MINIATURIZED FILTER DESIGN

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Abstract

In many applications, miniaturized filters are highly sought after. Planar filters are preferred as they can be fabricated using low-cost printed circuit technology. A conventional microstrip filter with parallel-coupled half-wavelength resonators is too large to be used in the mobile, personal communication systems. Thus, it is desirable to develop a new microstrip bandpass filter structure that is compact, planar, high and wide upper stop-band rejection.

Microstrip bandpass filters using novel resonator structures are presented. Firstly, a new class of compact microstrip filter loaded with low-characteristic impedance triangular stub is introduced. The respective filter upper stop-band selectivity is improved. Without the presence of cross coupling between non-adjacent resonators, one transmission zero is observed. Frequency tuning can be easily achieved by adjusting the reactance of the stub. Secondly, a new microstrip square open-loop resonator loaded with interdigital capacitive fingers is also proposed. Filters of this type with elliptic function and Chebyshev response are demonstrated. The advantage of the filter presented here not only features smaller in size but also has a wide upper stopband. Simulated and measured fabricated filter performances are presented. There is good agreement between experimental and full-wave electromagnetic (EM) simulation results.

On the other hand, a novel low temperature co-fired ceramics (LTCC) 3D bandpass filter structure was EM simulated based on LTCC manufacturing design rules. The
multilayer filter demonstrated the advantages of the LTCC's buried via process and size compactness, which is not possible to realize with the conventional topology.
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Chapter One

Introduction

1.1 Introduction

Most RF/microwave systems consist of many active and passive components that are difficult to design and manufacture with precise frequency characteristics. In contrast, RF/microwave passive filters can be designed and manufactured with predictable performance. As a result, filters are incorporated in the systems and are usually used to set the system frequency response. Almost all microwave receivers and transmitters require filter. The main filter functions are to reject undesirable signal frequencies outside the filter passband and to channelize or combine different frequencies signals. Filters can be fabricated from lumped elements (inductors, capacitors, and resistors), distributed elements (waveguide, microstrip, or any other medium) or a combination of both. Emerging applications such as wireless mobile and personal communication systems continue to challenge RF/microwave filters with more stringent requirements such as higher performance, smaller size, lighter weight, and low cost. For planar filters, microstrip filters are highly sought after and are preferred, as they can be fabricated using low-cost printed circuit technology. Thus, it is desirable to develop new microstrip bandpass filter structure that is compact and planar.

On the other hand, low temperature co-fired ceramics (LTCC) offers many advantages in achieving higher densities for RF/Microwave integrated circuits. LTCC technology allows the integration of multiple 3 dimensions (3D) passive components
within a single co-fired ceramics. Therefore, it is important to develop miniature RF filter that leverages new LTCC technologies and manufacturing processes.

1.2 Objectives

The objectives of this project are to study, design and develop miniaturized microstrip filters. Two classes of compact microstrip filters are proposed using new resonator structures. The filters not only feature smaller in size but also have a good stop band rejection. The design method is based on coupling coefficients of the inter-coupled resonators and the external quality factors of the input and output resonators. Characterization of the external quality factors and coupling coefficients can be extracted from Electromagnetic (EM) simulations. Design curves obtained using this method are plotted. From the design curves, both parameters related to the physical spacing can be graphically obtained, and the filter designed.

1.3 Scope

The scope of this project can be divided into three main parts. The first part involves the literature search of microstrip filter configurations and methods of filter miniaturization. While part two provides an overview of a general technique for designing coupled resonator filters. New microstrip resonator structures with improved in size and filtering responses are designed and simulated. Design curves are characterized to provide the needed external quality factors and coupling
coefficients to obtain the filter responses. The design methodology of the bandpass filter emphasizes the commercially available full-wave electromagnetic (EM) software simulation. Lastly, an overview of the multilayer LTCC process is presented. The importance in multilayer LTCC bandpass filter to meet the challenges of meeting size and performance are demonstrated.

1.4 Organisation

Chapter 2 provides an overview on the microwave resonators. Since the project is primarily devoted to planar microstrip filters. Emphasis is placed on planar microstrip resonator filters. A brief description of the microstrip-type filter configuration most commonly employed in various microwave components is presented.

Chapter 3 deals with open-loop microstrip resonator structures. Different novel miniaturized open-loop resonators are discussed. A new class of compact microstrip bandpass filter loaded with low-characteristic impedance triangular stub is proposed. Fundamental parameters are defined, together with circuit models. The design procedure for such resonator is presented. In general, coupling coefficients between two resonators are numerically determined by using full-wave electromagnetic (EM) simulations. The relationships between the coupling coefficient of two over-coupled resonators and the resonant peaks observed in the frequency domain by visualizing the resonant frequencies are established. These relationships provide the needed coupling coefficients to obtain the filter response. Design curves that provide the
external quality factor and coupling coefficient between two resonators are presented. Measured and simulated results agreed well.

Chapter 4 proposes another class of miniaturized open-loop microstrip resonator structure. The use of inter-digital capacitive fingers, which loaded within the loop area of the open-loop resonator, enables the presented bandpass filters to have feature smaller in size and also, a good spurious stop-band. Similarly, the coupling coefficient of two coupled resonators can be extracted from the information of the split resonant frequencies using a full-wave EM simulator. Measured and simulated results agreed well.

Chapter 5 discusses multilayer LTCC fabrication technology for microwave applications. LTCC bandpass filter using multilayer stack-up is investigated.

Chapter 6 concludes with discussions on the work done. Some other miniaturized filters techniques are suggested.
1.5 List of Contribution

Based on the works of this research, two papers were submitted and accepted for publication:


Chapter Two

Microstrip Resonators

2.1 Introduction

Microwave resonant structures are extensively used in a variety of applications, such as filters, oscillators and tuned amplifiers [1]. At low frequencies, resonant structures are composed of the lumped elements. As the frequency of operation increases, lumped elements in general cannot be used because the dimensions of lump resonator circuit become comparable to the wavelength and this may cause energy loss by radiation. At microwave frequencies, resonant structures using cavity resonators and microstrip resonators are commonly employed. Since the project is primarily devoted to planar microstrip filter circuits. Emphasis is placed on planar microstrip resonators only. The choice of microstrip resonators for filter design mainly attributed to easy of fabrication, low cost, lightweight, reproducibility and greater flexibility in the design. But its main drawback is its much higher insertion loss compared with the other types of resonators such as cavity and dielectric resonators. This results in difficulty in the design on low loss and high selectivity narrow band filters. Nevertheless, this type of resonator has received great attention from its application in a wide variety of microwave integrated circuit (MIC) components.
2.2 Microstrip Transmission Line

Various forms of planar transmission lines have been developed for use in MIC. The stripline, microstrip line, slot line and coplanar waveguide are some representative planar transmission lines. Microstrip line is one of the most popular types of planar transmission lines that will be described here. The geometry of a microstrip line is shown in Fig. 2.1. A conducting strip (microstrip line) with a width $W$ and a thickness $t$ is on top of a dielectric substrate that has a dielectric constant $\varepsilon_r$ and a thickness $h$. The bottom of the substrate is a conducting ground plane.

![Figure 2.1: Geometry of a microstrip transmission line [6].](image)

The microstrip line is an inhomogeneous transmission line. The field between the strip and the ground plane are not contained entirely in the substrate but extend within the two media namely, the air above and the dielectric below. Therefore, the mode propagating along the microstrip is not purely transverse electromagnetic (TEM) but quasi-TEM. Extensive literature dealing with analytical and numerical solutions of this medium [2] exists. For design purposes, many simple closed-form empirical formulations have been reported [3]. For fast approximation of the microstrip
characteristic impedance, we considered the effect of conducting strip thickness \( t = 0 \) in the calculation. The calculation of the microstrip characteristic impedance can also be obtained using commercial Computer-Aided Design (CAD) software [4,5].

For \( t = 0 \), the closed-form expression that provide accuracy better than 1% are given as follows:

For \( \frac{w}{h} \leq 1 \):

\[
\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{1}{2} \left\{ \frac{1 + 12 \frac{h}{w}}{1} \right\}^{-0.5} + 0.04 \left\{ 1 - \frac{w}{h} \right\}^2, \tag{2.2.1}
\]

\[
Z_c = \frac{60}{\sqrt{\varepsilon_{re}}} \ln \left( \frac{8h}{w} + 0.25 \frac{w}{h} \right), \tag{2.2.2}
\]

For \( \frac{w}{h} \geq 1 \):

\[
\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{1}{2} \left\{ \frac{1 + 12 \frac{h}{w}}{1} \right\}^{-0.5}, \tag{2.2.3}
\]

\[
Z_c = \frac{120\pi}{\sqrt{\varepsilon_{re}}} \left( \frac{w}{h} + 1.393 + 0.677 \ln \left( \frac{w}{h} + 1.444 \right) \right), \tag{2.2.4}
\]

where \( \varepsilon_r \) is the dielectric constant and \( \varepsilon_{re} \) is the effective dielectric constant of a microstrip line respectively.
The loss components of a single microstrip line loss include conductor loss, dielectric loss and radiation loss [6]. The estimation of the attenuation produced by the conductor loss is given by

\[ \alpha_c = \frac{8.686 R_s}{Z_c w} \text{ dB/unit length,} \]  

(2.2.5)

in which \( Z_c \) is the characteristic impedance of the microstrip of the width \( w \), and \( R_s \) is the surface resistivity of the conductor. For a conductor

\[ R_s = \sqrt{\frac{\omega \mu_0}{2\sigma}}, \]

(2.2.6)

where \( \sigma \) is the conductivity, \( \mu_0 \) is the permeability of the free space, and \( \omega \) is the angular frequency. Eq. (2.2.5) is only valid for large strip widths because it assumes that the current distribution across the microstrip is uniform, therefore it would overestimate the conductor loss for narrower microstrip lines. In practical situations, surface roughness and fabrication tolerances can increase the attenuation above theoretical values.

The attenuation due to the dielectric loss can be determined by

\[ \alpha_d = 8.686\pi \left( \frac{\varepsilon_r - 1}{\varepsilon_r + 1} \right) \frac{\varepsilon_r}{\varepsilon_{re}} \frac{\tan \delta}{\lambda_s} \text{ dB/unit length.} \]

(2.2.7)

The total attenuation is

\[ \alpha = \alpha_c + \alpha_d. \]

(2.2.8)
For most microstrip substrates, conductor loss is much greater than dielectric loss.

The upper limit of the operating frequency of a microstrip line is determined by the lowest value of the two frequencies. They are surface wave spurious and higher order modes.

The surface wave mode is given by

\[ f_s = \frac{c \tan^{-1} \varepsilon_r}{\sqrt{2\pi h (\varepsilon_r - 1)}}. \]  \hspace{1cm} (2.2.9)

The first higher-order mode is given by

\[ f_c = \frac{c}{\sqrt{\varepsilon_r (2w + 0.8h)}}. \]  \hspace{1cm} (2.2.10)

2.3 Coupled Microstrip Lines

Coupled microstrip lines are widely used for implementing microstrip filters and directional couplers [7]. Fig. 2.2 shows the cross section of a pair of coupled microstrip lines. The two microstrip lines of width w are placed side by side with a separation s on a dielectric substrate above a ground plane. The microstrip lines can be arranged in parallel or edge-coupled configuration. This coupled line structure supports two quasi-TEM modes. They are the even and odd-mode excitations. Fig. 2.3 illustrates the field distribution for the even and odd-modes on coupled microstrip lines [7]. In the even-mode excitation, both microstrip conductors are of the same potential and on which the same currents exist. The odd-mode excitation corresponds to the conductors being at opposite potential. Thus the currents on the two conductors are equal in amplitude but of opposite sign. The even and odd modes have different
characteristic impedances. Their values become equal when the separation between the conductors is very large, that is the lines are uncoupled. Because coupled microstrip lines are not pure TEM modes, the velocities of propagation of the two modes are unequal. The different in phase velocity result in the microstrip coupled filter having an asymmetrical passband response, deteriorating the upper stopband performance and moves the second passband towards the center frequency [8]. To overcome this problem, capacitive compensation of phase velocity in parallel microstrip filters has been employed.

Figure 2.2: Cross section of a pair of coupled microstrip lines [6].

Figure 2.3: The field distribution for the even and odd modes on coupled microstrip lines [7].
In the static approach, the fundamental mode of wave propagation in a microstrip is assumed to be pure TEM. The characteristic impedances and the effective dielectric constants of the two modes can characterize the coupled microstrip lines. Fig. 2.4 (a) shows the capacitances representation of symmetrical coupled lines. $C_p$ denotes the parallel plate capacitance between the conductor strip and the ground plane. $C_m$ denotes the coupling capacitance between the two conductors. By dividing the circuit along the symmetry axis as shown in Fig. 2.4 (b), the capacitance $C_m$ can be divided into a series circuit of two capacitances, $2C_m$. 
Figure 2.4: (a) representation of capacitances of coupled microstrip lines (b) excitation of symmetrical coupled lines.

In the even-mode excitation, the symmetry plane PP’ as shown in Fig. 2.4(b) acts as magnetic wall (open circuit). The determination of the even-mode capacitance
reduces to finding the capacitance of either line with the plane of symmetry PP' by a magnetic wall as shown in Fig. 2.5 (a). This results in a great simplification of the problem.

Similarly, in the odd-mode excitation, the symmetry plane behaves as an electric wall (short circuit). The determination of the odd-mode capacitance reduces to finding the capacitance of either line by replacing the plane of symmetry by an electric wall as shown in Fig. 2.5 (b).
Figure 2.5: (a) Even-mode excitation (b) Odd-mode excitation.
The even and odd mode capacitances $C_e$ and $C_o$ are given by

\[ C_e = C_p, \text{ and} \]
\[ C_o = C_p + 2C_n. \quad (2.3.1) \]

The even and odd mode characteristic impedances $Z_{ce}$ and $Z_{co}$ can be obtained from the capacitances.

\[ Z_{ce} = \frac{1}{c\sqrt{C_e C_{oe}}}, \text{ and} \]
\[ Z_{co} = \frac{1}{c\sqrt{C_o C_{oo}}}, \quad (2.3.3) \]

where $C_{oe}$ and $C_{oo}$ are the even and odd mode capacitance for the coupled microstrip line configuration with air as dielectric. The effective dielectric constant $\varepsilon_{ree}$ and $\varepsilon_{reo}$ for even and odd mode due to the unequal phase velocities of the two modes are

\[ \varepsilon_{ree} = \frac{C_e}{C_{oe}}, \text{ and} \]
\[ \varepsilon_{reo} = \frac{C_o}{C_{oo}}. \quad (2.3.5) \]

An excitation of symmetric coupled line can be considered as the superposition of even and odd modes. The understanding of basic concepts for microstrip lines and coupled microstrip lines are useful for design of microstrip resonator filters.
2.4 Microstrip Resonator Structures

A microstrip resonator is any structure that is able to contain at least one oscillating electromagnetic field [6]. In general, microstrip resonators for filter designs may be classified as lump-element or quasi-lumped element resonators and distributed line or patch resonators. In this project, we focus on microstrip filters design using both quasi-lump element and distributed line resonators.

Quasi-lump elements can be realized using microstrip line short sections and stubs. The physical lengths must be smaller than a quarter of guided wavelength $\lambda_g$ at which they operate. They are commonly used to realize approximate lumped elements in microstrip filters structures. A good example is the stepped-impedance low-pass microstrip filters [9]. High and low impedance lines are used to approximate series inductors and shunt capacitors respectively.

The distributed line resonator is merely a transmission line that can be formed in various wavelengths [6]. The fundamental resonant frequency $f_0$ of the distributed line resonators can resonate at quarter-wavelength $\frac{\lambda_g}{4}$, half-wavelength $\frac{\lambda_g}{2}$ and one wavelength $\lambda_g$. For quarter-wavelength $\frac{\lambda_g}{4}$ resonators, they can resonate at other higher frequencies when

$$f = (2n - 1)f_0 \Rightarrow n = 1, 2, 3, ....$$

(2.4.1)

For half-wavelength $\frac{\lambda_g}{2}$, they can resonate at other higher frequencies when

$$f = nf_0 \Rightarrow n = 2, 3, ....$$

(2.4.2)
A ring resonator that is formed in a closed loop [10] is a one-wavelength resonator. The ring will resonate at its fundamental frequency \( f_o \) when its median circumference 
\[ 2\pi r \approx \lambda_g, \] where \( r \) is the medium radius of the ring. The higher resonant modes occur at 

\[ f = nf_o \Rightarrow n = 2,3, \ldots \]  

(2.4.3)

### 2.5 Microstrip Bandpass Filters Using Half-wavelength Resonator

In many microwave communications, microstrip coupled line filters are commonly used because of their small size, lightweight and low cost. Fig. 2.6 illustrates some commonly used microstrip coupled line filters configurations using quarter-wavelength \( \lambda_g /4 \) and half-wavelength \( \lambda_g /2 \) resonators. They are end-coupled, parallel-coupled, hairpin-pin, interdigital and combline filters. Both interdigital and combine filters are examples that make used of quarter-wavelength \( \lambda_g /4 \) resonators except the resonator length for combine filter is less than quarter-wavelength because of capacitance loading at the end [9]. The interdigital structure is formed by grounding the resonators on the alternating ends instead of grounding all resonators at the adjacent ends as with the combine. Both filters require ground via holes, which are not compatible with planar fabrication techniques. Thus, in this project, we constrain ourselves to dealings with only half-wavelength \( \lambda_g /2 \) resonators.

The end-coupled bandpass filter is formed with half-wavelength \( \lambda_g /2 \) resonators. The resonators are coupled by means of the capacitive gap between the resonator sections [9]. The disadvantages of this type of filter are large physical circuit size at low
frequency and difficult to realize the required large coupling capacitors when the bandwidth exceeds a few percent.

The parallel coupled bandpass filter is more compact than the end-coupled filter [11]. The half-wavelength $\lambda_g / 2$ resonators are arranged side by side instead of end-to-end. The coupling between the resonators occurs over a quarter-wave long side of each resonator. This arrangement gives larger coupling and the tolerance on the gap is less critical. This filter structure permits a wider bandwidth as compared to the end-coupled filter.

The hairpin filter can be considered to be a folded version of a half-wave parallel-coupled filter [12]. The "U" shape resonators arranged in alternating orientation is much more compact and the filter gives approximately the same performance as parallel-coupled filter. As the frequency increases, the guided wavelength $\lambda_g$-to-width ratio is smaller for a given substrate, so that folding the resonator becomes impractical. Hence, this type of resonator is more suitable for lower frequencies. Also, the effective length of the coupling sections is less than a quarter-wavelength, so the bandwidth is less than the parallel-coupled filter.
Figure 2.6: Typical microstrip coupled line bandpass filter: (a) end-coupled (b) parallel coupled (c) hairpin (d) interdigital (e) combline.
Chapter Three

Microstrip Filter Miniaturization Using Triangular Stub

3.1 Miniaturized Open loop and hairpin Resonator Filters

The expanding wireless and mobile communication systems have presented new challenges to the design of high performance miniature RF filters. Obviously, the size of the conventional planar filters with parallel coupled half-wavelength resonators described in section 2.5 are too large to be used in these systems, where their frequencies of operation fall within L-band (1-2 GHz) and S-band (2-4 GHz). Thus, size reduction has been an important issue in developing these RF filters.

Note that the hairpin filter in Fig. 2.6 make progress in size reduction from the parallel-coupled line filter. Further progress in size reduction is made by the compact miniaturized hairpin resonator [13], where the conventional hairpin resonator in Fig. 3.1(a) is miniaturized by loading a lump-element capacitor between the both ends of the resonator as in Fig. 3.1(b). The capacitance loading reduced the overall length of the resonator. In Fig. 3.1(c), the two arms of the U-shape microstrip line are further folded to form a pair of closely coupled lines to replace the lump capacitor. The improved structure has added advantages compared with the one in Fig. 3.1(b). It has the ability to applied to a much higher frequency range than that using the lump capacitor, because lump capacitor introduces larger circuit loss at higher frequency. Besides, the resonant frequency of the structure can be easily adjusted simply by adjusting the length of the coupled lines to achieve much wider frequencies tuning.
Several research works have been performed to further reduce the miniaturized hairpin resonator or the so called open loop resonator filters using cross couplings between nonadjacent resonators to achieve compact size and elliptic function response [14-17]. The elliptic or quasi-elliptic function filters are able to place transmission zeros near cutoff frequencies of passband so that higher selectivity with less resonators can be obtained. This type of filtering characteristics offers attractive for those systems where size and shape cutoff is important. In chapter four, a miniaturized bandpass filter with this type of filtering characteristic will be discussed.
3.2 New Class of Microstrip Miniaturized Filter Using Triangular Stub

It has been well known that conventional edge parallel-coupled filter and hairpin filter, which are constructed by identical resonators cascaded in alternating series orientation, could only realize Chebyshev and Butterworth responses. It has been noticed that the filter response shows steeper roll-off on the lower frequency side than on the higher frequency side [13,18]. The reason is that the even and odd mode velocities of the microstrip filter are unequal. In [19], a modified parallel-coupled microstrip filter is proposed to improve the filter upper stopband rejection and of the response symmetry.

In this project, we present a novel miniaturized microstrip resonator reactively loaded with triangular stubs. 4-pole resonators arranged in conventional alternating orientation bandpass filter was designed and fabricated. The respective filter upper stop-band selectivity is improved. Without the presence of cross coupling between non-adjacent resonators, one transmission zero is observed. Frequency tuning can be easily achieved by adjusting the reactance of the stub.

3.3 Resonator Structure

Fig. 3.2 shows the resonator structure, which differs from the miniaturised hairpin resonator [13]. The resonator consists of a square ring with the characteristic impedance of the microstrip line and with folded arms first loaded with short open stubs. It was found that further miniaturization of the resonator can be achieved by
symmetrically loaded with triangular stub. This is due to the fact that the shunt capacitive of the stubs reduces the resonant frequency of the resonator.

(a)         (b) (c)

Figure 3.2: (a) conventional square hairpin resonator (b) modified square hairpin resonator having short straight stubs (c) miniaturized square hairpin resonator loaded with triangular stubs.

The resonator can be analysed by considering a capacitively loaded lossless transmission line as in Fig. 3.3. $L$, $\beta_s$ and $Z_s$ are the length, the propagation constant and the characteristic of the unloaded line respectively. The electrical length is $\theta_s = \beta_s L = \frac{2\pi f_0 L}{\sqrt{\varepsilon_r}}/3 \times 10^8$. Consider $C_t$ is the loaded reactance due to triangle stub. The circuit analysis can be described by [20]

Figure 3.3: Equivalent circuit of capacitively loaded transmission line resonator.
The ABCD matrix of the circuit is given by

\[
\begin{bmatrix}
    A & B \\
    C & D
\end{bmatrix} = \begin{bmatrix}
    1 & 0 \\
    \frac{j \omega C_i}{2} & 1
\end{bmatrix} \begin{bmatrix}
    \cos \theta_s & jZ_s \sin \theta_s \\
    j \sin \theta_s & \cos \theta_s
\end{bmatrix} \begin{bmatrix}
    1 & 0 \\
    \frac{j \omega C_i}{2} & 1
\end{bmatrix},
\]  
\tag{3.3.1}

with

\begin{align*}
A &= D = \cos \theta_s - \frac{1}{2} \omega C_i Z_s \sin \theta_s, \\
B &= j Z_s \sin \theta_s, \\
C &= j \left[ \omega C_i \cos \theta_s + \frac{1}{Z_s} \sin \theta_s - \frac{1}{4} \omega^2 C_i^2 Z_s \sin \theta_s \right], \tag{3.3.2, 3.3.3, 3.3.4}\end{align*}

\(\omega = 2\pi f\) is the angular frequency. A, B, C and D are the network parameters of the transmission matrix, which also satisfy the reciprocal condition:

\[AD - BC = 1,\]

for

\[
\begin{bmatrix}
    V_1 \\
    I_1
\end{bmatrix} = \begin{bmatrix}
    A & B \\
    C & D
\end{bmatrix} \begin{bmatrix}
    V_2 \\
    -I_2
\end{bmatrix},
\]  
\tag{3.3.5}

Assume that a standing wave has been excited subject to the boundary conditions \(I_1 = I_2 = 0\). For no vanished \(V_1\) and \(V_2\), it is required that
\[
\frac{C}{A} = \frac{I_1}{V_2} \quad \text{for} \quad I_2 = 0, \tag{3.3.6}
\]
\[
= \frac{I_2}{V_2} \quad \text{for} \quad I_1 = 0,
\]
\[
A = \frac{V_1}{V_2} \Rightarrow I_2 = 0, \tag{3.3.7}
\]
\[
= \begin{cases} -1 & \text{for the fundamental resonance} \\ 1 & \text{for the first spurious resonance} \end{cases}.
\]

From expression (3.3.2)
\[
\cos \theta_{s0} - \frac{1}{2} \omega_0 C_s Z_s \sin \theta_{s0} = -1, \tag{3.3.8}
\]
\[
\cos \theta_{s1} - \frac{1}{2} \omega_1 C_s Z_s \sin \theta_{s1} = -1, \tag{3.3.9}
\]

where \( \theta_{s0} \) and \( \omega_{s0} \) denote the fundamental electrical length and the resonance frequency, and \( \theta_{s1} \) and \( \omega_{s1} \) denote the first spurious resonance electrical length and the resonance frequency. By substituting (3.3.8) and (3.3.9) into (3.3.4), and letting \( C = 0 \) according to (3.3.6), yield
\[
\frac{\omega_{s0} C_s}{2} [1 - \cos \theta_{s0}] = \frac{1}{Z_s} \sin \theta_{s0}, \tag{3.3.10}
\]
\[
\frac{\omega_{s1} C_s}{2} [1 + \cos \theta_{s1}] = \frac{1}{Z_s} \sin \theta_{s1}. \tag{3.3.11}
\]
These two eigenequation can further be expressed as

\[ \theta_{s0} = 2 \tan^{-1} \left( \frac{1}{\pi f_0 Z_s C_t} \right), \]

\[ \Rightarrow \tan \left( \frac{\theta_{s0}}{2} \right) = \left( \frac{1}{\pi f_0 Z_s C_t} \right), \]  \hspace{1cm} (3.3.12)

\[ \theta_{s1} = 2 \pi - 2 \tan^{-1} \left( \pi f_i Z_s C_t \right), \]

\[ \Rightarrow \tan^{-1} \left( \frac{\theta_{s1} - 2 \pi}{2} \right) = \pi f_i Z_s C_t. \]  \hspace{1cm} (3.3.13)

From (3.3.12) and (3.3.13), the fundamental resonant frequency \( f_0 \) and the first spurious resonant frequency \( f_1 \) can be determined. It can be seen from the two solutions that \( \theta_{s0} = \pi \) and \( \theta_{s1} = 2 \pi \) when \( C_t = 0 \), which is the case for the unloaded half-wavelength resonator. For \( C_t > 0 \), it can be shown that the resonant frequencies are shifted down as the loading capacitance is increased. We can achieve resonator miniaturization because the resonator occupies the same size but the resonant frequency is lower depending on the value of loading capacitance.

To provide a more accurate design with less design iterations leading to first-pass, commercially available EM simulator IE3D [21] which uses numerical technique method of moment (MOM) is utilized in the resonator structures and filters design in this project. Fig. 3.4 illustrates the variation of the fundamental resonance frequency due to the different reactance of the triangular stub, which is achieved by tuning the angle \( \alpha \) or \( b \) as shown in Fig. 3.3. The square resonator's \( L \) is chosen to be 7.35 mm and the width of the microstrip is determined by its characteristic impedance to 50 ohms. RT/Duriod substrate having a thickness of 0.635 mm and a dielectric constant
of 10.2 is used in the simulation. The simulated results in Fig. 3.4 reveals that the fundamental resonance frequency is shifted down as the "fan-angle" of the triangular stubs increases.

Figure 3.4: Simulated fundamental resonance frequency as a function of b (triangular stub).
Figure 3.5: Structure of the square microstrip resonator loaded with reactive triangular stub (bold-tie).

From Fig. 3.5 (as in Fig. 3.3), we consider $C_t$ as the total loaded reactance due to triangle stub, the fringing capacitance between open-end triangle stub and the adjacent microstrip line, and the mode capacitance of the folded-coupled lines. The input reactance $X_{in}$ corresponding to opposite potential excitation (odd mode excitation) along plane $rr'$ can be extracted by using EM simulator. The total loaded shunt capacitance $C_t$ is given by

$$X_{in} = \frac{1}{j2\pi f_0 \frac{C_t}{2}}.$$  \hspace{1cm} (3.3.14)

$C_t$ can be theoretical estimated by solving the following equation from (3.3.12)

$$\tan\left(\frac{\theta_0}{2}\right) = \left(\frac{1}{\pi f_0 Z_s C_t}\right).$$
Design parameters, EM extracted and theoretical calculated $C_t$ are tabulated in Table 1. As it can be seen that loading capacitance is increased, the resonant frequencies are shifted down as predicted. A plot between EM extracted and theoretical calculated $C_t$ against $b$ is shown in Fig. 3.6. The expected lower theoretical capacitance values as compared to the EM extracted values can be explained by the fact that the $C_t$ does not take into accounts of the fringing capacitance between open-end triangle stub and the adjacent microstrip line, and the mode capacitance of the folded-coupled lines. However, as $b$ decreases, the respective fringing capacitance is reduced which results in a smaller error.

Table 1
Design parameters for resonator with $L = 7.35$ mm

<table>
<thead>
<tr>
<th>$b$ [mm]</th>
<th>$f_0$ [GHz]</th>
<th>$\theta_s$ [deg.]</th>
<th>EM extracted $C_t$ [pF]</th>
<th>Theoretical $C_t$ [pF]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.75</td>
<td>1.44</td>
<td>32.87</td>
<td>15.50</td>
<td>14.98</td>
</tr>
<tr>
<td>2.75</td>
<td>1.38</td>
<td>31.50</td>
<td>17.58</td>
<td>16.35</td>
</tr>
<tr>
<td>3.75</td>
<td>1.32</td>
<td>30.14</td>
<td>19.96</td>
<td>18</td>
</tr>
<tr>
<td>4.75</td>
<td>1.26</td>
<td>28.77</td>
<td>22.32</td>
<td>19.70</td>
</tr>
<tr>
<td>5.75</td>
<td>1.22</td>
<td>27.85</td>
<td>24.80</td>
<td>21.05</td>
</tr>
</tbody>
</table>
3.4 4-pole Bandpass Filter Design

In the design we have chosen the size of the resonator to be $L \approx \frac{\lambda g}{13} = 7.35 \text{ mm}$, $b = 5.75 \text{ mm}$ at the resonance frequency $f_0 = 1.22 \text{ GHz}$ as illustrated earlier. A 4-pole Chebyshev 0.01dB ripple bandpass filter center at 1.22 GHz with a 3dB bandwidth of 6% or fractional bandwidth (FBW) of 0.06 has been designed and fabricated on a RT/Duriod substrate having a thickness of 0.635mm and a dielectric constant of 10.2. Bandpass filters may be defined by only three entities. The resonator structure, coupling between resonators (coupling coefficients) and coupling to the terminations (external quality factor). Design procedures based on these concepts can derived from
the lowpass prototype values, $g$ [9]. The lump values of the lowpass prototype filter
are found to be $g_0 = 1$, $g_1 = 0.7128$, $g_2 = 1.2003$, $g_3 = 1.3212$, $g_4 = 0.6476$ and $g_5 = 1.1007$. The coupling coefficients and external quality factor can be found to be $K_{12} = K_{34} = 0.0648$, $K_{23} = 0.0476$, $Q_e = 11.88$ by

$$Q_{en} = \frac{g_n g_{n+1}}{FBW},$$  

$$Q_{en} = \frac{g_n g_{n+1}}{FBW},$$  

$$K_{i,i+1} = \frac{FBW}{\sqrt{g_i g_{i+1}}} \text{ for } i = 1 \text{ to } n - 1,$$  

$$FBW = \frac{\omega_2 - \omega_1}{\omega_0},$$  

$$\omega_0 = \sqrt{\omega_1 \omega_2} \text{ or } f_0 = \sqrt{f_1 f_2},$$

where $n$ is the degree of the filter, $\omega_1$ and $\omega_2$ denote the passband-edge angular frequency and $\omega_0$ denotes the center angular frequency.

Using EM simulations, we then extracted the external quality factor $Q_e$ and coupling coefficient $K$ against the physical dimensions [6]. The filter is designed to have tapped line input and output couplings. The tapped line is chosen to have characteristic impedance that matches to terminating impedance 50 ohms. In Fig. 3.7, couplings or the external quality factor $Q_e$ are determined by the position of the microstrip tap location along the resonator denoted by $t$. The $Q_e$ is extracted based on the phase of S11 parameter at the excitation port of the resonator [6].
The $Q_e$ is given by

$$Q_e = \frac{f_0}{\Delta f_{\pm 90^\circ}},$$  \hspace{1cm} (3.3.20)

where $f_0$ and $\Delta f_{\pm 90^\circ}$ are the resonant frequency and the phase shift $\pm 90^\circ$ corresponds to the bandwidth of the input or output resonator. A plot of the phase of $S_{11}$ as a function of $\Delta f_{\pm 90^\circ}$ is given in Fig. 3.8. The design curve on Fig. 3.9 gives the value of external quality factor $Q_e$ as a function of $t$.

Figure 3.7: Input or output tapped-line coupling structure.
Figure 3.8: Phase of S11 response with t = 6 mm simulated from the structure shown in Figure 3.7.

Figure 3.9: External quality factor $Q_e$ as a function of t.
The value of coupling coefficient $K$ is characterized against the coupling spacing denoted by $s$ between two adjacent resonators with opposite orientations as shown in Fig. 3.10. Using EM simulation, two splitting resonant frequencies are used to determine the coupling coefficient of the resonator pair as illustrated in Fig. 3.11. The two split frequencies where $f_{p1}$ and $f_{p2}$ are the lower and higher split resonant frequencies of a pair of coupled resonators, which is related to the coupling coefficient $K_{ij}$ as [17]

$$K_{ij} = \frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2}.$$  \hspace{1cm} (3.3.21)

It has been shown that as the coupling space $s$ decreases the two resonant peaks move outwards and the depression in the middle deepens. This implies an increase in the coupling. Fig. 3.12 shows the characterized coupling coefficient as a function of physical spacing, $s$. 

Figure 3.10: Coupling structure.
Figure 3.11: Resonance response for the determination of coupling coefficient

Figure 3.12: Simulated coupling coefficient
3.5 Simulation and Measurement

The simulated frequency response of the 4-pole bandpass filter as a function of \( b \) is shown in Fig. 3.13. A transmission zero at the high frequency side is clearly observed. This is due to the transition point between the magnetic and electric couplings of the resonators. The frequency response is better than conventional hairpin resonator filter [18]. The 1.22 GHz filter for \( b=5.75\text{mm} \) was fabricated and measured. The optimized spacing designated value of \( K_{12} \) and \( K_{23} \) is 0.45 mm and 0.7 mm respectively. The measured upper stopband achieved good rejection of about 80 dB. The network analyzer dynamic range limits the upper passband attenuation. The passband loss is approximately 3.7 dB, which is mainly due to the conductor loss and the dielectric loss of the substrate. Fig. 3.14 shows the photograph of the fabricated filter.
Figure 3.13: simulated and measured frequency responses of the 4-pole bandpass filter:

(1) $b=2.75\text{mm}$

(2) $b=3.75\text{mm}$

(3) $b=4.75\text{mm}$

(4) simulated and measured responses of the filter with $b=5.75\text{mm}$

Figure 3.14: Photo of the fabricated filter.
Chapter Four

Miniaturization Open Loop Microstrip Filter loaded with Capacitive Inter-digital Fingers

4.1 Introduction

Microstrip line with both ends capacitively loaded with a pair of rectangular open-stubs [20] and a pair of triangular stub (chapter three) has demonstrated to reduce the circuit size. On the other hand, capacitively loaded microstrip and CPW loop resonator has been investigated in [22-23]. Both resonators removed the internal part of the central strip of a conventional microstrip/CPW half-wavelength resonator. This effectively turns the standard resonator into a loop resonator. The fundamental resonance frequency might be estimated by assuming that its mean circumference equals the guided wavelength provided that the loop strip width is much smaller than the width of the associated half-wavelength resonator. The effect of this process on the slow-wave is only small. To reduce the frequency of the resonator, it is loaded with capacitive fingers. The velocity reduction on this type of transmission line is control by the number of fingers within the narrow loop strip width. The resonator is a closed-loop structure. They can only be implemented in a few filtering configurations. Conventionally, parallel-coupled line or end-coupled line filters design approaches can be used in this case [24].
In order to enhance filter performance, various filter configurations including elliptic or quasi-elliptic response would be desired. We proposed to use the inter-digital capacitive fingers, which are loaded in the open-loop resonator to achieve cross-coupled slow-wave microstrip elliptic function and Chebyshev filters. In [22,23], the resonator is a transmission line modified with etched fingers to achieve slow-wave effect and miniaturization. Here, the attempt is to provide close capacitive coupling between two arms of the transmission lines using added fingers to create a new miniaturize resonator. This method of enhancing the capacitively loaded microstrip open-loop resonator enables the presented filters not only feature smaller in size but also has a good spurious stop-band.

4.2 Resonator Structure

Fig. 4.1(a) shows the conventional square open-loop resonator, the characteristic impedance of the transmission line determines the width of the microstrip. For a large open-end gap, the fundamental resonance is approximately a half-guided wavelength. When the gap is small, the length of the transmission line and the gap capacitance determines the resonance frequency. In order to reduce the resonance frequency further, inter-digital fingers loaded between the two arms of the transmission line are used to increase the capacitance loading as shown in Fig. 4.1(b). It must be further emphasized that the fingers are added-in as opposed to etched-off fingers within the width of the transmission line. The capacitance can be increased by increasing the number of fingers. It then can be expected that the resonance frequency of the loaded square open-loop resonator will be shifted down as the number of fingers increases.
For the determination of the reduction factor, Fig. 4.2 shows the full-wave IE3D simulated frequency responses of the capacitively loaded square open-loop resonators with \( N = 4, 6, 8 \), \( N \) the number of the fingers, and conventional square open-loop resonator. The substrate used RT/Duriod dielectric thickness of 1.27 mm with relative dielectric constant of 10.5. All resonators occupy the same surface area but as can be seen, the resonance frequency is shifted down from 2.47 GHz of the conventional square open-loop resonator to 1.54 GHz of the loaded resonator with 8 fingers. A reduction factor of \( \frac{2.47}{1.54} = 1.6 \) has been obtained, based on same surface area.

Figure 4.1: (a) conventional square open-loop resonator (b) capacitively loaded square open-loop resonator. Finger length (fl) = 4.5 mm, finger width (fw) and spacing (fs) = 0.3 mm.
4.3 Cross-Coupled Filter

There have been increasing demands for narrow band RF/microwave bandpass filters with high selectivity other than conventional Chebyshev filters in order to meet stringent requirements from wireless communications systems [6]. Cross-coupled filters using cross couplings between nonadjacent resonators have achieved compact size and elliptic function response [14-17,25]. The elliptic or quasi-elliptic function filters are able to place transmission zeros or attenuation poles near cutoff frequencies of passband so that higher selectivity with less resonators can be obtained.

Selective filters with a pair of transmission zeros may be realized by cross coupling a pair of nonadjacent resonators of the standard Chebyshev filter [26]. An approximate synthesis method described in [26] based on a lowpass prototype filter is shown in
Fig. 4.3. The rectangular boxes represent ideal admittance inverters with characteristic admittance J.

Figure 4.3: Lowpass prototype filter.
The element values for the Chebyshev filters are given by the well-known formulas

\[
g_1 = \frac{2 \sin \frac{\pi}{2n}}{\gamma}, \tag{4.2.1}
\]

\[
g_i, g_{i-1} = \frac{4 \sin \frac{(2i-1)\pi}{2n} \sin \frac{(2i-3)\pi}{2n}}{\gamma^2 + \sin^2 \left(\frac{(i-1)\pi}{n}\right)} for i = 1,2,\ldots,m, \quad m = \frac{n}{2}, \tag{4.2.2}
\]

\[
\gamma = \sinh \left(\frac{1}{n} \sinh^{-1} \frac{1}{r}\right), \tag{4.2.3}
\]

\[
r = \frac{1}{\sqrt{\frac{19}{10} - 1}}, \tag{4.2.4}
\]

\[
S = \left(\sqrt{1 + r^2} + r\right)^2 \text{ (the passband VSWR)}, \tag{4.2.5}
\]

\[
J_m = \frac{1}{\sqrt{S}} \text{ for } m \text{ even}, \tag{4.2.6}
\]

\[
J_{m-1} = 0 \text{ for Chebyshev filters}, \tag{4.2.7}
\]

where \(r\) is a passband ripple constant, \(L_R\) is the passband return loss in dB and \(n\) is the degree of the filter. The frequency response of the bandpass filter can be determined by using a well-known lowpass to bandpass mapping:

\[
\Omega = \frac{1}{FBW} \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right), \tag{4.2.8}
\]

\[
FBW = \frac{\omega_2 - \omega_1}{\omega_0}, \tag{4.2.9}
\]

\[
\omega_0 = \sqrt{\omega_2 \omega_1}, \tag{4.2.10}
\]
where $\Omega$ is the frequency variable that is normalized to the passband cut-off frequency of the lowpass prototype filter. $\omega$ is the frequency variable of bandpass filter, $\omega_0$ is the midband frequency, $\omega_1$ and $\omega_2$ denote the passband-edge and FBW is the fractional bandwidth.

In order to introduce zeros at $\Omega = \pm \Omega_a$, the required value of $J_{m-1}$ is given by

$$J_{m-1} = -\frac{J_{m}^{\prime}}{(\Omega_{a}g_{m})^2 - J_{m}^{\prime 2}},$$

(4.2.11)

where $J_{m}^{\prime}$ is a slightly changed value of $J_m$ as

$$J_{m}^{\prime} = \frac{J_m}{1 + J_m J_{m-1}}.$$ (4.2.12)

Introduction of $J_{m-1}$ mismatches the filter, and to maintain a good VSWR at midband it is necessary to change the value of $J_m$ according to the formula (4.2.4). No other elements of the original Chebyshev filter are changed.

In [25], a design method for this class of microstrip highly selective bandpass filters is introduced, including tables and formulas for accurate and faster filter synthesis. For less selective filters that require a larger $\Omega_a$, the element values can be obtained using the approximate synthesis procedure as described above.

The locations of the two finite frequency attenuation poles or transmission zeros of the bandpass filter can be calculated by [25]
It has been shown that the closer the attenuation poles to the cut-off frequency $\Omega = \pm \Omega_a \rightarrow 1$, the sharper the filter skirt and higher the selectivity. If $\pm \Omega_a \rightarrow \infty$, the bandpass filter response degenerates to the familiar Chebyshev response [25].

\[ \omega_{a1} = \omega_0 - \Omega_a FBW + \frac{\sqrt{(\Omega_a FBW)^2 + 4}}{2}, \text{ and} \]

\[ \omega_{a2} = \omega_0 - \Omega_a FBW + \frac{\sqrt{(\Omega_a FBW)^2 + 4}}{2}. \]

4.4 4-pole Cross-Coupled Filter Design

The new square open-loop resonator loaded with 8 fingers depicted in Fig. 4.1(b) is used to demonstrate a cross-coupled elliptic function response filter. The 4-pole filter has a center frequency of 1.54 GHz and a fractional bandwidth FBW of 4%. We required the stopband rejection of at least 20 dB at $\pm 60$ MHz from the center frequency. The filter was fabricated on RT/Duriod substrate thickness of 1.27mm with relative dielectric constant of 10.5. EM simulator IE3D was used to simulate and design the filter.

The 4-pole filter with a pair of attenuation poles at $\Omega = \pm \Omega_a$ is calculated to be $\pm 1.98$. Using the tabulated data in Table 2 as described in [25], for $\Omega_a = 2$ the lump-element values were determined as $g_1 = 0.95449$, $g_2 = 1.38235$, $J_1 = -0.16271$, and $J_2 = 1.06062$. The locations of the two finite frequency attenuation poles of the bandpass filter can also be determined by (4.2.5). Shown in Fig. 4.4 (a) and (b) is the cross-
coupled filter configuration and the equivalent circuit, where K14 denotes the cross coupling between the input and output resonators. The external quality factor $Q_e$ and the coupling coefficients can then be found by

$$Q_e = \frac{g_1}{FBW},$$  

(4.2.15)

$$K12 = K34 = \frac{FBW}{\sqrt{g_1 \cdot g_2}},$$  

(4.2.16)

$$K23 = \frac{FBW \cdot J_2}{g_2}, and$$

(4.2.17)

$$K14 = \frac{FBW \cdot J_1}{g_1}.$$  

(4.2.18)

Similarly, the external quality factor $Q_e$ is determined by the position of the microstrip tap location along the resonator denoted by t as described in section 3.4. The $Q_e$ is extracted based on the phase of S11 parameter at the excitation port of the resonator. The design curve on Fig. 4.5 gives the value of external quality factor $Q_e$ as a function of t.
Table 2: Element values of a 4-pole prototype ($L_R = -20$ dB)

<table>
<thead>
<tr>
<th>$\Omega_\alpha$</th>
<th>$g_1$</th>
<th>$g_2$</th>
<th>$J_1$</th>
<th>$J_2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.80</td>
<td>0.95974</td>
<td>1.42192</td>
<td>-0.21083</td>
<td>1.11769</td>
</tr>
<tr>
<td>1.85</td>
<td>0.95826</td>
<td>1.40972</td>
<td>-0.19685</td>
<td>1.10048</td>
</tr>
<tr>
<td>1.90</td>
<td>0.95691</td>
<td>1.39927</td>
<td>-0.18429</td>
<td>1.08548</td>
</tr>
<tr>
<td>1.95</td>
<td>0.95565</td>
<td>1.39025</td>
<td>-0.17297</td>
<td>1.07232</td>
</tr>
<tr>
<td>2.00</td>
<td>0.95449</td>
<td>1.38235</td>
<td>-0.16271</td>
<td>1.06062</td>
</tr>
<tr>
<td>2.05</td>
<td>0.95341</td>
<td>1.37543</td>
<td>-0.15337</td>
<td>1.05022</td>
</tr>
<tr>
<td>2.10</td>
<td>0.95242</td>
<td>1.36934</td>
<td>-0.14487</td>
<td>1.04094</td>
</tr>
<tr>
<td>2.15</td>
<td>0.95148</td>
<td>1.36391</td>
<td>-0.13707</td>
<td>1.03256</td>
</tr>
<tr>
<td>2.20</td>
<td>0.95063</td>
<td>1.35908</td>
<td>-0.12992</td>
<td>1.02499</td>
</tr>
<tr>
<td>2.25</td>
<td>0.94982</td>
<td>1.35473</td>
<td>-0.12333</td>
<td>1.0181</td>
</tr>
<tr>
<td>2.30</td>
<td>0.94908</td>
<td>1.35084</td>
<td>-0.11726</td>
<td>1.01187</td>
</tr>
<tr>
<td>2.35</td>
<td>0.94837</td>
<td>1.3473</td>
<td>-0.11163</td>
<td>1.00613</td>
</tr>
<tr>
<td>2.40</td>
<td>0.94772</td>
<td>1.34408</td>
<td>-0.10642</td>
<td>1.00086</td>
</tr>
</tbody>
</table>
Figure 4.4: (a) 4-pole cross coupled filter configuration (b) the equivalent circuit.
Figure 4.5: (a) input or output coupling structure (b) value of external quality factor $Q_e$ as a function of $t$. 
Fig. 4.6 shows the characterized three basic coupling structures encountered in the cross-coupled filter. The coupled structures result from different orientations of a pair of identical resonators which are separated by a spacing $s$. It is obvious that any coupling in those coupling structures is through fringing fields. The coupling structure in Fig. 4.6(a) is for electric coupling because the electric fringe fields are stronger at the open ends of the resonators. At the middle of the resonator of Fig. 4.6(b), the fringe field is mainly magnetic coupling. For the coupling structure in Fig. 4.6(c), the electric and magnetic fringe fields at the coupled sides may have comparative distributions so that both the electric and magnetic couplings occur. In this case, the coupling may be referred to as mixed coupling. The details formulation of such relationships for the coupled structures can be found in [17]. The coupling coefficient $K_{ij}$ of any pair of adjacent resonators can be determined by

$$K_{ij} = \pm \frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2},$$  \hspace{1cm} (4.2.19)$$

where $f_{p1}$ and $f_{p2}$ are the lower and higher split resonant frequencies of a pair of coupled resonators when they are decoupled from the rest. When the coupling space $s$ decreases, the two resonant peaks move outwards and the depression in the middle deepens, which implies an increase in the coupling. The (+) upper sign applies to $K_{12} = K_{34}$ and $K_{23}$, while the lower sign (-) applies to $K_{14}$. Fig. 4.7 shows the computed coupling coefficients for the three different coupling structures, namely, the electric, the magnetic and the mixed coupling.
Figure 4.6: Three basic coupling structures. (a) electric coupling (b) magnetic coupling (c) mixed coupling.
4.5 Simulation and Measurement

Fig. 4.8 shows the layout, simulated and measured performances of the filter. The spacing designated in the cross-coupled filter are as follows: $d_{12} = 1$mm, $d_{23} = 1.7$mm, $d_{14} = 1.3$mm and, the offset $d = (K_{14} - K_{23})/2 = 0.2$mm. The mixed coupling has little change against the offset $d$. Additionally, a 4-pole Chebyshev filter was designed. Fig. 4.9 shows the layout, simulated and measured performances of the filter. The elements of the Chebyshev could be synthesized using standard Chebyshev filter. The designated value of $d_{12}$ and $d_{23}$ in the Chebyshev filter is $1$mm and $1.3$mm respectively. It was found that frequency response of the filter could be effectively fine-tuned by adjusting the open gaps of the resonator 1 and 4. The effects on the
center frequency and bandwidth are negligible as long as the gap tuning is small. The measured passband insertion loss is \(-3\text{dB}\). The experimental and simulation results are in good agreement. Fig. 4.10 shows the simulated and measured spurious characteristics of the filters. Both filters exhibit a good rejection at second harmonic as predicted. The second passband is not harmonically related because of the interdigital fingers, which can be approximated as a semi-lumped capacitor. Fig. 4.11 shows the fabricated filters.

Figure 4.8: Simulated and measured performances of the 4-pole cross-coupled bandpass
Fig. 4.9: Simulated and measured performances of the 4-pole Chebyshev filter.
Fig. 4.10: Simulated and measured spurious responses of the cross-coupled and Chebyshev filters.

Figure 4.11: Photo of the fabricated filters.
Chapter Five

Multilayer Filters

5.1 Multilayer Low-Temperature Co-fired Ceramic (LTCC) Substrates

Miniaturized filters using multi-layer technology provides another dimension in the flexible design and integration. Although planar filters are preferred as they can be fabricated using low-cost printed circuit technology, however, multilayer low temperature co-fired ceramics (LTCC) offers many advantages in achieving higher densities for RF/Microwave integrated circuits. LTCC technology allows the integration of multiple passive components within a single co-fired ceramics, exploiting the third dimension, resulting in Multi-Chip Module (MCM). Multi-layer 3D embedded passive circuits such as filters and baluns have been demonstrated [27,28]. Compact lightweight RF front-end module using LTCC multi-layer packaging with integrated RF interconnects, buried passive circuits, and MMIC has become a demand [29]. Therefore, it is important to develop miniature RF filter that leverages the new LTCC technologies and manufacturing processes.

Fig. 5.1 illustrates the cross-sectional view of a LTCC multi-layer RF front-end MCM. In such MCM, deep embedded passive components such as capacitors and resonators are used very frequently. All of the bias, control and RF interconnections, along with buried passive RF circuits are integrated onto a single LTCC substrate. The connection between the ground plane of the surface mounted MMIC and the baseplate is typically accomplished with vias. Exposed RF interconnection layer of
the substrate which are typically microstrip and GCPW along with vertical staggered vias, are used to connect individual MMIC chips and embedded components. One major advantage of the LTCC is the ability to fabricate buried vias, which is not possible with current conventional multilayer printed circuit board process.

![Cross-sectional view of a LTCC multilayer RF front-end MCM.](image)

5.2 LTCC Tape Systems

The relatively low firing temperature of LTCC dielectrics allows high conductivity materials such as gold and silver to be included in the conductor metallization for the internal layers. The combination of the low dielectric loss tangent and high conductivity metallization results in a substrate media with excellent RF and microwave characteristics. The characterization of different LTCC materials and benchmarks its properties vs 96% and 99% Alumina, and PTFE and Polyester based organic dielectric materials for microwave applications have been reported in [30-32]. Here, we summarized some properties of the common commercial LTCC tape materials as shown in Table 3.
Table 3: LTCC Tape Systems based on LTCC Design Guidelines

<table>
<thead>
<tr>
<th></th>
<th>Dupont 951</th>
<th>Dupont 943</th>
<th>Ferro A6M</th>
<th>Heraeus CT2000</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Electrical Properties</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Dielectric Constant</td>
<td>7.85</td>
<td>7.5</td>
<td>5.9</td>
<td>9.1</td>
</tr>
<tr>
<td>Loss tangent</td>
<td>0.0045</td>
<td>0.001</td>
<td>0.002</td>
<td>0.002</td>
</tr>
</tbody>
</table>

| **Dimensional Characteristics** |            |            |           |                |
| Tape layer Thickness- Fired (mils) | 1.7, 3.8, 5.5, 8.5 | 4.5 | 3.7, 7.4 | 1.77, 3.45 |
| Shrinkage x, y          | 12.7% ± 0.3 | 9.5% ± 0.3 | 14.8% ± 0.2 | 11.5%         |
| Shrinkage Z             | 15% ± 0.5 | 10.3% ± 0.3 | 27% ± 0.5 | 14% |
| Colour                  | Blue       | Light Blue | White     | Blue          |

From the above loss tangent data, it is apparent that dupont 951 tape is suitable to be used in many wireless applications where the frequency range up to 5 GHz. Above 5 GHz to 10 GHz, the loss is high so design using the tape must be accounted in the application.

For microwave and low loss application frequency range from 20 GHz above, dupont 943, Ferro A6 and Heraeus CT2000 provide the highest performance. Ferro A6 is stable with uniform properties over broad frequency and environment range. It has been in the market for a long time. Dupont 943 and Heraeus CT2000 are new tapes released a few years back.
LTCC material systems are composed of glass/ceramic dielectric tape along with metallization pastes that are formulated to provide shrinkage characteristics similar to the dielectric. Processing of all the individual LTCC substrate layers is performed in parallel with the dielectric tape still in its “green state”. The process flow for the LTCC is shown in Fig. 5.2. Each layer of tape is blanked to size, and registration holes punched. Vias are formed in the dielectric tape by punching or drilling, while conductor traces and via fills are screen printed. When all circuit layers have been punched, printed and inspected, the layers are stacked, laminated and then the entire structure is fired in a single time. The process allows greater layer capability and reliability. This differs from traditional thick film technology where each layer is processed and fired sequentially, and the layer capability is process limited to typically 3 to 5 layers.
Figure 5.2: LTCC process flow
5.3 Design Trade-off and Considerations

While the use of LTCC technology is an attractive option, there are a number of issues that must be carefully considered as part of the design of a multilayer substrate that must support high frequency signals. Electrically, any potentially resonant structures must be eliminated from the substrate layout. This requires careful consideration of the physical spacing and proximity of the RF ground structures, bias connections, and control connections. Also, the formation of cut-out/cavity surrounded by ground via’s will result in resonant cavities. Cavity dimensions must therefore be chosen such that no waveguides modes can be supported in the band of operation.

Another key electrical consideration is RF transmission loss. Although the LTCC dielectric and metallization materials possess good low loss characteristics, actual RF losses can still be relatively high depending upon the thick-film metal deposition processes associated with the LTCC system. Typical thick film processes result in metallization patterns with high surface roughness and poor edge definition as compared to the thin film processes traditionally employed for microwave ceramic substrates. Due to the skin effect, this results in higher conductor (resistive) loss, which is the dominant RF loss mechanism in a multilayer LTCC substrate. An ideal RF layout would be such that the lengths of all inter-connecting transmission lines are minimized. Also, the RF ground plane spacing can be increased, which results in physically wider and therefore lower loss transmission lines. However, the maximum ground plane spacing is limited by factors such as waveguide moding, maximum substrate thickness, and substrate cost.
5.4 LTCC Multilayer Resonator

Consider the size of the IF filter centered at 1.2 GHz to be integrated into the MCM is constrained. For a dielectric constant of 5.9, the effective wavelength is approximately 26% and 30% longer compared to higher dielectric constant of 10.2 and 10.5 at the same frequency, respectively. The filter design using microstrip topology described in chapters 3 and 4 will not able to meet the size requirement. Hence, multi-layer resonator structure is needed to miniaturize the filter. Here, multi-layer inter-digital parallel plates capacitor is used to realize the capacitance loading effect so as to reduce the resonator’s size. The resonator’s folded arms are symmetrically loaded with rectangular multi-layer parallel plate’s capacitor connected by means of buried vias, which takes advantages of the LTCC’s buried via process. Fig. 5.3 shows the 3D exploded view of the resonator. Fig. 5.4 shows the different views of the multilayer stack-up of the resonator. The size of the resonator is 250 x 255 mil (6.35 x 6.48 mm). The electrical length corresponds to resonant frequency of 1.2 GHz is $\theta_s = 65$ degree. The variation of the fundamental resonance frequency due to the different $dy$ is characterized by EM as shown in Fig. 5.5, where $dy$ is the width of the parallel plate's capacitor. The simulated responses reveal that the fundamental resonance frequency is shifted down with increasing $dy$, which results in increasing capacitance loading. The LTCC tape system used is Ferro A6 with a dielectric constant of 5.9 (Table 3). The capacitance of the multilayer parallel-plate related to geometry can be expressed by equation [33].
\[ C = \frac{0.2249 \epsilon_r A(n-1)}{d} \text{ pF}, \]  

(5.4.1)

where

\( A = \text{area of plates in square inches,} \)

\( n = \text{no. of conductor layers, and} \)

\( d = \text{plate spacing} \)
Figure 5.3: Multilayer stack-up of the resonator: (a) planar view (b) front view (c) side view.
Figure 5.5: Simulated resonance frequency as a function of dy.

5.5 LTCC 4-pole Bandpass Filter

A 6-layer LTCC structure was designed. The ceramic tape has a dielectric constant of 7.8, with a fired layer thickness of 3.7 mils. Both buried and exposed conductors are gold. The design rules allowed a minimum conductor thickness/spacing of 5 mil and via diameter of 5.4 mil. In this design, the minimum conductor spacing and via diameter of 10 mil and 9 mil are used respectively, so as not to push the fabrication process to the limit. A 4-pole Chebyshev 0.01dB ripple bandpass filter center at 1.2 GHz with a fractional bandwidth (FBW) of 6% has been designed and simulated. Similarly, the external quality factor $Q_e$ of Fig.5.6 and coupling coefficient $K$ of Fig. 5.7 is extracted using EM as described in the previous sections. Two design curves are
obtained. The simulated frequency response of the 4-pole bandpass filter is shown in Fig. 5.8. The optimized spacing designated value of K12 and K23 is 11.8 mil and 20 mil respectively. Fig. 5.9 is the layout of the 3D LTCC 4-pole bandpass filter.
Figure 5.6: (a) input or output coupling structure (b) external quality factor $Q_e$ as a function of $t$. 
Figure 5.7: (a) coupling structure (b) simulated coupling coefficient
Figure 5.8: Simulated performances of the LTCC bandpass filter.

Figure 5.9: 3D view of the LTCC bandpass filter.
Chapter Six

Conclusion

6.1 Miniaturized Microstrip Bandpass Filters

Miniaturized microstrip bandpass filters using two novel resonator structures are presented. Miniaturization of the resonators is achieved using reactive or capacitive loading on the conventional open-loop resonator. Firstly, a new class of compact microstrip filter loaded with low-characteristic impedance triangular stub is introduced. Secondly, a new microstrip square open-loop resonator loaded with interdigital capacitive fingers is also proposed. It has been shown that the resonator structures can be accurately simulated using EM software. Both resonator structures are characterized. Design procedure for such resonator bandpass filters are discussed. Design curves, which provide the needed external quality factor and coupling coefficients to obtain the filters responses, are also characterized. Bandpass filters designs using the both resonator structures are simulated and measured. The fabricated filter performances are presented. There is good agreement between experimental and full-wave electromagnetic (EM) simulation results.

On the other hand, a novel low temperature co-fired ceramics (LTCC) 3D bandpass filter structure was investigated and EM simulated based on LTCC manufacturing design rules. The multilayer filter demonstrated the advantages of the LTCC's buried via process and size compactness, which is not possible to realize with the microstrip topology. Design curves based on LTCC resonator are also provided.
6.2 Suggestions for Future Works

Filters miniaturization using multilayer LTCC is showing an attractive option in the wireless communication systems. There are three obvious advantages. Firstly, miniaturization is possible by vertical stack-up. Secondly, it eliminates the need for a discrete filter, therefore reduces the component and assembly cost. Lastly, it allows filter integration into module within a single substrate for compactness.

New designs and structures for lump-element filters configuration can be explored. This type of filter is generally much smaller than the distributed elements filter. Capacitor and inductor could be implemented using parallel plates and a metallic strip respectively.
References:


