

## AN ABSTRACT OF THE THESIS OF

Kala Gururajan for the degree of Master of Science in Electrical and Computer  
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Title: New Configurations of Bandpass Filters in Single and Multilayer Environment

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Mobile and telecommunication industry has experienced tremendous growth in the recent past. Miniaturization and increased functionality have become necessary for all passive components in the system. Bandpass filters are critical passive components in any communication system and the existing designs in the crucial 1-10 GHz bandwidth suffer many drawbacks. For instance, widely used coupled line bandpass filters have undesirable harmonics at twice the center frequency and generally offer large footprints at these frequencies. Several other variations of bandpass filters have been reported in the literature. However, many of the existing filter designs do not take optimum advantage of the emerging technologies such as Low Temperature Co-fired Ceramic (LTCC) and Multichip Module Deposition (MCM-D), where compact geometries can be realized in a multi-level multi-conductor environment. Basically, design of high performance

bandpass filters needs to address two important aspects. First, an efficient way of suppressing the higher order harmonics needs to be embedded into the filter geometry while maintaining the passband performance. Second, the filter configurations need to be compact and suitable for the current multilevel fabricational technologies.

This thesis attempts to address both of these aspects by proposing new bandpass filter configurations in single and multilayered environments. A simple design procedure is developed for synthesizing a bandpass filter based on an equivalent admittance inverter parameter approach. This procedure can be applicable for a wide range of filter geometries. Several new techniques to tackle the issue of higher order harmonics are proposed. New configurations of compact filters are proposed in single and multilayer configuration with harmonic suppression. The challenges faced in designing these filters are discussed. Some of the challenges include overcoming the problem of interconnections between the different layers and reducing the design time for these configurations. To address these challenges, new, multilevel vialess filters are proposed, where the broadside coupling between layers is effectively used as a part of the filter characteristic to interconnect different layers. Further, an innovative algorithm to achieve first time successful design based on electromagnetic simulations has been proposed. The proposed new filter configurations have been validated with the full-wave electromagnetic simulation as well as measurement. This research on new configurations of efficient bandpass filter realization should prove useful for a wide range of applications in the frequency range of 1 and 20 GHz.

New Configurations of Bandpass Filters in Single and Multilayer Environment

By

Kala Gururajan

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Master of Science thesis of Kala Gururajan presented on June 03, 2004.

APPROVED:

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Kala Gururajan, Author

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This work is dedicated to my grandmother Thunga  
but for whom this journey was impossible,  
my parents Gururajan and Padma,  
to my sister Jaya and brother in law Guru,  
to my nephew Vaibha

## 1. INTRODUCTION

Microwave filters have been subject of intensive research efforts since World War II. The massive explosion in wireless and mobile applications coupled with the invention of new technologies has only increased innovations in this field by demanding increased performance. Among microwave filters, bandpass filters are components of all vital communication equipment in the cellular, satellite and radar communication. Several variations of bandpass filters such as coupled line filter, interdigital filters, waveguide filters etc have been previously reported in the literature for applications at microwave frequencies [1-4]. At high frequencies, bandpass filter implementation in planar transmission line technology is preferred due to compact size and low cost.

The typical electrical specifications of a bandpass filter are return loss, insertion loss and wide range of spurious free frequency range. Significant physical and design considerations include compact size and ease of synthesis. Higher densities and compact geometries have been the focus of attention in recent years with the advent of high speed mixed signal integrated circuits for mobile and wireless applications. Embedded passives in multi-layered media and passive component design using lumped elements are key techniques for achieving compact circuits leading to higher yields and lower costs. New system-on-package (SOP) technologies like low-temperature co-fired ceramic (LTCC) and Multi-chip module deposition (MCM-D) are well suited for multi-layer embedded passives and lumped elements to achieve higher performances and stringent specifications [5]. Thus, the development of bandpass filters well suited for higher frequencies and newer technologies depends on the lumped elements inductors and capacitors suited for these frequencies.

Several researchers have discussed a number of variations for planar capacitors[6,7]. For an application to MMICs, the existing designs offer several limitations including lack of accurate circuit models and poor figure of merit in terms of capacitance per unit area. Bandpass filter design using capacitively coupled resonators has been successfully applied in the past to design narrow band

filters beyond 10 GHz. Application of this concept to lower RF frequencies and /or with larger bandwidth has not been popular due to two main reasons. Larger bandwidths require higher capacitance values for the end sections of the filter and the existing lumped configurations could not offer the required values at microwave frequencies. Extension of the design concept to lower RF frequencies results in large resonator lengths, which are typically half a wavelength long. Since the existing capacitor configurations such as gap discontinuity and small interdigital capacitors do behave more like lumped elements at these frequencies, their presence does not significantly affect resonator lengths. These limitations have led to development of bandpass filters suited only for specific applications or limited in functionality.

Current bandpass filter designs offer reasonable performance but are ill-equipped to emerging market. Issues like miniaturization and added functionality have to be addressed. The first has been addressed in a number of ways. Jun Seok Park *et. al* proposed a novel coupled line bandpass filter using defected ground structure [8]. Menzel and Schwab proposed multilayer filter structures separated using GaAs [9]. The Iris coupled and dielectric loading of bandpass filters has been put forth by Amari *et. al* to reduce the size of filters [10]. These methods work well for specific cases and frequency bands.

The issue of higher-order harmonic suppression has attracted several researchers [11-12]. Improving the stop band characteristics of filters has become a prime factor in ascertaining the isolation between transmitter and receiver in duplexers. Rejection of spurious signals is enhanced by two means either by increasing the number of resonators or by adding transmission zeroes. Most of these methods work well for single layer structures only. Furthermore, they are aimed at suppressing the second order harmonic. In some cases, several iterations are required before arriving at a filter with optimum performance. Some of the recent work in the field that has an excellent scope for development is discussed here.

Txemo *et. al* proposed an innovative wiggly-coupled line structure to suppress second order harmonic [13]. This method does not require re-optimizations of filter parameters and classical design methodology could be applied. Recently, Quendo *et. al* proposed the integration of low pass filters in bandpass filters for out of band performance improvement of multilayer filters[14]. This method did not deteriorate the in band performance of the bandpass filter. The superior performance was achieved by improving the out of band characteristic of the low pass filter. This method also provided additional degrees of freedom to tune the response of the overall filter structure. Suggestions to reduce coupling between resonators and external circuits at frequencies at which spurious responses occur have been made and seen to work well. Lastly, much of the work on harmonic suppression is for narrowband filters.

Extensive work has been devoted to miniaturization of filters. The most relevant of these are discussed here. Nguyen put forth a spurline-based realization of ultra wideband bandpass filter for compact realization [15]. The method of folded line bandpass filter propounded by Settaluri *et. al* [16] results in extremely compact filters with a slightly reduced bandwidth, which can be easily compensated by taking into account the effect of multiple coupling . Other variations such as square stepped impedance resonator on microstrip have been reported to achieve compact footprint [17]. The general assumptions with most of these methods are that they are specific to a certain type of filter. The reduction in size is achieved at times with a trade off in performance.

The existing bandpass filter technology requires:

- Simpler design models, which are generic in nature
- Approach to develop compact filters that are suited to new age technologies
- systematic techniques to reduce harmonics

In this thesis, a generic approach to design new configurations of bandpass filters is presented. A specific design issue of miniaturization without performance degradation is explained with examples and measurements. Some techniques to suppress higher order harmonics have been demonstrated while retaining the simplicity of design and fabrication.

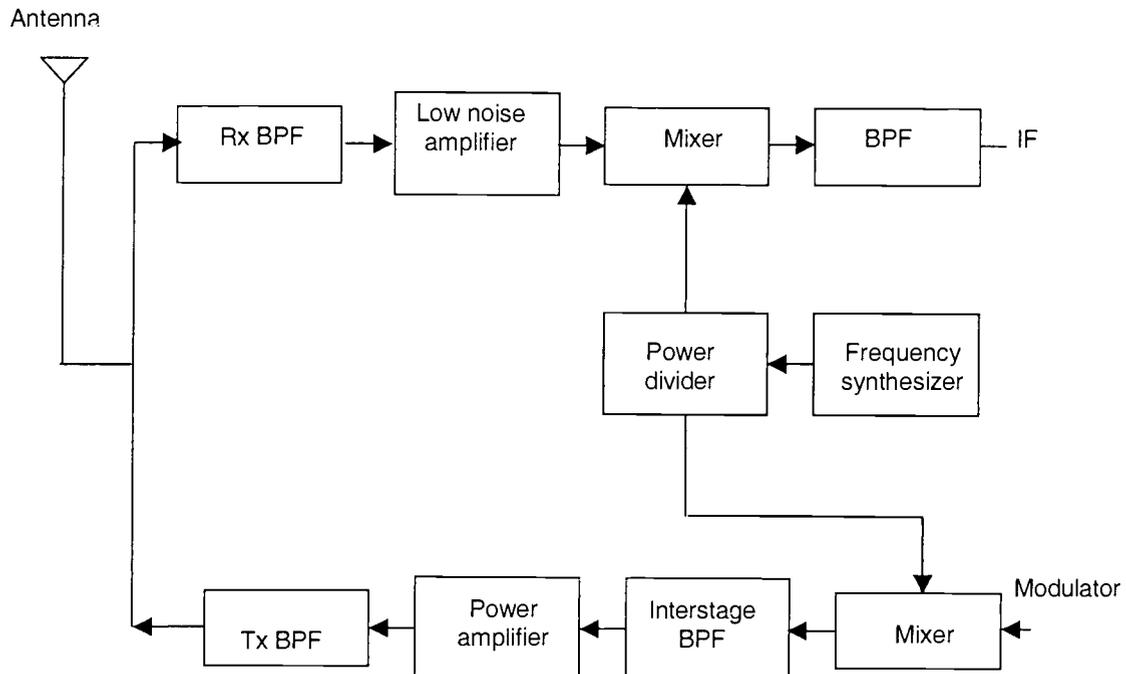
Before proceeding to the organization of thesis, a brief description of the typical transmission system is provided in the next section.

### **1.1 Typical Communication system**

To illustrate the importance of bandpass filters for communication system applications, a typical wireless transceiver system is shown in Fig1.1 [18].

The transmitter functions to up convert a pre-modulated signal. The interstage bandpass filter in the transmitter section is used to suppress the harmonics generated at the output of the mixer. The signal in the power amplifier also has several spurious signals, which are the harmonic components of the input signal. The transmitter bandpass filter (Tx-BPF) filters the harmonic components and the transmitted signal is finally transmitted through the antenna.

The receiver shown here employs a superheterodyne configuration, which is the most common form of communication systems. The low noise amplifier amplifies the received signal from the antenna after filtering the undesired signals using the receiver bandpass filter (Rx-BPF). This signal is then transferred through the second bandpass filter and to the intermediate frequency (IF) port after frequency conversion by the mixer. The IF signal is further amplified and converted to the base band signal after detection and modulation.



**Fig.1.1 Typical communication system**

In a receiver, a bandpass filter is typically used to prevent saturation of receiver front end due to leakage of output signal from the transmitter and to filter out interference signal components. It is important that the receiver bandpass filter provides high attenuation to interference signals with minimum pass band insertion loss.

The transmitter bandpass filter primarily reduces spurious radiation power from the transmitter to avoid interference with other systems. The dominant frequency components of these undesired signals are the second or third harmonics of the transmitting signal frequency. Yet another vital function of the transmitter bandpass filter is the attenuation of noise within the transmission signal at the receiver band and thus suppression of its level below the sensitivity point of the receiver. The transmitter bandpass filter must possess a wide stopband for spurious signal suppression while

maintaining a low pass band insertion loss. Thus having discussed the ideal properties of a bandpass filter we proceed to the contribution made in this thesis to achieving them.

## 1.2 Characteristics of a bandpass filter

It is established that bandpass filters form an important component in mobile and wireless communication systems. A typical bandpass filter response is shown in Fig.1.2.

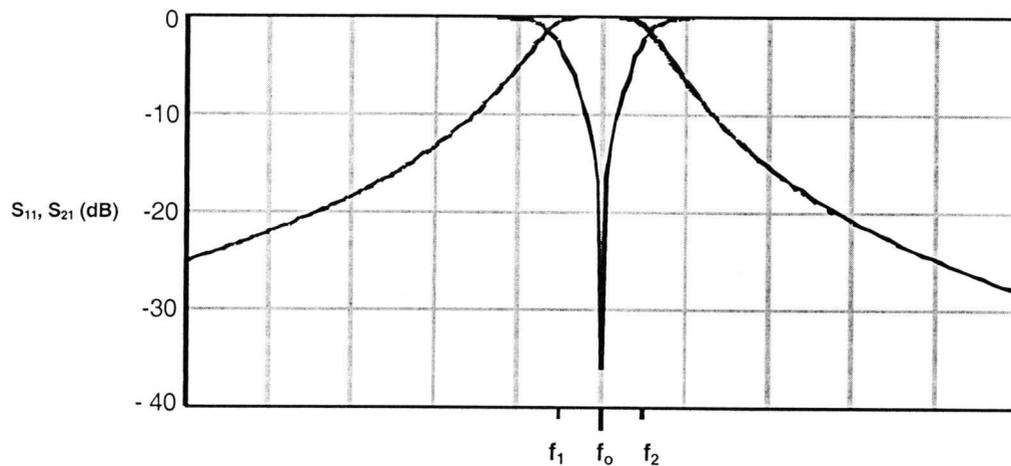


Fig.1.2 Bandpass filter response

Here  $f_0$  is the center frequency and  $f_2 - f_1$ , is the bandwidth.

The ideal bandpass filter provides infinite stopband attenuation and no passband loss. Depending on the response of filters in the passband and stopband, they can be classified as maximally flat (Butterworth) filters and equi-ripple (Chebyshev) filters.

## 1.3 Contributions and organization of thesis

The focus of research was to develop new configurations of filters specifically for the (1-10) GHz band. The aim was to develop a comprehensive method to design filters with superior performance

to suit a wide range of specifications. These new configurations are based on filter-sections with dominant capacitive coupling to provide superior band pass filter performance.

Zhu. *et. al*/ proposed larger versions of interdigital capacitors for application to coupled resonator filters at 1.575GHz. They could effectively take advantage of the distributed nature of the larger interdigital capacitors and reduce the resonator lengths. Based on this principle, several new configurations which can replace the conventional gap discontinuity and the interdigital capacitor in a capacitive coupled resonator bandpass filter is presented in this thesis.

In chapter 2, the theory and design equations underlying these bandpass filter designs are presented. An overview of conventional bandpass filter design procedure is discussed before the new admittance inverter approach is presented. This enables the readers to appreciate the simplicity of the new technique.

In chapter 3, we present a class of single-layered structures, suitable for bandpass filter applications. A systematic study is carried out to study various properties of these filter sections. Bandpass filters with different specifications are designed using these sections and their frequency response is presented. Full wave electromagnetic simulations have been carried out to validate the theory.

Multilayer structures have seen increased interest in recent years as they provide stronger coupling between different layers and transmission lines combining specific advantages to reduce system size or to provide additional functionality. Considering their importance, bandpass filters realized on a multilayered platform have been presented in chapter 4. These filter sections exhibit attractive features such as vialess interconnection between layers, increased capacitance etc. Two specific techniques of stub loading and spurline loading of bandpass filters to achieve harmonic suppression are explained.

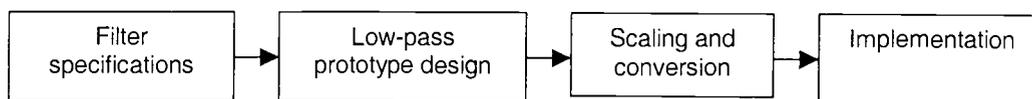
As the technological innovations advance, the complexity of systems has also been steadily increasing. Consequently, bandpass filters performance need to be accurately predicted. So far, full-wave electromagnetic simulations have generally been considered as means of validating the theoretical designs. Although, design optimization using full-wave simulations is offered in recent versions of commercially available software, they are not practical in terms of implementation due to the amount of time involved. At the same time, if successfully implemented, they can offer very accurate results as no quasi-static assumptions are involved in the simulations. A systematic procedure that can effectively utilize the full-wave EM simulations to synthesize new configurations of bandpass filters can be extremely useful. In Chapter 5, a new filter design technique based on efficient use of EM simulations is proposed.

In order to validate the proposed theory, bandpass filters were fabricated and measurement results are presented in Chapter 6 . Conclusion and some recommendations are put forth in Chapter 7.

## 2. FILTER DESIGN THEORY

A filter is an important block in microwave circuits and is among the first few circuit elements studied in any new technology. Filter theory has been a subject of investigation in the past and especially for the last two decades. This chapter presents an overview of the filter design theory relevant to this thesis. Section 2.1 explains the concept of the admittance inverters, with reference to the filter design theory. Section 2.2 discusses the conventional bandpass filter design. Section 2.3 presents the new approach to designing a bandpass filter - Admittance inverter design. Chapters 3 - 4 presents new filter configurations realized in single and multilevel topologies using the design procedure described earlier.

Traditional filter design procedures evolved in late 1930s. The two existing approaches to design of filters are a) image parameter method and b) insertion loss method. The image parameter method is relatively simple but has inherent disadvantages. The frequency response over the entire operating range cannot be specified and hence the filter design requires several iterations to achieve the desired response. On the other hand, the insertion loss method is a more modern method based on network synthesis techniques. This enables the complete specification of the frequency response. The complete design flow using the insertion loss method is shown in Fig.2.1.



**Fig. 2.1 Filter design by insertion loss method [19]**

The low pass filter prototype values are obtained based on the filter specifications. Several filter design hand books provide ready-to-use tables for the prototype values for a given set of filter

sections,  $N$ , type of the filter (Maximally flat or Chebyshev), passband ripple (if applicable) and fractional bandwidth required for the filter [20]. These prototype values are denoted as  $g_s$  for normalized values of frequency and impedance. Thus for a 3<sup>rd</sup> order filter, the element values are numbered from  $g_0$ , at the generator impedance to  $g_{n+1}$ , at the load impedance. The network elements alternate between series and shunt connections depending on  $g_0$  being a resistance or conductance.

The scaling and conversion techniques are applied to translate the prototype design to the desired response. The values so obtained are lumped elements. The modification for microwave circuits is provided by Richard's transformation and the Kuroda identities. The subject of microwave bandpass filters is quite extensive. Implementation of capacitively coupled resonator bandpass filter is dealt with in detail considering its relevance to the proposed design approach. The admittance inverter network forms the basis of capacitively coupled resonator bandpass filters.

### **2.1. Admittance inverter network**

Quarter wave lines when used in specific network configurations convert series elements to shunt and vice versa. These form the inverse of the load impedance or load admittance to achieve impedance inversion. The transmission line networks that can perform this function are called admittance or impedance inverters. The admittance inverters or J-inverters form the basis of our design procedure and we describe them in detail here. In its simplest form, a J inverter may be constructed using a quarter-wave transformer. The quarter wave line and its alternate implementation using a J-inverter are shown here in Fig.2.2.

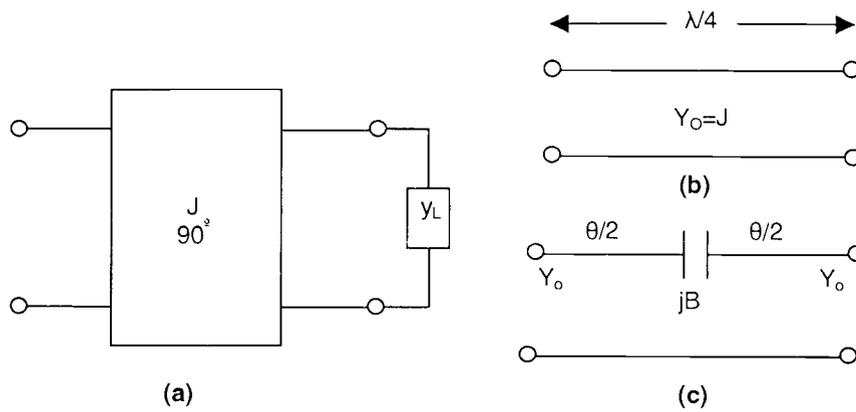


Fig.2.2 Admittance inverter and its alternate implementations

For the admittance inverter as shown in Fig 2.2(a)

$$Y_{in} = \frac{J^2}{Y_L} \quad (2.1)$$

For the  $90^\circ$  line as in Fig. 2.2(b)

$$J = Y_0 \quad (2.2)$$

For the lumped element implementation as in Fig. 2.2(c)

(2.3)

$$J = Y_o \tan \left| \frac{\theta}{2} \right|$$

$$B = \frac{J}{1 - \left( \frac{J}{Y_o} \right)^2}$$

$$\theta = -\tan^{-1} \frac{2B}{Y_o}$$

The transmission line length  $\theta/2$  is normally negative but in the design procedure here, it would be absorbed in the adjacent sections as will be seen in the forthcoming sections. The admittance inverter forms the basis of designing band pass filters from individual filter sections.

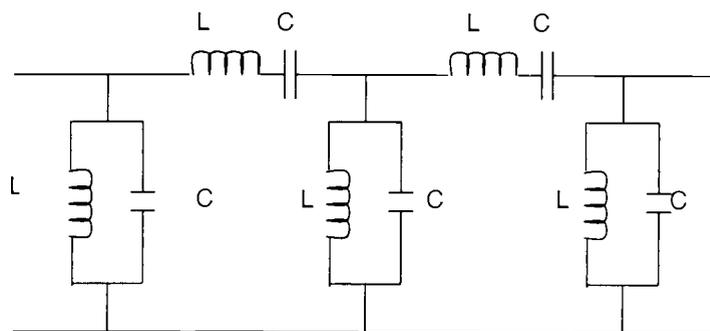
## 2.2. Conventional filter design

The conventional design approaches to implementation of a bandpass filter are:

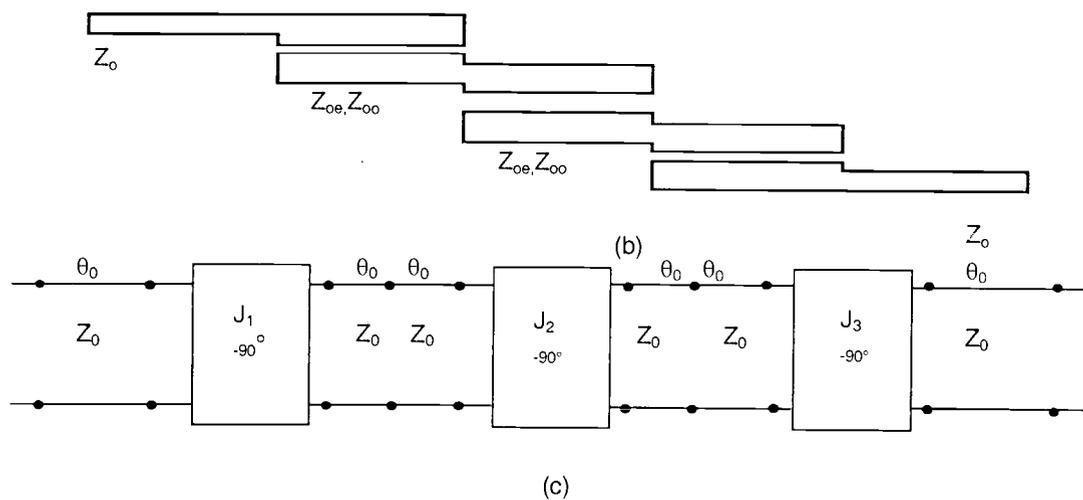
- Parallel coupled line bandpass filter
- Capacitive –gap coupled resonator bandpass filter

Due to their relevance both the methods is examined in detail here:

Parallel coupled line bandpass filter: - A popular implementation of the bandpass filter is cascading a number of coupled line sections realized on planar transmission configurations. This way, bandpass filters can be fabricated with ease in microstrip or stripline configuration, using capacitive coupling between the resonators. The lumped element representation of a bandpass filter is shown in Fig. 2.3(a). Fig.2.3(b) represents the equivalent transmission line model of the bandpass filter.



(a)



(c)

**Fig 2.3 (a) Lumped element representation of a bandpass filter (b) Layout of a three section coupled line bandpass filter (c) Representation using equivalent of each coupled line section**

The layout of the filter is shown in Fig 2.3(b). The sequential representation of the design is shown in Fig.2.4 and can be explained in terms of the following steps:

Step 1: For the given set of filter specifications such as number of resonators  $N$ , type of the filter, bandwidth and center frequency, get the lowpass filter prototype values [20].

Step 2: For each low pass prototype determine the corresponding J-inverter parameter using the equations

$$Z_o J_1 = \sqrt{\frac{\pi\Delta}{2g_1}} \quad (2.4)$$

$$Z_o J_n = \frac{\pi\Delta}{2\sqrt{g_{n-1}g_n}} \text{ for } n=2, 3, 4\dots N,$$

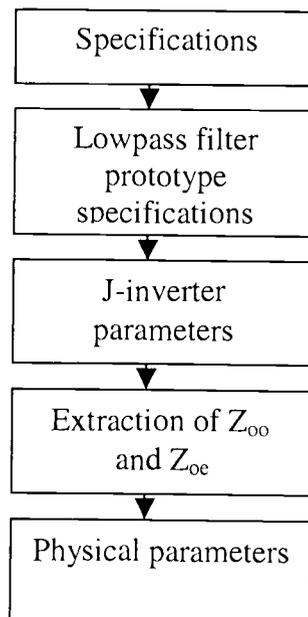
$$Z_o J_{N+1} = \sqrt{\frac{\pi\Delta}{2g_N g_{N+1}}}$$

Step 3: The even and odd mode characteristic impedance for each section is then found using the equations:

$$Z_{oe} = Z_o [1 + JZ_o + (JZ_o)^2] \quad (2.5)$$

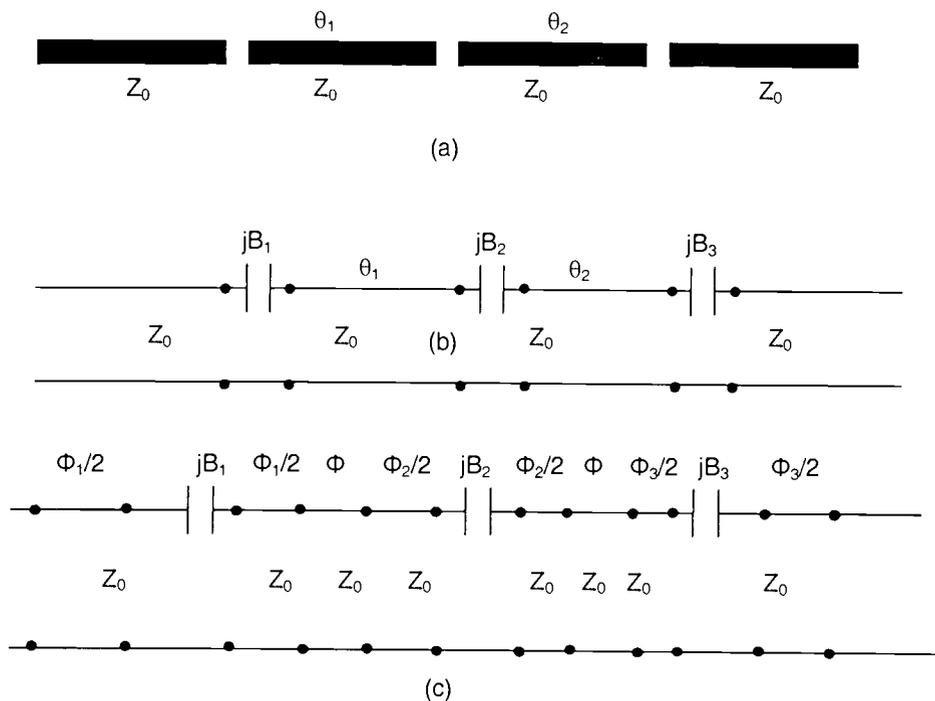
$$Z_{oo} = Z_o [1 + JZ_o + (JZ_o)^2]$$

Step 4: The physical parameters can be determined using equations given in [20].



**Fig.2.4 Design flow for the conventional parallel-coupled bandpass filter design**

Capacitive-gap coupled resonator bandpass filter:-The gap coupled resonator filter is fabricated using similar principle as the coupled line band pass filter. A gap coupled resonator filter of order  $N$  has  $N$  resonant sections and  $N+1$  gaps between them. Each of these gaps is modeled as a series capacitor. The resonator length  $\theta$  between two capacitive gaps is modeled as a sum of a line which is  $\lambda/2$  at center frequency and negative length transmission line sections on either side of the series capacitors. The series capacitors and the negative length of transmission line form the



**Fig. 2.5 (a) Representation of capacitive gap-coupled resonator filter (b) Transmission line model (c) Transmission line model with negative length sections forming admittance inverters [20]**

equivalence of an admittance inverter network. The steps involved in designing a bandpass filter using capacitively coupled resonators can be summarized as:

Step 1: Depending upon the type of filter obtain the  $g$  values of the low pass prototype filter.

Step 2: Calculate the admittance inverter parameter for each section using (2.4).

Step 3: Calculate the susceptances  $B_i$  for each section as

$$B_i = \frac{J_i}{1 - (Z_0 J_i)^2} \quad (2.7)$$

The capacitance  $C_i$  for the  $i^{\text{th}}$  gap is given by:

$$C_i = \frac{B_i}{\omega_0} \quad (2.8)$$

Step 4: The electrical length  $\Phi_i$  as

$$\phi_i = -\tan^{-1}(2Z_0 B_i) \quad (2.9)$$

This equivalence reduces the gap coupled resonator equivalent circuit to the same as that of a coupled line resonator bandpass filter circuit equivalent where the electrical length of the resonator section can be found as:

$$\theta_i = \pi - \frac{1}{2}[\phi_i + \phi_{i+1}] \quad (2.10)$$

### 2.3. Bandpass filter design using admittance inverter approach

The bandpass filter design proposed using admittance inverter approach is a simple approach and can be applicable to any type of bandpass filter section. The approach is independent of the physical configuration as it is based on the network model of the section rather than its physical implementation.

Let us consider the filter section connected between the input and output transmission lines as shown in Fig.2.6. The first step is to determine the susceptance parameters of this section. This can be done by dividing the filter section into a set of known geometries as indicated in the figure or by performing a full-wave Electromagnetic simulation for the entire structure.

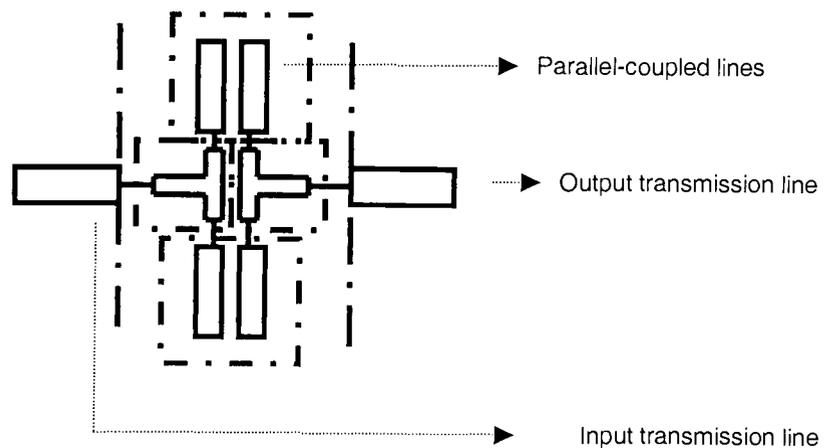


Fig. 2.6 Network model for an arbitrary filter section

The next step is to represent the filter section in terms of equivalent admittance parameter network as shown in Fig. 7. For symmetrical filter sections (such as the one shown in Fig. 2.6, the network shown in Fig. 2.7 will be identical to the filter section, provided the  $J$  and  $\phi$  are given by the equations (2.11).

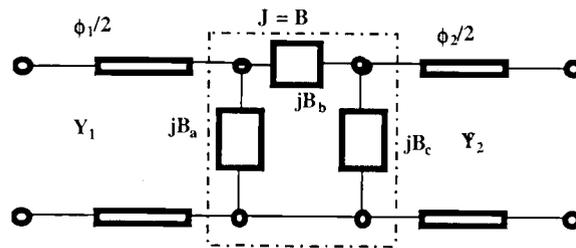


Fig.2.7 J-inverter network representation

$$J = \left| \tan \left( \frac{\phi}{2} + \arctan \left( B_p \right) \right) \right| \quad (2.11)$$

$$\phi = -\arctan \left( 2B_p + B_g \right) - \arctan \left( B_p \right)$$

On the other hand, for asymmetrical networks The J-Inverter parameters of these networks can be given by equations shown in (2.12) [1].

$$\frac{J}{\sqrt{Y_1 Y_2}} = \frac{\sin \left( -\phi_1/2 \right) + \overline{B_{11}} \cos \left( -\phi_1/2 \right)}{B_{12} \sin \left( -\phi_2/2 \right)} \quad (2.12)$$

$$\phi_1 = M_1 \pi + \tan^{-1} \left\{ \frac{2 \left( \overline{B_{11}} + \overline{B_{22}} |B| \right)}{1 + \overline{B_{22}}^2 - \overline{B_{11}}^2 - |B|^2} \right\}$$

$$\phi_2 = M_2 \pi + \tan^{-1} \left\{ \frac{2 \left( \overline{B_{22}} + \overline{B_{11}} |B| \right)}{1 + \overline{B_{11}}^2 - \overline{B_{22}}^2 - |B|^2} \right\}$$

Bandpass filter are realized by cascading the individual sections synthesized by the procedure shown above. It is of significance that when the sum of the interconnecting lengths between the pi networks is  $-\pi$  the length of resonator between the two sections is zero in accordance to equation 2.8. Thus, the filter consists only of cascaded sections with no additional resonator lengths between them, thereby making the structure extremely compact.

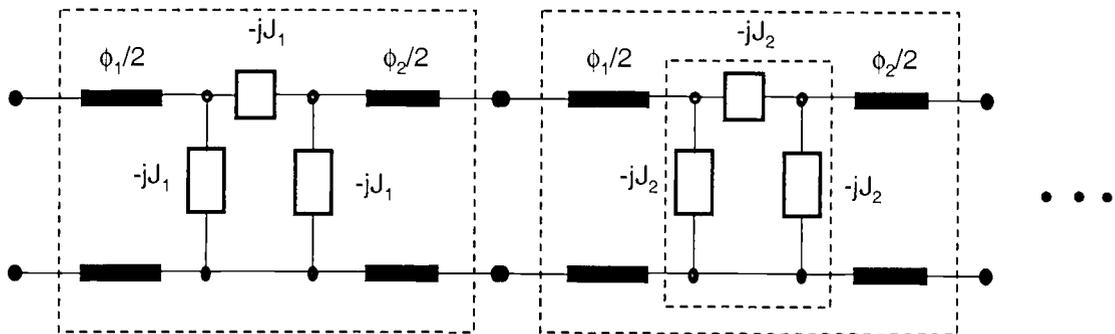


Fig. 2.8 Bandpass filter realization

### 3. NEW SINGLE-LAYER FILTER CONFIGURATIONS

In this chapter, new configurations of bandpass filters in single layer configuration, suitable for the frequency range of 1-10 GHz have been proposed. The 1-10 GHz frequency range is widely used for communication systems and design of passive components such as filters is particularly challenging in this band. Lumped configurations are more suitable to component designs below 1-2 GHz as parasitics can lead to poor performance beyond these frequencies. The distributed element approach using transmission lines leads to very large electrical lengths and may not be feasible for practical implementation. Designing bandpass filters, which exhibit compact footprint and yet take into account, the parasitic as well as distributed effects would be particularly attractive and will be highly suitable for this frequency range. The new configurations presented in this chapter will take these features into account without compromising the electrical characteristics.

Only recently, Zhu *et. Al* [1] proposed larger versions of interdigital capacitors (IDCs) for application to coupled resonator filters at low frequencies. They could effectively take advantage of the distributed nature of the larger IDCs and reduce resonator lengths. The overall filter lengths reported were smaller compared to the gap-coupled resonator design but are still considered large for practical implementation. Filter design using parallel-coupled filter sections would offer large footprint at these frequencies and also suffers from spurious harmonic response at twice the passband center frequency. This is attributed to the non-homogeneity of the medium, which results in the even, and odd mode phase velocities being unequal. This deteriorates the upper stopband performance and moves the spurious passband towards the center frequency. The existence of second harmonic for bandpass filters is particularly undesirable for applications in oscillators, amplifiers and receivers, in general. Phase velocity compensation techniques and more recently capacitive compensation techniques have been proposed by several authors to solve this problem [21]. In this chapter, several new single layer bandpass filter configurations have been proposed for

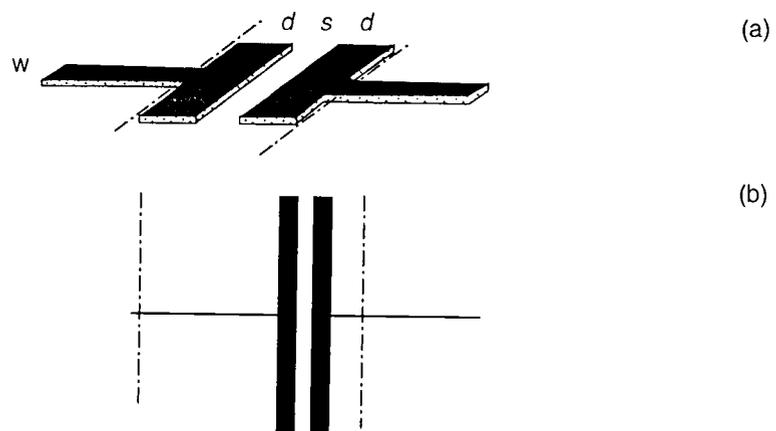
the 1 -10 GHz band using the equivalent-admittance inverter parameter approach described in chapter 2. The design procedure followed does not require the calculation of the equivalent circuit parameters in terms of L and C and takes into account the distributed nature of these structures. Some of the configurations offer the advantage of higher harmonic suppression, as described later in this chapter.

The proposed new filter configurations can be used to realize a wide range of fractional bandwidths. These filter sections are designed using the J-inverter method and are well suited to the emerging trend of miniaturization of devices. Several improvisations of the design have been made to reduce the size of these filters. Design techniques that altogether suppress the second harmonic have been put forth.

In the next section, we present an explanation of the various filter sections that can use as building blocks for the filters. We discuss the filters designed and their advantages in Section 3.2 - 3.3.

### 3.1. Filter sections

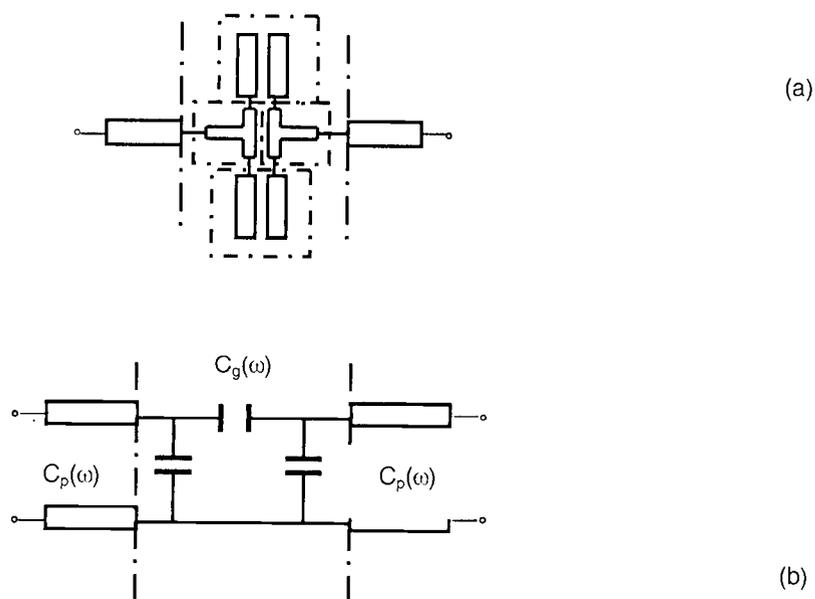
The filter sections on planar transmission lines can be classified into two categories (a) Quasi-lumped structures (b) Composite structures as shown in Fig.3.1. The Quasi-lumped structures are structures with low frequency dependence as shown in Fig. 3.1(a). These structures are modeled in terms of frequency dependent capacitances only.



**Fig 3.1 (a) Quasi Lumped Structure (b) Composite Structure**

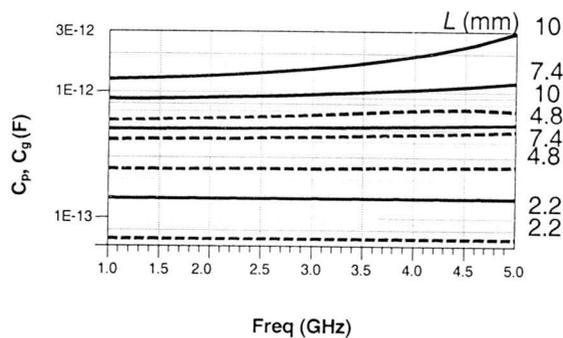
Composite structures are structures that have a high degree of frequency dependence. It is to be noted that the structure shown in Fig3.1 (b) has its reference plane shifted to include the length of transmission lines on either side. Modeling this structure requires a network of capacitances and inductances thereby the structure becomes highly frequency dependent. Such structures that require both inductances and capacitances for their accurate performance modeling are called composite structures.

The quasi-lumped structures are modeled in terms of a  $\pi$ -network of capacitances. For instance, the structure shown in Fig. 3.1(a) consists of two parallel conducting strips of length  $L$  and width  $d$  in edge-coupled configuration.

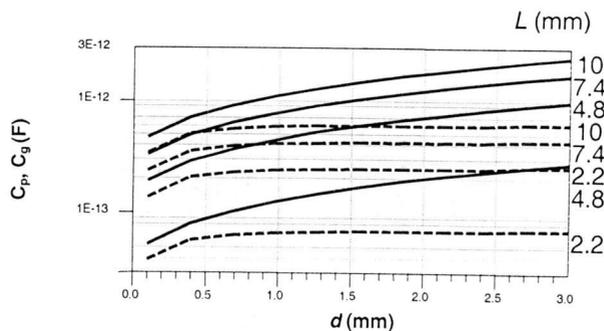


**Fig 3.2 (a) Network model using coupled transmission lines (b) Equivalent circuit**

For a conventional gap coupled structure the required high level coupling strength between the two open ends of uniform lines is difficult to realize with the conventional MIC fabrication and also the fabrication tolerance of an extremely small gap maybe out of reach. Since a pair of parallel stubs attached to the end of two lines, its coupling strength can be flexibly controlled within a relatively large range through an adjustment of the stub length. Extending its relevant stub length can also contribute to the compactness of each resonator. For large values of  $L$  and  $d$ , the structure cannot be considered as a single gap discontinuity and has to be modeled using coupled transmission lines as shown in Fig. 3.2(a). For the purpose of analysis, the equivalent circuit model is shown in Fig. 3.2(b). It may be noted that the assumption is valid when the operating frequency is below the self-resonant frequency of the structure.



(a)  $d = 1.2$  mm,  $s = 0.2$  mm

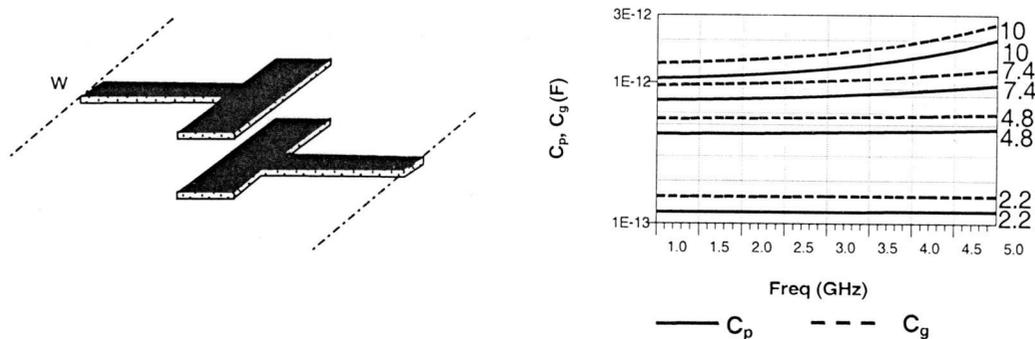


(b)  $s = 0.2$  mm,  $f = 1$  GHz

—  $C_p$     - - -  $C_g$

**Fig. 3.3 Equivalent circuit parameters for the structure shown in Fig. 3.1(a)**

Figs. 3.3 shows the variation of series and shunt capacitances for different values of  $L$ , as a function of frequency and strip width  $d$  respectively. The structure is realized on a microstrip with  $h=1.27\text{mm}$  and  $\epsilon_r = 10.2$ . As expected, the capacitance value is significantly higher than the conventional gap discontinuity, which is typically of the order of 0.02 pF to 0.07pF at 1 GHz for line width corresponding to  $50 \Omega$ . The other category of structures called the composite structures has a distributed nature. Due to this, their equivalent circuits consist of complex network of inductances and capacitances. Fig. 3.4(b) shows the variation of series and shunt capacitances for the composite structure shown in Fig. 3.4(a). These calculated network parameters will be used in the following section for the optimized design of a bandpass filter.

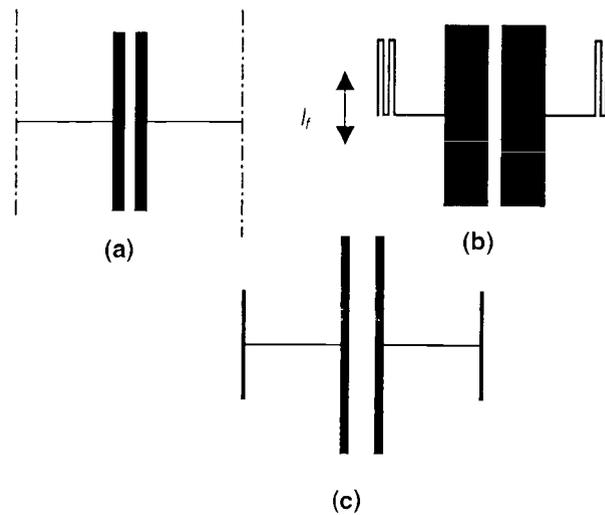


**Fig. 3.4(a) A composite structure (b) Variation of equivalent circuit parameters for structure**

The bandpass filters in any configuration for microwave applications have component sections each of which satisfy a particular value of  $J$  and  $\Phi$ . This section of thesis aims at presenting the possible filter section structures and their analysis.

Fig. 3.5(a)-(c) shows a few possible composite filter sections in single layer. Due to their distributed nature, these structures have equivalent circuit consisting of a network of capacitances and

inductances. Fig. 3.5(a) presents a structure that resembles a quasi-lumped capacitor with a simple variation. A short length of transmission line has been added on both sides and the reference planes are accordingly shifted. The two high impedance lines behave like series inductors at low frequencies and the structure would make the filter length shorter with better harmonic suppression.



**Fig. 3.5 Proposed composite structures**

In Fig.3.5 (b), the high impedance line is folded to form a multi-coupled line for better compactness. The effect of varying the folding length  $l_f$  is shown in Fig. 3.6. The structure is realized on a microstrip with  $h=1.27\text{mm}$  and  $\epsilon_r=10.2$ . Folded line is formed using a coupled 4-line section. Dimensions chosen for this geometry are  $L=8.1\text{mm}$ ,  $d=1.2\text{mm}$  and  $s=0.2\text{mm}$ . The width of the folded line sections  $w_f = 0.1\text{mm}$  corresponds to a characteristic impedance of  $110\Omega$ . The frequency response is plotted for different values of  $l_f$ . It can be seen that if  $l_f$  is varied the  $180^\circ$  crossing point for  $\Phi$  can be controlled to occur at any desired frequency. It may also be noted that this can be used to adjust the value of  $J$  giving an additional degree of freedom. The structure shown in Fig. 3.5(c) has the high

impedance line loaded with two open-ended stubs on either side. This can be effectively used to suppress higher order harmonics in a filter design.

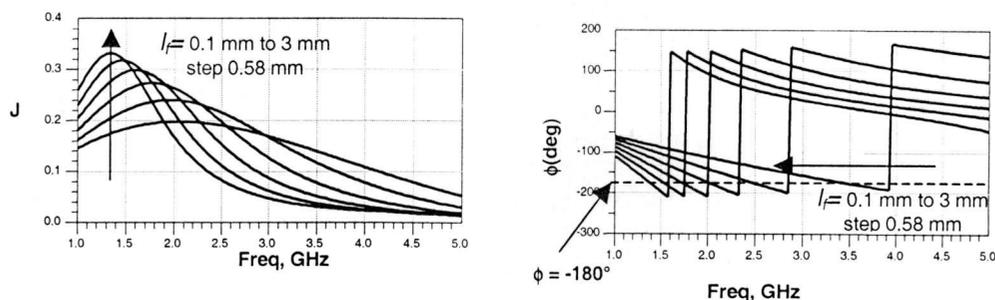
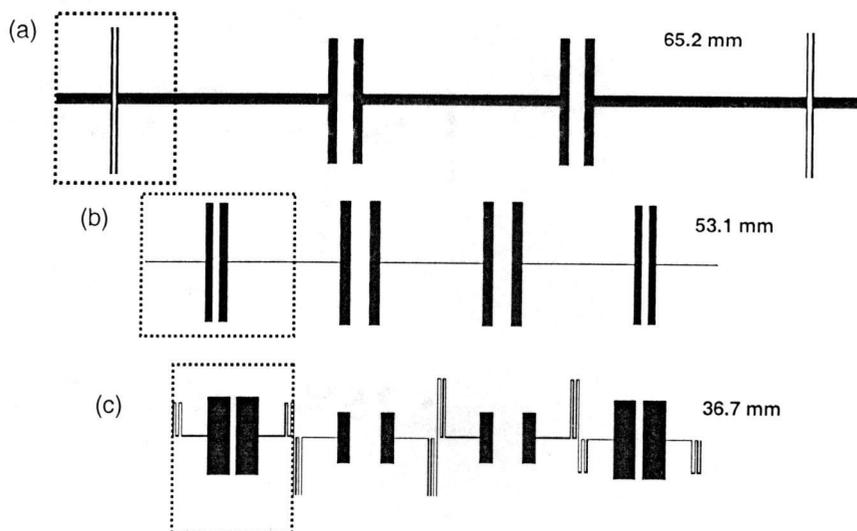
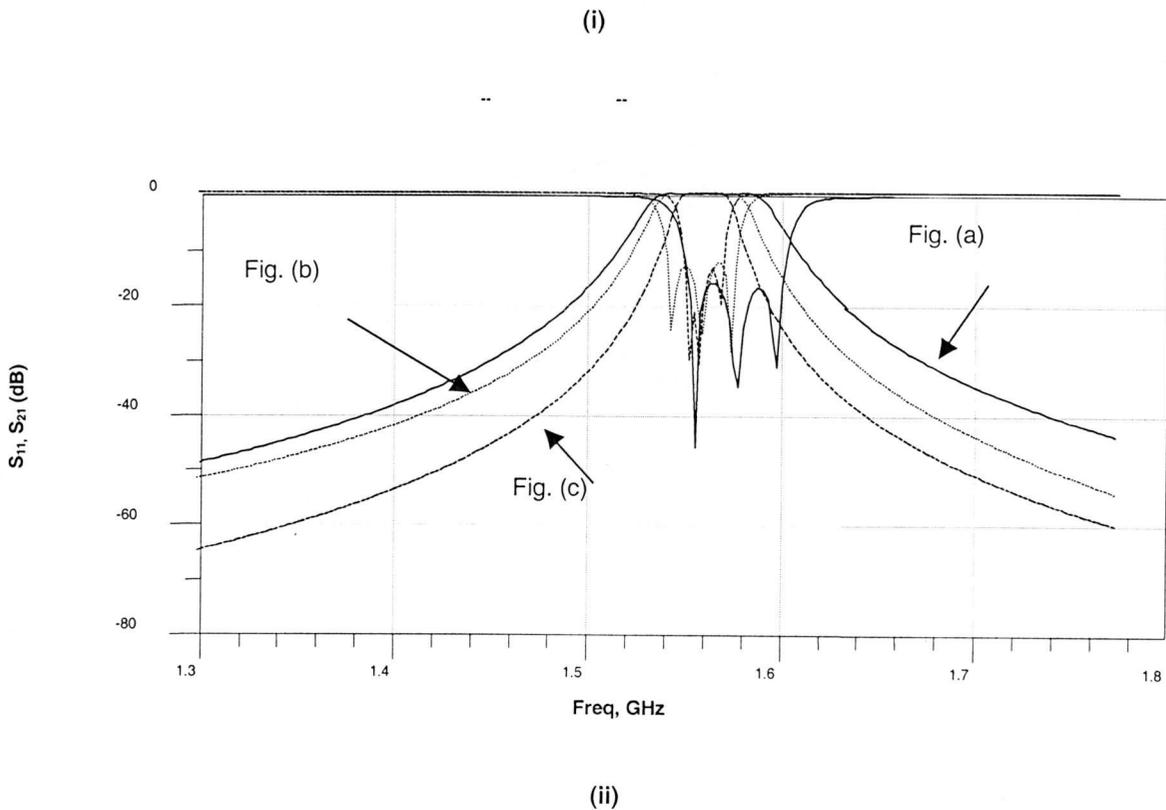


Fig. 3.6 Variation of  $J$  and  $\Phi$  for the structure in Fig.3.5 (b) as a function of frequency for different values of folded length,  $l_f$

### 3.2. Compact Filters

With the increasing advent of high-speed mixed integrated circuits for wireless and mobile applications, higher densities and compact geometries have been the focus of attention over recent years. High integration density and smaller size are becoming the norm of the industry for passive components. Thus new designs of bandpass filters suitable for miniaturization need to be proposed. In this section, we discuss the design of such compact single layer bandpass filters designed using the filter sections shown in Fig. 3.5. A single section of each filter is highlighted in Fig.3.7 (i).



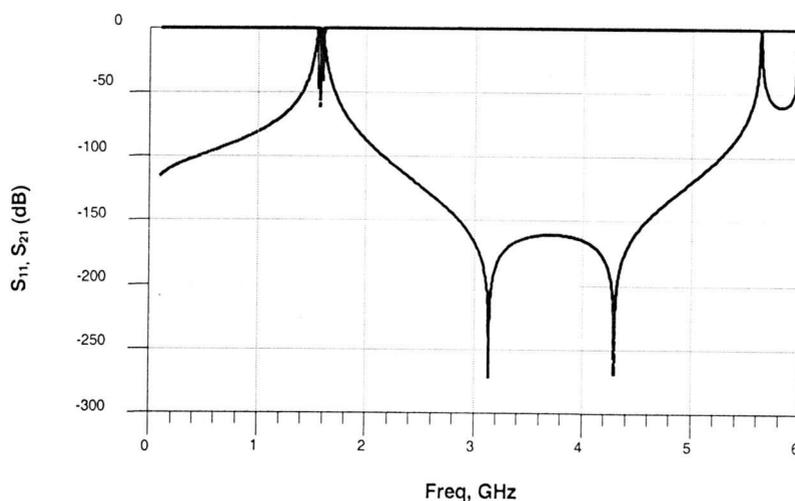


**Fig.3.7 (i) Footprint comparison of three new filter geometries (ii) simulated response**

To demonstrate the advantages of the proposed configurations, we have initially designed a 0.1dB ripple Chebyshev bandpass filter with the specifications as reported in [1] using the interdigital capacitors. The filter specifications are  $N=3$ ,  $f_0=1.575$  GHz, bandwidth=2.7% on a microstrip platform with  $h=1.27$ mm and  $\epsilon_r = 10.2$ . In [1], the reported overall filter length is 106.9 mm using the conventional design and 68.58 mm with inductive compensation. Fig.3.7 (i) presents the final footprint comparison for three new configurations. The filter showed in Fig.3.7 (i) (a) use the quasi-lumped capacitor shown in Fig.3.1(a) with the three resonators having a characteristic impedance of  $50\Omega$ . In the case of filter shown in Fig.3.7 (b), the  $50\Omega$  transmission line sections are eliminated by using the composite structure shown in Fig.3.5 (a). The filter in Fig.3.7(c) employs the folded line

version given in Fig. 3.5(b) for making the structure more compact. A comparison of lengths indicates that all new structures are shorter compared to the reported IDC filters [1]. The simulated filter response is shown in Fig.3.7 (ii). The folded line version exhibits slightly reduced bandwidth for the same specifications due to the multiple the placements of additional shunt stub sections[14].

Note that long and intensive simulations with the momentum is required for modeling the entire structure while our proposed techniques required negligible CPU time since the design is made with separate circuit elements.



**Fig. 3.8 Extended filter response of the structure shown in Fig3.7 (i)(b) demonstrating harmonic suppression.**

Of special interest is the extended response of the structure shown in Fig.3.7 (b) shown in Fig. 3.8. It can be seen that the second harmonic is completely suppressed. This can be attributed to the use of the section employing series inductors. The series inductance effect is due to the thin transmission lines added on both sides of the filter section and shifting the reference plane

accordingly. These high impedance lines behave as series inductors at low frequencies providing harmonic suppression. It has been observed that filters employing this technique exhibit complete second order harmonic suppression. The complete section is modeled as a network of capacitance with series inductance, which provides a low pass effect. It is of significance that this low pass effect is brought about by synthesizing the series inductance as a part of the filter section. Thereby there is no increase in length of the filter. In the past researchers have obtained this effect by cascading a low pass filter externally leading to a discernible increase in filter length and increased design complexity. A very desirable feature is that the negative lengths of the transmission line in the filter section itself absorb the resonator length between the filter sections making the filters extremely compact.

The new filter topologies presented here have been analyzed using network models. They give a wide range of bandwidths ranging from 2.7% to 15%. These filters exhibit compact footprint as well as excellent higher harmonic suppression while retaining the simplicity of design.

## 4. NEW VIALESS MULTILAYER FILTER CONFIGURATIONS

### 4.1. Overview

The recent system-on-package technologies such as multi-layer Low-Temperature Co-fired Ceramic (LTCC) and multi-chip module deposition (MCM-D) have become popular in RF design due to attractive features such as high integration density, better performance and reliability [22]. The low temperature co fired ceramic technology is multi layer thick film processing, which permits combining active and passive microwave components into monolithic module [23]. This enables microwave devices to be fabricated with high overall reliability while keeping the cost competitively low. The impetuous development of local area networks (LAN) and car radar systems caused the intensive investigation of LTCC properties at microwaves and millimeter waves. These experiments show the capability of multi-layer LTCC technology to be utilized at millimeter waves.

With the emergence of these new technologies, design of bandpass filters in multi-layered and multi-conductor environment has become popular due to increased scope of feasibility in the new environment. Several variations of filter designs have been reported suitable to the emerging technologies [24,25]. Many of the filter designs do not exploit the inherent advantages that the new technologies offer, which motivated the need for new filter configurations in multi-layer. In this chapter, several new via-less filter configurations, which are suitable for realization in multilayered technologies, are presented. In section 4.2, different multi-layered sections are discussed. The proposed configurations exhibit compact footprints. In section 4.3, we discuss specific techniques adopted to suppress higher order harmonics.

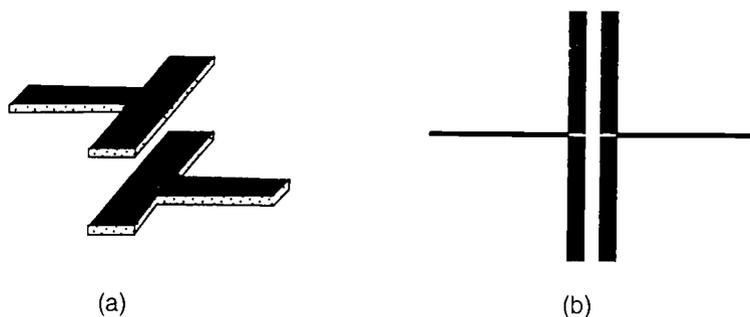
### 4.2. Filter sections

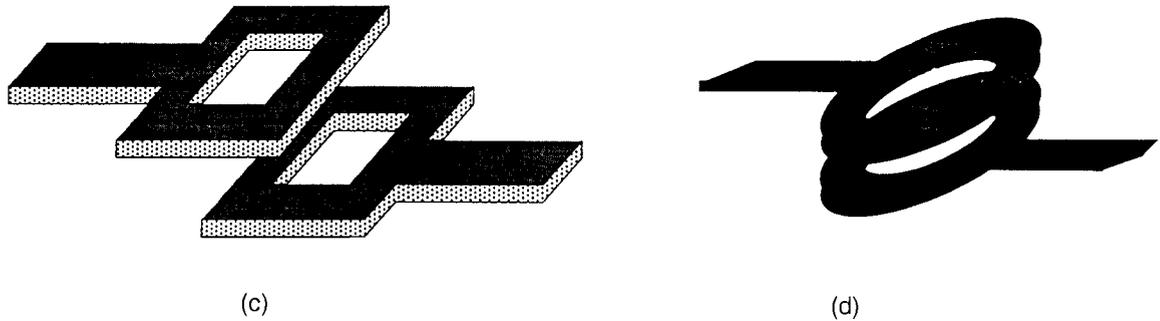
The bandpass filter consists of distinct sections, the performance of which is determined by the  $J$  and  $\Phi$ . The primary difference between single and multi-layer sections is that conducting traces exist on different dielectric layers. The complexity of design increases as issues like signal transition

from one layer to another, dielectric properties of each layer, etc., need to be addressed. The conventional means of signal transition from one layer to another in a multilayer structure is by means of a via. From a practical perspective, via is not a desirable feature due to the high parasitics and increased fabrication process. This disadvantage can be overcome by effectively utilizing the broadside coupling between the conductors on the different substrate layers to provide the path for signal transition. Thus, the via is eliminated and the multilayer geometry as an advantage of achieving higher values of  $J$ .

Some of the proposed multilayered filter section designs are shown in Fig.4.1. Depending on the dielectric layers, these structures can be adapted to obtain optimum advantage of the broadside coupling. One of the primary advantages in using these structures for filter sections is that the filter can be completely vialess. While designing a filter, a combination of these structures must be chosen so that the broadside coupling also serves to connect between layers without requiring a via, thereby taking advantage of the multilayer configuration without adding complexity to the design.

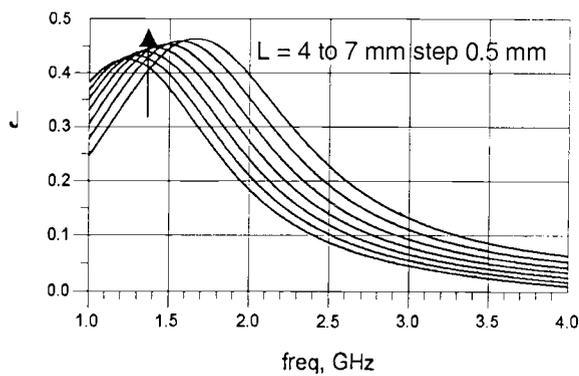
Each of these structures can be represented in terms of an admittance inverter network sandwiched between two negative length transmission lines as discussed previously. The expressions for calculating  $J$  from the network parameters are also dealt with in chapter 2.



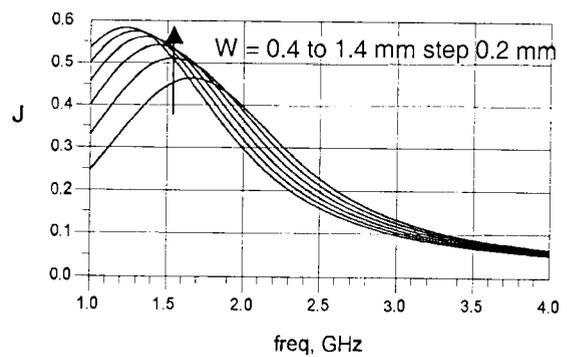


**Fig.4.1 Proposed configurations of multilayer filter sections**

The filter section shown in Fig. 4.1(a) can provide narrow as well as moderate bandwidths due to the wide variation of  $J$  resulting from the broadside coupling. The structure shown in Fig. 4.1(b) has thin length of transmission lines added on either side. These lines provide an inductive effect and act as harmonic suppressors. The structure in Fig. 4.1(c) has the square patch configuration. The two square patch traces are on different substrate levels and this gives excellent broadside coupling. The structure in Fig.4.1(d) is similar to the square patch structure. These family of structures exhibit some unique properties and their application to filter design will be discussed in detail.



$W=0.4\text{mm}$



$L=4\text{mm}$

#### Fig.4.2 Variation of J for the structure shown in Fig. 4.1(a)

The response of the filter may be predicted depending on the individual section characteristics. For instance, a section that provides a large range of J in most cases has severe harmonics. In this way, a pre-emptive measure may be adopted to obtain an optimum design. The following sections provide more information on these issues to the reader.

Fig.4.2 shows the variation of J for different values of length L and width W of the structure shown in Fig. 4.1(a). A uniform dielectric constant of  $\epsilon_r = 6.15$  with thicknesses 50 mil, 20 mil and 50 mil is chosen respectively.

### 4.3. Harmonic suppression

The telecommunication revolution and the emerging packaging technologies have brought about severe demands from microwave filters. The new multi-standard equipments have stringent requirements about possible interferences from nearby systems or internal components. Unfortunately, all distributed element filters suffer from higher order harmonics due to the periodic property of the resonators. It is important to suppress the higher order harmonics as high quality microwave filters play an extremely important role in designing communication systems such as cellular communications, radar equipment and global positioning systems. Many methods have been described in the past to overcome this problem based on the following ideas; push the spurious resonance responses far away from the specified bandwidth or to directly suppress the spurious response. However, most of these methods increase the losses and complexity of the systems. In fact, very few methods exist to suppress the higher order harmonics in a multilayer bandpass filter without degrading the performance of the filter. Since the new research and market trends are favoring multilayer filters, it is important to design methods to overcome the problems of higher order harmonics in these multilayer bandpass filters by simple means as suggested in section 4.3.1- 4.3.2. Fig 4.3(a)-(b) is an illustration of a simple approach to obtaining higher order

harmonic suppression without any noticeable increase in overall length of the filter. Fig. 4.3(b) shows a novel configuration, where a spur-line has been added on one side of the filter. Because of the bandstop property of the spur-line, this configuration can provide excellent higher order harmonic suppression. The characteristics and function of the spur-line is discussed in detail in coming sections.

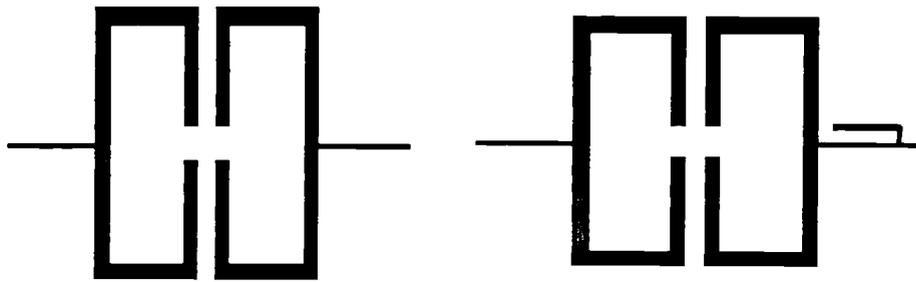
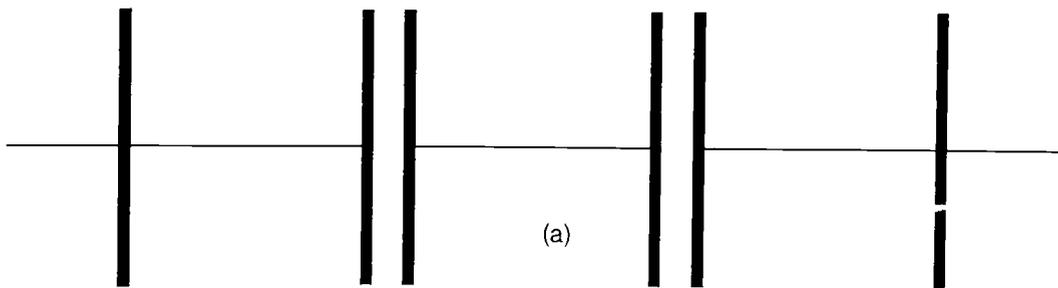


Fig 4.3 Illustration of harmonic suppression technique (a) before (b)after

#### 4.3.1 Stub loaded multilayer filters

In this section, we present bandpass filters with superior performance as compared to the conventional filters. These filters have been designed using the filter sections discussed earlier. A specific “stub-loading” technique has been implemented to suppress the higher order harmonic.



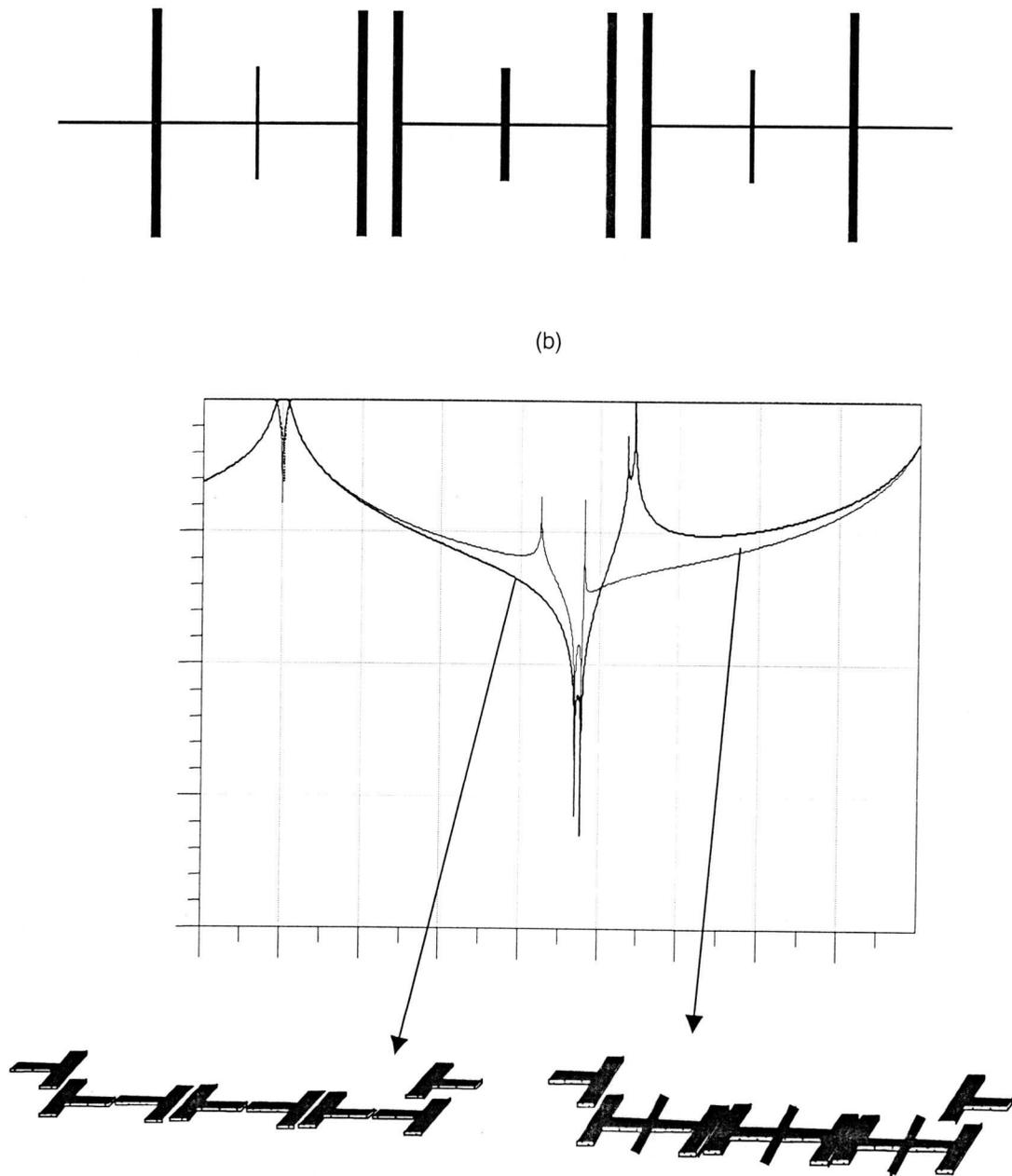


Fig.4.4 Response of a 3-resonator 2 GHz filter (a) without stub (b) with stub

Fig.4.4 presents two new configurations in three-layered topology. In the case of the filter shown in Fig.4.4(a), the end sections have been designed using a broadside-coupled quasi-lumped capacitor. The central sections have been designed using a single layer structure with a small variation. The ends of these sections have been loaded with stubs. It is seen that the filter-using stub exhibits excellent harmonic suppression. This filter was designed at a center frequency of 2 GHz and all harmonic components below 10 GHz were effectively suppressed.

### 4.3.2 Spur-line loaded filters

In this section, we discuss spurline loading of bandpass filters to suppress the higher order harmonics. The spurline resonator was first suggested by Cristal [26] for the particular type of resonator shown in Fig.4.4. Researchers have shown that this structure exhibits bandstop/lowpass characteristics. This configuration is assimilated in a bandpass filter section and the filter exhibits excellent higher order harmonic suppression. To the author's knowledge, it is for the first time that the spurline structure has been integrated to into a bandpass filter configuration to suppress higher order harmonics.

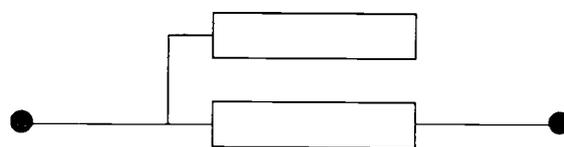
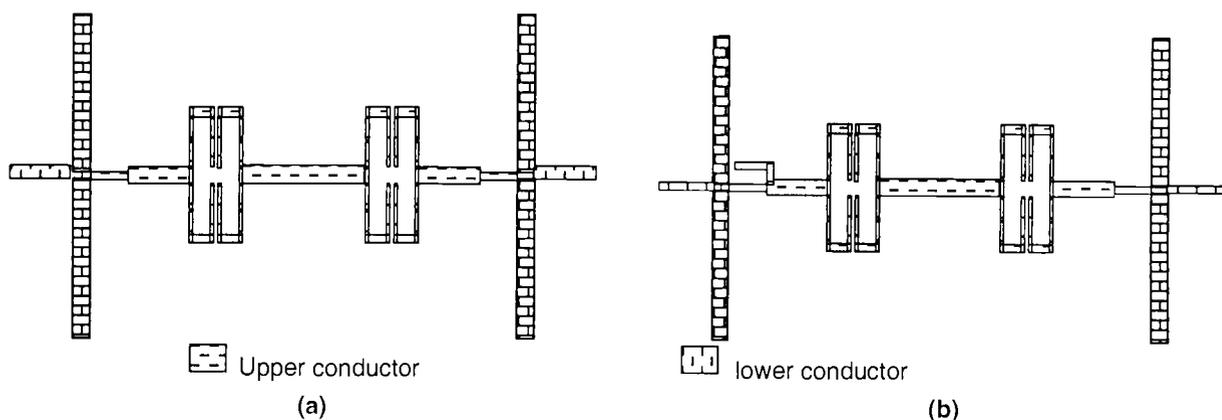


Fig 4.5 Spurline resonator structure

In this section, design methodology to include the spurline as a part of the bandpass filter has been carried out. The choice of the length of the spur is such that it throws a pole at the frequency, where the unwanted harmonic is occurring. Thus there is no degradation of the filter performance while the higher harmonic is suppressed. In order to illustrate the design procedure and

demonstrate the advantages, we present a comparative analysis of the filter performance designed for various cases.

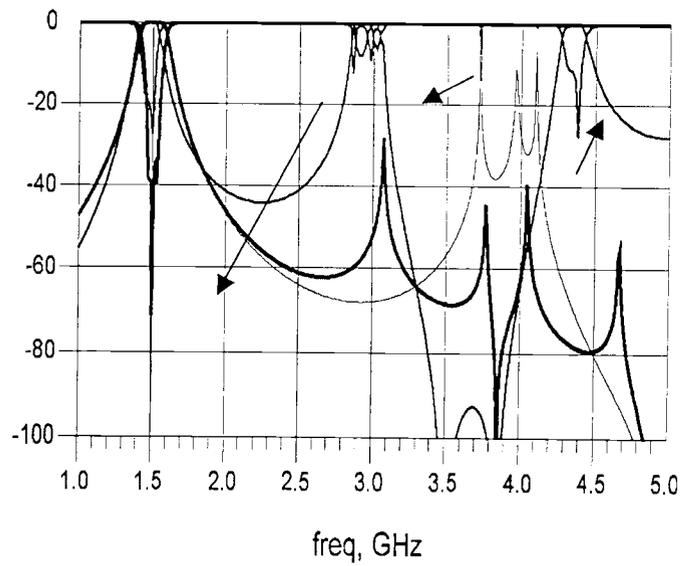
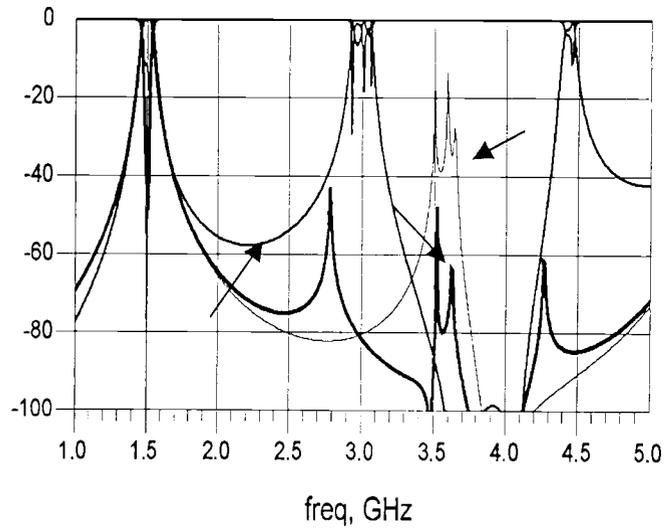
Designs are carried out with order  $N=3$ , center frequency  $f_0 = 1.5$  GHz on a three layered platform with a maximally flat response. The  $J$  values for the different percentage bandwidths are computed[20].

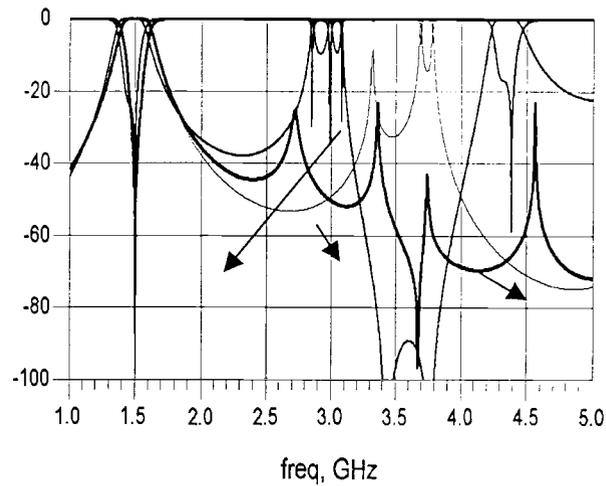


**Fig. 4.6 Footprint comparison of new geometries of filters with  $f_0 = 1.575$  GHz realized on three layers**

Fig. 4.6 shows the layout of the new vialess filters. The filter shown in Fig. 4.6(a) has been designed using the sections shown Fig. 4.1(a) and 4.3(a). Unlike the conventional parallel-coupled filters, this configuration offers second harmonic suppression as shown in Fig.4.7. The harmonic suppression achieved is a result of the new configuration employed in the multilayer configuration. The structure shown in Fig. 4.6(b) shows a novel spur-line loading of the multilayered bandpass filter. This has been realized using the structures shown in Fig.4.1(a) and 4.3(b). Spur-lines structures have been implemented as bandstop or lowpass filters but it is for the first time that they have been integrated into the bandpass filter design for harmonic suppression. It is interesting to note that the first three harmonics are suppressed as a result of the spur-line loading. The simulated response for the structures in Fig.4.6 is shown in Fig. 4.7(a), 4.7(b) and 4.7(c) for 5%, 10% and 15% bandwidths

respectively. For a center frequency of 1.5 GHz, harmonics up to 5 GHz have been suppressed to -40dB with an integrated spur-line configuration.

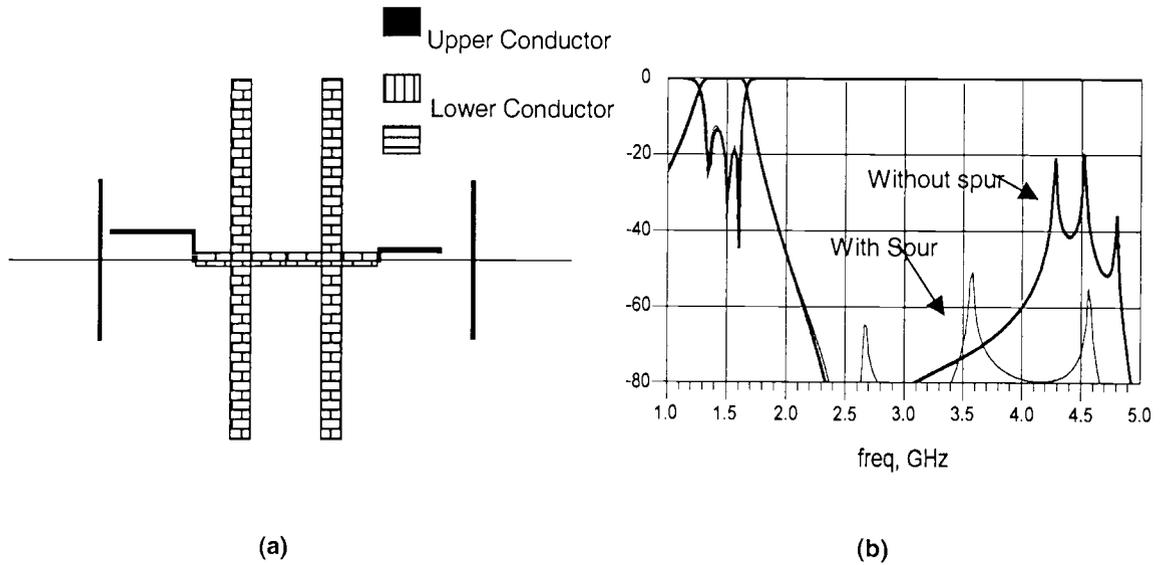




**Fig.4.7 Response comparison of new filter configurations with that of parallel-coupled line filters**

The response for the new filter configurations is compared with that of a conventional parallel-coupled filter designed on a microstrip platform. For each bandwidth, two designs were carried out with and without spur-line loading. It should be noted that there is no significant increase in the filter length as a result of the spur-line loading. It may be observed that better harmonic suppression is obtained for lower bandwidths as can be seen in the 5% case as compared to the 10% and 15%.

For further validation of the spur-line loaded bandpass filters, a four-layered 3-resonator bandpass filter is designed at 1.5 GHz using symmetrical sections for a 15% bandwidth. The filter has the first and last sections with spur-line loading. Fig.4.8 shows the filter layout and the frequency response of the filter. The analyzed response shows an excellent suppression of -50dB up to the third harmonic. In general, it was observed that multi-layered configurations offered better harmonic suppression compared to single layer structures.



**Fig.4.8 Response of a four-layered filter designed for 15% bandwidth**

One of the primary advantages in using these filter structures is that the filter can be completely vialess. Vias between interconnecting layers in filters can add complexity to the design by introducing parasitic effects. While designing a filter, a combination of component structures must be chosen so that the broadside coupling also serves to connect between layers without requiring a via, thereby taking advantage of the multilayer configuration without adding complexity to the design.

## 5. ELECTROMAGNETIC SIMULATION BASED FILTER DESIGN PROCEDURE FOR SQUARE PATCH VIALESS FILTERS

Several software packages exist for performing full-wave electromagnetic simulations at microwave frequencies. In recent years, increase in complexity of systems demand accurate predictability of component performances leading to greater dependence on results of simulation. The inherent disadvantage of these slow-running simulations is that they are highly time-consuming. The current models for these simulations may require several iterations for achieving the desired specifications. Fine tuning the physical parameters or a marginal change in the requirements of the system can lead to increased turn around time. With increased interest in the field of passive components, efforts are focused on developing new designs and techniques that are fast and efficient [27,28]. Considering the importance of bandpass filters in today's technology world, a design technique that that is fast and efficient with high degree of design accuracy can be potentially helpful. The technique should lead to fast running simulations with accurate results and must be adaptive to a wide range of specifications.

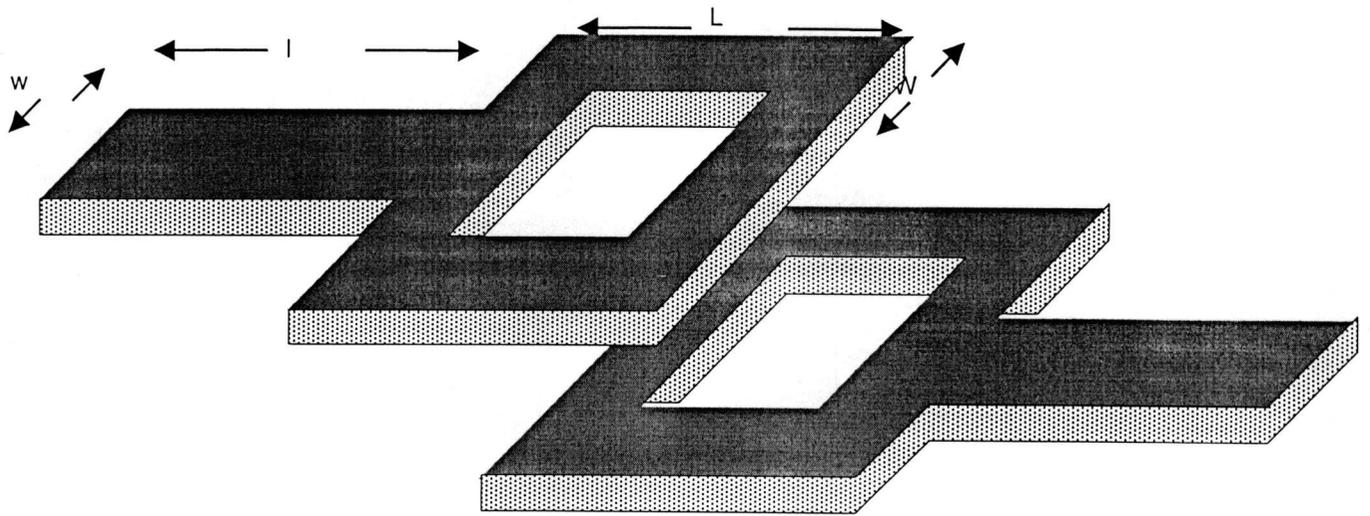
In this chapter, a dynamic design technique with a fast running simulation algorithm is proposed. The design theory discussed earlier in this work has been applied to a structure with unique properties leading to a novel design procedure. The procedure is based on conventional design methodology and does not require optimization of the physical parameters leading to a first time successful design. To demonstrate the potential of the proposed technique, designs are carried out for a wide variety of bandwidths as well as types of filter. Square-patch filter sections shown in Fig. 5.1 are considered for the purpose of demonstrating this technique. A closed form expression for extraction of design parameters is presented. To demonstrate the feasibility of the proposed approach, the design of filters of varying order has been carried out. Results have been validated by comparison with measurement.

### 5.1. Theory

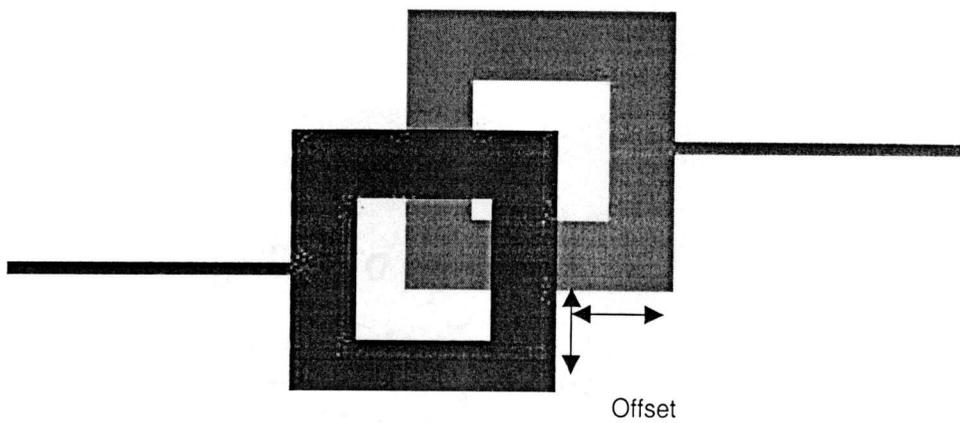
The filter design process using J-inverter network has several advantages but the drawback is it is as time consuming as the conventional design process if the network parameter extraction for the filter sections is based on full-wave electromagnetic simulation. This is due to the fact that this design process requires optimization of the dimension to obtain  $\Phi = 180^\circ$  for each section with a specific value of J. The optimization results in arduous simulation for each iteration. The requirement to reduce the time is that one of the parameters preferably  $\Phi$  remains constant. It is observed that an offset coupled line provides such a unique property. It is seen that as the offset between the coupled lines are increased the J value varies over a range while the  $\Phi$  remains close to  $180^\circ$ . Thus, based on this property, it is possible to derive a family of structures that can provide varying values of J with minimum deviations from the ideal value of  $180^\circ$  for the same dimensions. These structures are chosen to be broadside coupled multilayer structures in order to obtain a wide range of J in lieu of their greater coupling. The offset between the traces on the different layers is varied to obtain a wide range of J. Such structures eliminate optimization of filter sections to achieve the  $\Phi = 180^\circ$ . Thus, it is possible to design any type of filter with the same section just by varying the offset between the conductor layers. The offset can be varied in definite intervals along the X-axis, Y-axis or the XY-axis (diagonal offset). The choice of the degree of offset dictates the characteristic of the structure viz. the range of J, the sensitivity of  $\Phi$ , etc as is discussed in the forthcoming section.

### 5.2. Square patch

In order to demonstrate the feasibility of the procedure a specific case of square patch is presented. In this section, the properties of a square patch that can be used to design a wide variety of filters are presented. The structure consists of two square patches of length L and width W in broadside coupled configuration. The input and output lines are symmetric having length l and width w.



(a)

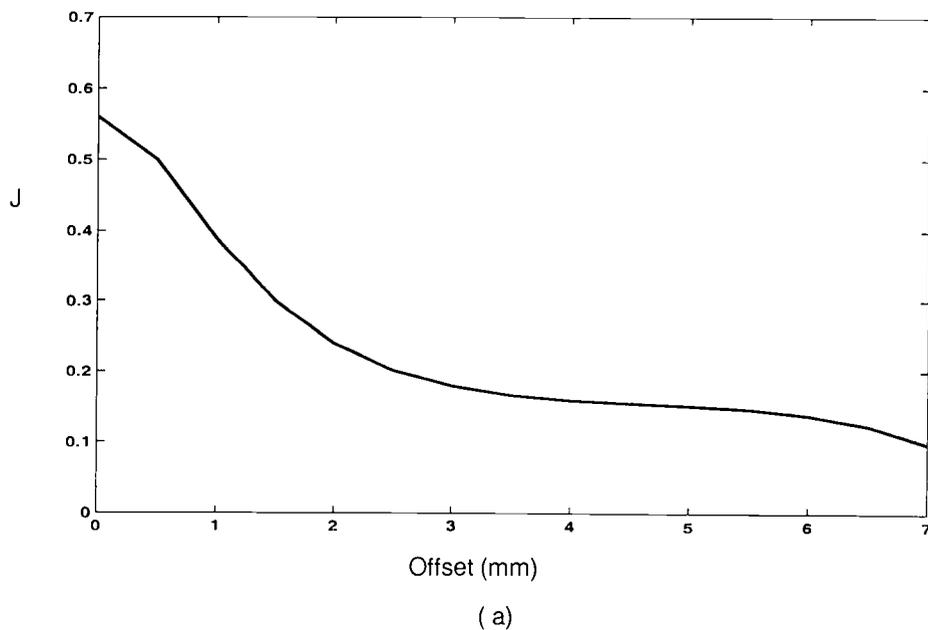


(b)

Fig.5.1 Schematic representation of the square patch (a) Side view (b) top view

Each set of physical dimensions exhibit a range of  $J$  with a relatively small deviation from the ideal value of  $\Phi$ . The range of  $J$  that can be obtained for any structure also depends on the substrate characteristics.

Fig.5.2 and 5.3 show the variation of  $J$  and  $\Phi$  for different values of  $L$ ,  $W$ ,  $l$  and  $w$ . Dielectric constants  $\epsilon_r$  of 2.2, 3.2 and 2.2 are chosen for the three-substrate layers of thicknesses 62, 30 and 62 mils respectively. It is observed that the marginal deviation of  $\Phi$  from the ideal value does not affect the overall filter performance. An extensive study of the structure showed interesting traits. These serve as guidelines in choosing the dimensions of the square patch for maximum flexibility in design. Smaller the value of  $L$ , larger the value of  $J$  achievable. This is a direct consequence of increasing  $l$  to achieve the ideal  $\Phi$  for the structure to start with.



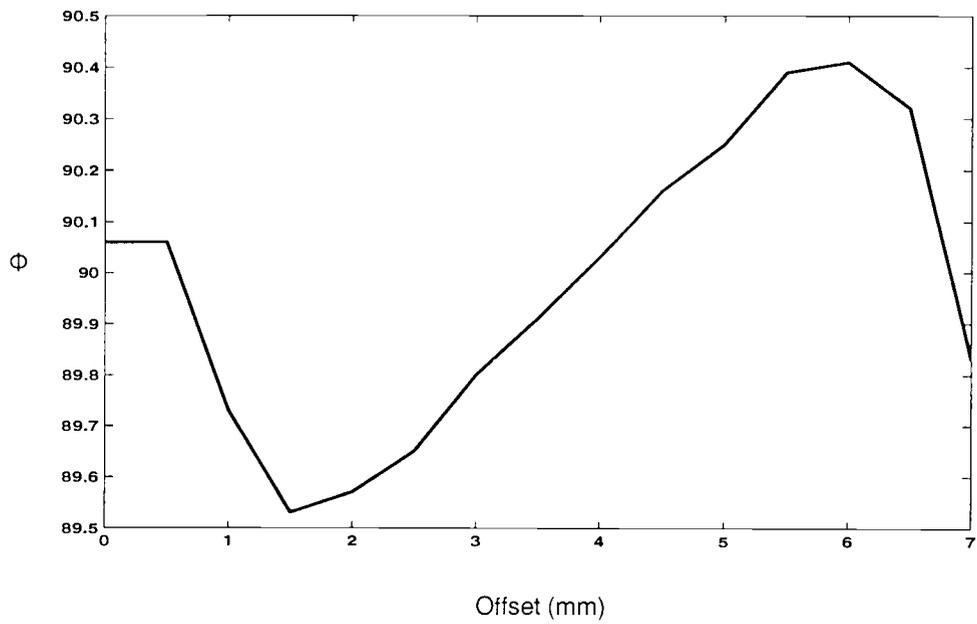
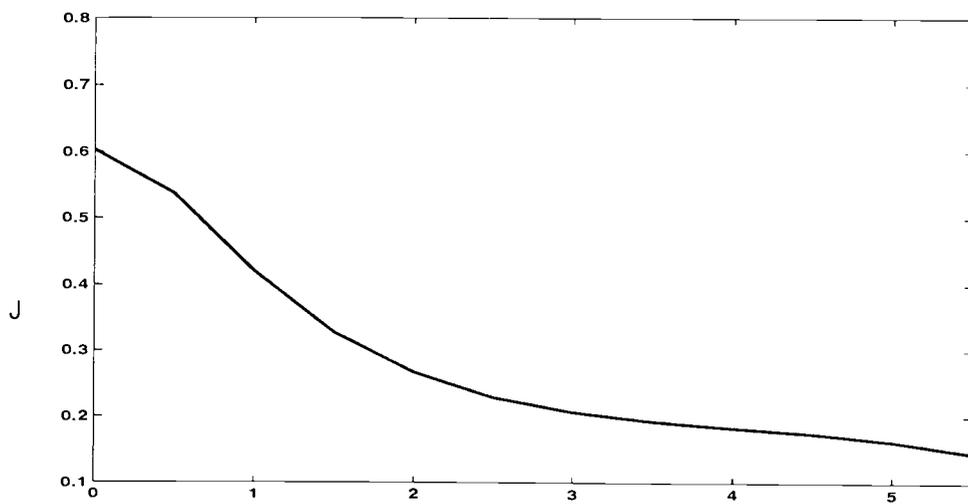
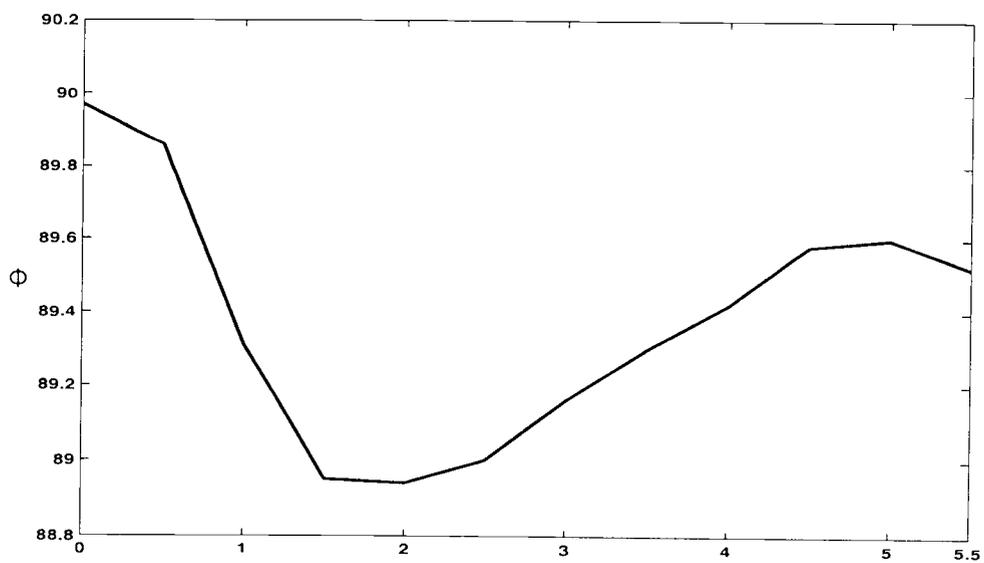


Fig 5.2 Variation of J and  $\Phi$  with offset for L=8 mm, W=0.5 mm, l=5.53 mm, w=0.55 mm



(a)



(b)

**Fig.5.3 Variation of  $J$  and  $\Phi$  with offset for  $L=7$  mm,  $W=0.5$  mm,  $l=6.66$  mm,  $w=0.5$  mm**

On the other hand, lower the dielectric constant, larger values of  $L$  required for achieving higher values of  $J$  and this in turn increases the sensitivity of  $\Phi$  to offset. Thus, there is a tradeoff between range of  $J$  achievable and sensitivity. Any particular specification can be achieved in multiple ways. A discerning designer will opt for the structure that provides maximum flexibility, minimum sensitivity and demands least board area. In general, it is observed that any change in widths  $W$  or  $w$  will have a severe effect on the structure behavior than the corresponding change in lengths  $L$  or  $l$ . Furthermore, the substrate spacing between the conductor layers must be chosen such that it gives sufficient coupling to achieve required values of  $J$ , retaining minimum variation of  $\Phi$  with offset.

### 5.3. Multilayer filter design

To demonstrate the potential of the proposed theory, a wide range of filters each having different specifications is chosen.

Table 5.1 gives the values of  $J$  required for different sections of these filters:

Filter Type $\rightarrow$ $\Delta$ $\downarrow$	Maximally flat	Chebyshev 0.1dB	Chebyshev 0.01dB
	10%	0.3963	0.3902
0.1111		0.1444	0.2011
15%	0.4854	0.4779	0.6120
	0.1666	0.2166	0.3016
20%	0.5605	0.5519	0.7067
	0.2221	0.2888	0.4021

**Table 5.1 Values of  $J$  for different filter specifications**

Fig.5.4 (a) shows the footprint of a maximally flat filter designed for a 10% bandwidth on a three layered substrate of 50,30 and 40 mil respectively and a dielectric constant  $\epsilon_r=6.15$ . Fig. 5.4b shows the filter response for the theoretical response as well as the full-wave electromagnetic simulation for the entire filter. It may be observed that both are in excellent agreement.

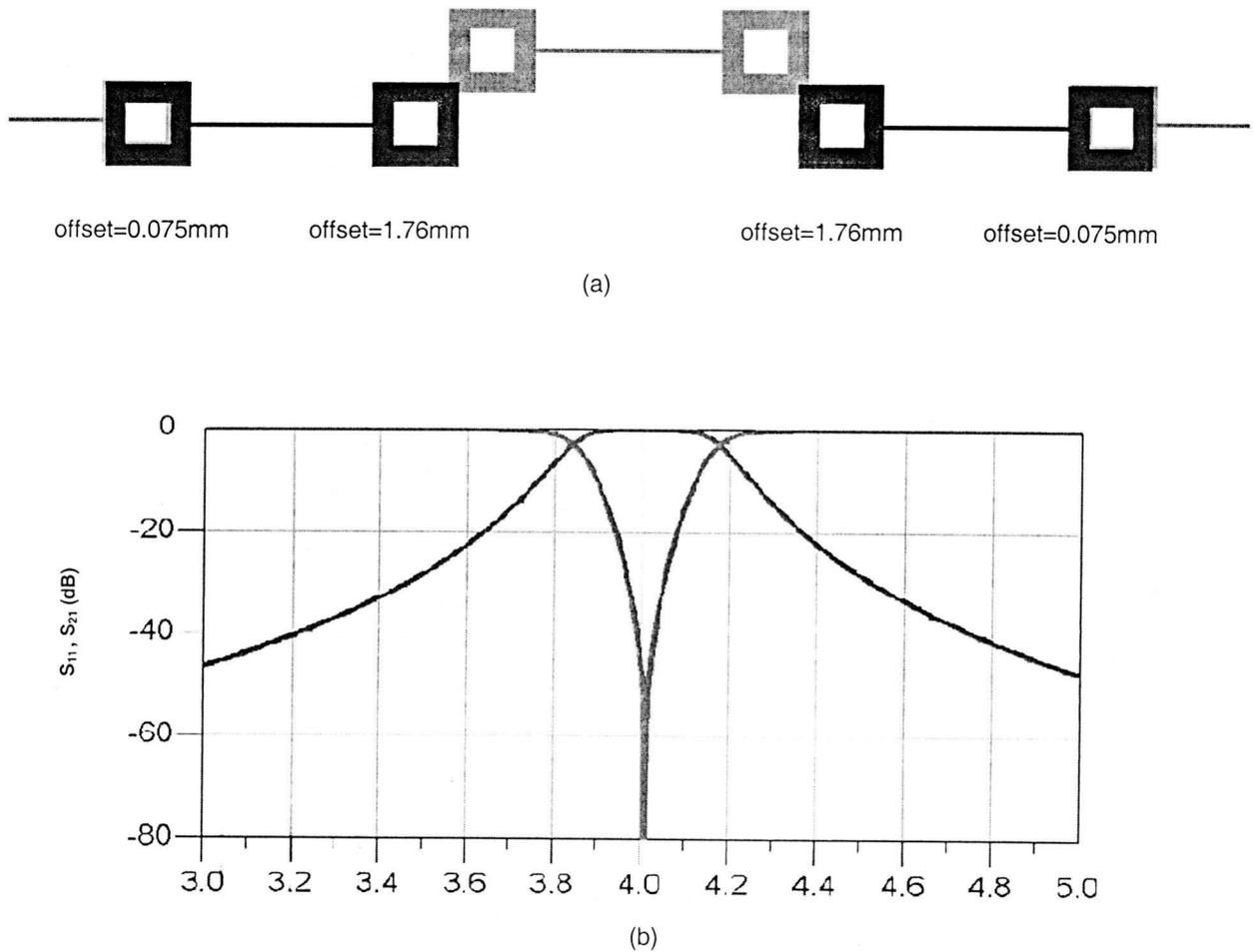


Fig 5.4 (a) Footprint of maximally flat filter with 10% bandwidth (b) response of the filter simulated using the proposed algorithm.

To exhibit the scope of the proposed theory, a single structure, which can be used for obtaining all the filters within the range shown in table 5.1, was designed. In order to achieve all of the above-mentioned values a square patch that provides a considerably large range of  $J$  is desired. Therefore, a substrate that has a fairly lower value of dielectric constant  $\epsilon_r$  is required. It is observed that lower bandwidths translate to higher structure sensitivities, demanding a structure that is inherently very stable. With these design considerations, a square patch of dimensions  $L=6$  mm,  $W=0.68$  mm,  $l=6.85$  mm and  $w=0.3$  mm and diagonal offset is found to be most suited with substrate specifications as earlier. The variation of  $J$  and  $\Phi$  for the structure is shown in Fig. 5.6. This data is obtained by running full wave electromagnetic simulations using HP momentum [29]. The distinct advantage of the proposed design technique is that full wave electromagnetic simulations need to be carried out only at discrete points to obtain the J-curve. These single point simulations are fast running and are extremely accurate. Thereafter, the design process does not require any electromagnetic simulations, thus eliminating the cumbersome process of time-consuming simulations. By curve fitting the data a closed for expression for offset  $os$  in terms of  $J$ , is obtained as [30]

$$os(mm) = -4859.5J^6 + 13395J^5 - 14928J^4 + 8512.7J^3 - 2566.1J^2 + 366.48J - 14.422$$

(5.1)

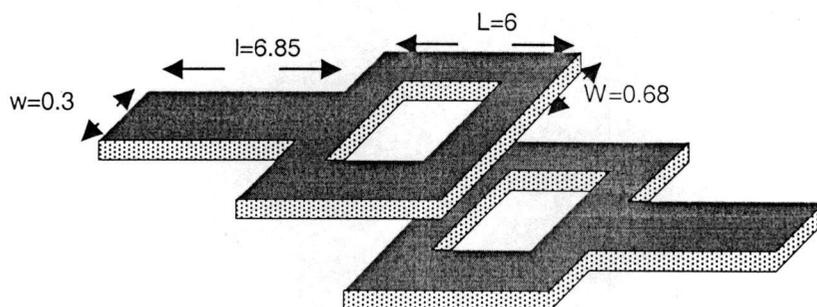
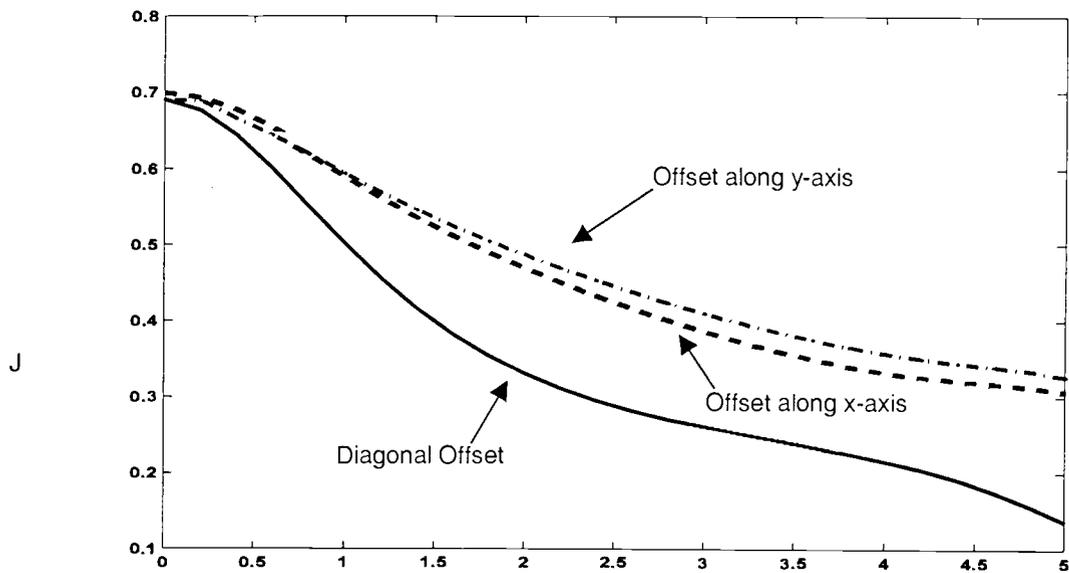
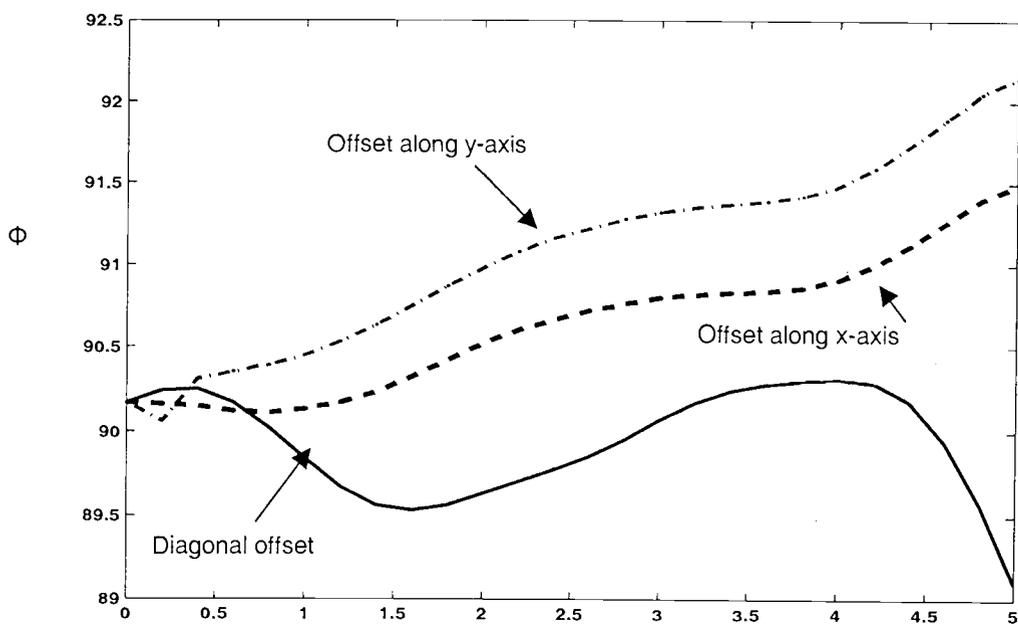


Fig. 5.5 Square patch- the single building block for all filters in table. 5.1.



Offset (mm)  
(a)



Thus with a few single point simulations accurate first time successful filter design can be carried out for a wide range of specifications. To elucidate the theory, the single structure shown in Fig. 5.5 is used as building block for all filters shown in Table 5.1. The comparison of theoretical and full wave electromagnetic simulation results carried out and they are in excellent agreement.

To demonstrate the design feasibility a 6-section filter structure with substrate thicknesses of 62, 30 and 32 mil is considered. A 6-section Chebyshev type filter with a passband ripple of 0.1dB at a center frequency of 2 GHz and with a desired fractional bandwidth of  $\Delta=0.15$  is designed. The required values of  $J$  for the different sections are 0.4533, 0.1879 and 0.1432. Fig.5.9 shows the footprint and the full wave electromagnetic simulation using momentum. The result shows excellent agreement with the predicted response proving the accuracy of the proposed technique.

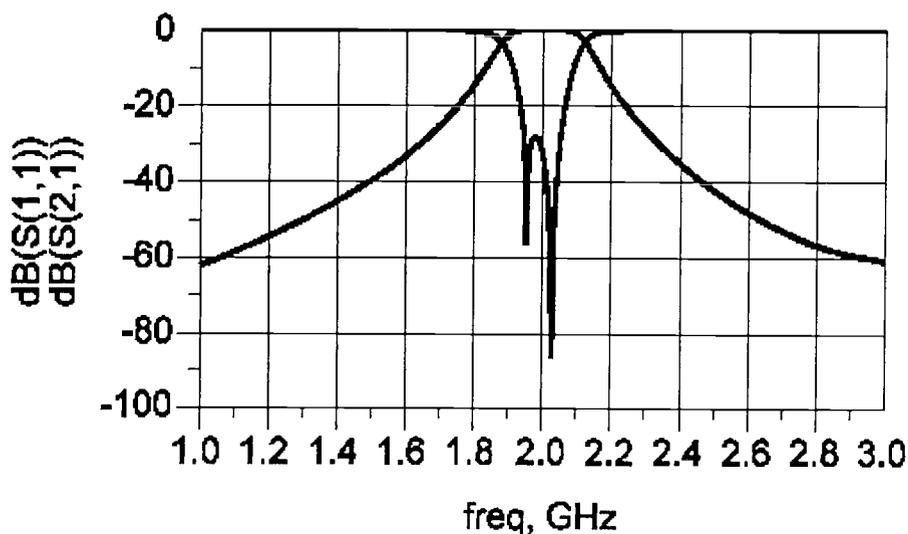


Fig. 5.7 Response of a 0.01dB ripple chebyshev bandpass filter with  $\Delta=0.10$ .

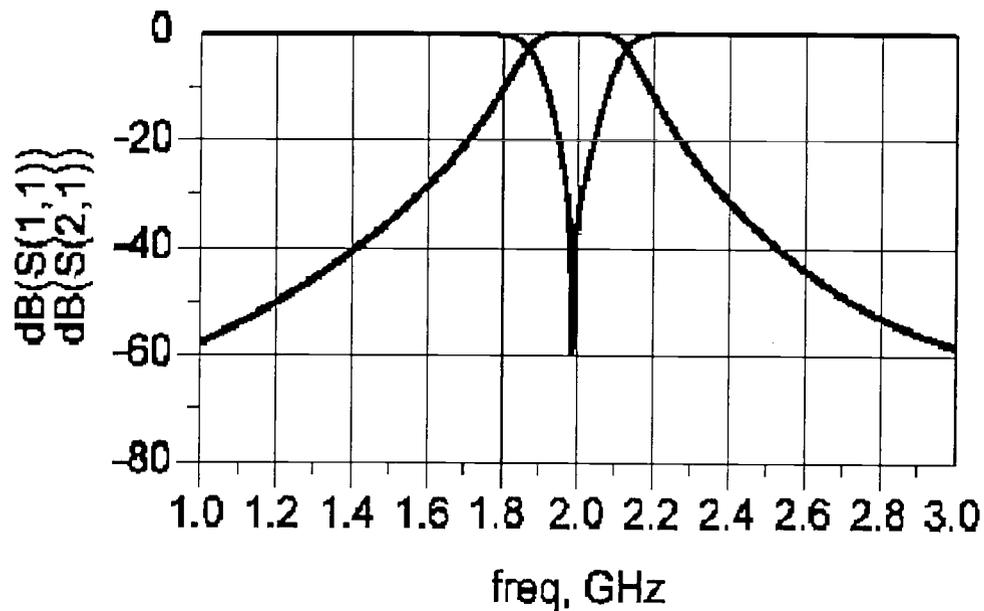
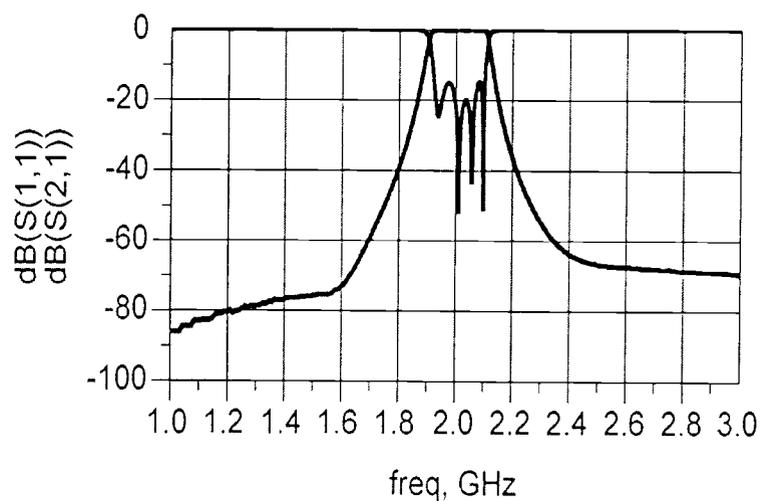


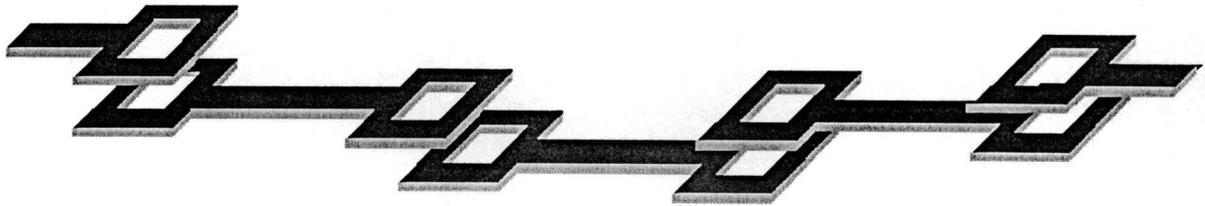
Fig. 5.8 Response of a maximally flat bandpass filter with  $\Delta=0.20$ .

A simple new design technique for efficient first time successful design of bandpass filters has been proposed. A theory to synthesize a variety of bandpass filters in minimum time has been presented. The proposed technique defies the conventional tradeoff between time versus accuracy of simulation. A simple closed form expression to determine the offset has been put forth. Several new configurations in multilayer have been designed using a single building block. The inherent advantage of these filters is that they are via less and hence suited for emerging packaging

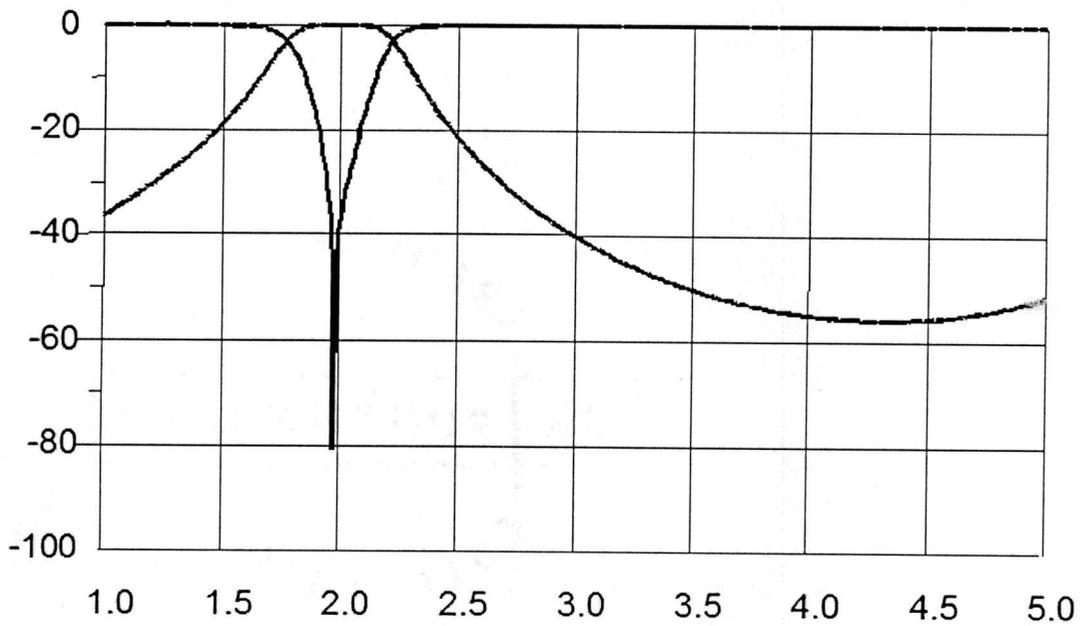
technologies. To demonstrate the design flexibility, a 6-section filter has been designed using the proposed technique.



Four-layered maximally flat filter was developed with a 20% fractional bandwidth. The dielectric constant was chosen as 6.15 with layer thicknesses of 40,10,30,40 mil respectively. The filter has been designed at a center frequency of 2GHz. The predicted response is superimposed on the fullwave electromagnetic response as shown in Fig.5.10.



(a)



(b)

**Fig. 5.10 (a) Footprint of the four-layered filter (b) response**

The technique exhibits a great potential, as it is fast running and accurate opening up avenues for electromagnetic simulations based design of microwave components.

## 6. IMPLEMENTATION AND RESULTS

In this chapter, we validate the proposed theory by fabricating some of the proposed filters. We explain how we progressed from the design phase to the fabrication phase and subsequently to the measurement phase. Since the fabrication depends on the tolerances of the facility, certain precautions were necessary in order to avoid compromise on filter performance. We explain these precautions and steps that lead to fabrication of the filters. Some interesting results are presented with relevant discussions.

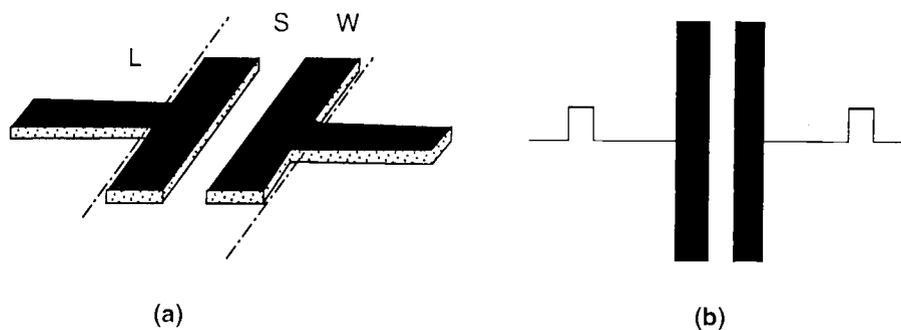
### 6.1 Stub loaded filter on single layer

In chapter 3, we discussed the implementation of filters on single layer and effective suppression of higher order harmonic using the stub loading technique.

In order to prove the validity of this approach a stub loaded third order bandpass filter with center frequency  $f_0 = 1.5$  GHz and bandwidth 10% was designed on a microstrip platform with  $h=0.7$ mm and  $\epsilon_r = 2.2$  was designed. The proposed fabrication was done at our in-house fabrication facility and hence RT 5880 duroid substrate was chosen. The measured response was to be validated with an accurate full wave electromagnetic simulation a center frequency of 1.5 GHz was chosen. An initial study carried out to prove the accuracy of simulations showed that at this frequency there are accurate models available to match mathematical and full wave simulations. At higher frequencies, the simulation tool showed a negligible frequency shift. Hence, a center frequency of 1.5 GHz was chosen.

The initial design process was as described in chapter 2. The  $J$  and  $\Phi$  were obtained for each of the four sections. The structure is symmetrical along the  $y$ -axis. The second and third sections were realized using a quasi-lumped structure as shown in Fig 6.1(a). The structure shown in Fig.6.1 (b) is

employed to realize the end sections of the filter. The short lengths of the thin transmission lines on either side of the structure in Fig.6.1 (b) behave like series inductors at high frequency because of their high impedance. This reduces the overall filter size and gives harmonic suppression. Additional degrees of freedom arise by varying the length of the folded line in this structure. In order to achieve an improved performance additional stub loading was carried out. The single stub was employed to effectively suppress the higher order harmonics in the filter design. The stub width was chosen to be 10 mil so that its addition does not affect the filter performance due to additional phase delay. The width of lines on either sides of the gap coupling is kept minimal so that the coupling between the tee junctions can be neglected.



**Fig.6.1 Sections of fabricated filter (a) end section (b) central section**

The predicted response is shown in Fig.6.2 (ii). The performance of the proposed filter structures is compared with the conventional parallel-coupled line filter. It is interesting to observe the extended response of the filter realized using structures in Fig.6.1 without any stub loading. The harmonics up

to 4GHz is effectively suppressed by using the high impedance line employed in the structure. The response exhibits a harmonic of about -20dB at 4.1 GHz and this can be effectively suppressed by stub loading. The stub loading further improves the performance by suppressing the sixth harmonic below -50 dB without any significant increase in the size of the filter.

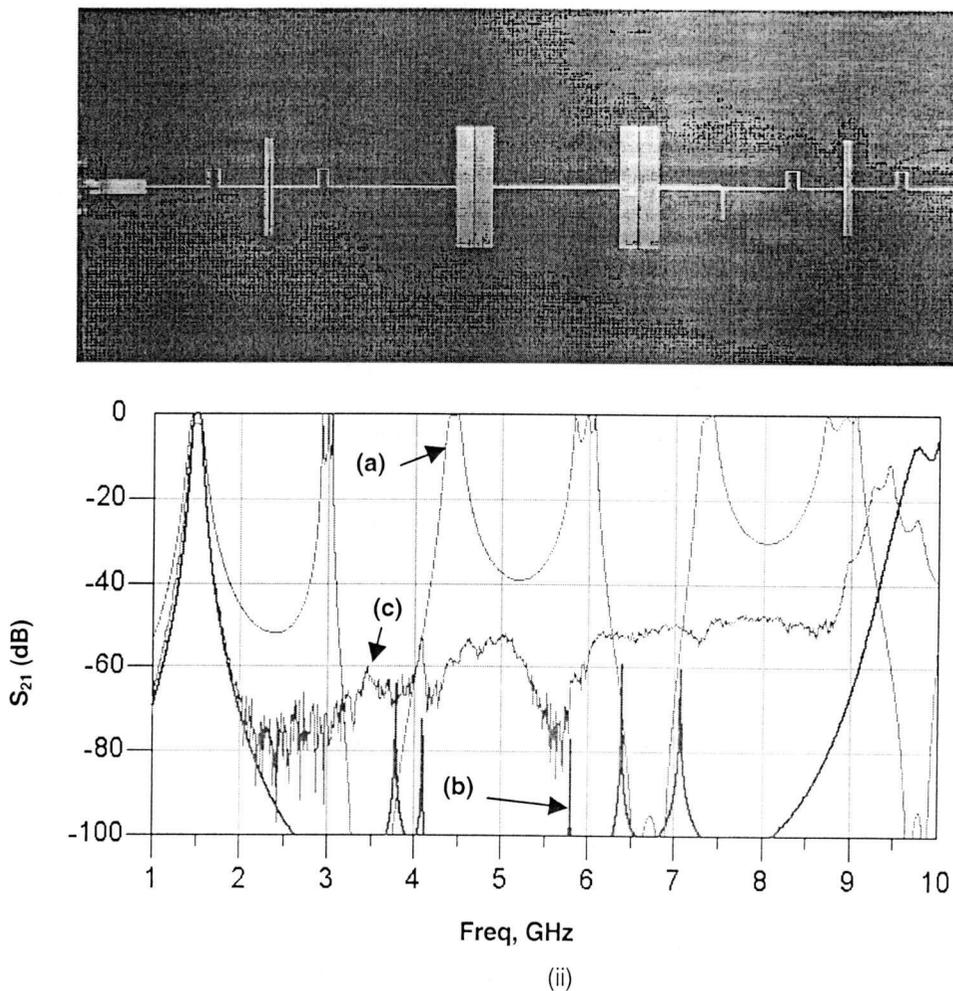


Fig.6.2 (i) Photograph of fabricated filter (ii) (a) Response of conventional filter (b) Predicted response(c) Measured response

The proposed structure was then fabricated on RT duroid 5880 substrate.

## **6.2 Measurement and testing**

The fabricated filter was tested on an HP 8722 vector network analyzer with a full two-port calibration up to 18GHz. The network analyzer is connected to a computer through a GPIB card to allow the capture of measurement results. These results were stored as .s2p files and plotted using commercial software.

The measured response is shown in Fig.6.2 (ii)(c). The measured response shows good harmonic suppression up to 9.4GHz, which is in good agreement with predicted results. The pass band performance satisfied all the specifications. The single layer bandpass filter fabricated above showed excellent performance. Nevertheless, it has an obvious drawback the full wave electromagnetic simulations required to proceed to the fabrication stage is time consuming. Furthermore, the structure being in single layer is unable to take maximum advantage of the attractive features offered by emerging technologies.

## 7. CONCLUSION

This chapter describes the highlights of the performed research. A summary and recommendations for future work have been provided.

### 7.1. Summary

The research was focused on developing new configurations of bandpass filters with distinct advantages. The groundwork was initiated with the study of existing filter theory. For instance, the insertion loss method and the disadvantages of the existing configurations were discussed. The J-inverter method was used for developing the new configurations of filters as it enabled accurate characterization for higher frequency and complex geometry.

As the research evolved, the drawback of considering symmetric configurations was overcome using generalized equations. The initial stages of research were focused on single layer symmetric filters that provided higher coupling coefficients. This led to more advanced techniques to control harmonics and improve performance. Efficient stub loading techniques were discussed and their usefulness was proved. A subsequent transition to multilayer structures was more involved. Besides suggestions for new configurations, a novel method of spur-line loading to suppress harmonics was suggested.

The ultimate goal was to reduce the turn around time in developing first time correct designs. The new design rule of the square patch achieved this goal with no accuracy trade-off. It is based on the unique property of the square patch to provide varying coupling coefficients with negligible change in the electrical length. The usefulness of this theory will lead to a whole gamut of applications.

## 7.2. Recommendations

With the above review of the work in the field of bandpass filters, it is clear that the focus is in achieving miniaturization and adding functionality to filters to obtain superior performance. Although this thesis proposed several new ideas in addition to a generic algorithm, a first of its kind –it opened avenues for a vast amount of additional work.

The J-inverter topology approach of designing filters can be extended to constant phase filters by a simple substitution of the low pass filter prototype element values with the constant phase delay filter prototype values.

With emerging multilayer technologies, the trend for stringent requirements would only increase. The bandpass filters proposed here could be extended to other class of filters such as low pass and high pass filters. Concerning the stub loading and spurline loading, efficient design formulations to compute these dimensions would be significant.

It would be an interesting study to implement the proposed single and multilayer structures to obtain varied characteristics. Another possible avenue is the integration of these compact filters to the proposed bandpass filters to obtain suppression over an ultra wide bandwidth. Additionally, new planar substrates may also be experimented with.

The new rule proposed in chapter 5 may just be the basis for a plethora of structures with special properties that would change the rules of filter design. For instance, it was observed that a circular ring also exhibited a similar property. An extensive theory that explains the cause-effect relationship of the electrical length insensitivity of these structures to offsets would be a valuable contribution.

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