. The tuning positions of screws can be well modeled by many popular 3D microwave software like HFSS. But for a 2D post-fabricated microstrip varactor tuned filter, the tuning components are usually the DC voltages provided on the tuning varactors. It is very hard to model the voltage directly in a 2D circuit based CAD tools.

The choice of varactors became very important for the microstrip tunable filter design. As reviewed in Chapter 2, different types of tuning components have their own strengths and weaknesses on the tuning speed, power requirement, tuning range and unloaded quality factor. The manufactory tolerances are also different for products from different manufacturing companies. One of the mostly used tuning components is the varactor diode which has a fast tuning speed and low cost. During the tuning process, a designer may choose to trust the relationship between the tuning varactor capacitance and the supplied DC voltage provided from manufacturing datasheet. Meanwhile, some extra DC and RF biasing components are implemented together with the varactors. It is normally necessary to build extra circuits to carry out sensitivity analysis to confirm the relationship between supplied DC voltage and the varactor circuit capacitance under testing frequency band. The relationship is often nonlinear and hard to be modeled using
CAD tools or general filter synthesis. Thus it can be difficult and expensive to implement efficient real-time optimizing algorithms to a post-fabricated varactor tuned microstrip filter because of the unpredictable and non-linear relations between tuning parameters and filter performance.

Fig. 4.1 Flow diagram of a normal real time tuning system for a varactor tuned microstrip filter

Fig. 4.1 shows a block diagram of traditional real-time computer-aided tuning system for varactor tuned microstrip filter. The DC tuning voltages for the varactor diodes are provided by a microcontroller which connects to a central computer. The response of the post-fabricated filter is measured by a vector-network analyzer (VNA) analyzer and loaded into the central computer as an output data file (touch stone file). After that, the file is read by a program and analyzed according to an analytical filter synthesis or expert tuning experience routine. The next voltage prediction is then
obtained through direct numerical optimization or determined by particular estimator like fuzzy logic system reviewed in Chapter 2 and sent to a microcontroller. For direct numerical optimization, because of the non-linear relations between filter performance and tuning parameters and some unpredictable tolerance, the gradient is often unavailable for high level optimizing algorithms. Thus, normally derivative-free optimization algorithms are used for this case, like the genetic algorithm (GA). These algorithms often require many iterations and result in an unguaranteed convergence. Some particular high level non-linear approximation system like fuzzy logic controller can significantly improve the efficiency and accuracy of the tuning. However, it often requires expert tuning experience and synthesis based analytical data information. Additionally, the fuzzy logic rules are required to be redesigned and reloaded into the database for different tuning goals for many designed structures.

In this Chapter, an efficient post-fabricated filter tuning method based on space mapping and an interpolation technique is proposed [51]. The control system of the post-fabricated tunable filter is only required to be connected to the central computer at the beginning of the tuning. after the mapping between the post-fabricated filter and the coarse model is created, the filter and control system are not required to be reconnected to the computer because all the optimizing process and voltage decisions can be carried out within the circuit based coarse model in any computer or mobile system. This will significantly improve the tuning efficiency because in most cases, the fabricated filter is
not required to be reconnected to a VNA analyzer and a central computer for optimizing.

The block diagram of the tuning system is shown in Fig. 4.2.

Fig. 4.2 Flow diagram of tuning system based on proposed method for a varactor tuned microstrip filter

As reviewed in Chapter 2, the space mapping technique introduces two models in the optimizing procedure, an accurate but computational-expensive fine model and an inaccurate but easy to evaluate coarse model. Since the fine model is meant to be an exact representation of the actual filter, the fabricated filter itself is considered to be the
fine model in the proposed method where measured results replace the traditional use of simulation results. A circuit based simulator, such as Keysight ADS, can be used as the coarse model to carry out the required optimization.

In the traditional space mapping technique, the parameters in the two models are required to share the same physical meaning. For a post-fabricated microstrip filter, it is difficult to directly model the varactor tuning voltages in the coarse model due to the unknown non-linear voltage/capacitance relationship of each varactor and the added parasitic elements introduced by the biasing circuit. In the proposed method, ideal capacitors are used to model the capacitance in the coarse model and an additional mapping between the ideal capacitors and the tuning voltages is established. Hence, during the tuning process, the implicit space mapping technique is used to map the two spaces between the coarse model and the fabricated filter and an explicit aggressive space mapping is used to update the mapping between tuning voltages and coarse model ideal capacitors.

4.2 Basic Concepts

For the proposed method, the set of tuning parameters in the fine model and the coarse model are $x_f$, $x_c$, where $x_f$ denotes a set of DC tuning voltages for the fabricated filter and $x_c$ denotes a set of ideal capacitances in the coarse model. The set of auxiliary parameters in the coarse model is $x_{aux}$. These parameters can be pre-assigned substrate characteristics or physical filter layout dimensions. Loss may be required to be modeled
in the coarse model as well. The measurement response of the fine model is $R_f(x)$ and the simulation response from the coarse model is $R_c(x, x_{aux})$.

### 4.3 Proposed Tuning Procedure

The proposed tuning procedure is as follows:

**Step 1:** Set up the tuning voltages for all tuning varactors to reasonable values such that the measured response is ready for parameter extraction. Record the voltages $\{x^0\}$ and the measured response $R_f(x^0)$. Create the microstrip circuit in the coarse model and use the parameter extraction technique to extract the required coarse model tuning and auxiliary parameters:

$$\{x^0_{ct}, x^0_{aux}\} = \arg \min_{\{x_{ct}, x_{aux}\}} \left\| R_c(x_{ct}, x_{aux}) - R_f(x^0) \right\|$$  \hspace{1cm} (4.1)

**Step 2:** Change the tuning voltages to new sets of values $\{x_{sens}^0\} = \{x^0 \pm \Delta x\}$ and record the measured responses $R_f(x_{sens}^0)$. In the coarse model, keep the auxiliary parameters from step 1 and optimize the corresponding tuning parameter to get the coarse model response similar to $R_f(x_{sens}^0)$. This is done to obtain a group of coarse model tuning parameters (ideal capacitance) for sensitivity analysis.

$$\{x_{sens}^0\} = \arg \min_{x_{sens}} \left\| R_c(x_{ct}, x_{sens}^0) - R_f(x_{sens}^0) \right\|$$  \hspace{1cm} (4.2)

The purpose for this step is to get a rough initial mapping, as it is difficult to achieve a fine match by only optimizing the ideal capacitors. It is important to note that if all the tuning varactors and bias circuits are the same such that their voltage capacitance
relationship are very similar, only one sensitivity analysis is required to be performed. If
the varactors are different, or there are fabrication and assembly tolerances that lead to
big variance between varactors, it is necessary to analyze all of them.

Thus, a more accurate initial Broyden, $B_0$, can be obtained by applying the Lagrange
interpolation technique to the values $(\{x_{ct}^0\}, \{x_{ct}^0\}, \{x_{ctx}^{sens}\}, \{x_{ct}^{sens}\})$ obtained in the first two
steps.

Step 3: Start from $\{x_{ct}^0, x_{aux}^0\}$, keep the auxiliary parameters from step 1 and optimize the
coarse model tuning parameters (ideal capacitors) until the simulation response achieves
the required tuning specifications. Obtain a new set of coarse model tuning parameters:

$$\{x_{ct}^{i}\} = \arg \min_{x_{ct}} \left\| R_c (x_{ct}, x_{aux}^0) - R_{\text{specification}} \right\|$$  (4.3)

Step 4: Since the tuning parameters in the coarse and fine models share different physical
meaning and have an uncertain nonlinear relationship, the initial Broyden obtained in
step 2 is used in the ASM algorithm to approximate the fine model tuning parameters,
which are the tuning voltages.

Step 5: If tuning specifications cannot be achieved from previous the steps, do parameter
extraction by only optimizing the auxiliary parameters to obtain:

$$\{x_{aux}^{i}\} = \arg \min_{x_{aux}} \left\| R_f (x_{aux}^i, x_{aux}) - R_f (x_{aux}^i) \right\|$$  (4.4)

In this step, the initial mapping obtained in step 1 between the two spaces of the
fabricated filter and the coarse model is updated.
Step 6: Keep the auxiliary parameters constant and optimize the coarse model tuning parameters until achieving the tuning specifications:

\[
\{x_{ir}^{i+1}\} = \arg \min_{x_{ir}} \left\| R_{i}\left(x_{ir}, x_{aux}^{i}\right) - R_{\text{specification}} \right\|
\]  \hspace{1cm} (4.5)

The fine model tuning parameters are approximated using aggressive space mapping by doing a rank-one update to the latest Broyden. Keep running steps 5 and 6 until the fabricated filter measurements achieve acceptable results.

It is important to note that the proposed method can only achieve tuning specifications within the filter’s tuning range. That is, the method cannot overcome the inherent design structure, components and material limitations.
Chapter 5 Application On a 4-pole Combline Microstrip Filter

In Chapter 4, a post-fabricated filter tuning method with space mapping and interpolation technique is proposed. In this Chapter, an example is given in detail to go through the tuning procedure and show the efficiency of the proposed tuning method [51].

To verify the tuning method proposed in Chapter 4, a four pole varactor-tuned microstrip combline filter is designed and fabricated. The fabricated filter is given in Fig.5.1. The dielectric material is chosen to be 62 mil thick FR4 due to its low cost and unpredictable electrical properties which help in demonstrating the performance of the proposed tuning algorithm. The tuning elements are four tuning varactor diodes (MA6H202). The filter is designed to have a center frequency of 1GHz and an absolute bandwidth of 200MHz at a return loss of 20dB, assuming the dielectric constant is 4.8. The aim is to verify that the method can efficiently estimate the optimum voltage combinations for the fabricated filter to satisfy tuning specifications within its tuning range.
5.1 Design of The Filter

The four pole combline filter is designed using the reflected group delay method, in which resonators are successively added and the group delay response of reflected signal is matched to calculated values from a corresponding low-pass prototype circuit. Compared to traditional filter design method, the reflected group delay method does not have the limitations to filter structures. The filter specifications can be achieved as long as the group delay goals can be satisfied. For example, it is very hard to design a bandpass combline filter with fractional bandwidth of more than 20% using traditional design methods. By using reflected group delay method, it is much easier and efficient to achieve a wider pass-band in the filter design.

The filter was designed and simulated by Sonnet EM simulator using an accurate cell size of 1mil x 1mil. The initial design geometry is shown in Fig.5.2, where 50Ω transmission lines are tapped into the filter to provide input and output couplings. All the resonators share the same electric length of 53 degree and width of 109 mil which is about 50Ω characteristic impedance. The filter is designed to be physically symmetric. The filter design parameters are the tab position $T_{in}$, capacitors $C_1$, $C_2$, and space between adjacent resonators $S1$ and $S2$. 

Fig.5.1 The fabricated 4-pole microstrip combline filter
Fig. 5.2. Geometry of the initial designed filter in Sonnet.

In the initial design, edge vias are used for resonator grounding. However, an edge via is difficult to manufacture. In this example, the filter is fabricated with Advanced Circuits [53], where the standard of grounding is identically provided by a 32 mil diameter
grounding pad and via. The PCB layout is given in Fig. 5.3. The change of the via type can lead to error in the approximation of the electric length and impedance for each resonator. Thus, it is considered to include correct via type during the filter design.

![PCB Layout of the fabricated filter](image)

**Fig. 5.3. PCB Layout of the fabricated filter**

To test the effect of different via types on the response of the designed filter, all of the edge vias in Sonnet EM simulator are changed to the standard 32 mil diameter grounding pads and vias without doing any other changes to filter dimensions. The physical
dimensions can be found in Table 5.1. A simulation results comparison is given in Fig. 5.4.

Table 5.1 Initial designed physical dimensions using edge vias

<table>
<thead>
<tr>
<th>Tin (mil)</th>
<th>C1 (pF)</th>
<th>C2 (pF)</th>
<th>S12 (mil)</th>
<th>S23 (mil)</th>
</tr>
</thead>
<tbody>
<tr>
<td>359.6679</td>
<td>2.6195</td>
<td>2.3905</td>
<td>21.8029</td>
<td>40.3817</td>
</tr>
</tbody>
</table>

Fig. 5.4. Comparison between the filters with same physical dimensions but different groundings: (a) edge via (blue and pink) (b) 32 mil diameter grounding pad (red and black)

From Fig.5.4 we can see the change of the grounding via lead to big difference of the designed simulation results. Hence, it is very important to keep the simulated geometry close to the fabricated one. A filter with a better layout assumption is redesigned, the final design layout and simulation response is given in Fig.5.5 and Fig.5.6. The physical
dimensions for modified design are shown in table 5.2.

Table 5.2 Modified physical dimensions using 32 mil diameter grounding pad

<table>
<thead>
<tr>
<th>Tin (mil)</th>
<th>C1 (pF)</th>
<th>C2 (pF)</th>
<th>S12 (mil)</th>
<th>S23 (mil)</th>
</tr>
</thead>
<tbody>
<tr>
<td>393.9894</td>
<td>2.6363</td>
<td>2.3358</td>
<td>21.112</td>
<td>39.1811</td>
</tr>
</tbody>
</table>

Fig. 5.5. Geometry of the modified filter geometry use 32 mil diameter grounding pad
Fig. 5.6 EM simulation results of the modified filter use 32 mil diameter grounding pad.

The physical layout of the tuning varactor is given in Fig. 5.5. The tuning varactor used in this example is chosen to be MA46H202 made by MA-COM technology solutions. The capacitance and DC voltage relations is shown Fig. 5.8, the varactors have a capacitance range from 0.5pF to 7.0pF over a voltage range from 20V to 0.5V. The listed quality factor at a frequency of 50MHz and a voltage of 4V is 2000.

From Fig. 5.8 we can see the manufacturing datasheet gives a logarithmic linear relations between the loaded DC voltage and the capacitance. Since the extra biasing components, whose layout is shown in Fig. 5.7, have unknown impedance within the filter tuning range, the relations between the voltage and the capacitance of the whole tuning part becomes unknown.
Fig. 5.7 Physical layout of the tuning varactor diodes and biasing circuits

![Diagram of varactor diodes and biasing circuits]

Fig. 5.8 Voltage to capacitance relations for the tuning varactor diode.

![Graph showing capacitance vs. reverse voltage for different varactor diodes]
In this example, the bias components include an RF choke inductor and a DC bias capacitor. The aim of the bias circuit is to make sure that there is no RF signal power goes into the DC control system while the DC control voltage can be loaded across the tuning varactor. In this way, within the designed filter frequency band, the inductors block all the RF signal from the filter to the control system and pass the DC voltage across the tuning varactor diode, the capacitors performed as short circuits to the RF signal and an open circuit to the DC voltages,

The voltage to capacitance relationship given in Fig.5.7 is not guaranteed to be correct because of manufacturing and assembly tolerances. Meanwhile, the biasing elements have unknown parasitics and losses within the testing frequency band. Thus it is very hard to predict the total capacitance and loss of the varactor and bias circuit when implemented with the filter. During the filter design, the whole circuit is modeled by ideal capacitors, So the post-fabricated test results could be very different from the design expectations.

Another unknown of the fabrication process is the dielectric material. For instance, in this example, the dielectric material is chosen to be FR-4, the dielectric constant, which can range from 4 to 5, is assumed to be 4.8. By doing experimental testing on the post-fabricated filter, it is found that the dielectric constant is around 4.4. It is known that the dielectric constant is very important in determining the resonator length, width and spacing between resonators. A wrong assumption in dielectric constant leads to a big center shift of the filter after the fabrication.
Hence, an efficient and fast post fabricated tuning method is required to carry out fine tuning to achieve specific tuning goals. Here, the proposed tuning method presented in Chapter 4 is applied to this post-fabricated filter.

5.2 Post Fabricated Tuning:

The post-fabricated tuning process is traditionally carried out by human technologist where the filter is directly connected to the VNA device and tuned. The tuning varactors are loaded with different DC controlling voltages. Take the combline filter fabricated in Section 5.1 as an example. Two power supplies are used to provide four of the controlling voltages to each of the varactors. The four voltages are roughly tuned until the measurement gives a response in which the $S_{21}$ response is flat in the pass band and the cut off skirt is clearly to observe. After this step the measurement gives a rough band pass filtering response near the pass band. Then the four voltages are slightly tuned to adjust the $S_{11}$ response. This often takes very long time because the $S_{11}$ response is much more sensitive. It is found from the experiment that by changing even 0.05 volts on any of the four voltages will show an effect on the return loss of the filter. It is very difficult and may take 4 to 6 hours for people without tuning experience to manually tune this conventional four pole filter to give a return loss of 18dB for the overall pass band. The manual tuning requires very strong filter background, tuning experience and patience.

The tuning method proposed in Chapter 4 provides a much easier, reliable and automated approach to the tuning of post-fabricated filter. In this section, detailed tuning
procedure of the fabricated 4-pole varactor-tuned combline filter is given to achieve different tuning specifications.

The DC voltages for all varactors are first set to 8 volts \( \{ x^0_p \} = \{8,8,8,8\} \) such that the fabrication measurement of the fabricated filter gives a rough filter like response shown in Fig.5.9. It is important to get a better initial response because it is used for the initial parameter extraction and implicit mapping. A better initial response will significantly reduce the number of space mapping iterations.

![Graph](image)

**Fig.5.9.** Measured results for a rough initial guess (All varactors set to 8 Volts). Blue curve is \( S_{21} \) response. Red curve is \( S_{11} \) response
A coarse model is then implemented in ADS. The coarse model layout is given in Fig.5.10, where ideal capacitors are used to model the capacitance of the varactors. As described in Chapter 4, the losses of the fabricated filter are required to be modeled in the coarse model in order to get a reasonable mapping, in this case, loss tangent of the substrate is used to model all the unpredicted insertion loss from material, fabrication and assembly tolerance.

Fig. 5.10. Coarse model circuit in Keysight ADS
Parameter extraction is then performed in ADS (step 1) by optimizing the coarse model to achieve the measured response shown in Fig.5.11. Thus, an initial mapping is established between the spaces of the fabricated filter and the coarse model. Values of capacitances $\{c^0\}$ and auxiliary parameters $\{x^0\}$ for later mapping updates are extracted. The extracted capacitances are given in Table 5.3.

Fig.5.11 Initial Parameter Extraction Results. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.

The next step is to do sensitivity analysis, in this example, the sensitivity analysis is performed for each of the varactors individually. Even though the varactors are the same type, from experimental results it is found that the voltage-capacitance relationship for
each varactor is quite different. This is caused by manufacturing and assembly tolerances. Thus, all the tuning varactor diodes are required to undergo sensitivity analysis. Hence, the DC voltage across each varactor is first set to 7 volts one by one while voltages for the other three varactors are kept to be 8 V ($\{x^\text{ext}_m\} = \{7,8,8,8\},\{8,7,8,8\}...$).

Parameter extraction is carried out using the coarse model by only optimizing the corresponding ideal capacitor to achieve a rough match to the measurements (step 2, Fig. 5). Please note the exact mapping is very hard to be achieved by only optimizing the capacitors. Our goal is to achieve a good match to the $S_{21}$ response and the positions of poles and zeros of the $S_{11}$ response such that the transfer and reflection polynomials of the two port network in the coarse model is close to the fabricated filter within the tested frequency band. Then the control DC voltage for each varactor is set to 9 V one by one while the other three voltages are kept to 8 V. Parameter extraction are carried out similarly, detailed parameter extraction results are shown in Fig.5.12 to Fig.5.19. (Blue and green curves are measurement results, red and pink curves are parameter extraction results)
Fig. 5.12. Parameter extraction for $x_{\text{fsense}}=[7,8,8,8]$. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.

Fig. 5.13. Parameter extraction for $x_{\text{fsense}}=[8,7,8,8]$. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.
Fig. 5.14. Parameter extraction for $x_{\text{fitense}} = [8, 8, 7, 8]$. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.

Fig. 5.15. Parameter extraction for $x_{\text{fitense}} = [8, 8, 8, 7]$. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.
Fig. 5.16. Parameter extraction for $x_{\text{ftpense}}=[9,8,8,8]$. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.

Fig. 5.17. Parameter extraction for $x_{\text{ftpense}}=[9,9,8,8]$. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.
Fig. 5.18. Parameter extraction for $x_{\text{fsense}} = [8, 8, 9, 8]$. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.

Fig. 5.19. Parameter extraction for $x_{\text{fsense}} = [8, 8, 8, 9]$. Blue and green curves are response of measured result. Red and pink curves are response of extracted result.
Hence, a group of eight capacitance values \( \{x_{ct}^{\text{new}}\} \) (two for each varactor) is extracted in the coarse model. Detailed capacitance values are given in table 5.3. The initial Broyden is then calculated with the help of the Lagrange interpolation technique in Matlab using the values \( \{x_{ct}^0\}, \{x_{ct}^0\}, \{x_{ct}^{\text{new}}\}, \{x_{ct}^{\text{new}}\} \) obtained from steps 1 and 2.

<table>
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<tr>
<th>( {x_{ct}} )</th>
<th>C1</th>
<th>C2</th>
<th>C3</th>
<th>C4</th>
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<td>2.1894586</td>
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</tbody>
</table>

The next step is to optimize the four ideal capacitors to achieve the desired specifications in the coarse model (step 3), then the varactor voltages are calculated using the Broyden obtained in step 3 and the aggressive space mapping technique (step 4). A number of different specifications with center frequencies ranging from 1.05GHz to 1.12
GHz each with an absolute bandwidth of 200MHz are tested with the optimum tuning solutions and measurement results comparison given in Table 5.4 and Fig. 5.20 to Fig.5.25. (red and pink curves are predicted results from the proposed method, Blue and green curves are measurement results.)

Table 5.4 Capacitance and tuning voltage for different specified center with absolute bandwidth of 200 MHz

<table>
<thead>
<tr>
<th>Center Frequency (GHz)</th>
<th>Varactor 1</th>
<th>Varactor 2</th>
<th>Varactor 3</th>
<th>Varactor 4</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>C (pF)</td>
<td>V (V)</td>
<td>C (pF)</td>
<td>V (V)</td>
</tr>
<tr>
<td>1</td>
<td>2.7966</td>
<td>6.8772</td>
<td>2.6447</td>
<td>7.2649</td>
</tr>
<tr>
<td></td>
<td>2.6356</td>
<td>7.5848</td>
<td>2.7915</td>
<td>6.995</td>
</tr>
<tr>
<td>1.03</td>
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<td>7.3454</td>
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</tr>
<tr>
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</tr>
<tr>
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</tr>
<tr>
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</tr>
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<td>1.8847</td>
<td>11.0185</td>
</tr>
</tbody>
</table>
Fig. 5.20. Comparison between the tuning estimation and measurement of 1st iteration results for $f_0=1.0\text{GHz}$ $BW=200\text{MHz}$. Red and pink curves are response of estimated result. Blue and green curves are response of measured result.

Fig. 5.21. Comparison between the tuning estimation and measurement of 1st iteration results for $f_0=1.03\text{GHz}$ $BW=200\text{MHz}$. Red and pink curves are response of estimated result. Blue and green curves are response of measured result.
Fig. 5.22. Comparison between the tuning estimation and measurement of 1st iteration results for $f_0=1.05\text{GHz}$ $BW=200\text{MHz}$. Red and pink curves are response of estimated result. Blue and green curves are response of measured result.

Fig. 5.23. Comparison between the tuning estimation and measurement of 1st iteration results for $f_0=1.08\text{GHz}$ $BW=200\text{MHz}$. Red and pink curves are response of estimated result. Blue and green curves are response of measured result.
Fig. 5.24. Comparison between the tuning estimation and measurement of 1st iteration results for $f_0=1.10\text{GHz}$ $BW=200\text{MHz}$. Red and pink curves are response of estimated result. Blue and green curves are response of measured result.

Fig. 5.25. Comparison between the tuning estimation and measurement of 1st iteration results for $f_0=1.15\text{GHz}$ $BW=200\text{MHz}$. Red and pink curves are response of estimated result. Blue and green curves are response of measured result.
From Fig.5.20 to Fig.5.25, it is obvious that the measurement results are very close to the coarse model simulation results especially within the center frequency ranging from 1.05GHz to 1.10GHz. The locations of the attenuation poles and zeros within filter pass band are well predicted. The dielectric constant is approximated to be around 4.4 such that best return loss can be achieved near the center frequency of 1.08GHz. However, the fabricated filter is a conventional combline filter structure, and the tuning range is very limited. Based on the testing result, the 20dB tuning range for an absolute bandwidth of 200MHz is around 10%. The combined tuning range results are shown in Fig.5.26 and Fig.5.27. It is obvious that this filter design gains a good tuning range for such a conventional combline structure. The insertion is less than 3dB and return loss is approximately 20dB for the given frequency range.
Fig. 5.26. Tuning range test for the 3dB insertion loss

Fig. 5.27. Tuning range test for 20dB return loss
Fig. 5.28. Measured results of the fabricated filter with different center frequencies each with a constant bandwidth of 200 MHz.

For different specifications given in Table 5.3, step 1 and step 2 are identical and only need to be performed once. As only one iteration was needed to reach the solutions given in Table 5.3, step 3 and 4 only needed to be performed once for each specification and step 5 and 6 did not need to be performed at all. This means that each of the tuning results given in Table 5.3 was reached by simply performing an optimization of the capacitance values using the coarse model (step 3) and a simple calculation (step 4) to get the tuning voltages across the varactors.

A second iteration is carried out to see if the tuning results can be further improved by taking more iterations. The results with center frequency of 1.0 GHz and 1.13 GHz
are taken as an example because it seems to be close to the edges of the tuning range of this combline filter with a desired return loss of 20dB. One more iteration was carried out and the results are compared to the first iteration as shown in Fig 5.29 and Fig 5.30. From these results, the result of second iteration slightly changed the return loss but not by much. The return loss level is very similar to the one from first iteration. That means the estimation converges to the best solution and the edges of tuning range for this combline filter is reached.

![Comparison between the results of first and second iteration for f0=1.0 GHz BW=200MHz. Red and pink curves are response of first iteration. Blue and green curves are response of second iteration](image)

Fig.5.29. Comparison between the results of first and second iteration for f0=1.0 GHz BW=200MHz. Red and pink curves are response of first iteration. Blue and green curves are response of second iteration
Fig. 5.30. Comparison between the results of first and second iteration for $f_0=1.13$ GHz BW=200MHz. Red and pink curves are response of first iteration. Blue and green curves are response of second iteration.
Chapter 6 Conclusion

In this thesis, a post-fabrication tuning method is proposed for varactor tuned microstrip tunable filter. The implicit space mapping technique is used to establish a mapping between the space of a post-fabricated tunable filter and a coarse model. An interpolation technique and aggressive space mapping is used to approximate the optimum combinations of tuning parameters for the post-fabricated filter. By applying this method, a post-fabricated tunable filter can be tuned to fit specific filter requirements within a few iterations without expert tuning experience. Meanwhile, an adaptive look up table is created during the tuning process. The tuning decision can be made directly in coarse model without any implementation and measurement to the filter and VNA.

A basic 4-pole varactor-tuned combline filter is designed, fabricated and tuned with the proposed method. By applying the tuning method, the measured response is successfully tuned from a bad response with return loss of 10dB to a very good 20dB response. Several different tuning goals are tested to verify the proposed tuning method. All the tuning goals are achieved within only one iteration.

During the tuning process, due to the precision limitation of multimeter, only 2 decimals are able to be read for voltage greater than 6 volts, it is very hard to get precise DC supplied voltage equals to approximated results which have 4 decimals. But from the comparison it is obvious that the tuned measurement results are very close to the approximated response in the coarse model. Thus, the proposed method is shown to be efficient and accurate enough for the tuning of a tunable filter within its tuning range.
In the recent years, more and more specially designed structures for realizing tunable filters and new tuning components are presented for different tuning goals. In this thesis, the popular 2D combline structure is tested with the proposed method. In the future, the method can be applied to all the presented tunable structures and tuning components to verify the efficiency, accuracy and limitations of the method. Since the method is based on space mapping technique, it is able to support both 2D and 3D tuning problem for any design structure as long as it can be implemented within a coarse model.
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