Miniaturized and ferrite based tunable filters in LCP and LTCC technologies for SoP applications

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ABSTRACT

Wireless systems with emerging applications are leaning towards small size, light-weight and low cost. Another trend for these wireless devices is that new applications and functionalities are being added without increasing the size of the device. To accomplish this, individual components must be miniaturized and the system should be designed to maximize the integration of the individual components. The high level of 3D integration feasible in system on package design (SoP) concept can fulfill the latter requirement.

Bandpass filters are important components on all wireless systems to reject the unwanted signals and reduce interference. Being mostly implemented with passive and distributed components, bandpass filters take considerable space in a wireless system. Moreover, with emerging bands and multiple applications encompassed in a single device, many bandpass filters are required. The miniaturization related to bandpass filters can be approached by three main ways: (1) at the component level through the miniaturization of individual bandpass filters, (2) at the system level through the use of tunable filters to reduce the overall number of filters, and (3) at the system level through the high level of integration in a 3D SoP platform. In this work we have focused on all three aspects of miniaturization of band pass filters mentioned above.

In the first part of this work, a low frequency (1.5 GHz global positioning system (GPS) band) filter implemented through 3D lumped components in two leading SoP technologies, namely low temperature co-fired ceramic (LTCC) and the liquid crystal polymers (LCP) is demonstrated. The miniaturized filter is based on a second order topology, which has been

modified to improve the selectivity and out-of-band rejection without increasing the size. Moreover, for the case of LCP, the filter is realized in an ultra-thin stack up comprising four metallization layers with an overall thickness of only 100 μ m. Due to its ultra-thin structure, the LCP filter is ten times smaller size as compared to the filters reported in published work. The filter is flexible and, therefore, suitable for conformal applications.

In the second part of this work, relatively higher frequency (Ku band) distributed bandpass filter is presented which can be tuned through an applied magnetic field. This has been realized in a relatively new LTCC tape with magnetic properties, known as ferrite LTCC. Traditionally, magnetically tunable filters require large external electromagnets or coils, which are non-integrable to typical planar circuit boards and are also inefficient. To demonstrate high level of integration, completely embedded windings realized in multiple layers of LTCC have been used instead of the external coils. As a result, the presented bandpass filter is several orders of magnitude smaller that the reported ones. Aside from reducing the size, the embedded windings based design is more efficient than the external coils because it can avoid the demagnetization effect (fields lost at air-ferrite interface) and thus require much smaller bias fields for tunability.

Though the embedded windings bring in a number of advantages as mentioned above, the currents passing through these windings generate considerable heat which can influence the performance of the microwave structure (bandpass filter in our case). This has never been studied before fro Ferrite LTCC based designs with embedded windings. In this work, the effect of the heat generated by these windings has been investigated. It has been found that this self-heating effect influences the tunability of the filter considerably so it must be estimated at the design stage. Therefore, a strategy to simulate this effect has been developed. The resultant simulations agree well with the measurements verifying the simulation strategy. The designs presented in this work demonstrate the feasibility of realizing highly integrated, miniaturized and tunable filters in SoP platform which are very suitable for modern and futuristic small form factor and slim wireless devices.

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Chapter 1

Introduction

1.1 Motivation

Modern wireless devices are characterized by multi-functionality. A typical smart phone, for instance, has a wide touch-screen, multiple digital cameras, multiple transceivers (WiFi, Bluetooth, GSM, GPS), a large battery and a complicated digital part. All these functionalities are required to be implemented within a small size. For the radio frequency (RF) front end, such increasingly high degree of integration requires incredibly small components. Moreover, when the different RF sub-systems are positioned close to one another to save space, self-interference becomes significant and needs to be reduced without sacrificing the small size. This means that the requirements for bandpass filters to have higher rejections is ever increasing. The focus of this thesis, is the design of miniaturized bandpass filters with increased selectivity.

Miniaturization of bandpass filters can be achieved by three main techniques, which have all been utilized in this work: (1) Miniaturization can be achieved by reducing the sizes of individual bandpass filters. (2) A typical RF system has several bandpass filters; therefore, miniaturization can be achieved in a system level by using a single, tunable filter to replace all these filters. (3) Miniaturization at a system level can be achieved by integrating more components (including filters) within the same available volume. The later technique has led to the introduction of the system on package (SoP) concept, which is explained next.

1.1.1 The system on package (SoP)

Most of the electronic systems commercially available use printed circuit boards (PCB) to provide interconnectivity between system components. In such systems, components have multiple packages that do not provide any added functionalities. The integrated circuit (IC), for instance, starts with a silicon die which needs to be packaged for system integration as shown in Fig. 1.1 (a). The packaged IC itself is mounted on the PCB, which is packaged again. This redundant packaging is avoided in the SoP concept by utilizing a unified package, where all components are directly integrated (Fig. 1.1 (b)). This system design provides optimal integration; therefore, system-level miniaturization [14].

Historically, most of the miniaturization efforts have focused on reducing the size of the transistor in the IC. ICs, however, occupy only about 10 % of the system's volume while other components such as passives, antennas, batteries etc., fill about 90% of the system [15]. Therefore, a better approach is to miniaturize all system components and not just the IC by utilizing a unified package with all components integrated within. Moreover, instead of using the package as a mere holder, functionalities can be added to it to further reduce the size. For instance, a bandpass filter or an antenna can be implemented within the already used package, which saves considerable space. To achieve this, a packaging technology that has 3D integration capabilities and excellent RF characteristics is desirable.

There are two main 3D technologies used for the SoP: low temperature co-fired ceramic (LTCC) [16, 17], and multi-layer liquid crystal polymers (M-LCP) [18–20]. LTCC is a ceramic based, mature technology, which has become an industrial standard. M-LCP, on the other hand, is an organic-based technology, which has very promising RF as well as packaging characteristics. In this thesis, the potential of these two technologies for the SoP, where components are implemented within the package, has been investigated. Miniaturized



Figure 1.1: Comparison between conventional PCB-based systems and the SoP. (a) PCB-based system with various components mounted (b) System on Package (SoP) with components integrated.

bandpass filters have been designed and tested in both LTCC and LCP, and a considerable size reduction has been obtained.

1.1.2 Miniaturization of bandpass filters

Modern wireless systems are required to have multiple RF connectivities, which poses great challenges on the designs of RF transceivers as high isolation is required between the different RF signals occupying close frequency bands. Also, the wireless spectrum is getting more dense due to the increasing demand for wireless services. The global mobile data has increased 81% in 2013 with over half a billion mobile devices and connections added [21]. For the RF receiver this means more stringent requirements for its selectivity as well as outof-band rejection. The key component in any RF front end, which sets its selectivity and rejection is the bandpass filter. Filters are used in wireless transceivers to reject the noise and interfering signals before they reach the non-linear devices such as amplifiers and mixers [22]. The design of miniaturized filters at low frequencies (below 10 GHz) is challenging because of the inverse relationship between size and frequency. In this work miniaturized and highly selective filters have been designed and tested in both LTCC and LCP. The problem of the system miniaturization through the design of filters has been approached in two main ways: (1) The miniaturization of individual filters. (2) The miniaturization of the system by reducing the number of filters, which has been achieved by using tunable filters. It is also worth mentioning here that all the filters in this work have been designed for the SoP platform; therefore, additional miniaturization is inherently achieved due to the high integration capabilities introduced by the SoP.

1.2 Objectives

The goals of the this work are listed below:

- Design of bandpass filters for the SoP technologies with considerable size reduction as compared to the literature. Selectivity and out-of-band rejection needs to be improved without a considerable increase in the size.
- Utilization of LCP's ultra-thin capabilities to design ultra-thin bandpass filters for the GPS band, which are suitable for slim and flexible devices
- Design of magnetically tunable and miniaturized bandpass filters for the SoP platform.

1.3 Challenges

The challenges of this work can be summarized as follows

• At low microwave frequencies (below 10 GHz), the miniaturization of bandpass filters, which are based on distributed components, is particularly challenging because their sizes are inversely proportional with the frequency. Discrete lumped components are used but they can not be implemented in the package of the system and has to be surface-mounted, which increases their cost and reduces their reliability.

- Improving bandpass filters selectivity without increasing their sizes is challenging because improving the selectivity requires, either increasing the order of the filter or the use of transmission zeros. In both cases more components are added to the design which increases the size.
- The use of transmission zeros to improve the selectivity results in a deteriorated outof-band rejection. Improving the out-of-band rejection by adding more transmission zeros also results in larger sizes.
- Tunable filters, which are based on ferrite materials need external windings for biasing. These windings are usually very large, which makes their miniaturization and integration very difficult.
- The external windings require high currents because some of the magnetic field is dissipated at the interface between the coil and the dielectric, which is known as the *demagnetization effect*. The high currents requirement is difficult to achieve in small, battery powered devices.
- The commercial ferrite LTCC tape used in this work was not developed for microwave frequencies; therefore, no proper characterization data are available and the characteristics of the material is not optimized for applications at the microwave frequencies.

1.4 Contributions

The contributions of this work are listed below:

1. A library of lumped component have been implemented in M-LCP in the thinnest stack-up reported with thickness of 100 μ m. Moreover, similar components have been designed and tested in LTCC for the sake of comparison.

- 2. An LTCC-based miniaturized bandpass filter for the global positioning system, has been designed and tested, which demonstrates miniaturization, improved selectivity and out-of-band rejection.
- 3. For comparison purposes, the same bandpass filter is re-designed in the M-LCP technology. This design demonstrates a considerable size reduction as compared to the literature due to the ultra-thin capabilities of M-LCP.
- 4. A ferrite LTCC-based tunable and miniaturized bandpass filter at fifteen GHz, has been designed and tested.
- 5. The effect of the heat generated from the embedded windings (self-heating effect) in the ferrite LTCC filter, has been measured and characterized.

1.5 Publications

- E. Arabi, and A. Shamim "The effect of self-heating on the performance of a tunable filter with embedded windings in a ferrite LTCC package", in IEEE Transactions on Components, Packaging, and Manufacturing January, 2015
- E. Arabi and A. Shamim "3D RF SoP Technologies: LTCC versus LCP", in Wiley's Microwave and Optical Technology Letters (MOTL), February. 2015.
- E. Arabi, F. A. Ghaffar and A. Shamim "Tunable Bandpass Filter Based on Partially Magnetized Ferrite LTCC with Embedded Windings for SoP Applications", in IEEE Microwave and Wireless Component Letters (MWCL), October, 2014
- E. Arabi, T. Vähä-Heikkilä, M. Lahti, and A. Shamim "A 3D Miniaturized High selectivity Bandpass filter in LTCC Technology", in IEEE Microwave and Wireless Component Letters (MWCL), Jan. 2014.

- E. Arabi and A. Shamim "3D Lumped Components and Miniaturized Bandpass Filter in an ultra-thin M-LCP for SoP Applications", in Progress in Electromagnetics Research C (PIERS), October, 2013.
- 6. E. Arabi and A. Shamim, "A Planar and Tunable Bandpass Filter on a Ferrite Substrate with Integrated Windings", *International Microwave Symposium (IMS)*, 2015.
- E. Arabi, G. Mckerricher, and A. Shamim, "Comparison of filters: inkjet printed on PEN substrate versus a laser-etched on LCP substrate", *European Microwave confer*ence, 2014.
- 8. E. Arabi and A. Shamim, "A miniature LCP-based quadrature hybrid coupler for GPS system on package (SoP)," *European Microwave Conference 2012.*
- E. Arabi and A. Shamim, "High Q, miniaturized LCP-based passive components, and filter design for SoP applications" in the European Microwave Conference, Manchester, 2011.



Chapter 2

Background and Literature Review

In this work, multiple bandpass filters have been designed and tested, which are presented in chapters 5 and 6. All these designs are based on the SoP platform, which offers tremendous miniaturization opportunities due to the 3D integration capabilities as discussed in the previous chapter [15]. Two miniaturization schemes have been considered in this work. (1) At the design of component level i.e. the filter itself and , (2) at the system level by reducing the number of filters required through a tunable filter approach. In this chapter, a review of the previously reported miniaturized as well as tunable bandpass filters for the SoP technologies, is presented.

2.1 Miniaturized, lumped-based bandpass filters

Implementing bandpass filters on the package of the system already results in miniaturization. Further miniaturization can be achieved by the optimal design of the bandpass filter. In this work the frequency targeted is 1.5 GHz, which is the frequency of the global positioning system (GPS). At this frequency, filters that are based on distributed components are large; therefore lumped components can be used, which can be conveniently implemented within the 3D technologies such as LTCC and LCP.



Figure 2.1: Illustration of the fabrication processes of the LTCC technology.

2.1.1 Miniaturized Bandpass Filters in LTCC

LTCC is a ceramic based, packaging technology that provides 3D capabilities. It has excellent packaging as well as RF characteristics for the SoP realization, which are shown in table 2.1. The permittivity of typical LTCC tapes for microwave applications is relatively low (around 5.5 [23]). This relatively low permittivity results in less parasitic capacitance between adjacent components, which means that the components can be positioned closer to one another resulting in system-level miniaturization. The loss tangent of LTCC is small (about 0.002 [23]), which makes LTCC excellent for RF circuits that require low losses such as filters. The temperature coefficient of the dielectric constant of LTCC (τ_{ϵ_r}) is (about -48 ppm/°C) [23] which indicates a stable operation across a wide temperature range. Being ceramic, LTCC is hermetic with zero moisture absorption which makes it an excellent packaging material.

The main steps in the fabrication of LTCC-based 3D modules are illustrated in Fig. 2.1. The fabrication steps are in this order, [16]: (1) the preparation of the green LTCC tape, (2) the drilling of the via-holes using Lasers or mechanical drills, (3) the realization of the metallized patterns by screen printing, (4) the lamination and firing of the different LTCC layers, and (5) the integration of surface mounted components and ICs.

Review of Bandpass filters in LTCC

Many lumped-based miniaturized filters have been reported in the literature in LTCC [1, 2, 2]24–29, 29–31]. In [2, 24] filters at 1.3 GHz and 2.5 GHz are presented with relatively large bandwidths of 27% and 15%, respectively. These filters demonstrate good insertion loses; however, their roll-off factors at the upper band are low. In [27], a bandpass filters at 2.4 GHz is presented in ten layers of LTCC, which shows a good roll-off factor at the upper band. However, the size of this filter is relatively large and its roll off factor at the lower band is low. In [25], a bandpass filter is presented at 1.55 GHz with good insertion loss and out-of-band rejection. This filter, however, does not use transmission zeros; therefore, the roll off factor is low. [28] demonstrates a bandpass filter with a very high roll off at the upper-band but its size is considerably large and the out-of-band rejection is low. In the bandpass filter of [1], high roll off factors at both the upper and lower bands have been achieved as shown in Fig. 2.2 in a very small size; however, the out-of-band rejection of the filter is low and its insertion loss is 5 dBs which is too high for most applications. Also, it appears that the two transmission zeros have been positioned too close to the passband, which has resulted in the improved roll-off clearly at the expense of the insertion loss. In general, it can be observed that most of the presented designs are implemented in large numbers of LTCC layers (average of 7) which increases their thicknesses as well as their cost. Also, the trade-offs between the size, the roll-off factor, the out-of-band rejection and the insertion loss are clearly observed. These trade offs are discussed next.

To further discuss the trade-offs presented in the previous paragraph, the problem of reducing the size of the bandpass filter is analysed. The first step in reducing the size of the filter is by reducing its order because lower order filters require less components [1,2,25,27]. When the order of the filter is reduced, however, its roll-off factor decreases. A high roll-off is a major requirement in today's congested spectrum; therefore, the roll-off needs to be improved without increasing the size. To increase the roll-off of low order filters, the concept of transmission zeros (TZ) is used. A transmission zero, as its name implies, is a zero in



Figure 2.2: The bandpass filter presented in [1]. (a) A photograph of the fabricated module, (b) the 3D structure and, (c) The measured and simulated results.

the transfer function of the filter's network, which occurs at a frequency near the passband. In [2, 30] a transmission zeros is realized by adding a capacitor in parallel with the filter as shown in Fig. 2.3 (a). The addition of the capacitor creates a second, parallel path for the RF energy, which can have a 180° phase difference with the main path at the right frequency; therefore creates a short circuit at the termination. As shown in Fig. 2.3(c), two transmission zeros have been obtained resulting in an increased roll-off factor. However, this has also resulted in a deteriorated out-of-band rejection as is clearly observed.



Figure 2.3: The bandpass filter presented in [2]. (a) The topology. (b) The 3D structure. (c) The measured and simulated results.

Two transmission zeros can also be obtained by connecting a capacitor in series with the filter as in [3] (Fig. 2.4(a)). In this case the addition of the capacitor provides a ground path for the RF energy which again becomes a short circuit at the right frequency. As shown in Fig. 2.4(c), two transmission zeros have been obtained in this case. The same problem of low out-of-band rejection can also be observed in this case especially at the lower band. At the higher band, the transmission zero is positioned away from the passband which results in a high out-of-band rejection, but the roll-off factor is low, which clearly highlights the trade-off between the out-of-band rejection and the roll-off factor in this case.



Figure 2.4: The bandpass filter presented in [3]. (a) The topology. (b) the 3D structure. (c) The measured and simulated results.

Summary

It can be concluded that the addition of transmission zeros improves the selectivity but the our-of-band rejection deteriorates. If the transmission zero is positioned away from the passband to improve the out-of-band rejection, the roll-off factor deteriorates. This problem can be addressed by adding more transmission zeros, for example by inductive coupling [31] or capacitive coupling [29]; however, the size in this case increases due to the additional components. In chapter 5 of this document, the problem of increasing the roll off and the

Technology	LTCC [23] [32]	LCP [33]
Dielectric Constant	$5.7 \sim 9.1$	$2.9 \sim 3.2$
Loss Tangent	0.002	0.0025
Tensile Strength [psi]	250	31000 [33]
Thermal Conductivity [W/mK]	2 - 4.4	0.2 [33]
CTE $[ppm/^{\circ}C]$	Fixed (7) [16]	$\begin{array}{c} \text{Controllable} (0 \ - \ 40) \\ [20] \end{array}$
Moisture Absorption [%]	0	$0.02 \sim 0.04$ [33]
Metallization Method	Screen Printing	Electrodiposition
Processing Temperature	$(850 - 950)^{\circ} C$	280° C
Panel Size	$4" \times 4"$	$18" \times 24"$
Min. layer thickness	$50 \ \mu \mathrm{m}$	$25 \ \mu \mathrm{m}$
Flexible?	Not	Yes, for small thickness
Recyclable?	Not	Recyclable

 Table 2.1: Comparison between LCP and LTCC

out-of-band rejection without increasing the size is addressed.

2.1.2 Miniaturized Bandpass Filters in LCP

LCP is a relatively new technology as compared to LTCC. It is an organic-based material with many attractive RF and packaging characteristics as shown in table 2.1. LCP has a permittivity of 2.9, which is much lower than the permittivity of LTCC. It has a value of τ_{ϵ_r} of about -42 ppm/°C [34] making it also stable across all practical temperature ranges. One of the main advantages of LCP is the ability to engineer its coefficient of thermal expansion between 0 - 40 ppm/°C [35] which is an attractive characteristic for SoPs, where multiple materials are integrated on the same system. As for the packaging side, LCP has a moisture absorption of 0.02% to 0.04% [36], which is very low making LCP's packages quasi-hermetic. It also have very low oxygen and water vapour transmission rates [36]. One of the attractive



Figure 2.5: Illustration of the fabrication processes of the LCP technology.

features of LCP is that it can be produced in ultra-thin layers, which can be used to fabricate ultra-thin 3D structures.

The main steps in the fabrication of multi-layer LCP (M-LCP) are (Fig. 2.5): (1) the preparation and pre-heating of the LCP tapes, (2) the realization of the conductive patterns on top and bottom of the individual LCP tapes using wet etching, (3) the drilling of the cavities and integration of the ICs, if any, (4) the lamination and firing of the different LCP layers, and (5) the drilling and metallization of the via holes.

Review of Bandpass filters in LCP

Because LCP is a relatively new technology, its fabrication process has many challenges particularly for larger number of layers. Therefore, very few filter designs have been reported in the literature in M-LCP. Among these designs [37] and [38] are optimized for miniaturization. In [37] an ultra-wideband filter is presented in five layers of M-LCP with an overall thickness of 0.6 mm. In [38] a miniaturized bandpass filter for the VHF band is designed in five M-LCP layers. To improve the roll-off, this design has been based on a five order topology; therefore, its size is large $(56.7 \times 13.75 \times 0.4)$ mm³. In [4], a bandpass filter is presented in a new and experimental organic technology called RXP. To improve the roll-off, transmission zeros have been used but as expected, the out-of-band rejection deteriorates. To improve the out-of-band rejection, a shunt inductor at the input has been used to add more transmission zeros. Unfortunately, this affected the size of the filter considerably because the size of the added inductor is larger than the size of the filter itself as shown in Fig. 2.6 (a).

It is clear that all the previously reviewed designs have not provided a simultaneous solution for the roll-off and out-of-band rejection problems in a compact size. Also, these designs have thicknesses $\geq 400 \ \mu m$ micrometers; therefore are not suitable for flexible electronic applications. In chapter 5 of this work, the design of LCP bandpass filter in an ultra-thin stack-up is presented.



Figure 2.6: The bandpass filter presented in [4]. (a) A photograph of the fabricated filter. (b) The results of the filter.

2.2 Tunable Bandpass Filters

The use of a single tunable filter to replace a switchable bank of filters can result in a systemlevel miniaturization. Tunable filters can be realized by many techniques. Variable capacitors in the form of varactor diodes are typically used with interdigital filters [39–42]. The varactor diodes are used to load the transmission lines, which are the filter's resonators. By changing the value of the capacitance of the varactor diode, the electrical length and; therefore, the resonant frequency of the resonators can be controlled. One of the main drawbacks of this type of filters is the high losses of the solid-state varactor diodes at high microwave frequencies (above 10 GHz). Also, the diodes are usually surface mounted components [43]; therefore, they violates the true SoP concept which is based on embedded thin film components.

Tunability can also be realized by using ferroelectric materials. The most commonly used ferroelectric material is barium strotium titanate (BST) [7,44–48]. The permittivity of a ferroelectric material changes when a direct electric field is applied across it; therefore, if a capacitor is realized with a ferroelectric as its dielectric, its capacitance can be controlled by an externally applied electric field. This capacitor can then be used to load the transmission lines of tunable filters as discussed in the previous paragraph. BST-based bandpass filters are featured by their fast switching times, low power consumption and low insertion losses. For the realization of the BST material, however, photolithographic methods are used, which makes the integration of these designs into an SoP platform difficult.

Mechanically tunable filters were the first generation of tunable filters which are clearly not suitable for modern miniaturized wireless applications. However, with the introduction of micro-electromechanical (MEMS) fabrication techniques, the same ideas behind the classical mechanical tunability can be re-used in a miniaturized fashion. The major characteristics of MEMS-based tunable filters are their extremely low losses. However, their switching times are large, their fabrication and packaging costs are high, and their long term reliabilities are an issue.

2.2.1 Tunable bandpass filters in LTCC

Generally, microwave components in LTCC should be comprised of completely embedded conductors and via-holes for a true SoP realization. If surface mounted components are used instead, this implies that true advantages of functional packaging has not been availed. As mentioned in the previous section, most of the reported tunable filters employed surface mounted components, particularly for their tunability. The same trend is observed in the LTCC medium as well; that is, surface mounted components are used for tunability. In [5], varactor diodes have been mounted on top of an LTCC filter as shown in Fig. 2.7. The design is compact and the achieved tunability is high. However, the use of surface mounted components limit the integrability of this module.



Figure 2.7: The bandpass filter presented in [5]. (a) The topology. (b) a photograph of the fabricated filter after mounting the components. (c) The measured and simulated results.

In [6], a piezoelectric cantilever has been used as a variable capacitor as shown in Fig. 2.8. A commercially available PZT actuator has been mounted on top of an already fabricated LTCC circuit. This design shows a good tunability and relatively low voltage requirement. However, the PZT actuator has been manually mounted on the LTCC clearly restricting the integration of this design within a true SoP.

In [7] a varactor diode embedded in LTCC is presented as shown in Fig. 2.9 (a). The varactor diode is realized by filling a via-hole with a BST material modified for screen printing and low cintering temperature, which allows it to be co-fired with the LTCC. The capacitance of the fabricated varactor changes from 1.47 to 0.97 pF for a bias voltage change of 0 to 200 V as shown in Fig. 2.9(b). This varactor, however, has never been employed in a bandpass filter to demonstrate tunability.

In [8] the piezoelectric material lead-zirconate-titanate (PZT) has been co-fired directly on LTCC as shown in Fig. 2.10. Screen printing has been used to first deposit a layer of gold on the LTCC then a layer of PZT. It has been found, however, that the characteristics of



Figure 2.8: The tunable bandpass filter presented in [6]. (a) A cross section of the fabricated LTCC. (b) The tunability of the fabricated varactor diode.



Figure 2.9: The embedded BST varactor diode presented in [7]. (a) A cross section of the fabricated LTCC. (b) The tunability of the fabricated varactor diode.



Figure 2.10: A cross section of the LTCC which has been fired on the PZT in [8]. The gold electrode is also illustrated.

the PZT deteriorate after co-firing with LTCC. Again, no tunable bandpass filter has been demonstrated here.

Summary

From the review above it can be clearly concluded that even though there are many attempts to realize tunable filters on LTCC, there has been no realization of a fully embedded design for true SoPs. Most of the designs presented have relied on surface mounted components, which require additional cost for mounting the components and reduce the reliability. The attempts to screen print PZT or BST directly on LTCC have produced promising results but no tunable component has been reported.

2.2.2 Magnetically tunable bandpass filters

Tunability of bandpass filters can be obtained by using magnetic materials. The permeability of magnetic materials change when a static magnetic field is applied across it. Several techniques to realize tunable bandpass filters have been reported in the literature. Following is a brief review.
Tunable filters based on the ferromagnetic resonance (FMR) frequency

The ferromagnetic resonance of the ferrite material is the frequency at which the magnetic dipoles resonate; therefore, the material absorbs the RF signal and acts as a resonator. Several designs utilizing the FMR frequency have been reported [9,49–52]. In [9], a magnetically tunable bandpass filter is presented in yittrium iron garnet (YIG) which is a ferrite material suitable for microwave applications (Fig. 2.11). To reduce the losses, the filter is implemented using a high temperature superconductor (HTS). An external electromagnet has been used to provide the bias field as shown in Fig. 2.11 (a). The design achieves good tunability with low insertion loss. However, since HTS is used, the whole design needs to be measured in a cryogenic environment. Also, the deign requires a high magnetic field of 1000 Oe.



Figure 2.11: The magnetically tunable filter presented in [9]. (a)The structure of the filter and the electromagnet. (b) The measured results of the filter.

In [49–51] YIG films realized by epitaxial growth on a gadolinium gallium garnet (GGG) substrate are used as shown in Fig. 2.12. A meander microstrip line is used to enhance the tunable bandwidth and peak absorption. As can be seen in Fig. 2.12 (a), the filter is first fabricated on a GaAs substrate, then the YIG-GGG is integrated. The bias filed in this case has also been obtained by an external electromagnet. The design achieves good tunability; however, the magnetic field required is large (around 6000 Oe).



Figure 2.12: The magnetically tunable filter presented in [9]. (a)The structure of the filter. (b) The measured results of the filter.

Tunable filters based on partially magnetized ferrites

To utilize the FMR frequency of the ferrite material for tunability, the material needs to be biased in saturation, which requires high magnetic fields and therefore high dc currents. Another approach is to operate the ferrite material in the partially magnetized state, where tunability can be achieved by much less magnetic fields. Few designs using this approach have been reported [10, 11, 53]. In [10, 53], bandpass filters realized by electroplating copper on YIG substrates are presented as shown in Fig. 2.13. The bias field has been provided by an external electromagnet and was oriented transverse to the direction of the RF propagation. Therefore, a downwards frequency shift has been observed. Due to the operation in the partial magnetization state, the presented filter demonstrates good tunability using a much lower magnetic field as compared to the FMR case. Good insertion loss and high power handling capability has also been demonstrated. However, the use of the external electromagnet makes it difficult to integrate such design within a miniaturized SoP.

In [11] the same technique as mentioned above has been used to realize a tunable filter on a YIG substrate. In this case however, the magnetic field, which has been produced by an external electromagnet, is applied perpendicular to the direction of propagation. Therefore,



Figure 2.13: The magnetically tunable filter presented in [10]. (a) The structure of the filter. (b) The measured results of the filter.

an upwards frequency shift has been observed. Like the previous case, this design demonstrates good tunability with a relatively low magnetic field due to the operation in the partial magnetization state.



Figure 2.14: The magnetically tunable filter presented in [11]. (a) A photograph of the fabricated filter. (b) The measured results of the filter.

Tunable filters based on electric field

All the designs presented so far use electromagnets to produce the magnetic field necessary for tunability. These electromagnets require a continuous supply of dc currents to produce the magnetic fields. In [12] a piezoelectric material coupled to a magnetic material has been used as shown in Fig. 2.15 (a). When the piezoelectric material is deformed as a result of an applied electric field, it generates stress which is transformed to the ferrite material. the permeability of ferrite materials change when they are submitted to stress. As a result, tunability have been achieved without passing any current as shown in Fig. 2.15 (b). However, the achieved tunability is relatively low and the voltage required is high (3000 V/cm).



Figure 2.15: The magnetically tunable filter presented in [12]. (a)The structure of the filter. (b) The measured results of the filter.

In [13] a similar, yet simpler, method is proposed as shown in Fig. 2.16 (a). In this case a ferrite film coupled to a piezoelectric transducer is used to perturb the fringing fields of the filter and cause tunability. Much lower voltages are required in this case to achieve tunability. However, the design is complex and requires high cost packaging techniques. Moreover, such designs are not reliable due to the existence of vibrating parts. Therefore; it is difficult to integrate such designs with a 3D SoP platform at a reasonable cost.

Magnetically tunable bandpass filters on LTCC

Recently, an LTCC tape with magnetic properties has become commercially available [32]. This ferrite LTCC tape is fabricated by dispersing magnetic powder in an organic matrix. When the green tape is fired, the resultant ceramic material has ferrite characteristics. If



Figure 2.16: The magnetically tunable filter presented in [13]. (a)The structure of the filter. (b) The measured results of the filter.

the 3D capabilities of the ferrite LTCC is utilized, embedded windings can be realized inside the LTCC substrate; therefore, the large coils or electromagnets, which are typically used, can be avoided and the size can be reduced considerably. The internally generated magnetic field from the windings can be used to tune microwave structures implemented in the ferrite substrate. Even though, embedded windings in ferrite LTCC have been demonstrated in [54], no tunable bandpass filter has been reported on this technology. This approach is addressed in chapter 6.

Summary

From the review above, it can be concluded that magnetically tunable bandpass filters that are based on the FMR frequency require high magnetic fields, thus, high currents, which can not be produced and handled in miniaturized wireless devices. The current requirement has been reduced considerably by operating the ferrite material in the partially magnetized state. Even in this case, all the reported designs have used electromagnets or coils to produce the magnetic field required for biasing the ferrite substrate. The need for the external electromagnet have been avoided by the use of piezoelectric transducers coupled to the ferrite material. However, in this case the resultant designs either requires high voltages, or can not be integrated. Ferrite LTCC tapes are available commercially, however, no bandpass filter has been reported on them before. The magnetic tunability in the SoP platform is the subject of chapter 6.

Chapter 3

Bandpass Filter Design Theory

In this chapter, the basic theory behind the design of bandpass filters is introduced. Due to the extremely large literature of filter theory, however, only lumped-based and edgecoupled filters are discussed. Lumped-based filter theory is used in chapter 5 in the design of miniaturized filters while edge-coupled filter theory is utilized in chapter 6, where the design of magnetically tunable filters is presented. Derivations as well as examples are presented to better explain the concepts.

3.1 Introduction

A bandpass filter is a two port network with frequency selectivity properties. Signals at a certain frequency band, known as the *passband*, passes with minimal attenuation, whereas signals at other bands, known as *stopbands*, are highly attenuated. Filers are characterized by their voltage transfer function which is the voltage across the load divided by the voltage at the input of the network. At microwave frequencies, it is very hard to measure voltages and currents; therefore, the scattering parameters (S-parameters) are used instead. S-parameters

deal with incident and reflected waves and can be defied as follows for a 2-port network

$$V_1^- = S_{11}V_1^+ + S_{12}V_2^+ \tag{3.1}$$

$$V_2^- = S_{21}V_1^+ + S_{22}V_2^+ \tag{3.2}$$

or in matrix form

$$\begin{bmatrix} V_1^-\\ V_2^- \end{bmatrix} = \begin{bmatrix} S_{11} & S_{21}\\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} V_1^+\\ V_2^+ \end{bmatrix}$$
(3.3)

where V^- and V^+ are the incident and reflected voltage waves, respectively. For bandpass filters, S_{11} , which is often expressed in decibels (dBs) gives an indication of how much of the energy at the input of the filter is absorbed and how much is being reflected. This is referred to as the return loss L_R and is expressed in dBs as follows

$$L_R = -20 \log(|S_{11}|) \tag{3.4}$$

 S_{21} , which is the same as S_{12} for reciprocal networks, is a measure of how much of the energy already absorbed in the network is transferred to the output end and how much is dissipated within the filter itself. This is referred to as the insertion loss (IL), which is defined as

$$IL = -20 \log(|S_{21}|) \tag{3.5}$$

The insertion loss is only significant within the passband. A common practice is to report the minimum insertion loss, which occurs at the peak of the passband. The insertion loss is a very important measure of the performance of a bandpass filter, especially if the filter is for an RF receiver because it affects the noise figure of the overall receiver. The insertion loss is sometimes defined as the power loss ratio in dBs. The power loss ratio (P_{LR}) is defined as

$$P_{LR} = \frac{\text{The power available from the generator}}{\text{The power delivered to the load}} = \frac{1}{1 - |\Gamma(\omega)|^2},$$
(3.6)

where $\Gamma(\omega)$ is the reflection coefficient seen looking towards the input of the filter.

3.2 Bandpass Filter Design Specifications

The first step in a bandpass filter designed is to determine its specifications, which are typically application specific. Some of the most important characteristics and performance merits for a bandpass filter are briefly discussed next.

3.2.1 Frequency of operation (f_c)

The frequency of operation is the most important factor in the design of the filter. For ultra-low frequencies (<20 MHz), active filters are very common because at such frequencies the required inductors are extremely large and expensive. As the frequency is increased above few hundreds of MHz, the gains of operational amplifiers deteriorate and lumped based filters become very common. As the frequency is increased above few GHz, lumped components can not be used because their physical sizes become close to the wavelength; therefore, distributed lines are used in the form of microstrip, stripline or co-planar striplines. Above millimeter-wave frequencies, the dimensions of transmission lines become too small to realize therefore silicon-based and substrate integrated waveguide (SIW) technologies are used to realize bandpass filers at these frequencies [55, 56]. Tus, based on the frequency of operation, suitable choice of technology is very important.

3.2.2 Insertion loss

The insertion loss, as is explained in the previous section, is an indication of the amount of energy dissipated between the input and output of the filter circuit. It is,; therefore, an important performance merit especially for the case of receivers, where a filter is typically positioned after the antenna to provide selectivity. The loss of this filter affects the sensitivity of the whole receiver therefore it is characterized by its minimum insertion loss.



Figure 3.1: An illustration of the characteristics and performance merits of the bandpass filter.

3.2.3 Operational Bandwidth

Typically the 3dB bandwidth is used which is defined as the band between the two frequencies at which the value of $|S_{21}|$ is 3dBs below its minimum value (Fig. 3.1). This bandwidth is often normalized to the center frequency and given in the form of a percentage ratio known as the fractional bandwidth (FBW). Filters are classified into narrow-band, wide-band, and ultra-wide band based on the achieved fractional bandwidth.

3.2.4 Out-of-band rejection

The out-of-band rejection is a measure of the rejection for unwanted signals. It is usually specified in dBs at a defined offsets. For instance an out-of-band rejection of 50 dBs at 200 MHz offset means that a signal 200 MHz away from the edge of the passband is attenuated by 50 dBs or more. i.e its power is attenuated 100,000 times.

3.2.5 Roll-off performance

The roll-off of the filter defines how sharply the filter switches from the passband to the stopband. An ideal filter has an instantaneous transition form passband to stopband. Obviously, this is not physically realizable with finite energy components. A typical filter usually have what is know as transition bands which are the bands that the filter introduces attenuation less than the specified value as shown in Fig. 3.1. The transition band starts from the edge of the passband to the frequency at which the specified attenuation is reached. The roll-off performance is measured by the roll-off factor which is a measure of the slope of $|S_{21}|$ and is measured in dBs/octave or dBs/decade as shown in Fig. 3.1. To evaluate the roll-off factor, the width of the transition band is calculated in Hz and base-two logarithm is taken to get the width in Octavs, while the base-ten logarithm is taken to produce the width in decades. The roll-off factor is then evaluated as the ratio between the attenuation and the width of the stopband in either octaves or decades.

3.2.6 Phase performance

For some applications the signal passing through the filter should not experience variable phase delays. This requirement is usually expressed in terms of the linearity of the phase within the passband. In mathematical terms, the phase that the signal undertakes is given by

$$\phi = Arg\{S_{21}\},\tag{3.7}$$

where ϕ represents the phase difference between the input and the output of the filter and is measured in Radians. The first derivative of the phase with respect to the frequency gives the group delay in the filter as follows:

$$t_g = \frac{d\phi}{d\omega},\tag{3.8}$$

where t_g is the group delay in seconds and ω is the angular frequency in Rad/Sec. t_g is the time signals take to propagate through the filter network. Depending on the application, there could be restrictions on this number. For ultra-wide band applications, for instance,

this number is required to be constant with respect to frequency for the whole passband so that signals do not experience distortion.

3.2.7 Size

With modern applications, miniaturization is a requirement for many applications. Traditionally, only the planar size of the filter is specified. However, with the 3D, multilayer technologies the complete volume of the circuit is important. The techniques and technologies utilize to reduce the size of the filter are discussed in details in chapter 5 of this document.

3.2.8 Power handling capability

For high power applications, a situation mainly occurs at the transmitter section, the bandpass filter is required to handle high powers. The power handling is usually specified by the maximum current and voltage the filter network can handle without a change in performance.

3.3 Design of lumped-based bandpass filters

The design of filters based on lumped inductors and capacitors form the base of almost all filter theory. After a lumped-based filter circuit is reached, transformations are used to realize the filter using transmission lines or waveguide. With the advances of the 3D multi-layer technologies, filters can be realized in lumped components, which are highly miniaturized. Two main methods are used in the design of lumped-based filters: the image parameter and the insertion loss method. The image parameter method is based on trial and error and can not guarantee the realization of a specific filter function. Therefore, only the insertion loss method is discussed here, which goes through the following steps

1. An insertion loss *magnitude* function is formulated based on the required filter specifications. This function must be realizable.

- 2. From the magnitude function $(|S_{21}|)$, the complete S_{11} function is calculated from which the input impedance (Z_{in}) seen looking at the filter's input is calculated.
- 3. The normalized low-pass prototype ladder network is synthesized from the input impedance.
- 4. The input and output impedances are scaled as required by the application.
- 5. The center frequency of the normalized lowpass filter is scaled to the required frequency.
- 6. The low pass prototype is converted to a bandpass prototype.
- 7. Additional transformations might be done to improve the topology

Because of the maturity of the subject, extensive literature is available. Therefore these steps are discussed only briefly in the following sections. Also the design procedure of the edge-coupled structure, which is used in the design presented in chapter 5, is explained in details.

3.3.1 Insertion loss magnitude function formulation

In most of the cases the specifications of the filter are formulated in the power loss ratio P_{LR} rather than the insertion loss. The conditions that make P_{LR} realizable are driven from the fact that in order to realize this function using passive lumped components, the network must be causal. Therefore, $|\Gamma(\omega)|^2$ is an even function in ω and can be expressed as follows [22]

$$|\Gamma(\omega)|^2 = \frac{M(\omega^2)}{M(\omega^2) + N(\omega^2)}$$
(3.9)

where M and N are real polynomials. Substituting this in (3.6) gives

$$P_{LR} = 1 + \frac{M(\omega^2)}{N(\omega^2)}$$
(3.10)

which is the condition for P_{LR} to be realizable.

Several physical realizations of (3.10) have been introduced. Here we will discuss only two of them: the maximally flat and the equal ripple or Chebyshev.

Maximally flat

The maximally flat P_{LR} of the lowpass prototype is defined by



Figure 3.2: The gain function $(|S_{21}|)$ of a maximally flat realization for various filter orders.

$$P_{LR} = 1 + \left(\frac{\omega}{\omega_c}\right)^{2N} \tag{3.11}$$

where N is the order of the filter, which is the number of inductors and capacitors. ω_c is the cutoff frequency at which the loss is 3 dB. A plot of $|S_{21}|$ for this case is shown in Fig. 3.2 for orders of one to four. It can be seen that as the order is increased, the roll-off increases and the performance in the passband becomes more flat, hence the name maximally flat.

Equal ripple

The other important realization of (3.10) is the equal ripple or Chebyshev function, which is defined as follows

$$P_{LR} = 1 + k^2 T_N^2 \left(\frac{\omega}{\omega_c}\right),\tag{3.12}$$



Figure 3.3: The gain function $(|S_{21}|)$ of an equal ripples (Chebyshev) realization for various filter orders.

where $1 + k^2$ is the ripple level in linear units, T_N is a Chebyshev function of order N. A plot of the gain function for this realization is shown in Fig. 3.3 for orders of one to four. The equal ripple realization has a much better roll-off as compared to the maximally flat one. However, the maximally flat has a stable gain throughout the passband unlike the equal ripple which oscillates between the zero dBs and the three dBs levels. Moreover, an ideal maximally flat filter has a linear phase response unlike equal ripple filters.

3.3.2 Calculation of the transfer function from the magnitude function

In the previous section two realizations of the filter's power loss ratio have been calculated. P_{LR} is a scalar quantity, which is the same as the transfer function of the filter and equals $|S_{21}|^2$ for the case of input and output matching, which is assumed here. In order to synthesize the filter the input impedance Z_{in} seen looking towards the filter needs to be determined. Z_{in} can be calculated from S_{11} . The problem is then calculating S_{11} from $|S_{21}|(\omega)$. These steps are driven next [57]. $|S_{11}|^2$ can be written as follows for the case of a lossless network

$$|S_{11}|^2(\omega) = 1 - |S_{21}|^2(\omega)$$
(3.13)

If the function $h(s^2)$ is defined as

$$h(s^2) = S_{11}(s) \ S_{11}(-s) \tag{3.14}$$

where s is the Laplace variable, a complex frequency variable of the form $s=\sigma+j~\omega.$ Then

$$|S_{11}|^2(\omega) = h(-\omega^2) \tag{3.15}$$

Now, for the case of the maximally flat $|S_{21}|$ is defined as the reciprocal of P_{LR}

$$|S_{21}|^2(\omega) = \frac{1}{1+\omega^{2N}} = \frac{1}{1+(-1)^N(-\omega^2)^N}$$
(3.16)

From (3.13) $|S_{11}|^2(\omega)$ is

$$|S_{11}|^2(\omega) = \frac{\omega^{2N}}{1+\omega^{2N}} = \frac{(-1)^N (-\omega^2)^N}{1+(-1)^N (-\omega^2)^N}$$
(3.17)

The function $h(s^2)$ can be used to represent $|S_{11}|^2$ in the s-domain as follows

$$h(s^{2}) = \frac{(-1)^{N}(s^{2})^{N}}{1 + (-1)^{N}(s^{2})^{N}} = \frac{(-s)^{N}(s)^{N}}{1 + (-1)^{N}(s^{2})^{N}} = S_{11}(s) \ S_{11}(-s).$$
(3.18)

The poles of this function can be found from its denominator by solving the following equation

$$1 + (-1)^N (s^2)^N = 0 (3.19)$$

or

$$(-1)^{N}(s^{2})^{N} = -1 = e^{j(2k-1)\pi}, \qquad k = 0, 1, \dots, 2N - 1 \qquad (3.20)$$

which gives

$$s_k = e^{j[(2k+N-1)/2N]\pi}, \qquad k = 0, 1, ..., 2N$$
 (3.21)

Now, let us express s_k as $s_k = \sigma_k + j\omega_k$ with the following values

$$\sigma_k = \cos\left(\frac{2k+N-1}{2N}\pi\right) = \sin\left(\frac{2k-1}{N}\frac{\pi}{2}\right)$$
(3.22)
(3.23)

$$\omega_k = \sin\left(\frac{2k+N-1}{2N}\pi\right) = \cos\left(\frac{2k-1}{N}\frac{\pi}{2}\right).$$

In the s-plane, these poles fall in a circle with unit radius and center at the origin. Half of them are on the left-half plane and the other half on the right-half plane. For S_{11} to represent a stable and causal system, all the poles should fall in the left-half plane. Therefore, the poles in the right-half are assigned to $S_{21}(-s)$ and the ones in the left-half are assigned to $S_{21}(s)$. As an example let us consider a maximally flat lowpass filter with N = 5 then from (3.21)

$$s_k = e^{j[(k+2)/5]\pi},\tag{3.24}$$

which gives the following poles in the left-half plane: $e^{3/5\pi}$, $e^{4/5\pi}$, e^{π} , $e^{6/5\pi}$, and $e^{7/5\pi}$. $S_{11}(s)$ can be constructed as

$$S_{11}(s) = \frac{s^5}{(s - e^{3/5\pi})(s - e^{4/5\pi})(s - e^{\pi})(s - e^{6/5\pi})(s - e^{7/5\pi})},$$
(3.25)

which can be simplified as

$$S_{11}(s) = \frac{s^5}{(s+1)(s^4 + \sqrt{5}s^3 + 3s^2 + \sqrt{5}s + 1)}$$
(3.26)

The input impedance of the filter can be calculated as follows

$$Z_{in}(s) = \frac{1 + S_{11}(s)}{1 - S_{11}(s)},\tag{3.27}$$

which is given as follows for the maximally flat filter with N = 5

$$Z_{in}(s) = \frac{1 + (1 + \sqrt{5})s + (3 + \sqrt{5})s^2 + (3 + \sqrt{5})s^3 + (1 + \sqrt{5})s^4 + 2s^5}{1 + (1 + \sqrt{5})s + (3 + \sqrt{5})s^2 + (3 + \sqrt{5})s^3 + (1 + \sqrt{5})s^4}$$
(3.28)

or approximately

$$Z_{in} = \frac{1 + 3.24s + 5.24s^2 + 5.24s^3 + 3.24s^4 + 2s^5}{1 + 3.24s + 5.24s^2 + 5.24s^3 + 3.24s^4}$$
(3.29)

3.3.3 Synthesis of the low-pass ladder network from the input impedance



Figure 3.4: Lowpass L-C ladder network

The ladder circuit of series inductors and shunt capacitors shown in Fig. 3.4 can be constructed from the input impedance by expressing the input impedance as follows

$$Z_{in} = sL_1 + \frac{1}{sC_2 + \frac{1}{sL_3 + \frac{1}{sC_4 + \frac{1}{sL_5 + \dots}}}},$$
(3.30)

which can be achieved through consecutive division and inversion. As an example let us consider the maximally flat impedance produced in the previous section. Since the degree of the numerator is higher than the degree of the denominator, then the first element can be chosen as an inductor with direct division as follows

$$Z_{in} = \frac{1 + 3.24s + 5.24s^2 + 5.24s^3 + 3.24s^4 + 2s^5}{1 + 3.24s + 5.24s^2 + 5.24s^3 + 3.24s^4} = 0.617s + \frac{1 + 2.6s + 3.2s^2 + 2s^3}{1 + 3.24s + 5.24s^2 + 5.24s^3 + 3.24s^4}$$
(3.31)

from which it follows that $L_1 = 0.617H$. For the second part, the degree of the numerator is less than the degree of the denominator; therefore this term is inverted and the result of the long division gives a shunt capacitor

$$\frac{1+3.24s+5.24s^2+5.24s^3+3.24s^4}{1+2.6s+3.2s^2+2s^3} = 1.622s + \frac{1+1.6s+s^2}{1+2.6s+3.2s^2+2s^3},$$
(3.32)

which gives $C_2 = 1.622F$. This process is repeated until the remainder of the division is unity which is the terminating resistance. The remaining values are $L_3 = 2H$, $C_4 = 1.55F$, and $L_5 = 0.69H$.

Tables for the lowpass prototypes exist in the literature [58] for various realizations and orders. The components are typically referred to as g_i where i = 0 to N + 1. g_0 and g_{N+1} refer to the generator and load and are typically 1 Ohm. A typical lowpass prototype is illustrated in Fig. 3.5.



Figure 3.5: A typical lowpass filter prototype. The dual of this circuit can also be used giving equal performance.

3.3.4 Impedance Scaling

All the lowpass prototypes defined earlier have unity terminating impedances. To transform the generator and load impedances to any practical value, the values of all elements in the circuit should be scaled as follows [59]

where Z_0 is the new system impedance. It is worth mentioning here that the previous equation is valid only if the load and generator impedances are unity. As an example, the new values of the components in the maximally flat example from the previous section that give a termination of 50 Ω are $L_1 = 30.85H$, $C_2 = 0.03244F$, $L_3 = 100H$, $C_4 = 0.031F$, and $L_5 = 34.5H$

3.3.5 Frequency Transformation

The lowpass prototypes discussed so far are all normalized to a cutoff frequency of 1 rad/sec. To transform this to a practical cutoff frequency the following equation can be used

$$\begin{array}{rcl}
L & \rightarrow & L/\omega_c \\
C & \rightarrow & C/\omega_c
\end{array} \tag{3.34}$$

The values of the components in the previous example that give a cutoff frequency of 1.5 GHz are $L_1 = 3.2nH$, $C_2 = 3.44pF$, $L_3 = 10.6nH$, $C_4 = 3.29pF$, and $L_5 = 3.66nH$. The

complete filter circuit is illustrated in Fig. 3.6.



Figure 3.6: Lowpass L-C ladder network

3.3.6 Lowpass to bandpass transformation

The prototype lowpass filter designed in the previous sections can be transformed to a bandpass filter by substituting the series inductors with an inductor and a capacitor in series, and the shunt capacitor with an inductor and a capacitor in parallel. The values of the new components are related to the values of the lowpass components by the following equations:

$$L_{s} = \left(\frac{1}{FBW\,\omega_{0}}\right)L\tag{3.35}$$

$$C_{s} = \frac{FBW}{\omega_{0}L},$$

and

$$C_{p} = \left(\frac{1}{FBW\omega_{0}}\right)C$$

$$L_{p} = \frac{FBW}{\omega_{0}C},$$
(3.36)

where FBW is the fractional bandwidth defined as

$$FBW = \frac{f_2 - f_1}{f_0},\tag{3.37}$$

where f_2 and f_1 represents the edges of the passband, f_0 , and ω_0 represent the center frequency. It is worth mentioning here that the previous equations take into account the frequency transformation; therefore, the values of C and L represents a lowpass prototype with unity cut off frequency. As an example the lowpass filter in the previous section is converted into a bandpass filter with a center frequency of 1.5 GHz and a FBW of 0.1 (10%) as shown in Fig. 3.7. The response of this circuit is shown in Fig. 3.8.



Figure 3.7: Schematic of the bandpass filter example.



Figure 3.8: Magnitude response of the bandpass filter example.

3.3.7 Topology transformation



Figure 3.9: Impedance and admittance inverters.

Even though the topology of the bandpass filter of Fig. 3.7 provides a very good approximation to the exact filter specifications, it is not suitable for realization using embedded components mainly because of the series inductors. This topology can be converted to allshunt inductors topology using impedance inverters as shown in [58]. Impedance (K) and Admittance (J) inverters can be defined as shown in Fig 3.9. Admittance inverters can be used to transform the topology of Fig. 3.7 to a one with shunt inductors only as shown in Fig 3.10. The values for the inverters can be calculated as follows [58]

$$J_{01} = \sqrt{\frac{Y_0 \omega_0 C_{pi} F B W}{g_0 g_1}}$$

$$J_{i,i+1} = Y_0 F B W \omega_0 C_{pi} \sqrt{\frac{1}{g_i g_{i+1}}}$$

$$J_{n,n+1} = \sqrt{\frac{Y_0 \omega_0 C_{pi} F B W}{g_n g_{n+1}}},$$
(3.38)

and

$$L_{pi} = \frac{1}{\omega_0^2 C_{pi}}$$
(3.39)

where g_i are the values of the normalized low pass filter prototype. Either L_{pi} or C_{pi} can be chosen arbitrarily and the other calculated.

A realization of the admittance inverter in terms of lumped components is shown in Fig.



Figure 3.10: Schematic of the bandpass filter with modified topology.

3.11 below. The negative capacitances can be absorbed in the other capacitors connected in shunt. As an example, the bandpass prototype of the previous sections is transformed into all-shunt inductors topology. The lowpass prototype values are $g_0 = 1$, $g_1 = 0.617$, $g_2 = 1.622$, $g_3 = 2$, $g_4 = 1.55$, $g_5 = 0.69$, and $g_6 = 1$ as calculated in section 3.3.3. Let us assume a constant value for all L_{pi} of 1 nH then C_{pi} are all 11.2579 pF. From (3.38) $J_{01} = 0.01854$, $J_{12} = 0.0106$, $J_{23} = 0.00589$, $J_{34} = 0.00602$, $J_{45} = 0.01025$, and $J_{56} = 0.0175$. From which the values of the inverter's capacitance C are $C_{01} = 5.255pF$, $C_{12} = 1.1253pF$, $C_{23} = 0.62505pF$, $C_{34} = 0.6394pF$, $C_{45} = 1.0886pF$, and $C_{56} = 3.8703pF$. When combining the shunt capacitors, the final circuit of Fig. 3.12 is reached, where $C_1 = 9.3958pF$, $C_2 =$ 0.5075pF, $C_3 = 9.9934pF$, $C_4 = 9.5299pF$, and $C_5 = 9.2747pF$. The response of this filter with ideal lumped components is shown in Fig. 3.13.



Figure 3.11: A realization of the admittance inverter in terms of capacitors or inductors only.



Figure 3.12: 5th order filter example with the modified topology.



Figure 3.13: Response of the 5th order filter example with modified topology.

3.4 Design of edge coupled bandpass filter in microstrip technology

The edge-coupled filter is a very popular filter structure that can be implemented in either microstrip or strip-line technologies. a typical structure of the edge-coupled filter is shown in Fig. 3.14. It consists of N + 1 coupled sections which are $\lambda_g/4$ each, where λ_g is the guided wavelength for the microstrip mode.

The analysis of this filter can be started from its building block, which is a network of coupled lines as shown in Fig. 3.16 (a). The analysis of these lines was reported by Johns and Bolljahn using the odd-even modes [60]. The coupled lines are treated as a four port network. The even mode is shown in Fig. 3.15(a) and (b) and the odd mode is shown in Fig.



Figure 3.14: Layout of a typical edge-coupled filter of order N. The filter has N + 1 coupled sections, $\lambda_g/4$ long each.

3.15 (c) and (d). The currents i_1 and i_3 are used to excite the even mode and the currents i_2 and i_4 are used to excite the odd mode. The characteristic impedance for the even and odd modes are Z_{0e} and Z_{0o} respectively, while the characteristic impedance of each of the lines is Z_0 .



Figure 3.15: Even and Odd modes used to analyse the edge-coupled transmission lines.

The voltage at port 1 can be calculated from the superposition of all four cases of Fig. 3.15. For the case of Fig. 3.15 (a) the input impedance at port 1 is that same as that of port 2 and is calculated as the input impedance of a transmission line terminated at an open circuit, which can be given as:

$$Z_{in} = -jZ_{0e}\cot(\beta l) \tag{3.40}$$

The voltages along either of the lines are equal and can be calculated as a function of z as follows

$$v_{a1}(z) = v_{b1}(z) = V_e^+ \left[e^{-j\beta(z-l)} + e^{j\beta(z-l)} \right]$$
(3.41)

where V_e^+ is the magnitude of the standing voltage wave on the line. The voltage at ports 1 or 2 is then

$$v_{a1}(0) = v_{b1}(0) = 2V_e^+ \cos\beta l = i_1 Z_{in}$$
(3.42)

from (3.40)

$$v_{a1}(z) = v_{b1}(z) = -jZ_{0e} \frac{\cos\beta(l-z)}{\sin\beta l} i_1$$
(3.43)

Likewise, the voltages on ports 3 and 4 can be calculated as

$$v_{a3} = v_{b3} = -jZ_{0e} \frac{\cos\beta z}{\sin\beta l}$$
 (3.44)

The analysis for the odd configuration can be done in a similar way giving

$$v_{a2} = -v_{b2} = -jz_{0o} \frac{\cos\beta(l-z)}{\sin\beta l} i_2$$
(3.45)

$$v_{a4} = -v_{b4} = -jz_{0o} \frac{\cos\beta z}{\sin\beta l} i_4 \tag{3.46}$$

The overall current on each port is the superposition of all four cases of Fig. 3.15

$$I_{1} = i_{1} + i_{2}$$

$$I_{2} = i_{1} - i_{2}$$

$$I_{3} = i_{3} - i_{4}$$

$$I_{4} = i_{3} + i_{4}$$
(3.47)

which can be solved for the mode currents to give

$$i_{1} = \frac{1}{2} (I_{1} + I_{2})$$

$$i_{2} = \frac{1}{2} (I_{1} - I_{2})$$

$$i_{3} = \frac{1}{2} (I_{4} + I_{3})$$

$$i_{4} = \frac{1}{2} (I_{4} - I_{3})$$
(3.48)

The over all voltage at port 1 (V_1) can be written as

$$V_1 = (v_{a1} + v_{a2} + v_{a3} + v_{a4})|_{z=0}$$
(3.49)

using (3.43), (3.45), (3.44), (3.46), and (3.48)

$$V_{1} = -j \left(Z_{0e} + Z_{0o} \right) \frac{\cot\theta}{2} + -j \left(Z_{0e} - Z_{0o} \right) \frac{\cot\theta}{2} + -j \left(Z_{0e} - Z_{0o} \right) \frac{\csc\theta}{2} + -j \left(Z_{0e} + Z_{0o} \right) \frac{\csc\theta}{2}$$
(3.50)

where $\theta = \beta l$ is used. (3.50) gives the first raw of the impedance matrix (Z-matrix). Since the structure is completely symmetrical, all other components can be written as follows



Figure 3.16: The equivalence between coupled transmission lines (a), and a J inverter between two transmission lines (b).

$$Z_{11} = Z_{22} = Z_{33} = Z_{44} = -j \left(Z_{0e} + Z_{0o} \right) \frac{\cot\theta}{2}$$

$$Z_{12} = Z_{21} = Z_{34} = Z_{43} = -j \left(Z_{0e} - Z_{0o} \right) \frac{\cot\theta}{2}$$

$$Z_{13} = Z_{31} = Z_{24} = Z_{42} = -j \left(Z_{0e} - Z_{0o} \right) \frac{\csc\theta}{2}$$

$$Z_{14} = Z_{41} = Z_{23} = Z_{32} = -j \left(Z_{0e} + Z_{0o} \right) \frac{\csc\theta}{2}$$
(3.51)

The impedance parameters for a two port network of coupled lines can be extracted from this four port network by letting ports 2 and 4 terminate in an open circuit, which means that I_2 and I_4 become zero giving the new two port network at ports 1 and 3 with the same values as in (3.51).

To design the bandpass filter an equivalence between the coupled lines sections and the lumped-based topology of Fig. 3.10 is established. The equations presented previously for that topology can then be used for this filter. This treatment was first introduced by Cohn [61] and is explained here briefly.

The analysis starts by establishing an equivalence between the coupled-line network (Fig. 3.16 (a)), and the network of inverters and transmission lines shown in Fig. 3.15 (b). The ABCD matrix of the circuit of (b) can be found as the cascade of a J inverter between two transmission lines and is given by

$$\begin{bmatrix} A \\ C \end{bmatrix} = \begin{bmatrix} \left(JZ_0 + \frac{1}{JZ_0}\right)\cos\theta\sin\theta & j\left(JZ_0^2\sin^2\theta - \frac{\cos^2\theta}{J}\right) \\ j\left(\frac{1}{JZ_0^2}\sin^2\theta - J\cos^2\theta\right) & \left(JZ_0 + \frac{1}{JZ_0}\right)\cos\theta\sin\theta \end{bmatrix}$$
(3.52)

The following two equations are achieved when Z_{21} and the propagation constant of the two networks of Fig. 3.15 (a) and (b) are equated as $\theta \to \pi/2$.

$$Z_e = Z_0 \left[1 + JZ_0 + (JZ_0)^2 \right]$$
(3.53)

$$Z_0 = Z_0 \left[1 - JZ_0 + (JZ_0)^2 \right]$$
(3.54)

Now the complete bandpass filter can be constructed as shown in Fig. 3.17 (c). As seen on the figure the filter is constructed from inverters and sections of transmission lines with length of 2θ or $\lambda/2$. As $\theta \to \pi/2$ this transmission line section is equivalent to a shunt network of inductors and capacitors as shown in Fig. 3.17 (d). An equivalence between these two networks in the vicinity of $\theta = \pi/2$ gives the following two relations

$$L = \frac{2Z_0}{\pi\omega_0} \tag{3.55}$$

$$C = \frac{\pi}{2Z_0\omega_0} \tag{3.56}$$

These equations can be used together with the relations for the complete filter circuit shown in (3.38) and (3.39) to give the design equations for the edge-coupled filter as follows

$$J_{01} = Y_0 \sqrt{\frac{\pi FBW}{2g_0 g_1}}$$

$$J_{i,i+1} = \frac{\pi FBW}{2} \sqrt{\frac{1}{g_n g_{n+1}}}$$

$$J_{n,n+1} = Y_0 \sqrt{\frac{\pi FBW}{2g_n g_{n+1}}}$$
(3.57)

The design process then is strait forward, which starts from (3.57) by calculating the J-inverter values form the lowpass prototype values and the filter's specifications. Equation (3.53) and (3.54) can then be used to calculate the even and odd impedances for each section of the filter. Finally, physical dimensions for the filter sections are calculated either from nomograms as in [62] or using a computer software.



Figure 3.17: A layout of a typical edge-coupled filter of order N. The filter has N + 1 coupled sections, $\lambda_g/4$ long each.

Chapter 4

Lumped components for SoP Technologies

Bandpass filters that are based on distributed components are very large at low microwave frequencies (below 10 GHz). To reduce the size at these frequencies, lumped components can be used. Traditionally, lumped components are realized in the form of discrete components that are mounted on the surface of the printed circuit board (PCB). With the development of the 3D packaging technologies such as LTCC and LCP, lumped components can be implemented in a miniaturized and 3D fashion. Compared to the discrete surface mounted components, 3D components offer smaller sizes and lower costs. Using such components, bandpass filters can be implemented within the package of the system in a SoP structure, which reduces the cost and size (both in a system level as well as a component level). LTCC and LCP, based components have been presented in the literature [63, 64] showing good qualities and small form factors. LCP sheets can be produced in an ultra-small thickness; therefore, multilayer structures that are ultra-thin and flexible can be implemented. In this chapter, the effects of reducing the thickness of the substrate on the performance of lumped components is studied through the implementation of an ultra thin LCP stack-up with a total thickness of 100 μ m as shown in Fig. 4.1(c). For the sake of comparison, similar components in LTCC and standard (*thick*) LCP have also been designed and fabricated according to the stack-ups presented in Fig. 4.1 (a) and (b). The analysis and comparison results presented in this chapter should help in the design of the lumped components as well as in the selection of the technology.



Figure 4.1: Illustration of the three stack-ups used in this work. (a) 4 layers of LTCC with total thickness of $400\mu m$. (b) The equivalent of (a) in LCP. (c) An ultra-thin LCP stack-up with 4 layers and a total thickness of only $100\mu m$

4.1 Inductors

Inductors are important components in all RF systems. They are especially important for filters, matching networks and amplifiers. The main figures of merit of inductors are the effective inductance (L_{eff}), and the quality factor (Q-factor). With reference to the basic inductor model of Fig. 4.2, the effective inductance can be estimated as follows:

$$L_{\text{eff}} = \frac{L}{(1 - \omega^2 L C_P)}$$

$$(4.1)$$

$$L$$

$$R$$

$$M$$

Figure 4.2: A simple inductor model with parasitic and inter-turn capacitance and resistive losses.

where L_{eff} is the effective inductance in Henries, L is the inherent inductance in Henries, and C_P is the shunt parasitic capacitance in Farads, and R represents the resistive losses in Ohms (ignored in (4.1)). As can be deduced from (4.1), at a frequency of $\omega_{SRF} = 1/\sqrt{LC_P}$ the effective inductance becomes very large. This frequency is commonly known as the self resonant frequency (SRF) of the inductor. For microwave applications, inductors can only be utilized below the SRF. Above the SRF, the inductor becomes a capacitor with a very low and unstable value.

The Q-factor of an inductor is a measure of how much energy is stored within the inductor due to its magnetic field and how much is dissipated in its resistance. There are many definitions for the quality factor of an inductor [65] among which

$$Q = \frac{\Im(Z_{in})}{\Re(Z_{in})} \tag{4.2}$$

is the most suitable for microwave applications, where $\Im(.)$ and $\Re(.)$ refers to the imaginary and real parts respectively. This equation can also be re-written as

$$Q = \frac{\omega L_{\text{eff}}}{R}.$$
(4.3)

It is worth mentioning that the Q-factor as represented in (4.3) is directly proportional to the frequency. This will be compared to the Q-factor of the capacitors in the following section. The Q-factor of the inductor is a very important figure of merit for a bandpass filter because it directly affects the insertion loss.

One of the main objectives of the SoP structure, is the integration of passives within the package. Both LTCC and LCP offer very good characteristics for inductors [63, 64]. First, their dielectric constants are low (2.9 for LCP, and 5.7 for LTCC) which reduces the parasitics and inter-turn capacitance. This, in turn, increases the SRF, allowing for wider bandwidths. Second, both technologies have low loss tangents (0.002 for LCP, and 0.0025 for LTCC), which means inductors with high Q-factors can be produced.



Figure 4.3: Illustration of spiral inductors (a) A circular inductor with one and a half turns (b) A circular inductor with two and a half turns (c) A rectangular inductor with one and a half turns (b) A rectangular inductor with two and a half turns . The inductors are a single port network with the other end shorted. A transition from microstrip to CPW is also included to facilitate probe-measurement.

There are two main types of inductors: line and spiral. Line inductors are used to obtain very low effective inductances, while spiral inductors are used when higher effective inductances are required. There are two main types of spiral inductors: circular and rectangular. The main design parameters of spiral inductors are the width of the line W, the spacing between the lines s, and the outer diameter of the inductor d_o as shown in Fig. 4.3, where both rectangular and circular spirals are presented.

4.1.1 Measured Results of Inductors in ultra-thin LCP

A total of 30 inductors have been fabricated by Metro Circuits in Rochester, New York [66]. In Fig. 4.5 a photograph of the fabricated panel is presented. These components have been measured using Anristu vector network analyser (ME7828A) and Cascade, Summit
12000 probe-station, with $150\mu m$ pitch size. First, the S-parameters are measured, then the Z-parameters are extracted assuming 50Ω termination (the impedance at the probe tips). From the Z-parameters the effective inductance and Q-factor are calculated. The Q-factor has been calculated using (4.2), while the effective inductance is calculated using:

$$L_{\text{eff}} = \frac{\Im(Z_{in})}{2\pi f}.$$
(4.4)



Figure 4.4: Measured and simulated inductance and quality factor for an inductor with one and a half turns and an outer radius of 1.1 mm implemented in the ultra-thin LCP. (a) Inductance and, (b) Q-factor.

In Fig. 4.4, the measured and simulated inductance and Q-factor are compared for an inductor with one and a half turns implemented in the ultra-thin LCP. It is observed that the measured inductance is slightly higher than the simulated one. This observation is common for all the other inductors. The measured Q-factor, on the other hand, is observed to be considerably lower than the simulation. As detailed in [67], the measurement of Q-factors using VNAs becomes less accurate as the value of the Q-factor increases; therefore, rapid fluctuations are observed when the Q-factor is plotted, as shown in Fig. 4.4 (b). To make the plot more readable, the data is filtered using MATLAB, which also adds to the uncertainty of the measured Q-factor. Also plotted in Fig. 4.4 is the simulation with a surface roughness of $R_a = 0.5 \mu m$, which has been measured using Dektak 102 Surface profylometer. It can be

observed that the surface roughness has a very minor effect on the inductance as well as the quality factor which is in accordance with the analysis in [68], where it has been reported that the surface roughness does not have a significant effect at frequencies below 10 GHz.



Figure 4.5: A photograph of the fabricated lumped components in M-LCP.

The measured values for all the fabricated inductors are shown in table 4.1, where Q-factors as high as 50 have been measured. Higher Q-factors can be obtained by increasing the thickness of the substrate. For a thickness of 400 μ m (the stack-up of Fig. 4.1(b)), a Q-factor as high as 150 has been achieved. This demonstrates the a trade-off between the low thickness of the substrate; therefore flexibility, and the Q-factor of the inductors. It is worth mentioning that the Q-factors reported in table 4.1 have been obtained from the VNA measurement of S_{11} , which is not an accurate method to measure the Q-factor as discussed in [67]. Therefore, the actual values for the Q-factors are expected to be higher.

4.1.2 Effect of the Ground Plane on the Q-factor of inductors

To investigate the effect of the proximity of the ground plane on the quality factor of inductors, two identical inductors have been fabricated. One of them has a complete ground and the other has its ground partially-removed as shown in Fig. 4.6 (a) and (b), respectively. Measured results of the effective inductance and Q of both grounded and ungrounded induc-

#	Type	Ν	R_o	W	S	Q_{max}	SRF	L_{eff}
			[mm]	$[\mu m]$	$ $ [μm]		[GHz]	[nH]
1	Circ.	0.5	1.1	150	50	30@ 4.5	> 5	1.8
2	Circ.	0.5	2.5	150	50	27@0.5	5	3.5
3	Circ.	2.5	1.1	150	70	30@1	3.3	7
4	Circ.	2.5	2.5	150	70	22@0.5	1.3	N/A
5	Circ.	1.5	1.2	150	75	28@1	4.5	5
6	Circ.	1.5	2.5	150	50	25@0.75	2	19
7	Circ.	1.5	1.2	150	100	24@1	4	6
8	Circ.	1.5	2.5	100	100	23@0.75	2	25
9	Circ.	1.5	1.1	75	50	10@1	4	4
10	Circ.	1.5	1.1	150	50	12@1.2	4.5	2
11	Circ.	1.5	1.1	75	50	16@1	4.2	5
12	Circ.	1.5	1.1	150	50	20@1	4.7	3
13	Circ.	1.5	1.1	75	50	24@1	4.2	7
14	Circ.	1.5	1.1	150	50	30@1	5	4.7
15	Circ.	1	1.1	150	50	25@0.9	> 5	2
16	Circ.	1	1.1	150	50	17@1	> 5	2
17	Circ.	1	1.1	150	50	12@1	> 5	2
18	Circ.	1	2.5	150	50	25@0.9	3	7
19	Circ.	2	1.1	150	50	30@1	3.7	6
20	Circ.	2	1.1	150	50	18@1.1	4	4
21	Circ.	2	1.1	150	50	12@1	3.7	2.5
22	Circ.	2	2.5	150	50	25@0.5	1.5	N/A
23	Circ.	1.5	1.5	100	100	30@0.25	2.6	23
24	Circ.	1.5	1.5	100	100	35@0.25	1.25	N/A
25	Citrc.	2.5	2.5	70	70	55@0.25	1.2	N/A
26	Circ.	0.5	2.5	150	50	35@0.25	3	16
27	Rect.	0.5	1.1	150	50	30@0.25	> 5	7
28	Rect.	2.5	1.1	150	70	27@0.7	2.6	10
29	Rect.	1.5	1.2	100	75	27@1	3.5	7
30	Rect.	1	1.1	100	100	25@1	5	4

 Table 4.1: Measured results of LCP Inductors



Figure 4.6: (a) An illustration of an inductor with two and a half turns.(b) The same inductor in (a) with the ground under it partially-removed. (c) Photograph of a one and a half turns inductor in LCP. (d) A Photograph of a one and a half turns inductor in LTCC, which is similar to the one in (c).

tors are shown in Fig. 4.7. As shown in the figure, when the ground is partially removed, the Q-factor increases (doubles) at low frequencies then deteriorates at higher frequencies. The inductance, on the other hand, increases by approximately 3 times. This observation is common for all the three inductors presented in Fig. 4.7 (a), (b), and (c), which indicates that it is independent on neither the number of turns nor the outer diameter. The improved inductance and Q-factor of the inductors with partially removed ground at low frequencies can be attributed to the reduction in the eddie currents, while the deterioration of the Qfactor at higher frequencies can be attributed to the reduction in the SRF. The reduction of the SRF when the ground is partially-removed can be attributed to the increased stray capacitance as well as the inductance of the ground feed line.

4.1.3 Comparison between LCP and LTCC inductors

An inductor with one and half turns has been implemented in LCP and LTCC according to the stackups of Fig. 4.1 (a) and (c). The inductances and Q-factors of both the LCP and LTCC implementations are shown in Fig. 4.8 (a) and (b), respectively. A third implementations, referred to as thick LCP, which is based on the stackup of Fig. 4.1 (b), has been included in simulations only for completeness of the comparison.



Figure 4.7: Comparison between inductors with complete and partially-removed grounds in terms of inductance in nH, and Q-factors (smoothed). (a) Measured inductances and Q-factors of inductors with half a turn and diameter of 2.2mm. (b) Measured inductances and Q-factors of inductors with one and a half turns and outer diameter of 2.4mm. (c) Measured inductances and Q-factors of inductors of inductors with two and a half turns and outer diameter of 5mm.



Figure 4.8: Plots of the inductance (in nH) and Q-factor of one and a half turns inductors with radii of 1.1mm.

Thick LCP and LTCC inductors show very close inductances, but the LTCC inductor has a lower SRF. This is mainly because LTCC has higher dielectric constant which results in increased stray capacitances and, therefore, reduced SRF. The ultra-thin M-LCP inductor, on the other hand, has approximately 50% lower inductance. This is because the ground in the case of the ultra-thin inductors is four times closer to the inductor than it is for the case of the thick designs, which results in increased eddie currents induced in the ground. The magnetic field produced by these eddie currents adds destructively to the main field which results in reduced inductance as compared to the other stack-ups. The Q-factors of the inductors are plotted in Fig. 4.8 (b). The thick LCP inductor has a higher Q-factor as compared to LTCC inductor. Again, this can be attributed to the lower ϵ_r of LCP, which results in lower stray capacitance and; therefore, lower losses. The ultra-thin inductor, on the other hand, has almost 50 % less Q as compared to LTCC which is due to the closer proximity of the ground plane as explained previously. It is worth mentioning that the Q-factors for the ultra-thin LCP and LTCC are not very accurate because they have been calculated from the S_{11} measurements [67]; therefore the actual Q-factors for these cases are expected to be higher.



Figure 4.9: Plots of the characteristics of LTCC, thick, and thin LCP inductors as a function of the ratio between the spacing between the lines (S) and the width of the line (W). (a) Inductance at 0.5GHz, and (b) self resonant frequency (SRF).

The results of the comparison are summarized in Fig. 4.9. LTCC inductors have slightly higher inductances than the thick LCP ones. Ultra-thin LCP inductors, on the other hand, have the smallest inductances, but the largest SRFs. This indicates a clear trade-off between the amount of inductance and Q on one hand, and the SRF and thickness on the other hand. The Q of ultra-thin inductors can be considerably improved especially at lower frequencies if the ground is partially removed as shown in Fig. 4.7.

4.2 Capacitors

Capacitors are very important components in RF and microwave systems. They are used in circuits such as matching networks, oscillator tanks, and as dc blocks. The quality factor of a capacitor can be defined as [65]

$$Q = \frac{1}{\omega CR} \tag{4.5}$$

where the losses of the substrate are ignored. It is clear form the definition of the Q in (4.5) that it is inversely proportional to the frequency, unlike the case of inductors where it is

directly proportional to the frequency. Therefore, at low frequencies, the Q of the tank and resonator is dominated by the Q of the inductor, while at high frequencies it is dominated by the Q of the capacitor. A mid point exists around 10 GHz where they equally contribute to the overall Q. At low microwave frequencies, therefore, capacitors are only characterized by their capacitance and self resonant frequencies. In this section capacitors fabricated in LTCC, thick and thin LCP are presented and compared.



Figure 4.10: Illustration of the VID or parallel plate capacitor. (a) Schematic of rectangular and circular VID capacitors. (b) A photograph of an LCP capacitor. (c) A photograph of an LTCC capacitor.

4.2.1 Capacitor Design and Optimization

There are two main types of capacitors; parallel plate capacitors, also commonly referred to as vertically interdigitated (VID) capacitors, and planar, interdigital capacitors. VID capacitors are composed of multiple plates aligned vertically, with interchanging plates coupled together (Fig. 4.10). If only two plates are used, the capacitor is also commonly referred to as metal insulator metal or MIM capacitor. The capacitance of such capacitor can be estimated by the classical formula (4.6).

$$C = \epsilon A/d \tag{4.6}$$

where C is the capacitance in Farad, ϵ is the absolute permeability of the dielectric between the plates, A is the area of the plates in m^2 , and d is the vertical distance between the plates in meters. If more than two parallel plates are used, the total effective capacitance can be estimated as

$$C = (N-1)\epsilon A/d \tag{4.7}$$

where N is the number of parallel plates. As shown in (4.7) the capacitance roughly multiplies when the number of parallel plates is increased, this highlights the advantages of 3D integration since the increase in the capacitance comes with no cost at all in the planar area. In this work many different VID capacitors are fabricated in the ultra-thin LCP stack-up. As for the case of inductors, similar designs are presented in LTCC and thick LCP for the sake of comparison. Moreover, a new type of capacitor is presented, which resulted from the combination of horizontal and vertical interdigital capacitors and is referred to as vertical and horizontal interdigitated (VHID) (Fig. 4.12).

4.2.2 Measured results of capacitors in the ultra-thin LCP

The measured results of four VID capacitors of the type shown in Fig. 4.10 (a) in the ultrathin LCP stackup of Fig. 4.1 (c) with different sizes are shown in Fig. 4.11. It is clear that the capacitance increases almost exponentially with planar size. Increasing the planar size of the parallel plates, however, will not increase the capacitance indefinitely, because the SRF decreases as the size is increased as clearly observed in the figure, limiting the usable bandwidth of the capacitor.

VHID capacitors in LTCC

The VHID capacitor introduced previously and illustrated in Fig. 4.12 can be used to introduce an additional degree of freedom. It starts with a planar interdigital capacitor in the top layer. On the next lower layer, polarities of interdigital fingers are reversed to keep the



Figure 4.11: Illustration of the effect of size on the effective capacitance and SRF. Measured capacitance for 3-layered rectangular VID capacitors with different sizes in ultra-thin LCP.



Figure 4.12: Illustration of the different realizations of the VHID capacitor. (a) Standard VHID capacitor. (b) The same as (a) but with circular profile. (c) More symmetrical design but with two additional via holes. (d) Design with higher SRF, but lower effective capacitance.

parallel fingers at different polarities. This approach is repeated for multiple layers to form a complete capacitor. One of the advantages of this capacitor is symmetry in comparison to the conventional VID [4]. Another advantage presented in this work for the first time is the ability of the VHID capacitor to trade capacitance with SRF and substrate thickness. For large substrate thickness where the SRF is usually high enough the VHID capacitor offers higher capacitance per unit area as compared to VID. This is illustrated in Fig. 4.13 for the case of LTCC. For ultra-thin substrates, however, when the SRF is critical, the VHID capacitor offers higher SRFs as compared to the VID capacitor.



Figure 4.13: Measured capacitance of two square VID capacitors of lengths of 1.5mm and 2.5mm, and their VHID counterparts implemented in LTCC (Fig. 4.1(a)).

VHID capacitors in ultra-thin LCP

For the case of LCP, where the thickness of the substrate is small, the SRF decreases rapidly limiting the frequency of operation. In this case, the VHID capacitor can be used to provide slightly higher SRFs as compared to VID. A comparison between VHIDs and VIDs implemented in ultra-thin LCP is shown in Fig. 4.14. When the size of the capacitor is small, the VHID capacitor offers higher SRF but with lower capacitance. When the size increases, on the other hand, the VID capacitor prevails with higher capacitance and higher SRF.

In Fig. 4.15 a comparison between the different realizations of the VHID capacitors of Fig. 4.12 is illustrated. The capacitor of type (a) shows a superior performance in terms of SRF when the size of the capacitor is small. When the size is large type (d) emerges as having much higher SRF than all other types, but the decrease in capacitance is considerable limiting the use of this type. Type (c) is very comparable to the conventional VID in both cases with added fabrication complexity.

In conclusion, the VHID capacitor's performance is compared to the VID capacitor for

two main cases. (1) When the thickness of the substrate is relatively large like the case of LTCC, the VHID provides higher capacitance per unit area but its SRF is lower than the VID. (2) When the thickness of the substrate is ultra-thin like the case of LCP, the VHID capacitor offers higher SRF but its capacitance per unit area is less than the VID. From this discussion the VHID capacitor provides an additional degree of freedom for further optimization of the design.



Figure 4.14: Comparison between the conventional VID capacitor and the VHID capacitor for various sizes of circular capacitors implemented in the ultra-thin LCP.



Figure 4.15: Comparison between the different types of the VHID capacitor illustrated in Fig. 4.12. (a) with length of $600\mu m$. (b) with length of $1500\mu m$.

4.2.3 Comparison between LCP and LTCC VID capacitors

In this section a comparison between VID capacitors implemented in the LTCC, thick LCP and ultra-thin LCP is presented according to the stack-ups of Fig. 4.1. As for the case of the inductor the stackup of Fig. 4.1 (b) for thick, LTCC like, LCP is included in simulations only to complete the comparison while capacitors in the other two stackups have been fabricated and tested. A comparison between the VHID capacitors in LTCC and LCP can already be deduced from Fig. 4.13 and Fig. 4.14, respectively as discussed in the previous section.

Measured capacitance of the LTCC as well as M-LCP are presented in Fig. 4.16. It is observed that the LTCC capacitor has approximately two times more capacitance per unit area as compared to the thick M-LCP capacitor, which corresponds well to the difference in the dielectric constant between the LTCC and LCP used in this work according to the classical formula. The ultra-thin M-LCP capacitor, on the other hand, has the highest capacitance per unit area which is slightly higher than the LTCC capacitor even thought the dielectric constant of LCP is approximately half that of LTCC. This is due to the reduced thickness of the M-LCP substrate, which also complies well with equation (4.7).



Figure 4.16: Comparison between three circular VID capacitors with a radius of 0.6mm implemented in LTCC and two stack-ups of M-LCP according to Fig. 4.1.

In Fig. 4.17, a summary of the comparison is presented for two different radii. It can be concluded that LTCC provides more capacitance per unit area and therefore, smaller size as compared to thick LCP because of the higher dielectric constant of LTCC. However, since ultra-thin stack-ups are possible for LCP, it can provide higher capacitance than LTCC if the ultra-thin stackup is used as shown in Fig. 4.17 (a). For the case of the SRF, the thick LCP capacitors have the highest SRFs, especially for smaller sizes (Fig. 4.17(b)). The ultra thin M-LCP capacitors, on the other hand, show slightly higher SRFs as compared to LTCC. As the radius is increased, however, all three technologies approach each other and the SRFs generally deteriorate.

In this chapter, lumped inductors and capacitors in LTCC and LCP have been presented with a focus on ultra-thin LCP because LTCC and thick LCP have already been reported in the literature. A comparison between the LTCC, thick and ultra-thin LCP have also been presented. In the following chapter, bandpass filters based on these components are presented, which highlights the miniaturization and performance potential of such components.



Figure 4.17: Plots of the characteristics of LTCC, thick and thin LCP capacitors. (a) Capacitance at 0.5GHz, and (b) self resonant frequency (SRF).



Chapter 5

Bandpass Filter Design in SoP technologies

In this chapter, the lumped components presented in the previous chapter are used to design a miniaturized bandpass filter for the GPS application. The GPS band is at 1.5 GHz, which is particularly good for demonstrating the miniaturization of bandpass filters because the size of transmission lines at this frequency is large. Also, selectivity and out-of-band rejection are very important in this band because of the tight placement of the nearby commercial bands. Self interference rejection is another important factor because of the multiple radio transceivers typically integrated in the modern wireless systems. Adding transmission zeros to the bandpass filter improves the selectivity by increasing the rolloff factor but, as a consequence, the out-of-band rejection deteriorates. In this chapter, a design that simultaneously tackles the problems of size, selectivity, and out-of-band rejection, is presented. First, the proposed design topology is presented and analysed to extract the design information. Next, the implementations of the bandpass filter in LTCC and LCP are presented. Finally, the simulated and measured results of the fabricated modules are presented and discussed.



Figure 5.1: (a) The topology of the second order filter with mutually coupled inductors. (b) The response of the topology in (a). (c) The proposed, modified topology with a capacitor in parallel and an inductor in series. (c) The response of the new topology with the capacitor only. (e) The response of the new topology with both the capacitor and inductor.

5.1 The analysis of the Proposed topology

In this work, a bandpass filter is designed in the smallest order (the second order) to reduce the number of components and, thus the size. However, a second order filter does not provide the selectivity required by most modern applications as shown in Fig. 5.1(b). The topology presented in Fig. 5.1 (a) is used as a starting point, which has two inductively coupled resonators [2]. The inductive coupling is used in this case because it allows the inductors to be closely positioned, which results in a smaller size. This topology has been improved by adding a capacitor in parallel to introduce two transmission zeros and increase the selectivity of the filter as shown in Fig. 5.1 (d) [2]. Even though, the selectivity is increased in the this topology, the out-of-band rejection deteriorates. To improve the out-of-band rejection at the upper band an additional transmission zero is proposed in this work by simultaneously adding an inductor in series and a capacitor in parallel with the circuit of the filter as shown in Fig. 5.1 (c). As a result, three transmission zero are obtained as shown in Fig. 5.1 (e). The required inductance L_L can be manipulated, as discussed later, to have a value equals to the inductance of a via-hole; therefore, it can be implemented using a via-hole, which saves considerable space. Typically, the via holes are already used to provide the ground path for the resonators and they can be used for the series inductance which means the out-of-band rejection can be improved without adding any new component. In the following section, the theoretical analysis of the proposed topology is presented to extract the design information.

5.1.1 Theoretical Analysis

In this section, the proposed topology presented in Fig. 5.1 (c), which provides three transmission zeros, is analysed. The main objective of the analysis is to obtain relations between the positions of the three transmission zeros and the values of the parallel capacitor (C) and the series inductor (L_L). These relations are expected to help in the design of bandpass filters by allowing control over the positions of the three transmission zeros.

The analysis can be started from the topology of Fig. 5.1 (a), which can be synthesized from the low pass prototype values (g_i) using the relations driven for the modified topology of Fig. 3.10. For the middle J-inverter, the inductor equivalent network of Fig. 3.11 is used instead of the capacitor network with the negative inductors absorbed in the resonator inductors as shown in chapter 3. To this end, the circuit of Fig. 5.2 (a) is reached, which is much easier to synthesize than the actual topology. To establish an analytical relation between the two topologies, the coupled resonator network of Fig. 5.2 (b) is first analysed by writing its currents and voltages as follows:



Figure 5.2: (a) Bandpass filter resulted from the modified topology driven in chapter 3 with inductor J-inverter network. (b) Mutually coupled inductors network. (c) and (d) Networks equivalent to (b).

$$v_1 = -\left(L_1 \frac{di_1}{dt} + M \frac{di_2}{dt}\right) \tag{5.1a}$$

$$v_2 = -\left(M\frac{di_1}{dt} + L_2\frac{di_2}{dt}\right) \tag{5.1b}$$

from which the Z-parameters of the network can be written as

$$Z = j\omega \begin{bmatrix} L_1 & M \\ M & L_2, \end{bmatrix}$$
(5.2)

where a sinusoidal time-variation is assumed and M is the mutual coupling between the inductors, which is related to the coupling coefficient (k) by

$$M = k\sqrt{L_1 L_2}.\tag{5.3}$$

From the Z-parameters of (5.2), an equivalent T-network with the same Z-parameters can be constructed as shown in Fig. 5.2(c). This T-network can be related to the II network of Fig. 5.2(d) using the delta-star transformation to yield the following relations:

$$L'_{1} = \frac{(L_{1} - M)(L_{2} - M) + M(L_{1} - M) + M(L_{2} - M)}{(L_{2} - M)}$$
(5.4a)

$$L'_{2} = \frac{(L_{1} - M)(L_{2} - M) + M(L_{1} - M) + M(L_{2} - M)}{(L_{1} - M)}$$
(5.4b)

$$L_{mm} = \frac{(L_1 - M)(L_2 - M) + M(L_1 - M) + M(L_2 - M)}{M}$$
(5.4c)

To analyse the effect of the series inductance (L_L) on the performance of the filter, the transmission impedance (Z_{21}) of the filter's network is calculated by successively using the delta-star transformation as shown in Fig.5.3. First, the resonator elements C_1 and L'_1 are combined to form the impedance Z_p as shown below:

$$Z_p = \frac{\frac{1}{j\omega C_1} \times j\omega L'_1}{\frac{1}{j\omega C_1} + j\omega L'_1} = \frac{L'_1}{C_1(-\frac{j}{\omega C_1} + j\omega L'_1)}$$
(5.5)

Second, the Delta network composed of, Z_p , Z_p and L_{mm} is transformed to a Star network with elements Z_{s1} , Z_{s1} , and Z_{s2} as follows:

$$Z_{s1} = \frac{(j\omega L_{mm} \times Z_p)}{2Z_p + j\omega L_{mm}} = \frac{j\omega LL_{mm}}{C_1(-\frac{j}{\omega C_1} + j\omega L)(j\omega L_{mm} + \frac{2L}{C_1(-\frac{j}{\omega C_1} + j\omega L)})}$$
(5.6a)

$$Z_{s2} = \frac{Z_p^2}{2Z_p + j\omega L_{mm}} = \frac{L^2}{C_1^2 (-\frac{j}{\omega C_1} + j\omega L)^2 (j\omega L_{mm} + \frac{2L}{C_1 (-\frac{j}{\omega C_1} + j\omega L)})}$$
(5.6b)



Figure 5.3: Schematic of the circuits resulting from consecutive application of the delta-star transformation.

Next Z_{s1} is combined with C_2 to form Z_{T1} , while Z_{s2} is combined with L_L to form Z_{T2} as shown in Fig. 5.3(c).

$$Z_{T1} = Z_{s1} + \frac{1}{j\omega C_2} = -\frac{j}{\omega C_2} + \frac{j\omega L L_{mm}}{C_1(-\frac{j}{\omega C_1} + j\omega L)(j\omega L_{mm} + \frac{2L}{C_1(-\frac{j}{\omega C_1} + j\omega L)})}$$
(5.7a)

$$Z_{T2} = j\omega L_L + Z_{s2}$$

= $j\omega L_L + \frac{L^2}{C_1^2 (-\frac{j}{\omega C_1} + j\omega L)^2 (j\omega L_{mm} + \frac{2L}{C_1 (-\frac{j}{\omega C_1} + j\omega L)})}$ (5.7b)

Finally, the Deta network composed of Z_{T1} , Z_{T1} , and C is transformed to a star network with elements Z_{ss1} , Z_{ss1} , and Z_{ss2} as follows:

$$Z_{ss1} = \frac{\frac{Z_{T1}}{j\omega C}}{2Z_{T1} + \frac{1}{j\omega C}} = -\frac{j\left(-\frac{j}{\omega C_2} + \frac{j\omega LL_{mm}}{C_1(-\frac{j}{\omega C_1} + j\omega L)(j\omega L_{mm} + \frac{2L}{C_1(-\frac{j}{\omega C_1} + j\omega L)})}\right)}{\omega C\left(-\frac{j}{\omega C} + 2\left(-\frac{j}{\omega C_2} + \frac{j\omega LL_{mm}}{C_1(-\frac{j}{\omega C_2} + j\omega L)(j\omega L_{mm} + \frac{2L}{C_1(-\frac{j}{\omega C_1} + j\omega L)})}\right)\right)}(5.8)$$

$$Z_{ss2} = \frac{Z_{T1}^2}{2Z_{T1} - \frac{j}{\omega C}} = \frac{\left(-\frac{j}{\omega C_2} + \frac{j\omega LL_{mm}}{C_1(-\frac{j}{\omega C_1} + j\omega L)(j\omega L_{mm} + \frac{2L}{C_1(-\frac{j}{\omega C_1} + j\omega L)})}\right)^2}{-\frac{j}{\omega C} + 2\left(-\frac{j}{\omega C_2} + \frac{j\omega LL_{mm}}{C_1(-\frac{j}{\omega C_1} + j\omega L)(j\omega L_{mm} + \frac{2L}{C_1(-\frac{j}{\omega C_1} + j\omega L)})}\right)\right)}(5.9)$$

 Z_{ss2} is then combined with Z_{T2} to form the transmission impedance Z_{12} as follows:

$$Z_{12} = Z_{T2} + Z_{ss2}$$

$$= j\omega L_L$$

$$+ \frac{L^2}{C_1^2 (-\frac{j}{\omega C_1} + j\omega L)^2 (j\omega L_{mm} + \frac{2L}{C_1 (-\frac{j}{\omega C_1} + j\omega L)})}{(-\frac{j}{\omega C_2} + \frac{j\omega LL_{mm}}{C_1 (-\frac{j}{\omega C_1} + j\omega L) (j\omega L_{mm} + \frac{2L}{C_1 (-\frac{j}{\omega C_1} + j\omega L)})})^2$$

$$+ \frac{j\omega LL_{mm}}{-\frac{j}{\omega C_2} + 2(-\frac{j}{\omega C_1} + \frac{j\omega LL_{mm}}{C_1 (-\frac{j}{\omega C_1} + j\omega L) (j\omega L_{mm}) + \frac{2L}{C_1 (-\frac{j}{\omega C_1} + j\omega L)})})$$

$$= j(-2CL - CL_{M} + 2CC_{1}L^{2}\omega^{2} + 2CC^{2}L^{2}\omega^{2} + C_{2}^{2}L^{2}\omega^{2} + 4CC^{2}LL_{L}\omega^{2} + 2C_{2}^{2}LL_{L}\omega^{2} + 2CC_{1}LL_{M}\omega^{2} + 2CC_{2}LL_{M}\omega^{2} + C_{2}^{2}L_{L}L_{M}\omega^{2} - 4CC_{1}C_{2}L^{2}L_{L}\omega^{4} - 2C_{1}C_{2}^{2}L^{2}L_{L}\omega^{4} - 2CC_{1}C_{2}L^{2}L_{M}\omega^{4} - 2CC_{1}C_{2}L^{2}L_{M}\omega^{4} - 4CC_{1}C_{2}LL_{L}L_{M}\omega^{4} - 2CC_{2}^{2}LL_{L}L_{M}\omega^{4} - 4CC_{1}C_{2}LL_{L}L_{M}\omega^{4} - 2CC_{2}^{2}LL_{L}L_{M}\omega^{4} - 2CC_{1}^{2}C_{2}L^{2}L_{L}L_{M}\omega^{6} + 2CC_{1}C_{2}^{2}L^{2}L_{L}L_{M}\omega^{6} + C_{1}^{2}C_{2}^{2}L^{2}L_{L}L_{M}\omega^{6})/(C_{2}\omega(-1 + C_{1}L\omega^{2})(-4CL - 2C_{2}L - 2CL_{M}) - C_{2}L_{M} + 2CC_{1}LL_{M}\omega^{2} + 2CC_{2}LL_{M}\omega^{2} + C_{1}C_{2}LL_{M}\omega^{2}))$$

$$(5.10a)$$

while Z_{11} and S_{21} are calculated as follows:

$$Z_{11} = Z_{12} + Zss1$$

$$\begin{split} &= j\omega L_{L} + \frac{L^{2}}{C_{1}^{2}(-\frac{j}{\omega C_{1}}+j\omega L)^{2}(j\omega L_{mm}+\frac{2L}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)})} \\ &- \frac{j(-\frac{j}{\omega C_{2}}+\frac{j\omega L_{mm}}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)(j\omega L_{mm}+\frac{2L}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)})})}{\omega C(-\frac{j}{\omega C}+2(-\frac{j}{\omega C_{2}}+\frac{j\omega L_{mm}}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)(j\omega L_{mm}+\frac{2L}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)})})) \\ &+ \frac{(-\frac{i}{\omega C_{2}}+\frac{j\omega L_{mm}}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)(j\omega L_{mm}+\frac{2L}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)})})^{2}}{-\frac{j}{\omega C}+2(-\frac{j}{\omega C_{2}}+\frac{j\omega L_{mm}}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)(j\omega L_{mm}+\frac{2L}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)})})) \\ &+ \frac{(-\frac{i}{\omega C_{2}}+\frac{j\omega L_{mm}}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)(j\omega L_{mm}+\frac{2L}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)})})^{2}}{(-\frac{j}{\omega C}+2(-\frac{j}{\omega C_{2}}+\frac{j\omega L_{mm}}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)(j\omega L_{mm}+\frac{2L}{C_{1}(-\frac{j}{\omega C_{1}}+j\omega L)})}) \\ &+ \frac{2Z_{0}Z_{12}}{(Z_{11}+Z_{0})^{2}-Z_{12}^{2}} \\ &= (100\omega(-2CL-CL_{mm}+2CC_{1}L^{2}\omega^{2}+2CC_{2}L^{2}\omega^{2}+C_{2}^{2}L^{2}\omega^{2}+4CC_{2}LL_{L}\omega^{2} \quad (5.11b) \\ &+ 2C_{2}^{2}LLL\omega^{2}+2CC_{1}LL_{mm}\omega^{2}+2CC_{2}LL_{mm}\omega^{2}+2CC_{2}L_{L}L_{mm}\omega^{2}+C_{2}^{2}L_{L}L_{m\omega}\omega^{2} \\ &- 4CC_{1}C_{2}L^{2}L_{L}\omega^{4}-2C_{1}C_{2}^{2}L^{2}L_{L}\omega^{4}-CC_{1}^{2}L^{2}L_{mm}\omega^{4}-2CC_{1}C_{2}L^{2}L_{mm}\omega^{4} \\ &- 4CC_{1}C_{2}LL_{L}L_{mm}\omega^{4}-2CC_{2}^{2}L^{2}L_{L}L_{mm}\omega^{4}-2CC_{1}C_{2}^{2}L^{2}L_{L}L_{m}\omega^{4} \\ &+ 2CC_{1}C_{2}^{2}L^{2}L_{L}L_{mm}\omega^{6}+C_{1}^{2}C_{2}^{2}L^{2}L_{L}L_{mm}\omega^{6}))/((2jL+jL_{mm}-200CL\omega-100C_{2}L\omega \\ &- 100CL_{mm}\omega-50C_{2}L_{mm}\omega-jC_{1}LL_{mm}\omega^{2}-jC_{2}LL_{mm}\omega^{2}+100CC_{1}LL_{mm}\omega^{3} \\ &+ 100CC_{2}LL_{mm}\omega^{3}+50C_{1}C_{2}LL_{mm}\omega^{3})(1+50jC_{2}\omega-C_{1}L\omega^{2}-C_{2}L\omega^{2}-2C_{2}L_{L}\omega^{2} \\ &- 50jC_{1}C_{2}L\omega^{3}+2C_{1}C_{2}LL_{L}\omega^{4}))) \end{split}$$

If $|Z_{21}| = 0$, no transmission between the input and output can occur; therefore, a transmission zero occurs. The denominator of $|Z_{21}|$ in this case is of no importance and the only condition for transmission zeros to occur is:

$$|Z_{21}|_n = 0 \tag{5.12}$$

where $|Z_{21}|_n$ is the numerator of the magnitude of Z_{21} given in a simplified form as:

$$|Z_{21}|_n = A \omega^6 + B \omega^4 + C \omega^2 + D = A (\omega^2)^3 + B (\omega^2)^2 + C \omega^2 + D$$
(5.13)
where,

$$A = (L_L L_M L^2 C_1 C_2) (2C(C_1 + C_2) + C_1 C_2)$$

$$B = -(L + L_{mm}) (2C_1 C_2 L L_L) (2C - C_2 + 1) - 2CC_2 L L_{mm} (C_1 L + C_2 L_L)$$

$$-CL^2 L_{mm} (C_1^2 + C_2^2)$$

$$C = 2CL (C_1 + C_2) (L + L_{mm}) + C_2^2 (L^2 + 2L L_L + L_L L_{mm}) + 2CC_2 L_L (2L + L_{mm})$$

$$D = -2CL - CL_{mm},$$

which is a sixth order, even polynomial in the frequency variable ω . This polynomial can be written as a function of ω^2 which means the six roots form positive-negative pairs. When the three negative roots are discarded, we end up with three possible real positive values for ω , which are the frequencies at which transmission zeros occur. The roots of the polynomial can be obtained graphically as plotted in Fig. 5.4, where $|Z_{21}|_n$ is plotted for various values of the parallel capacitor and the series inductor. The corresponding plots of $|S_{21}|$ are also included (using (5.11) (b) to highlight the effect of the transmission zeros on the performance of the bandpass filter. For the case of C = 0, no transmission zeros occur regardless of the value of L_L . When C is not zero and $L_L = 0$, only two transmission zeros occur which can be slightly controlled by the value of C. Increasing C brings the frequencies of the transmission zeros closer, which increases the roll-off factor but the out-of-band rejection at the upper stop band increases. The out-of-band rejection at the lower stop band is slightly affected because of the trivial transmission zero at dc. When L_L and C are not zeros, three transmission zeros occur, which can be controlled by either L_L or C. As C is increased the frequency band between the two transmission zeros at the upper band shrinks, which improves the rejection at this band. An important observation is that these transmission zeros occur only for very



Figure 5.4: Plots of the magnitude of the numerator of the transmission impedance $(|Z_{21}|_n)$ and the corresponding $|S_{21}|$ for the bandpass filter for values of C ot 0 fF, 60 fF, and 80 fF, and for a series inductance (L_L) of 0 pH, 90 pH, and 300 pH.

small values of L_L . As L_L is increased a critical point is reached where the frequencies of the transmission zeros at the upper stop band become the same. If L_L is increased above this critical value, the transmission zeros vanish. However, even then a relatively high rejection around the frequency of the transmission zeros is observed. To investigate the dependence of the TZs on L_L further, the analytical solution of the cubic polynomial can be used, which is given as follows [69]:

$$\omega_i = y_i - \frac{B}{3A} \qquad i = 1, 2, 3 \tag{5.14}$$

where ω_i are the roots of the polynomial given by (5.13) and the substitutes y_i are given by

$$y_{1} = U + V$$

$$y_{2} = -\frac{1}{2}(U + V) + \frac{j\sqrt{3}}{2}(U - V)$$
 (5.15a)

$$y_{3} = -\frac{1}{2}(U + V) - \frac{j\sqrt{3}}{2}(U - V)$$

where U and V are functions given by

$$U = \sqrt[3]{-\frac{b}{2} + \sqrt{\Delta}}$$
$$V = \sqrt[3]{-\frac{b}{2} - \sqrt{\Delta}}$$
(5.15b)

(5.15c)

and a, b and Δ are functions fiven by

$$a = \frac{1}{3} \left(3\frac{C}{A} - \left(\frac{B}{A}\right)^2 \right)$$
$$b = \frac{1}{27} \left(2\left(\frac{B}{A}\right)^3 - 9\left(\frac{BC}{A^2}\right) + 27\frac{D}{A} \right) \qquad (5.15d)$$
$$\Delta = \frac{b^2}{4} + \frac{a^3}{27}$$

from (5.15b), it can be seen that if $\Delta > 0$ the polynomial will have only one real root. If

 $\Delta = 0$, the polynomial will have three real roots but at least two of them are equals. If $\Delta < 0$, the polynomial will have three real and unequal roots. To see the effect of L_L and C on Δ , it is plotted as a function of L_L for various values of C in Fig. 5.5. It can be seen that for any value of C there is always a maximum (critical) value of L_L above which Δ becomes positive. This critical value of L_L can be increased by increasing C. Therefore the value of C can be increased to get more reasonable values for L_L .



Figure 5.5: Plot of the Δ function as defined in (5.15d) as a function of the series inductor (L_L) and for various values of the parallel capacitor C. if $\Delta > 0$ only one TZ exists. If $\Delta = 0$, three TZ exist but at least two of them are equals. If $\Delta < 0$, three, unequal TZs exist.

5.1.2 Summary of analysis

From the analysis presented in Fig. 5.4 and 5.5, the availability and positions of the three transmission zeros can be controlled from the values of L_L and C. For the case of L_L a critical minimum point exists above which the two transmission zeros at the upper band are lost. The range of this critical value can be increased by increasing the value of C as illustrated in Fig. 5.5. However, as shown in Fig. 5.4, consistent increase in the value of C results in the transmission zeros at the upper-band to shift away from one another, which reduces the rejection in the band between them. The analytical results and relations presented in this section are useful in the prediction and control of the frequencies of the transmission zeros. In the following section these techniques are used in the design and implementation of a bandpass filter in LTCC.

5.2 Bandpass filter implementation in LTCC

5.2.1 3D implementation

A bandpass filter has been designed for the GPS band with a centre frequency of 1.57 GHz and a bandwidth of 100 MHz (FBW of 6%). The first step in the design is to synthesize the circuit of Fig. 5.1(b) based on ideal, lumped components. To improve both the roll-off and the out of band rejection, the TZs are required to be at 1.1 GHz, 1.8 GHz, and 2.7 GHz. To find the values of L_L and C that produce these values, first the inductance of the via-hole has been estimated using electromagnetic simulation to be 0.79 nH. This value has been assigned to L_L and the information in Fig. 5.4 and 5.5 have been used to estimate C to be 0.135 pF. The values of the other components have been calculated as detailed in chapter 3 and the beginning of this chapter and are: $C_1=2.58$ pF, $C_2=0.96$ pF, $L_1=2.88$ nH, and k =0.18.

After the values of the ideal lumped components have been theoretically synthesized,

the individual 3D components have been optimized using the EM simulator high frequency structure simulator (HFSS). For the coupling structure, a novel completely planar structure has been used as shown in Fig. 5.6 (a). To design this coupled inductor network, the coupling coefficient k and the inductance L_1 have been optimized simultaneously. This has been achieved by varying the outer diameter of the inductors and the horizontal spacing between them until the required values of L_1 and k are simultaneously obtained. The coupling coefficient k, has been optimized by changing the spacing between the inductors, and the inductance has been optimized by varying the width and diameter of the inductors.



Figure 5.6: (a) 3D structure of the bandpass filter. (b) The topology of the bandpass filter.

The other component that needed to be individually optimized is the capacitor. The VHID capacitor introduced in chapter 4 has been employed which, beside being symmetrical, has higher capacitance per unit area. This means smaller size as compared to the conventional parallel plate capacitor, when implemented in LTCC as discussed in chapter 4. The improved size capabilities of the VHID capacitors is illustrated in Fig. 4.13, where the measured capacitance of VHID and VID capacitors of identical sizes are presented. The filter



Figure 5.7: A photograph of the fabricated prototype in LTCC.

presented here has been simulated with both VID and VHID capacitors and the one with VHID capacitors demonstrated around 3% size reduction. Finally, the feedback capacitor C has been implemented in the form of an interdigitated capacitor with four fingers, and the grounding via-hole produces the necessary inductance for L_L as described previously. After the individual components have been optimized, the complete bandpass filter is assembled in only 4 layers to minimize the thickness and cost as shown in Fig. 5.6.

5.2.2 Measurements and results

A prototype of the bandpass filter has been fabricated in VTT Electronics, Finland, as shown in Fig. 5.7. Four layers of Ferro[®] A6 LTCC, with a dielectric constant of 5.9 and a loss tangent of 0.002 have been used. The measurement has been performed using a vector network analyser (Anritsu ME7828) and $150\mu m$ pitch probes.

The measured results of the bandpass filter are plotted in Fig. 5.8 with the EM simulation (using $HFSS^{(R)}$) and the circuit simulation (using $ADS^{(R)}$). Even though the circuit simulation assumes ideal components without neither losses nor parasitics, it matches the measurement and EM simulation very well. A measured out-of-band rejection greater than 20 dB has been attained up to almost 4 GHz and almost 10 dB up to 8 GHz, which is due to the third transmission zero obtained by the proposed topology. A measured insertion loss of 2.7 dBs is obtained, which matches the EM simulated value exactly. A slight frequency shift is noticeable between the EM simulation and the measurement, which is attributed to the

fabrication tolerance of the dielectric constant of the Ferro[®] A6 (±0.2 [23]). Also, since the frequencies of the two TZs at the upper band are very sensitive to the value of L_L , which has been implemented through a via hole, a fabrication tolerance in this via hole may have resulted in the slight shift of the TZs observed.

In this section a bandpass filter realized in LTCC has been presented, which is based on a proposed topology that provides three transmission zeros; therefore, achieves high selectivity and out-of-band rejection. The three transmission zeros have been implemented by adding a capacitor in parallel as well as an inductor in series. A via-hole has been utilized to provide the required inductance of the series inductor; therefore, the third transmission zeros has been obtained without any added component. The measured and simulated results agree very well, which proofs the concept. This same filter topology is implemented in an ultra-thin LCP stack-up in the following section and the LTCC and LCP realizations are compared with the literature in Table 2.1, which is discussed later.



Figure 5.8: Measurement, EM and circuit simulations of the BPF.

5.3 Bandpass filter in LCP

As discussed in chapter 4, LCP is available in ultra thin sheets [33], which have been used to fabricate lumped components in an ultra-thin LCP stack-up. In this section a bandpass filter is presented in this ultra-thin LCP, which is based on the same topology presented in the previous section for LTCC.the presented filter demonstrates the potential of LCP for ultra-thin and flexible modules. This filter and the LTCC filter share the same design specifications. However, due to the ultra-thin nature of the LCP used, the following minor modifications have been introduced:

- Two layers have been used to implement the input capacitor (C_1) of the ultra-thin LCP design, while three layers have been used for the case of LTCC. This is because ultra-thin LCP has much higher capacitance density. Using only two layers for the capacitor for the case of LCP enables the inductors to expand on the layer above the capacitor (Layer 1), which added more compactness to the design.
- To improve the Q-factor of the inductors in the ultra-thin LCP case, the ground under them has been partially removed as shown in Fig. 5.9 (c).

5.3.1 Measurement results of the LCP bandpass filter

A prototype of the LCP bandpass filter, which has been fabricated by Metro Circuits in the USA [66], is shown in Fig. 5.10. The filter has been measured using a vector network analyser (Anritsu ME7828) and $150\mu m$ pitch probes. Measured and simulated results are compared in Fig. 5.11 (a). The filter exhibits an out of band rejection of -25 dBs @ 100 MHz from the pass band for both sides. Also, due to the third TZ, the out of band rejection stays below -25 dBs up to 3 GHz. The initial simulations of the filter show an insertion loss of 2.3 dBs, which is expected according to the low loss of the LCP material. The measurements, however, revealed an insertion loss 3 dBs more than the value of the simulations as shown



Figure 5.9: (a) Schematic of the BPF. (b) 3D structure of the mutually coupled inductors. (c) 3D structure of the bandpass filter.



Figure 5.10: (a) A Photograph of the fabricated M-LCP bandpass filter where a miss-alignment between the top two layers can be seen. (b) Illustration of the flexibility of the LCP filter.

in Fig. 5.11 (a). To investigate the reasons for this increased loss, firstly, the roughness of the LCP material has been measured and included in the simulation. The results show that the surface roughness has a negligible effect on the insertion loss of the filter; secondly, the fabricated modules have been examined closely and a miss-alignment between the top two layer of about 100 μ m has been observed, which can be clearly seen in Fig. 5.10 (a). When this miss-alignment is included in the simulations, an upwards frequency shift is obtained. The insertion loss, however, remains unchanged. Finally, the finishing layer typically added to protect the copper top layer, has been included in the simulation. The updated simulation results shown in Fig. 5.11 (b) matches well with the measurements, which indicates that the finishing layer is the major cause of the increased insertion loss. The effect of the finishing layer on the insertion loss of the filter is discussed in the next subsection.



Figure 5.11: Measured and simulated results of the bandpass filter in ultra-thin LCP. (a) Measurements and original simulation. (b) Measurements and corrected simulations (including the miss-alignment and the ENIG finish layer).

Investigation of the effect of the finishing layer on the insertion loss of the filter

Finishing layers are used on PCBs with copper traces to prevent the copper from oxidation. For this particular prototype, the finishing layer used is electro-less nickel immersion gold
(ENIG), which consists of a layer of nickel (~ $9\mu m$), and a thin layer of gold (~ $0.1\mu m$) on top of the original copper metallization. The conductivities of nickel and gold are $1.43 \times 10^{-5} S/m$ and $5.8 \times 10^{-5} S/m$, respectively. The skin depth at 1.5 GHz for nickel and gold are (~ $0.34\mu m$) and (~ $2\mu m$) respectively. Therefore, most of the current flows in the nickel layer rather than the copper and the gold, and thus increases the resistivity by almost 6 times, leading to an increased insertion loss. Also, since nickel has a relative permeability of about 600, it decreases the fundamental frequency of the resonators because it acts as a magnetic core for the inductors. In Fig. 5.12 (a), however, an upwards frequency shift is observed. This can be explained by the effect of the miss-alignment presented in the previous section (can be clearly observed in Fig. 5.10 (a)), which results in smaller capacitors and; therefore, upwards frequency shift. Obviously, the effect of the miss-alignment exceeds the effect of the nickel layer, which is expected since the miss-alignment is considerable (100 μ m).



Figure 5.12: Simulated results of the LCP filters with different thicknesses. (a) Original simulations as compared to simulations with the ENIG layer and the miss-alignment (filter thickness is 100 μ m). (b) Simulated results with and without ENIG finishing layer for a filter with thickness of 400 μ m.

In Fig. 5.12 (b) a similar filter but with four times the thickness (400 μ m) is simulated with and without the ENIG finishing layer. As expected, the finishing layer results in an

Ref.	Tech.	f_0	Vol.	Roll. Up.	Roll. Low.	IL	FBW *	Lay-		
		(GHz)	$\times \frac{10^5}{\lambda_g^3}$	(dB/100MHz)*		dB		ers		
[24]	LTCC	1.3	2.71	12	5.5	1	27	8		
[25]	LTCC	1.55	4.7	8	13.8	2.8	6	8		
[26]	LTCC	2.4	5.52	1.9	7.5	1.8	12	9		
[2]	LTCC	2.5	5.64	12.5	7.7	1	15	4		
[4]	RXP	2.4	1.45	2	11.25	2.3	8	4		
[27]	LTCC	2.44	4.69	2.1	10	1.7	14	10		
[28]	LTCC	2.45	16	4	20	2.2	15	3		
[70]	M-PCB	0.81	38	40	48	1.13	15	4		
[29]	LTCC	3.5	NA	1.6	5	2.7	21	NA		
[1]	LTCC	1.55	1.44	50	34	5	6.6	8		
This Work	LTCC	1.57	1.5	16.7	10	2.7	5	4		
This Work	M-LCP	1.57	0.1	8.7	9.3	2.8	5	4		
* The roll-off & FBW have been estimated from the literature.										

Table 5.1: Comparison between this work and the literature.

upwards frequency shift and a deteriorated insertion loss. It can be seen, however, that the effect of the finishing layer on the insertion loss is less for this case (1.7 dBs) as compared to the ultra-thin case where a difference of 3 dBs is observed (Fig. 5.12 (a)). The nickel-based finishing layer has a greater effect on the ultra-thin filter design because the thickness of the nickel layer in this case is almost 10% of the thickness of the substrate, while it is only 2% of the substrate for the case of the thick filter. Because the effect of the finishing layer is less for thicker stack-ups, it has never been observed and reported before. In this work a bandpass filter is reported in an ultra-thin structure for the first time; therefore, the increased effect of the nickel-based finishing layer has not been considered at the design level. To avoid the increased loss and frequency shift of the finishing layer, it is recommended to use a finishing layer that does not involve metals with low conductivity and high permeability like nickel.

5.3.2 Comparison between the LTCC, LCP filters and the literature

In table 5.1, a comparison between the filters introduced in this chapter and other state of the art filters is presented. It can be seen that apart from [1], which was published after the publication of this work, the proposed LTCC filter design has the highest roll-off factor at the upper band and is amongst the smallest reported designs. Also, the roll-off factor at the lower band is still good (10 dB/100 MHz) contrary to some of the other designs. For the case of LCP, the design provides good roll-offs at both the upper and lower bands and is the smallest reported, being ten times smaller than the filters reported in literature. This is mainly due to the ultra-small thickness of the LCP tapes used in this work.

Even though [1] has been published after our work and has high roll off factors, its insertion loss is 5 dBs which is too high for receiver front ends. The high insertion loss indicates that the transmission zeros of this design are positioned too close to the passband which is why the filter achieves such high roll off factors.

An important aspect of the designs presented in this chapter is the low number of layers used which is only four layers. As can be observed from the table, the average number of layers for the other designs is about seven layers. Reducing the number of layers has a direct impact on the cost of the module because the cost is almost exponentially related to the number of layers. Moreover, reducing the number of layers results in a reduced thickness, which is important in today's slim devices. The designs presented in this chapter demonstrate a high level of miniaturization and, given their design for the SoP platform, they can introduce even higher miniaturization at system level due to their ability to be part of a highly integrated system.



Chapter 6

Magnetically tunable bandpass filter in Ferrite LTCC

Miniaturization is an important requirement for modern wireless systems. The overall size of a wireless system can benefit from the miniaturization of the bandpass filter as well as by reducing the number of filters in the system. This can be done by utilizing tunable filters, which can operate efficiently on multiple bands. If the tunable filter is implemented in the SoP platform, the level of miniaturization enhances as the components can be closely placed in a 3D fashion. In the previous chapter the miniaturization of an individual filter through the design and 3D implementation has been discussed. This chapter addresses the other approach of tunable filters but still in the SoP platform.

6.1 Introduction

Tunable filters can be realized by many techniques using varactor diodes [43], ferroelectric materials [44], and micro-electromechanical (MEMS) techniques. Its difficult to implement the above mentioned techniques in a completely embedded SoP platform; therefore, they can not truly benefit from the miniaturization of the SoP concept. In this chapter, the first tunable bandpass filter, which is completely embedded within multiple vertically integrated

layers of LTCC, is reported. To obtain tunability, a functional LTCC material is used instead of the typical LTCC tapes, which are used as 3D packages only. The LTCC tape used is ESL 40012 from ElectroScience[®] [32], which is produced by dispersing magnetic powder in an organic matrix. The permeability of a magnetic material changes when a direct magnetic (magneto-static) field is applied along the material. By controlling the magnitude of the applied magneto-static field the change in the permeability can be controlled and tunability can be obtained because the change in the permeability affects the velocity of the electromagnetic waves within the material.



Figure 6.1: Illustration of the size reduction introduced by the embedded windings. (a) A typical electromagnet (b) The windings embedded in the ferrite LTCC substrate and connected by viaholes.

Traditionally, magnetically tunable filters use large electromagnets or coils as shown in Fig. 6.1(a). The filters which use these electromagnets can not be integrated with miniaturized wireless systems [71]. Also, the air gap between the electromagnet and the magnetic substrate creates high magnetic losses, which increase the current required to obtain the magneto-static field. In this work, these electromagnets have been avoided and embedded windings are used instead as shown in 6.1(b). These windings are implemented in multilayer LTCC and are connected using via-holes. The use of these windings in the design of the bandpass filter introduces considerable miniaturization and makes the design perfectly suitable for an SoP platform. Because the windings are implemented inside the LTCC substrate, the magnetic losses are eliminated and the effect of the loss of magnetic fields at the air interface, which is known as demagnetization, is minimized.

Typically, magnetically tunable filters are based on the tunability of the ferromagnetic resonance (RMR) frequency [51]. To tune the FMR frequency, however, large magnetostatic fields are required, which can not be produced and handled by the miniaturized wireless systems. In this work two main approaches are used: (1) The large and bulky electromagnets have been replaced with the embedded windings, which results in a large size reduction. (2) Tunability is achieved by operating the ferrite material in the partially magnetized state to reduce the field requirement. The tunable filter presented in this chapter can be readily integrated in wireless systems that require tunability and miniaturization. In the following section the theory behind the tunability of ferrite-based filters is presented briefly. Next the reported design is presented in details.

6.2 Theory of Ferrite-Based Tunable Filters

Magnetism generally falls into 4 main categories: diamagnetism, paramagnetism, ferromagnetism (and anti-ferromagnetism), and ferrimagnetism. We will be concerned only about the last category but all four categories are discussed briefly next.

Diamagnetism

Diamagnetism exists in all elements and arises from the orbital motion of electrons, which produces an extremely small and negative susceptibility. Diamagnetism is a negligibly weak effect.

Paramagnetism

Unlike Diamgnetism, Paramagnetism arises from the spin motion of electrons because Paramagnetic materials have unpaired electrons which are free to align with an externally applied magnetic field giving rise to a small and positive susceptibility. Because the electron has a charge, its spin about its own axis produces a magnetic moment. However, the vector summation of these magnetic moments in Paramagnetic materials is zero even with an externally applied fields, therefore magnetic characteristics can only be observed at very low temperatures.

Ferromagnetism

In Ferromagnetic materials, unpaired electrons also exist producing a total magnetic moment, which can be observed even without any applied magnetic field. The reason of the strong magnetic characteristics of Ferromagnetic materials is the existence of domains, which are atoms with parallel magnetic moments. Examples of Ferromagnetic materials are Iron and Cobalt.

Ferrimagnetism

Ferrimagnetic materials have similar characteristics to ferromagnetic materials. However, they have different magnetic ordering. Examples of Ferrimagnetic materilas are Ferrites and Garnets.

Ferrites

Ferrites are Ferrimagnetic materials but they have much higher electrical resistivities making them suitable for microwave applications. They are ceramic like materials with resistivities around 10^{14} more than metals and permittivities around 15. They are made by cintering a mixture of metallic oxides and they all have the common chemical composition of MO. Fe₂O₃, where M is a divalent metal such as manganese, magnesium, iron, zinc etc.

6.2.1 Derivation of the Tensor Permeability

Because ferrites are anisotropic materials, their permeabilities are expressed in the form of matrices, which are known as tensor permeabilities. The elements of the tensor permeability can be calculated from a simple microscopic model [72] [73]. As previously stated, the magnetic effects in ferrimagnetics are associated with the spin movement of the electrons. The electron can be described by a number of properties including its charge $-e = -1.602 \times 10^{-19}$ C, its mass $w = 9.107 \times 10^{-31}$ Kg, and its angular momentum \vec{J} which is equals in magnitude to $\frac{1}{2}\hbar$ or 0.527×10^{-34} J . s, where \hbar is Planck's constant divided by 2π . The magnetic dipole moment of the electron (m) equals to one Bohr magnetic dipole is referred to as the gyromagnetic ratio γ which is

$$\gamma = \frac{m}{J}.\tag{6.1}$$

If the electron undergoes a direct magnetic field $(\vec{B_0})$, torque (\vec{T}) will be exerted on the electron which is given by

$$\vec{T} = \vec{m} \times \vec{B_0} = \mu_0 \vec{m} \times \vec{H_0}. \tag{6.2}$$

This torque causes the dipole moment to precess about an axis parallel to $\vec{B_0}$. The rate of change of the angular momentum gives the torque or

$$\frac{d\vec{J}}{dt} = \vec{T} = \mu_0 \vec{m} \times \vec{H_0}$$
(6.3)

or

$$\frac{d\vec{m}}{dt} = -\mu_0 \gamma \vec{m} \times \vec{H_0},\tag{6.4}$$

which is the equation of motion for \vec{m} and the negative sign has been included because \vec{m} and \vec{J} have opposite directions. This equation can be written in terms of its rectangular coordinate components as follows:

$$\frac{dm_x}{dt} = -\mu_0 \gamma m_y H_0 \tag{6.5a}$$

$$\frac{dm_y}{dt} = -\mu_0 \gamma m_x H_0 \tag{6.5b}$$

$$\frac{dm_z}{dt} = 0. \tag{6.5c}$$

Equation (6.5) can be used to obtain equations on m_x only and m_y only as follows:

$$\frac{d^2 m_x}{dt^2} + \omega_0^2 m_x = 0 ag{6.6a}$$

$$\frac{d^2m_y}{dt^2} + \omega^2 m_y = 0, \tag{6.6b}$$

where

$$\omega_0 = \mu_0 \gamma H_0 \tag{6.7}$$

is called the Larmour frequency. Microwave signals at this frequency are largely attenuated as the ferrite material resonates and the RF energy is absorbed by the magnetic dipoles. It is worth mentioning that mks units have been assumed. If Gaussian units are used, μ_0 can be omitted because it has a numerical value of unity in the Gaussian units.

If we assume that there are N electron spins per unit volume the magnetization can be defined as

$$\dot{M} = N\vec{m} \tag{6.8}$$

and consequently the equation of motion can be written as

$$\frac{d\vec{M}}{dt} = -\mu_0 \gamma \vec{M} \times \vec{H}.$$
(6.9)

 \vec{H} in the previous equation is the internal applied magnetic field.

Now the analysis for the case of a small RF magnetic signal superimposed on the dc magnetic signal can be performed. We will assume that the static magnetic field is z-directed and that the magnitude of the RF signal is small compared to the static magnetic field. The total magnetic field can then be written as:

$$\vec{H_t} = H_0 \,\hat{z} + \vec{H}$$
 (6.10)

where H_0 is the magnitude of the dc magnetic field and \vec{H} is the RF magnetic field which is assumed to be small $(H \ll H_0)$. The total magnetization inside the ferrite material will then be:

$$\vec{M}_t = M_s \,\hat{z} \quad + \quad \vec{M} \tag{6.11}$$

where M_s is the saturation magnetization of the material and \vec{M} is the additional magnetization due to the applied field. Substituting (6.10) and (6.11) in (6.9) gives:

$$\frac{dM_x}{dt} = -\mu_0 \gamma M_y (H_0 + H_z) + \mu_0 \gamma (M_s + M_z) H_y, \qquad (6.12a)$$

$$\frac{dM_y}{dt} = \mu_0 \gamma M_x (H_0 + H_z) - \mu_0 \gamma (M_s + M_z) H_x, \qquad (6.12b)$$

$$\frac{dM_z}{dt} = \mu_0 \gamma M_x H_y + \mu_0 \gamma M_y H_x.$$
(6.12c)

Since a small RF signal is assumed, the higher powers of M and H can be ignored reducing (6.12) to:

$$\frac{dM_x}{dt} = -\omega_0 M_y + \omega_m H_y, \qquad (6.13a)$$

$$\frac{dM_y}{dt} = \omega_0 M_x - \omega_m H_x, \qquad (6.13b)$$

$$\frac{dM_z}{dt} = 0. \tag{6.13c}$$

where $\omega_0 = \mu_0 \gamma H_0$ and $\omega_m = \mu_0 \gamma M_s$, which is the ferromagnetic (FMR) frequency. Microwave signals at or near this frequency are highly attenuated. Equation (6.13) can be solved to give the following equations:

$$\frac{d^2 M_x}{dt^2} + \omega_0^2 M_x = \omega_m \frac{dH_y}{dt} + \omega_0 \omega_m H_x$$
(6.14a)

$$\frac{d^2 M_y}{dt^2} + \omega_0^2 M_y = -\omega_m \frac{dH_x}{dt} + \omega_0 \omega_m H_y$$
(6.14b)

which are the equations of motions for the saturated case with small RF signal. If a sinusoidal time dependence is assumed for the RF signal, the RF steady-state form of (6.14) reduces to the following phasor equations:

$$(\omega_0^2 - \omega^2)M_x = \omega_0 \omega_m H_x + j\omega \omega_m H_y \tag{6.15a}$$

$$(\omega_0^2 - \omega^2)M_y = -j\omega\omega_m H_x + \omega_0\omega_m H_y$$
(6.15b)

which can be written in a matrix form as follows:

$$\vec{M} = [\chi]\vec{H} = \begin{bmatrix} \chi_{xx} & \chi_{xy} & 0\\ \chi_{yx} & \chi_{yy} & 0\\ 0 & 0 & 0 \end{bmatrix} \vec{H}$$
(6.16)

where the elements of the susceptibility matrix ([χ]) are given by:

$$\chi_{xx} = \chi_{yy} = \frac{\omega_o \omega_m}{\omega_0^2 - \omega_2} \tag{6.17a}$$

$$\chi_{xx} = -\chi_{yy} = \frac{j\omega\omega_m}{\omega_0^2 - \omega_2}.$$
(6.17b)

The magnetic field and the magnetic flux density can be related by:

$$\vec{B} = \mu_0(\vec{M} + \vec{H}) = [\mu]\vec{H},$$
(6.18)

where the tensor permeability is given by:

$$[\mu] = \mu_0([U] + [\chi]) = \begin{bmatrix} \mu & j\kappa & 0 \\ -j\kappa & \mu & 0 \\ 0 & 0 & \mu_0 \end{bmatrix}$$
(6.19)

with elements defined as follows:

$$\mu = \mu_0 (1 + \chi_{xx}) = \mu_0 \left(1 + \frac{\omega_0 \omega_m}{\omega_0^2 - \omega^2} \right)$$
(6.20a)

$$\kappa = -j\mu_0 \chi_{xy} = \mu_0 \frac{\omega \omega_m}{\omega_0^2 - \omega^2}$$
(6.20b)

Effects of losses

The tensor permeabilities presented so far are simplified because they do not take into account various practical effects such as damping, magnetic anisotropy and de-magnetization. The damping can be represented mathematically in the equation of motion in the Landau-Lifshitz form as follows [72]:

$$\frac{d\vec{M}}{dt} = -\mu_0 \gamma \vec{M} \times \vec{H} + \lambda \left[\frac{(\vec{H}.\vec{M})\vec{M}}{M^2} - \vec{H} \right]$$
(6.21)

where λ is a damping factor which is the inverse of the relaxation time associated with the motion of the magnetic dipoles. The elements of the tensor permeability can be updated based on (6.21) to include the losses as follows:

$$\mu_{re} = 1 + \frac{\omega_M T^2 \omega_0 [(\omega_0 T)^2 - (\omega T)^2 + 1]}{\Delta}$$
(6.22a)

$$\mu_{im} = \frac{\omega_M T[(\omega_0 T)^2 + (\omega T)^2 + 1]}{\Delta}$$
(6.22b)

$$\kappa_{re} = \frac{\omega \omega_M T^2 [(\omega_0 T)^2 - (\omega T)^2 - 1]}{\Delta}$$
(6.22c)

(6.22d)

$$\kappa_{im} = -\frac{2\omega\omega_M\omega_0 T^3}{\Delta} \tag{6.22e}$$

$$\Delta = \left[(\omega_0 T)^2 - (\omega T)^2 - 1 \right]^2 + 4(\omega_0 T)^2$$
(6.22f)

$$T = \frac{\gamma}{\omega\lambda} = \frac{2}{\gamma\Delta H} \tag{6.22g}$$



Figure 6.2: Illustration of the demagnetization effect

where the re and im subscripts refer to the real and imaginary parts, respectively. As expected, when the losses are included the elements of the tensor permeability become complex. The losses of ferrite materials is commonly represented as ΔH which is referred to as the magnetic line width of the ferrite material. ΔH is directly related to T from (6.22g).

Effects of Demagnetization

The previous formulas for the tensor permeability all depends on the *internal* static field. This field equals the *applied* field only for ideal and infinite samples. For practical cases, however, the internal magnetic field will neither be uniform nor equals to the applied field. Uniformity can only be assumed if the sample is small compared to the wavelength, or is lossy. To relate the internal field to the applied field, the demagnetization has to be taken into account. The demagnetization arises from the discontinuity of the magnetic lines at the interface between the sample and the air. When a homogeneous specimen is placed in a uniform magnetic field, it will become polarized. The magnetic dipoles at the edge create a magnetic field opposing the applied one as shown in Fig. 6.2. The internal magnetic field will then be given by:

$$H_i = H_0 - N(4\pi M) \tag{6.23}$$

where the opposing internal field $(N4\pi M)$ is proportional to the magnetization (M) with Nbeing the constant of proportionality. N is a vector with the condition $N_x + N_y + N_z = 1$, from which it can be deduced that for a sphere specimen $N_x = N_y = N_z = \frac{1}{3}$. If the following updated values for the dc and RF magnetic fields are used:

$$\vec{H}_{0i} = \vec{H}_0 - 4\pi (\vec{N}.\vec{M}_s) \tag{6.24a}$$

$$\vec{H}_i = \vec{H} - 4\pi(\vec{N}.\vec{M})$$
 (6.24b)

to replace the ones in (6.10), the following values for the tensor permeability are reached:

$$\mu_{xx}^{eff} = 1 + \frac{\omega_M \left[(\omega_0 + j/T) + N_y \omega_M \right]}{\omega_r^2 - \omega^2 + (2j/T) \left\{ \omega_0 + \left[(N_x - N_y)/2 \right] \omega_M \right\}}$$
(6.25a)

$$\mu_{yy}^{eff} = 1 + \frac{\omega_M \left[(\omega_0 + j/T) + N_x \omega_M \right]}{\omega_r^2 - \omega^2 + (2j/T) \left\{ \omega_0 + \left[(N_x - N_y)/2 \right] \omega_M \right\}}$$
(6.25b)

$$\kappa_{yx}^{eff} = -\kappa_{xy}^{eff} = \frac{j\omega\omega_M}{\omega_r^2 - \omega^2 + (2j/T) \left\{\omega_0 + \left[(N_x - N_y)/2\right]\omega_M\right\}}$$
(6.25c)

where $\omega_0 = \gamma (H_0 - N_x \omega_M)$ and ω_r is the updated resonant frequency given by Kittel as [74]:

$$\omega_r = \sqrt{\left[\gamma H_0 + (N_x - N_z)\omega_M\right] \left[\gamma H_0 + (N_y - N_x)\omega_M\right]} \tag{6.26}$$

and H_0 is the externally applied magnetic field.

Other effects

Aside from the demagnetization, in practice, the magnetic material is not usually homogeneous due to the existence of preferred magnetic directions. This effect is known as magnetic anisotropy and causes the internal magnetic field to deviate from the externally applied one. Another effect arises from the polycrystalline nature of most of the commercially available ferrites. As compared to a single crystal specimen, a polycrystalline specimen has much more losses which are reflected in a wider absorption line. However, the fabrication and handling of single-crystal or crystallized ferrites are hard, therefore polycrystalline ferrites are commonly used.

Tensor Permeability for Partially Magnetized Ferrites

The values for μ and κ defined so far are all derived for saturated ferrite substrates. Therefore, they do not give accurate results for partially magnetized substrates. For such cases, the following empirical formulas introduced by Rado, Green and Sandy [75], [76] can be used instead:

$$\mu' = \mu'_0 + (1 - \mu'_0) \left(\frac{M}{M_s}\right)^{\frac{3}{2}},$$
(6.27a)

$$\mu_0' = \mu_0 \left(\frac{1}{3} + \frac{2}{3} \left(1 - \left(\frac{\gamma 4\pi M_s}{\omega} \right)^2 \right)^{\frac{1}{2}} \right), \qquad (6.27b)$$

$$\kappa' = \gamma 4\pi M/\omega, \tag{6.27c}$$

where the accent is used for the partially magnetized case. The tensor permeability for this case is given by:

$$[\mu'] = \begin{bmatrix} \mu' & j\kappa' & 0 \\ -j\kappa' & \mu' & 0 \\ 0 & 0 & \mu'_z \end{bmatrix}$$
(6.28)

where μ'_z is calculated by

-

$$\mu_z' = \mu_0^{\prime \left[1 - \frac{M}{M_s}\right]^{\frac{5}{2}}}.$$
(6.29)

-

It is worth mentioning here that the elements in equation (6.27) are empirical formulas and no physical bases are claimed for any of them.

The tensor permeability in (6.28) is derived for a z-directed bias field. The tensor for an arbitrarily directed field as shown in Fig. 6.3 is driven in [77] and is given by:

These updated values can be used with any of the formulas presented in the previous sections for the saturated case.

6.2.2 Microstrip Resonators on Ferrite substrates

In this section, the tunability of a microstrip resonator implemented on a ferrite substrate is presented. In this case, the resonator is a half-wave microstrip line, which is the basic building block for the edge-coupled bandpass filter discussed later. Therefore, the theoretical results presented here are compared with the measurements and electromagnetic (EM) simulations



Figure 6.3: Coordinate system and angles for the tensor permeability of arbitrary bias direction of (6.30)

later in this chapter.

Microstrip lines on isotropic materials can be synthesized from the well-known empirical formulas introduced by Wheeler [78–80] and improved by Hammerstad and Jensen [81]. However, for anisotropic substrates such as ferrites where the permeability is described in the form of a tensor, these formulas need to be updated. Since an analytical solution for the microstrip problem does not exist, using the tensor permeability directly is not possible. Instead, an effective permeability μ_{eff} can be defined that take into account the magnetization state of the substrate. Once μ_{eff} is defined, the substrate can be assumed both isotropic as well as homogeneous and the new permeability can be used to calculate the various characteristics of the microstrip line.

For the biasing of the ferrite substrates, two cases arise, which are illustrated in Fig. 6.4. In the first case, the magneto-static field is applied parallel to the direction of propagation of the microstrip mode. For the second case, the magneto-static field is applied transverse to the direction of propagation of the microstrip mode. For the parallel case, the effective permeability of the ferrite material is observed to decrease with increased bias resulting in an



Figure 6.4: Different orientations of the magneto-static bias field applied on the microstrip substrate.

upwards frequency shift [11,82]. For the transverse case, however, the effective permeability increases with the bias resulting in a downwards frequency shift [10,71]. In this work we are only interested in the transverse case because the embedded windings produce transverse magneto-static fields. For this case the formulas for the effective permeability in partially magnetized substrates by Pucel and Massé [83,84] can be used. For the case where no magnetic field is applied, (6.27 b) can be used to obtain the unbiased effective permeability. As the magneto-static field is applied, however, the following formula can be used to estimate the effective permeability of the magnetized material [84]:

$$\mu_{\text{eff}} = \frac{\mu_2' - \kappa'^2}{\mu'} \quad \frac{1}{1 - \frac{1}{7}\sqrt{\frac{h}{w}} \left[\frac{\kappa'}{\mu'}\right]^2 \ln\left(1 + \frac{\mu'}{\mu'^2 - \kappa'^2}\right)},$$
(6.31)

where μ' and κ' are the tensor elements for the partially magnetized case, which have been defined in (6.27), h/w is the ratio of the height of the substrate to the width of the resonator line. Equation (6.31) above is again an empirical formula with no physical basis claimed. The effective permeability for the case of ferrite LTCC at 15 GHz is shown in Fig. 6.5 in relation to the normalized magnetization, where (6.31) has been used with M_s of 4000 Gauss and a FMR frequency of 11.2 GHz.



Figure 6.5: Theoretical values for the effective permeability as well as the resonant frequency for the microstrip line on a partially magnetized substrate(a) Effective permeability from (6.31) (b) Calculated resonant frequency.

6.2.3 Effects of Heat on Ferrite Materials

It is well known that the saturation magnetization of magnetic materials, M_s , depends on the temperature. As the temperature increases, M_s decreases until it reaches zero at the Curie temperature (T_C), at which the material becomes completely non-magnetic (Paramagnetic) [85]. This temperature dependence can be described mathematically by Bloch's law in the form of a simple formula given by [86]



Figure 6.6: Theoretical results of the effects of the temperature on μ_{eff} and the center frequency of a microstrip resonator. (a) Center frequency (b) μ_{eff} .

$$M_s(T) = M_s(0) \left(1 - \left(\frac{T}{T_C}\right)^{\frac{3}{2}} \right),$$
 (6.32)

where $M_s(0)$ is the saturation magnetization when the temperature is at 0° K. The decrease in M_s with temperature affects the effective permeability of the substrate and, thus, resonant frequencies of filters implemented on it. To illustrate the effects of heat on μ_{eff} and the center frequency of a microstrip resonator, equations (6.32) and (6.31) can be used to give μ_{eff} and the center frequency as functions of the magnetization level M/M_s and the normalized temperature T/T_c . Plots of μ_{eff} and the center frequency are given in Fig. 6.6 (a) and (b), respectively. As the temperature increases the resonant frequency decreases and the tunability also decreases until it vanishes completely at T_c .

6.3 Bandpass Filter Design

As a proof of concept, an edge-coupled bandpass filter has been designed for the ferrite-LTCC substrate. The complete filter design is illustrated in Fig. 6.7. The design of the filter consists of two main parts: the RF filter, and the dc windings. The RF part consists of the filter, which is a second order edge-coupled design in microstrip configuration. The dc part consists of the RF windings and is located below the ground plane of the RF circuit. Since the ground plane is a good conductor, the tangential components of the RF electric field become zero beyond the skin depth ($1\mu m$ in this case). Therefore, the RF electric and magnetic fields are not affected by any structure beyond the ground plane. For the case of the dc magnetic field generated by the windings, no electric field exists because the magnetic field is static. Therefore, the ground plane does not pose any barrier for the dc field, which exists on the RF substrate located above the windings. Such structure enables the independent design of the RF and dc parts of the filter, which simplifies the design process considerably.



Figure 6.7: An illustration of the bandpass filter with embedded windings.

6.3.1 RF edge-coupled Filter Design

The design of the edge-coupled filter has been performed as detailed in chapter 3, which starts by identifying the center frequency and the fractional bandwidth. Since this design is a proof of concept, no application-specific characteristics exist, therefore the center frequency and FBW can be selected arbitrarily. A FBW of 6 % has been selected because it is narrow enough to avoid the dispersive effect of the ferrite substrate on the performance of the passband. This FBW is also not too narrow to increase the insertion loss of the filter. The selection of the FBW has been verified by EM simulations. The criteria for selecting the center frequency is discussed next.



Figure 6.8: The total achievable tunability for a microstrip line in relation to the resonant frequency, which is normalized to the FMR frequency.

Selection of the center frequency

As shown in the previous section, ferrite materials *resonate* at the frequency of the precession of the magnetic dipoles, which is referred to as the ferromagnetic resonance (FMR) frequency. The FMR can be estimated by Kittel's equation (6.26) [74]. In this case the demagnetization factors N_x , N_y and N_z can be ignored since the magnetic field is generated internally by the windings. From which Kittel's equation is reduced to

$$\omega_r = \omega_M = \gamma 4\pi M_s. \tag{6.33}$$

If γ is taken as 2.8MHz/Gauss and the $4\pi M_s$ for the ferrite LTCC is 4000 Gauss the above equation gives the FMR directly in units of MHz which is 11200 MHz (or 11.2 GHz). For RF frequencies at or near this frequency, the ferrite material absorbs the RF energy with quality of absorption depending on the line width (Δ H) of the material. Therefore, bandpass filters with passbands near the FMR frequency exhibit high insertion losses. Below the FMR frequency the ferrite material typically has high losses, which are known as low field losses. Therefore, for optimal loss performance, the filter should be designed for a frequency as high as possible from the FMR frequency.

From (6.27b) it can be concluded that the permeability range, which is proportional with

the total achievable tunability, depends on the frequency of operation. The total tunability of a microstrip line (deduced from (6.31)) is plotted in Fig. 6.8 in relation to the normalized frequency. It can be seen that most of the tunability is achieved in the vicinity of the FMR frequency. At 3 times the FMR frequency hardly any tunability is achieved. From the above discussion it is clear that for the design of the bandpass filter to be optimized for low insertion loss, the operating frequency should be larger than the FMR frequency and as far from it as possible. However, in order to optimize the design for the largest possible tunability, the operating frequency should be chosen as close to the FMR frequency as possible. In light of these conclusions, the operating frequency has been selected to be 15 GHz which is about 1.5 times the FMR frequency. It is worth mentioning that equation (6.31)), which is used to produce the data of Fig. 6.8 has a singularity at the FMR frequency, therefore, the values of the elements of the tensor permeability can not be obtained at the FMR using this equation.

Design and simulation of the Filter for the unbiased case

The ferrite LTCC tape used in this work is the ESL 40012 with a dielectric constant of 13.5 and an electric loss tangent of 0.004 [54]. The magnetic permeability is calculated from (6.27b) to be 0.65 at 15 GHz. Using these values and the filter specifications of center frequency of 15 GHz and FBW of 6%, a second order Chebyshev design is performed. For the microstrip resonators, a hairpin structure has been selected instead of the strait line. The use of hairpin resonators results in a compact design with a lower planar aspect ratio (more square design). This last characteristic is very important because it enables the construction of square bias windings below the filter, which produces a more uniform magnetic field as compared to a rectangular structures. The design is performed as detailed in chapter 3 [58, 59]. Next it is optimized using EM simulations (CST microwave studio). The EM results of the filter are shown in Fig. 6.9. Due to the electric loss tangent of the ferrite material, an insertion loss of about 2.3 dBs can be observed in the pass-band. It is worth mentioning here that this simulation model takes into account the dispersive nature of the

magnetic material because the material is defined as a tensor material in the EM simulator. Instead of simulating with a constant permeability, equation (6.27) is used in the simulation to model the frequency dependence of the permeability.

It is worth mentioning here that ten layers of LTCC have been provided for fabrication. Some of these layers need to be assigned to the RF filter and the rest to the windings. The number of layers assigned to the RF filter defines the substrate thickness for the microstrip structures. Coupled microstrip lines of thicker substrates are wider and have larger spacings which are good characteristics for fabrication. However a thicker RF substrate leaves fewer layers for the windings, which results in lower number of turns. A lower-turn coil requires more current to produce the same magneto-static field as compared to a higher-turn coil. Therefore, a trade-off between fabrication tolerance and dc current requirement arises. In this work after doing EM simulations, it has been found that four layers for the RF filter provides reasonable dimensions, which leaves six layers for the windings as shown in Fig. 6.7.



Figure 6.9: EM simulations of the hairpin bandpass filter.

6.3.2 Magneto-static simulations of the windings

The magnetic bias is provided by a solenoid winding composed of nine turns in each layer and occupies six layers in total as shown in Fig. 6.7. The electro-static simulations of these windings (shown in Fig. 6.10(b)) shows that the generated magnetic field is not uniform. Therefore, careful simulation of this non-uniform field needs to be performed. To do this, the RF substrate, which is the area of interest, is divided or discretized into sub-layers. The RF simulations have revealed that a total of four discretized layers is sufficient to accurately simulate the non-uniform magnetic field. For each sub-layer, magnetio-static simulations have been performed for different dc-current values. The magnetic fields are then averaged over center lines extending along the substrate parallel to X, Y and Z. This resulted in three average values for the magneto-static fields for each layer. These values are then used as vector components to define the magnitude and direction of the magneto-static field on each layer. This magneto-static field is assumed to exist uniformly throughout each layer. The results of this process, which is shown in Fig. 6.11, have revealed that for the bottom most layer (layer seven) the magnetic fields are mainly z-directed. The X and Y components of the magneto-static fields for this layer are negligibly small. Therefore, for simplicity, the average field on this layer is assumed to be z-directed. The magnitude of the magneto-static field of layer 7 is shown in Fig. 6.10, where both the magnetic field as well as the magnetization are plotted. For a total bias current of 300 mA, the magnetization can be increased from zero to about 3500 Gauss which is about 88% of magnetization. The magnetic field required for such magnetization is about 12 Oe only. Such low field requirements is possible because of the embedded windings, where the demagnetization is reduced to a negligible value.

For the next layer up (layer eight) the magneto-static fields are also z-directed but the magnitude of the magnetization is 70% the magnitude for layer seven. For the top two layers (layers nine and ten) the magneto-static fields in the z direction have negligibly small magnitudes. Therefore, for these two layers, the fields are assumed to have a direction tangential to the filter (in the XY plane). The simulations also revealed that the X and Y



Figure 6.10: Magneto-static simulation results. (a) Average magneto-static simulation of layer 7. (b) 3D illustration of the magneto-static simulation of the windings.

components of the fields are almost equal and are 60% the value of layer seven for the case of layer nine and 50% the value of layer seven for the case of layer ten (the top most layer). These results are illustrated in Fig. 6.11 and will be used in the following section in the EM simulation of the filter.

6.3.3 EM Simulation of the Filter

In this section the EM simulation of the filter is presented. As explained in the previous section and Fig. 6.11, the RF substrate is sub-divided into four substrates. Each of these substrates is defined as a separate material in the EM simulator (CST EM Studio). These materials are defied as tensor permeability materials, which enables the simulation of the frequency and magnetization dependences. For the two bottom layers (seven and eight), where the magnetic field is z-directed, the standard tensor permeability of (6.28) with elements defined in (6.27) for the partially magnetized case is used. For the two top layers (nine and ten), where the magnetic field is in the X-Y and 45° with the X (or Y) axis, the tensor permeability of (6.30) is used with θ =90° and ϕ =45°. The effect of the bias current is reflected in the simulation by making the magnetization (M) a variable, which is related



Figure 6.11: Illustration of the magnitudes and directions of the magneto-static fields in the four discretized substrates.

to the dc bias current from the electromagnetic simulation. This way, full EM simulation results are obtained for every bias current which are shown in Fig. 6.12.



Figure 6.12: Electromagnetic simulations of the bandpass filter using the tensor permeability. (a) $|S_{21}|$ (b) $|S_{11}|$.

6.3.4 Measurement and Discussions

A prototype of the bandpass filter has been fabricated by the technical research center of Finland (VTT) which is shown in Fig. 6.13. The design module has a total volume of $5 \times 5 \times 1.1 \text{ mm}^3$ which is several orders of magnitude smaller than the conventional ferritebased filters if the external electromagnet used to bias them is taken into account. A network analyser (Anritsu ME7828) with $150 \mu m$ pitch size probes has been used to perform the measurement of the filter. Two dc probes have been used to fed the bias current from a dc current source. The measured and simulated results are shown in Fig. 6.14 as compared to the simulations presented in the previous section. It can be seen that the filter achieves about 4% of tunability for a bias current of 280 mA. Above this value further tunability is possible; however, the insertion loss starts to deteriorate. An insertion loss of 2.3 dBs is achieved, which has been almost constant through out the tuning range. It can also be observed that the measurement and simulation agree very well at low bias currents. When the bias current is increased, however, the measurement shows more tunability that predicted by the simulation. This is attributed to the effects of heating which is treated in details in the text section. The change of the FBW with the bias current is illustrated in Fig. 6.15, where it is clearly seen that the FBW stays almost constant throughout the full tuning range. Close agreement between the measurements and the simulation can also be observed.



Figure 6.13: A photograph of the fabricated tunable filter.



Figure 6.14: Measured and simulated results of the tunable filter (a) $|S_{21}|$ (b) $|S_{11}|$.

6.3.5 Comparison between this filter and the literature

In Table 6.1, a comparison between this filter and other reported designs is presented. It can be concluded that this filter is considerably smaller than all the other designs because it is the only one with the embedded bias windings. Due to this, the fields lost at the air-ferrite interface (demagnetization effects) are minimized; therefore this filter provides the highest tunability per unit field which is 5 times greater than the second largest reported in [71]. Because of its compactness, this filter can be directly integrated within small and portable wireless systems unlike the other designs.

6.4 The effects of self-heating

Since the windings used to generate the magneto-static fields are densely packed inside the LTCC substrate, heat will be generated when a dc current is passed through them. This heat affects the characteristics of the ferrite substrate resulting in more downwards frequency shift as discussed in section 6.2.3. To verify this more, Fig. 6.16 shows the initial measured and simulated $|S_{21}|$ results of the bandpass filter for bias currents of 0 mA and 300 mA. As



Figure 6.15: The simulated and measured FBW of the filter in relation to the dc bias current.

Ref.	\mathbf{f}_0	IL	Bias Fields	Tuning per Oe	\mathbf{Size}^*	Integrable			
[71]	6.2	1.25	100 Oe	0.06%	Large	No			
[11]	9	1.4	200 Oe	0.04%	Large	No			
[51]	4	5	3160 Oe	0.03%	Large	No			
[52]	19.5	5	1400 Oe	0.005%	Large	No			
This Work	15	2.3	12 Oe	0.33%	Small	Yes			
* This comparison considers the overall size with bias windings.									

Table 6.1: Comparison between this work and the literature.

the substrate is biased with 300 mA, the simulation predicts much less tunability than the amount measured. This difference is mainly due to the effects of the heat generated by the windings. These effects are considerable as is clear from the figure and should be considered in the design. It can also be observed that at 300 mA the insertion loss increases slightly (0.4 dBs more than the unbiased state). This is because the operating frequency gets closer to the FMR of the material (FMR=11.2 GHz). In the following sections, we characterize the self heating effects and update the simulation model to take into account these effects. The new simulation results match the measurement very well.



Figure 6.16: Highlight of the frequency shift caused by the internally generated heat.

6.4.1 Measurement of the Joule heating in the ferrite substrate

Measurements of the filter presented in the previous section revealed higher tunabilities than predicted by simulations especially at higher bias currents. This discrepancy has been attributed to self-heating effects. To validate this assumption, first the measurement of the Joule heating is performed by a digital thermometer (Omega HH501 DK). The tip of the probe (thermocouple) of the thermometer is fixed on the top surface of the ferrite substrate by a teflon tape. As the dc current is passed through the windings, the rise in temperature is recorded in relation to time as shown in Fig. 6.17(a). Because the thermocouple is on the top surface of the substrate, the measured temperature is expected to be slightly less that the actual temperature inside the substrate. This measurement shows that the temperature rises rapidly and reaches about $240^{\circ}C$ in less than one minute for the case of 240 mA, which confirms that the amount of heat generated is considerable. It can also be observed that the temperature reaches about 90 % of its steady state value 40 to 50 seconds after the dc current source is switched on. The time for the substrate to reach its thermal steady state can be assumed to be 50 seconds.

The steady-state temperature (recorded 50 seconds after the application of the current) is plotted in relation to the current in Fig. 6.17(b). The temperature increases with the



Figure 6.17: Measurement of the temperature rise in the ferrite substrate. (a) Temperature rise on top of the LTCC substrate in relation to the time for different values of the dc current (b) Measured temperature in relation to the dc current. The measurement is taken on the surface of the ferrite substrate 50 seconds or more after the dc current is applied.

current following a power law. This behaviour is in accordance with Joule's first law, which states that the energy generated by passing a dc current through a conductor is directly proportional to the resistance of the conductor and the square of the current. It can be seen that the temperature exceeds $250^{\circ}C$ for currents greater than 250 mA. In the following section, the tunability due to self heating is isolated from the one due to magnetic biasing.

6.4.2 Verification of the effects of self-heating on the performance of the filter

To verify the effects of self-heating on the performance of the filter it is important to isolate the tunability generated by the heat from that generated by the magnetic bias of the substrate. To remove the temperature effect, measurements are taken when the substrate is at room temperature. To achieve this, the measurements must be taken quickly before the substrate is heated. In Fig. 6.18 $|S_{21}|$ for two different measurement sets are shown. The first measurement (case (I)) has been performed immediately (< 5 seconds) after the current source has been turned on. In this case, the temperature of the substrate is assumed to remain close to room temperature during the measurement. The second measurement (case (II)) has been performed 50 seconds or more after the current source has been turned on. The temperature in this case is assumed to reach its steady state value. As expected, a considerable difference between the two measurement cases is clear, especially at the higher current. Since the alignment of the magnetic dipoles responsible for the magnetic tuning occurs in the order of μs [50], the difference between these two measurements can be attributed mainly to the increase in temperature. In the following section the tunability due to only heating the substrate (no bias) is investigated.



Figure 6.18: Measured $|S_{21}|$ for a dc current bias of 120 mA as well as 280 mA. Two different measurement cases are plotted: case (I), where the measurement has been taken less than 5 seconds and case (II), where the measurement has been taken 50 seconds or more after the dc current is applied.

6.4.3 Measurement of the effects of external-heating on the tunability of the filter

The effects of increased temperature are investigated in this section. The chuck of the probe-station (Cascade Summit 12000) is heated from $0^{\circ}C$ to $190^{\circ}C$ using a thermal system (ESPEC, ETC-200L Thermal System) and the S-parameters are measured with no bias current through the windings. The results are shown in Fig. 6.19, where a frequency shift of about 1 GHz is observed for the total temperature range, which is equivalent to a temperature



Figure 6.19: Measured $|S_{21}|$ of the filter for chuck temperatures of $25^{\circ}C$, $50^{\circ}C$, $100^{\circ}C$, $150^{\circ}C$ and $190^{\circ}C$. No bias current is applied.

tunability of about (10-13)MHz/° C. It is also clear that most of the change occurs between 0°C and 150° C with very little change occurring above 150°C. It is worth mentioning that the minimum insertion loss of the filter remains almost constant.

The temperature range in the previous measurements is quite large and would not be encountered under normal operation conditions. However, as reported in the previous section, the temperature rise in the substrate due to the Joule heating is considerable. Therefore, the temperature effects should be considered in the design of the filter. In the following section the simulation of this effect is investigated.

6.4.4 Simulation of the self-heating effect

The simulation model described in section (6.3.3) does not take into account the effects of self-heating. As a consequence, this model does not give accurate simulation results especially at high bias currents as illustrated in Fig. 6.16. Therefore this model needs to be updated to take into account the effects of self-heating.

To include the effects of the temperature, the simulation model illustrated in Fig. 6.20 (b) is introduced. The saturation magnetization as a function of the bias current $(M_s(I_{dc}))$ needs to be determined. In this study, the filter itself is used to characterize the ferrite


Figure 6.20: A block diagram for the simulation strategy of the tunable filter, which takes into account the effect of temperature on M_s .

substrate. The simulation strategy used here can be summarized in five main steps as shown in Fig. 6.20 (b), which are outlined below:

- Step 1: Magneto-static simulations are performed for the windings to obtain the magnetization as a function of the bias current $(M(I_{dc}))$.
- Step 2: The measured S-parameters of Fig. 6.19 for variable temperature values and zero bias current are used as input to the EM simulator. M, is kept at zero while M_s is varied until the center frequency in the simulation and the measurement match. The values of M_s that give the match are recorded in relation to the temperature to yield a discrete relationship between M_s and the temperature (M_s(T)).
- Step 3: Since the data obtained in step 2 above are discrete (shown as red squares in Fig. 6.21), a continuous relation is obtained by curve fitting using the sine function given by:

$$M_s = a \sin(bT + c), \tag{6.34}$$

where T represents the temperature, a, b, and c are constants, which are determined by curve fitting in a least squares error sense. In Fig. 6.21, both the measured data



Figure 6.21: Saturation magnetization (M_s) in relation to the temperature. The data represented by squares are extracted from the measured data, while the data represented by the solid line are obtained by curve fitting.

and the fitting function are plotted.

- Step 4: The measured steady-state temperature of the substrate for each bias current value, which is shown in Fig. 6.17(b), is used with $M_s(T)$, which is obtained from steps 2 and 3, to get a direct relationship between M_s and the bias current $(M_s(I_{dc}))$. The resulting data are plotted in Fig. 6.22 together with M, which is obtained from the magnetostatic simulations.
- Step 5: The simulation model is used with both M and M_s as functions of the bias current which are obtained from steps 1 to 4 above. This way, the S-parameters of the filter can be simulated for any value of the dc current, taking into account both the magnetic effect, which is reflected in M, and the temperature effect of the current, which is reflected in M_s .

The final simulation results are compared with the measurements in Fig. 6.23 which indicate clear agreement. The measured and simulated tunabilities are compared in Fig. 6.24 where it can be observed that the simulation model is able to predict the tunabilities of the filter for both measurement cases, demonstrating that the model can indeed predict the effect of the temperature increase. The small difference between the simulation and the measurement can be attributed to the slight error in the measurement of the temperature,



Figure 6.22: Saturation magnetization (M_s) and magnetization (M) in relation to the bias current.

which has been measured on the top surface rather than inside the substrate. Nevertheless, the data are very close and the simulation model should provide useful information for the design of the filter.



Figure 6.23: Measured and simulated S-parameters of the filter with temperature effects taken into account in the simulation. (a) S_{21} . (b) S_{11}



Figure 6.24: Quick and steady state measurements of the tunability of the bandpass filter as well as the simulations where the temperature effect is taken into account and when it is not. The theory extracted from (6.31) is also plotted.

6.5 Conclusion

In this chapter a tunable bandpass filter has been presented in the ferrite LTCC. To the best of the author's knowledge, this filter demonstrates the first tunable filter on ferrite LTCC for the SoP platform. The problem of large and bulky electromagnets used in magnetically tunable designs has been avoided by implementing windings inside the LTCC substrate, which reduced the size and current requirements. As a result the reported filter has a total size of $(5\times5\times5\times1.1)$ mm³, which is several orders of magnitude smaller than the reported designs. The heat results from passing dc currents in the embedded windings (self-heating) has also been studies and characterized. The heat effect has also been included in the EM simulation by estimating its effect on the saturation magnetization which becomes a variable in the EM simulator. The new EM simulation matches the measurements very well, which verifies the simulation model.

Chapter 7

Conclusion

Miniaturized wireless devices are everywhere nowadays. These devices are outlined by small sizes and multiple functionalities. In this work, bandpass filters for these devices have been presented with size reduction as a major objective. Three main approaches have been followed: (1) Reducing the size of the individual filter to reach miniaturization at a component level. (2) Replacing the switchable bank of filters typically used by a single tunable filter to get miniaturization at a system level. (3) Increase the total number of components integrated in the system by realizing the filters in a 3D SoP platform. To reduce the size of individual filters a lumped components approach has been used. Before implementing the filter, individual lumped components have been investigated with focus on an ultra-thin LCP stack-up. Next, these lumped components have been used to design a miniaturized bandpass filter for the GPS band (1.57 GHz). For the second approach, magnetic tunability has been used in the design of a tunable filter. The relatively new ferrite LTCC technology has been utilized to realize the filter in a compact, 3D SoP platform.

7.1 Lumped components in ultra-thin LCP

In chapter 4, lumped components have been presented in an ultra-thin LCP tack-up, which has four metal layers and is 100 μ m thick. For inductors, reducing the thickness of the substrate is found to have resulted in a reduced Q-factor. To improve the Q-factor, the ground under the inductors have been partially remove, which increases the Q-factor by nearly 100%. For the case of the 3D capacitors, it has been found that reducing the thickness of the stack-up reduces the size because it increases the capacitance per unit area. A new type of capacitor, referred to as VHID, which combines the horizontal and vertical interdigital capacitors, have been analyzed. For the case of LTCC, where the stack-up is relatively thick, it has been found that the proposed VHID capacitor provides higher capacitance as compared to the conventional VID capacitor. For the case of the ultra-thin LCP, on the other hand, the VHID capacitor demonstrates higher SRFs as compared to the VID capacitor. Therefore, the VHID capacitor provides an additional degree of freedom for the designer to deal with the trade-off between the capacitance and the SRF. For the sake of comparison, inductors and capacitors in LTCC have been implemented. It has been found that the ultra-thin LCP inductors have lower quality factors as compared to the LTCC inductors because of the closer proximity of the ground plane. The ultra-thin LCP capacitors, on the other hand, demonstrate higher capacitances per unit area as compared to the LTCC capacitors, which is mainly due to the reduced thickness of the ultra-thin LCP stack-up.

7.2 Lumped based bandpass filters

In chapter 5, the presented lumped components in LTCC and ultra-thin LCP have been used in the design of miniaturized bandpass filters. In addition to the use of lumped components to realize the filter, the size has been reduced further by utilizing the 3D nature of LTCC and LCP and by reducing the order of the filter. A new topology with a series inductor and a parallel capacitor has been used to increase the selectivity and out-of-band rejection of the filter. The proposed topology has three transmission zeros, two of which improve the roll-off factors of the filter at the upper and lower bands, while the third transmission zero increases the out-of-band rejection at the upper band. The value of the series inductor has been manipulated, using a theoretical model, to have a value of an inductance of a via hole. Therefore, the via-hole used to provide the ground path for the resonators has been utilized to provide the series inductor, which has resulted in an improved out-of-band rejection without adding any new component. As a proof of concept, A miniaturized filter in LTCC has been designed with an overall size of $(5 \times 3.8 \times 0.4)$ mm³. The measured results show roll-off factors of 16.7 and 10 dB/100MHz at the upper and lower bands, respectively. Also, an out-of-band rejection of 20 dBs is maintained up to almost three times the center frequency.

The same topology presented above has been realized in the ultra-thin LCP. In addition to the improved selectivity and out-of-band rejection, the size of the LCP filter is an order of magnitude smaller than the reported ones. Moreover, due to the extremely thin stack-up, this filter demonstrates flexibility, which makes it suitable for conformal applications that require small sizes.

7.3 The tunable bandpass filter

In chapter 6, a tunable bandpass filter has been presented in the SoP, which can be used to reduce the total number of filters and, therefore, the size of the system. The ferrite LTCC technology has been utilized to design the tunable filter for the SoP platform. Magnetically tunable filters are typically biased by external electromagnets which are large and inefficient. The current requirement of the electromagnets is usually high because of the loss of magnetic fields at the air-dielectric interface. These limitations have been avoided by using miniaturized 3D windings, which are implemented inside the LTCC substrate; therefore, reduce the losses and minimize the demagnetization. As a result, the presented filter has 4% of tunability with a magnetic field of only 12 Oe which is an order of magnitude less that what has been reported in the literature. Also, the complete filter (including the windings) occupies a size of $(5 \times 5 \times 1.1)$ mm³, which is several orders of magnitude smaller than the reported designs. Such huge miniaturization demonstrates the potential of magnetically tunable filters on the ferrite LTCC technology.

As presented above, the use of embedded windings has resulted in a huge size reduction and a reduced bias current. However, when a dc current is passed through the windings the temperature of the substrate starts to increase due to the Joule heating. The increase in the temperature, in turn, results in an increase in the tunability. The Joule heating has been characterized and studies in chapter 6. It has been found that the temperature rise within the substrate can reach 250°C for a current of 300 mA, which is considerable. Therefore, it has been concluded that the heating effect should be considered in the design phase. For this purpose, a simulation strategy has been developed to estimate the effects of heat. The strategy is based on estimating the temperature dependence of the saturation magnetization M_s , which becomes an input to the simulation model. The new simulations and measurements agree very well which verifies the simulation strategy. This simulation strategy should help in the design of more reliable and stable ferrite LTCC filters.

7.4 Future work

The bandpass filter presented in chapter 5 in the ultra-thin LCP shows higher insertion loss that estimated by the simulations. This increased loss has been attributed to the finishing layer used to protect the copper clad. It has been found from simulations that the finishing layer increases the insertion loss and slightly lowers the center frequency. These effects of the finishing layer have been explained by the conductivity and permeability of the nicked, which is used in this specific finish. Since finishing layers are commonly used with copper PCBs, it is recommended to further investigate their effects and confirm the findings of this work by fabricating and measuring similar structure with and without the finishing layer.

Since ferrite LTCC-based tunable devices have been reported only recently, good characterizations of the ferrite LTCC tapes are required. The few articles reporting ferrite LTCC characterization have not included characteristics such as the switching speeds, cycling and reliability. Therefore, it is recommended to perform further characterizations of the ferrite LTCC material.

The self-heating of the embedded windings studied in chapter 6 is found to be considerable. Two approaches might be used to solve the heating problem; the first approach is to reduce the generated heat. One way to do this is by optimizing the windings for low current and high heat dissipation. To optimize the windings for low current, their resistances should be reduced, which can be done by increasing the width and thickness of the lines. For instance, if the thickness of the metal is increased by a factor of 4 (by increasing the number of screen printing runs), and the width is increased by a factor of 2, the resistance per unit length of the windings will be decreased by a factor of 8. of To optimize the windings for high heat dissipation, the Joule heating of the windings should to be simulated and optimized using a multi-physics simulator. The windings can be connected to a large metallic layer coupled to a heat sink to dissipate the heat. Microfluidic systems, which can be readily implemented in LTCC, might also be used for a better cooling.

In the second approach, the generated heat can be eliminated completely if the ferrite material is operated in the latching mode, where pulses of current are applied to the windings to latch the ferrite substrate on different magnetization states instead of the continuous current. Operation in the latching mode, however, requires the ferrite material to have a high remanent value. The remanence of the ferrite is a material property, which can be controlled and enhanced by material synthesis techniques.



References

- G. Brzezina and L. Roy, "Miniaturized, Lumped-Element Filters for Customized System-on-Package L-Band Receivers," *Components, Packaging and Manufacturing Technology, IEEE Transactions on*, vol. 4, no. 1, pp. 26–36, Jan 2014.
- [2] L. K. Yeung and K.-L. Wu, "A compact second-order LTCC bandpass filter with two finite transmission zeros," vol. 51, no. 2, pp. 337–341, 2003.
- [3] C.-F. Chang and S.-J. Chung, "Bandpass filter of serial configuration with two finite transmission zeros using LTCC technology," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 53, no. 7, pp. 2383–2388, 2005.
- [4] S. Hwang, S. Min, M. Swaminathan, V. Sundaram, and R. Tummala, "Thin-Film High-Rejection Filter Integration in Low-Loss Organic Substrate," *Components, Packaging* and Manufacturing Technology, IEEE Transactions on, vol. 1, no. 8, pp. 1160 –1170, aug. 2011.
- [5] K. Kageyama, K. Saito, H. Murase, H. Utaki, and T. Yamamoto, "Tunable active filters having multilayer structure using LTCC," *Microwave Theory and Techniques*, *IEEE Transactions on*, vol. 49, no. 12, pp. 2421–2424, Dec 2001.
- [6] M. Al-Ahmad, R. Maenner, R. Matz, and P. Russer, "Wide piezoelectric tuning of LTCC bandpass filters," in *Microwave Symposium Digest, 2005 IEEE MTT-S International*, June 2005, pp. 4 pp.–.

- [7] T. Tick, V. Palukuru, M. Komulainen, J. Perantie, and H. Jantunen, "Method for manufacturing embedded variable capacitors in low-temperature cofired ceramic substrate," *Electronics Letters*, vol. 44, no. 2, pp. 94–95, 2008.
- [8] M. Hrovat, J. Holc, S. Drnovšek, D. Belavič, J. Bernard, M. Kosec, L. Golonka, A. Dziedzic, and J. Kita, "Characterization of PZT thick films fired on LTCC substrates," *Journal of Materials Science Letters*, vol. 22, no. 17, pp. 1193–1195, 2003. [Online]. Available: http://dx.doi.org/10.1023/A%3A1025388214955
- [9] M. Tsutsumi, T. Fukusako, and H. Shimasaki, "Magnetically tunable superconductor filters using yttrium iron garnet films," *Magnetics, IEEE Transactions on*, vol. 31, no. 6, pp. 3467–3469, 1995.
- [10] X. Yang, J. Wu, S. Beguhn, T. Nan, Y. Gao, Z. Zhou, and N. Sun, "Tunable Bandpass Filter Using Partially Magnetized Ferrites With High Power Handling Capability," *Microwave and Wireless Components Letters, IEEE*, vol. 23, no. 4, pp. 184–186, 2013.
- [11] S. Gillette, A. Geiler, Z. Chen, Y. Chen, T. Arruda, C. Xie, L. Wang, X. Zhu, M. Liu, S. Mukerjee, C. Vittoria, and V. Harris, "Active tuning of a microstrip hairpin-line microwave bandpass filter on a polycrystalline yttrium iron garnet substrate using small magnetic fields," *Journal of Applied Physics*, vol. 109, no. 7, pp. 07A513–07A513–3, 2011.
- [12] A. Tatarenko, V. Gheevarughese, and G. Srinivasan, "Magnetoelectric microwave bandpass filter," *Electronics Letters*, vol. 42, no. 9, pp. 540–541, 2006.
- [13] G.-M. Yang, O. Obi, G. Wen, and N. Sun, "Design of Tunable Bandpass Filters With Ferrite Sandwich Materials by Using a Piezoelectric Transducer," *Magnetics, IEEE Transactions on*, vol. 47, no. 10, pp. 3732–3735, 2011.
- [14] R. Tummala and M. Swaminathan, System on Package: Miniaturization of the Entire System. McGraw-Hill, 2008.

- [15] R. Tummala, "SOP: what is it and why? A new microsystem-integration technology paradigm-Moore's law for system integration of miniaturized convergent systems of the next decade," Advanced Packaging, IEEE Transactions on, vol. 27, no. 2, pp. 241 – 249, may 2004.
- [16] M. U. P. Kulke, R. Rittweger and C. Gunner, "LTCC multilayer ceramic for wireless and sensor applications," *English translation from Production von Leiterplatten und Systemen (PLUS), IMST GmbH*, Dec 2001.
- [17] J.-H. Lee, G. DeJean, S. Sarkar, S. Pinel, K. Lim, J. Papapolymerou, J. Laskar, and M. Tentzeris, "Highly integrated millimeter-wave passive components using 3-D LTCC system-on-package (SOP) technology," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 53, no. 6, pp. 2220 – 2229, june 2005.
- [18] M. Tentzeris, J. Laskar, J. Papapolymerou, S. Pinel, V. Palazzari, R. Li, G. DeJean, N. Papageorgiou, D. Thompson, R. Bairavasubramanian, S. Sarkar, and J.-H. Lee, "3-Dintegrated RF and millimeter-wave functions and modules using liquid crystal polymer (LCP) system-on-package technology," Advanced Packaging, IEEE Transactions on, vol. 27, no. 2, pp. 332 – 340, may 2004.
- [19] K. Jayaraj, T. Noll, and D. Singh, "A low cost multichip packaging technology for monolithic microwave integrated circuits," Antennas and Propagation, IEEE Transactions on, vol. 43, no. 9, pp. 992–997, sep 1995.
- [20] D. Thompson, O. Tantot, H. Jallageas, G. Ponchak, M. Tentzeris, and J. Papapolymerou, "Characterization of liquid crystal polymer (LCP) material and transmission lines on LCP substrates from 30 to 110 GHz," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 52, no. 4, pp. 1343 – 1352, april 2004.
- [21] "Cissco's Global Mobile Data Trafic Forecast Update 2013 2018," http://www.cisco. com/, accessed September, 2014.

- [22] D. M. Pozar, Microwave and RF Design of Wireless Systems. Wiley, 2001.
- [23] "Ferro's website," http://www.ferro.com, [October, 2013].
- [24] B. Yuan, W. Yu, H. Sun, and X. Lv, "A bandpass filter using LTCC System-on-Package (SOP) Technology," in *Microwave Technology and Computational Electromagnetics*, 2009. ICMTCE. International Conference on, nov. 2009, pp. 257 –260.
- [25] G. Brzezina, L. Roy, and L. MacEachern, "Design Enhancement of Miniature Lumped-Element LTCC Bandpass Filters," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 57, no. 4, pp. 815–823, april 2009.
- [26] Y.-S. Dai, H.-S. Zhang, G.-P. Qi, D.-L. Lu, Z.-L. Tao, G.-X. Qian, and H. Wang, "LTCC bandpass filter for Bluetooth application with dual transmission zeros," in *Microwave* and Millimeter Wave Technology, 2008. ICMMT 2008. International Conference on, vol. 1, april 2008, pp. 284 –286.
- [27] C.-F. Chang and S.-J. Chung, "Bandpass filter of serial configuration with two finite transmission zeros using LTCC technology," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 53, no. 7, pp. 2383–2388, 2005.
- [28] H.-H. Huang, C.-H. Chen, and T.-S. Horng, "Quasi-lumped bandpass filter with sharp transition edge and wide stopband rejection," *Electronics Letters*, vol. 49, no. 7, pp. 479–480, 2013.
- [29] H.-H. Huang and T.-S. Horng, "Capacitive-coupled resonator-based bandpass filter with controllable multiple transmission zeros at stopband," *Microw. Opt. Technol. Lett.*, vol. 55, no. 7, p. 15631565, 2013.
- [30] I. Awai, A. Kundu, and T. Yamashita, "Equivalent-circuit representation and explanation of attenuation poles of a dual-mode dielectric-resonator bandpass filter," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 46, no. 12, pp. 2159–2163, Dec 1998.

- [31] C.-W. Tang, Y.-C. Lin, and C.-Y. Chang, "Realization of transmission zeros in combline filters using an auxiliary inductively coupled ground plane," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 51, no. 10, pp. 2112–2118, Oct 2003.
- [32] "ESL's website," http://www.electroscience.com/, [October, 2013].
- [33] "Rogers Corporation Website," http://www.rogerscorp.com/index.aspx, accessed August 22, 2012.
- [34] D. Thompson, J. Papapolymerou, and M. Tentzeris, "High temperature dielectric stability of liquid crystal polymer at mm-wave frequencies," *Microwave and Wireless Components Letters, IEEE*, vol. 15, no. 9, pp. 561 – 563, sept. 2005.
- [35] D. C. Thompson, "Characterization and design of liquid crystal polymers (LCP) based multilayer RF components and packages," *PhD Thesis, Electrical and Computer Engineering, Georgia Institute of Technology*, may 2006.
- [36] D. D., "Developements in Clear High Barrier Packaging," in *Flexible Packaging Con*ference 2004, mar. 2004.
- [37] S. Qian and J. Hong, "Miniature Quasi-Lumped-Element Wideband Bandpass Filter at 0.5-2-GHz Band Using Multilayer Liquid Crystal Polymer Technology," *Microwave Theory and Techniques, IEEE Transactions on*, vol. PP, no. 99, pp. 1–9, 2012.
- [38] S. Qian and J. Hong, "A compact multilayer liquid crystal polymer VHF bandpass filter," in *Microwave Conference (EuMC)*, 2013 European, Oct 2013, pp. 1207–1210.
- [39] A. Brown and G. Rebeiz, "A varactor-tuned RF filter," Microwave Theory and Techniques, IEEE Transactions on, vol. 48, no. 7, pp. 1157–1160, Jul 2000.
- [40] M. Makimoto and M. Sagawa, "Varactor Tuned Bandpass Filters Using Microstrip-Line Ring Resonators," in *Microwave Symposium Digest*, 1986 IEEE MTT-S International, June 1986, pp. 411–414.

- [41] Y.-H. Shu, J. Navarro, and K. Chang, "Electronically switchable and tunable coplanar waveguide-slotline band-pass filters," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 39, no. 3, pp. 548–554, Mar 1991.
- [42] I. Hunter and J. Rhodes, "Electronically Tunable Microwave Bandpass Filters," Microwave Theory and Techniques, IEEE Transactions on, vol. 30, no. 9, pp. 1354–1360, Sep 1982.
- [43] K. Kageyama, K. Saito, H. Murase, H. Utaki, and T. Yamamoto, "Tunable active filters having multilayer structure using ltcc," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 49, no. 12, pp. 2421–2424, Dec 2001.
- [44] J. Nath, D. Ghosh, J.-P. Maria, A. I. Kingon, W. Fathelbab, P. Franzon, and M. Steer, "An electronically tunable microstrip bandpass filter using thin-film barium-strontiumtitanate (bst) varactors," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 53, no. 9, pp. 2707–2712, Sept 2005.
- [45] S. Courreges, Y. Li, Z. Zhao, K. Choi, A. Hunt, and J. Papapolymerou, "A Low Loss X-Band Quasi-Elliptic Ferroelectric Tunable Filter," *Microwave and Wireless Components Letters, IEEE*, vol. 19, no. 4, pp. 203–205, April 2009.
- [46] J. Papapolymerou, C. Lugo, Z. Zhao, X. Wang, and A. Hunt, "A Miniature Low-Loss Slow-Wave Tunable Ferroelectric BandPass Filter From 11-14 GHz," in *Microwave Symposium Digest*, 2006. IEEE MTT-S International, June 2006, pp. 556–559.
- [47] G. Subramanyam, A. Zaman, N. Mohsina, F. Van Keuls, F. Miranda, and R. Romanofsky, "A narrow-band ferroelectric tunable bandpass filter for K-band applications," in *Microwave Conference*, 2000 Asia-Pacific, 2000, pp. 938–941.
- [48] V. Pleskachev and I. Vendik, "Tunable microwave filters based on ferroelectric capacitors," in Microwaves, Radar and Wireless Communications, 2004. MIKON-2004. 15th International Conference on, vol. 3, May 2004, pp. 1039–1043 Vol.3.

- [49] C. Tsai, G. Qiu, H. Gao, L. Yang, G. P. Li, S. Nikitov, and Y. Gulyaev, "Tunable wideband microwave band-stop and band-pass filters using YIG/GGG-GaAs layer structures," *Magnetics, IEEE Transactions on*, vol. 41, no. 10, pp. 3568–3570, 2005.
- [50] C. Tsai and G. Qiu, "Wideband Microwave Filters Using Ferromagnetic Resonance Tuning in Flip-Chip YIG-GaAs Layer Structures," *Magnetics, IEEE Transactions on*, vol. 45, no. 2, pp. 656–660, 2009.
- [51] Y. Murakami, T. Ohgihara, and T. Okamoto, "A 0.5-4.0-GHz Tunable Bandpass Filter Using YIG Film Grown by LPE," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 35, no. 12, pp. 1192–1198, 1987.
- [52] M. Popov, D. Murthy, I. Zavislyak, and G. Srinivasan, "Magnetic field tunable 18-36 GHz dielectric bandpass filter," *Electronics Letters*, vol. 48, no. 2, pp. 98–99, January 2012.
- [53] X. Yang, J. Wu, S. Beguhn, Z. Zhou, J. Lou, and N. X. Sun, "Novel C-band tunable bandpass filter with low bias magnetic fields using partially magnetized ferrites," in *Microwave Symposium Digest (MTT)*, 2012 IEEE MTT-S International, 2012, pp. 1–3.
- [54] A. Shamim, J. Bray, N. Hojjat, R. Elasoued, and D. Baillargeat, "Microwave and Magnetostatic Characterization of Ferrite LTCC for Tunable and Reconfigurable SiP Applications," in *Microwave Symposium, 2007. IEEE/MTT-S International*, June 2007, pp. 691–694.
- [55] S. Sun, J. Shi, L. Zhu, S. Rustagi, and K. Mouthaan, "Millimeter-Wave Bandpass Filters by Standard 0.18- μm CMOS Technology," *Electron Device Letters, IEEE*, vol. 28, no. 3, pp. 220–222, March 2007.
- [56] S. W. Wong, K. Wang, Z.-N. Chen, and Q.-X. Chu, "Design of Millimeter-Wave Bandpass Filter Using Electric Coupling of Substrate Integrated Waveguide (SIW)," *Microwave and Wireless Components Letters, IEEE*, vol. 24, no. 1, pp. 26–28, Jan 2014.

- [57] F. F. Kuo, Network Analysis and Synthesis. Wiley, 1960.
- [58] Microwave Filters, Impedance Matching Networks and Coupling Structures. McGraw-Hill, 1980.
- [59] J. S. Hong and M. J. Lancaster, Microstrip Filters for RF/Microwave Applications. Wiley, 2001.
- [60] E. Jones and J. T. Bolljahan, "Coupled-Strip-Transmission-Line Filters and Directional Couplers," *Microwave Theory and Techniques, IRE Transactions on*, vol. 4, no. 2, pp. 75–81, April 1956.
- [61] S. B. Cohn, "Parallel-Coupled Transmission-Line-Resonator Filters," Microwave Theory and Techniques, IRE Transactions on, vol. 6, no. 2, pp. 223–231, April 1958.
- [62] S. Cohn, "Shielded Coupled-Strip Transmission Line," Microwave Theory and Techniques, IRE Transactions on, vol. 3, no. 5, pp. 29–38, October 1955.
- [63] A. Sutono, D. Heo, Y.-J. Emery Chen, and J. Laskar, "High-Q LTCC-based passive library for wireless system-on-package (SOP) module development," *Microwave Theory* and Techniques, IEEE Transactions on, vol. 49, no. 10, pp. 1715 –1724, oct 2001.
- [64] W. Yun, V. Sundaram, and M. Swaminathan, "High-Q Embedded Passives on Large Panel Multilayer Liquid Crystalline Polymer-Based Substrate," Advanced Packaging, IEEE Transactions on, vol. 30, no. 3, pp. 580–591, aug. 2007.
- [65] I. Bahl, "Lumped Elements for RF and Microwave Circuits," in Artech House, Norwood, MA, USA, 2003.
- [66] "Metro Circuits website," http://www.metrocircuits.com/, accessed September, 2014.
- [67] W. Kuhn and A. Boutz, "Measuring and Reporting High Quality Factors of Inductors Using Vector Network Analyzers," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 58, no. 4, pp. 1046–1055, april 2010.

- [68] A. Horn, J. Reynolds, and J. C. Rautio, "Conductor profile effects on the propagation constant of microstrip transmission lines," in *Microwave Symposium Digest (MTT)*, 2010 IEEE MTT-S International, May 2010, pp. 868–871.
- [69] R. S. Burington, Mathematical Tables and Formulas. Handbook Publishers, Inc., 1933.
- [70] K. Cao, Q.-W. Li, and C.-H. Cheng, "A Fourth-Order Lumped-Element Bandpass Filter Constructed on Multilayer Substrates," *Microwave and Wireless Components Letters*, *IEEE*, vol. 24, no. 11, pp. 745–747, Nov 2014.
- [71] X. Yang, J. Wu, J. Lou, X. Xing, D. Oate, G. Dionne, and N. Sun, "Compact tunable bandpass filter on YIG substrate," *Electronics Letters*, vol. 48, no. 17, pp. 1070–1071, 2012.
- [72] B. Lax and K. Button, *Microwave ferrites and ferrimagnetics*, ser. Lincoln Laboratory publications. McGraw-Hill, 1962.
- [73] R. Soohoo, *Microwave magnetics*. Harper & Row, 1985.
- [74] C. Kittel, "On the Theory of Ferromagnetic Resonance Absorption," Phys. Rev., vol. 73, pp. 155–161, Jan 1948. [Online]. Available: http://link.aps.org/doi/10.1103/ PhysRev.73.155
- [75] J. J. Green and F. Sandy, "Microwave characterization of partially magnetized ferrites," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 22, no. 6, pp. 641–645, 1974.
- [76] G. T. Rado, "Theory of the Microwave Permeability Tensor and Faraday Effect in Nonsaturated Ferromagnetic Materials," *Phys. Rev.*, vol. 89, pp. 529–529, Jan 1953.
 [Online]. Available: http://link.aps.org/doi/10.1103/PhysRev.89.529

- [77] G. Tyras, "The permeability matrix for a ferrite medium magnetized at an arbitrary direction and its eigenvalues (correspondence)," *Microwave Theory and Techniques, IRE Transactions on*, vol. 7, no. 1, pp. 176–177, 1959.
- [78] H. Wheeler, "Transmission-Line Properties of Parallel Strips Separated by a Dielectric Sheet," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 13, no. 2, pp. 172–185, Mar 1965.
- [79] H. Wheeler, "Transmission-Line Properties of Parallel Wide Strips by a Conformal-Mapping Approximation," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 12, no. 3, pp. 280–289, May 1964.
- [80] H. Wheeler, "Transmission-line properties of a strip on a dielectric sheet on a plane," Microwave Theory and Techniques, IEEE Transactions on, vol. 25, no. 8, pp. 631–647, Aug 1977.
- [81] E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-Aided Design," in *Microwave symposium Digest*, 1980 IEEE MTT-S International, May 1980, pp. 407–409.
- [82] G. F. Dionne and D. Oates, "Tunability of microstrip ferrite resonator in the partially magnetized state," *Magnetics, IEEE Transactions on*, vol. 33, no. 5, pp. 3421–3423, 1997.
- [83] R. Pucel and D. Masse, "Microstrip Propagation on Magnetic Substrates Part I: Design Theory," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 20, no. 5, pp. 304–308, May 1972.
- [84] D. Masse and R. Pucel, "Microstrip Propagation on Magnetic Substrates Part II: Experiment," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 20, no. 5, pp. 309–313, May 1972.

- [85] J. L. Melchor and P. H. Vartanian, "Temperature Effects in Microwave Ferrite Devices," *Microwave Theory and Techniques, IRE Transactions on*, vol. 7, no. 1, pp. 15–18, 1959.
- [86] N. Ashcroft and N. Mermin, Solid State Physics. Philadelphia: Saunders College, 1976.