Low-Cost Compact Spiral Inductor Resonator Filters for System-In-a-Package

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ABSTRACT

In this paper, compact spiral inductor resonators filters, which are embedded in the package substrate, using spiral inductor type resonator are proposed. The design is based on the self-resonance frequency of the spiral inductors and electromagnetic coupling effects among resonators. The filters were built and measurement results have good agreement with the simulation data. The proposed filter design technique is generalized and can be implemented in different multilayer arrangement. The new compact inductor resonator filters exhibit a significant size reduction compared to the existing distributed filter topology. The inductor resonator filter designs are based on full-wave electromagnetic analysis. A lumped element equivalent circuit based on the simulated data of the filter is also derived for integration in SPICE type simulator. The compactness of the newly developed bandpass filter makes the design and integration of bandpass filters attractive for further development and applications in System-In-a-Package (SIP).
I. INTRODUCTION

Recent advances in integration technology and device performance paved the way for higher level of System integration On Chip (SOC) or In Package (SIP) [1]. The new wireless and mobile communication systems use miniature Radio Frequency (RF) module design technologies to satisfy low-cost and compact size. These requirements are critical for some specific application such as bandpass filter to reduce the cost and size of mobile system. Planar filters, such as parallel-coupled filters [2], hairpin filters [3]-[5], and elliptic function filter [6]-[8], would be preferred since they are compatible with low cost printed circuit technology. However, they require large real state area because each resonator needs quarter wavelength or half wavelength to get resonance effects. A multilayer filter based on ceramic was introduced as a multilayer LC chip filter to reduce its size [9]. Embedded passives in multilayer have gained importance for offering an attractive solution to the implementation of off-chip passive components. Combline RF bandpass filter integrated into FR4/Epoxy based multilayer substrates has been realized to satisfy low-cost and compact realization [10]. However, capacitive loaded combline RF bandpass filter uses external capacitor to reduce the filter size. In addition, the performance of the filter is affected by the tolerance of external capacitors. Integrated passive components on package substrate provide new opportunities to achieve compact hybrid-circuit design in a single package. The System-In-Package (SIP) technology not only provides interconnects to both digital and RF circuits, but also includes a unique feature of building integrated passive components. However, planar bandpass filters, which are based on microstrip or coplanar waveguide, may not be suitable for SIP due to their relatively large real estate. Most of today embedded filters in SIP are made on Low Temperature Co-fired Ceramics (LTCC) technology, which eliminates Surface Mount Technology (SMT) and reduces the size of the filter [11], [12]. While LTCC filter can reduce lateral size due to relatively high
dielectric constant, it may not represent the most economical solution. With an increased interest in embedded passive circuits in RF module, Raghu et al. reported design methodology of compact multilevel folded-line bandpass filter [13], [14].

This paper describes a general design methodology for embedded passive components with emphasis on bandpass filter. Newly developed bandpass filters using spiral inductor type resonators are designed without any external component. Proposed bandpass filters are designed based on low-cost laminate substrate \( \varepsilon_r = 4.2, \tan \delta = 0.009 \). Typical design rules for laminate substrate technology are used in the proposed new design for SIP. This paper describes the design and analysis of the filter performance and is divided in 6 sections. Section II introduces the concept of the spiral inductor resonator and its behavior at the self-resonance frequency. The design techniques for spiral resonators filter are provided in section III. Section IV discusses the synthesis and optimization of spiral resonator geometry to satisfy the target center frequency and bandwidth. Possible variations of the proposed filter using arbitrary and multilayer arrangements for the resonators are shown in section V. In the same section, the arrangement of inductor type resonators is also shown. The proposed design is flexible and can be optimized for a given available real estate on the carrying substrate for specified center frequency and bandwidth. This means that the multilayer inductor resonator bandpass filters can be used to occupy the available space on package substrate using an arbitrary shape of bandpass filter. Finally, section VI presents the measured filter responses and compares them with the predicted simulated responses.
II. SPIRAL INDUCTOR RESONATORS

II. A. Self-resonance frequency

Integrated passive components improve package and MCM efficiency, enhance electrical and high frequency performance by reducing the parasitics, and eliminate surface mount assembly procedure which in turn improves the yield and reliability due to the reduced solder joint failures. Especially, embedded inductor is one of the key components that determines the performance of every RF/microwave circuit, particularly in voltage control oscillator (VCO), power amplifier, low noise amplifier and filter. The inductors on silicon substrates have usually limited Q values [15][16] and RF front-end module such as the power amplifier often require off-chip matching and biasing components due to their poor Q factor and current-handling capability limitation [17]. The low cost multilayer laminate technology has been proved to be an alternate process to LTCC as an extremely cost effective, high density RF/microwave packaging solution with embedded passives [18]. The design and use of compact and high Q inductors with the desired inductance is critical to ensure successful circuit operation. Fig. 1 shows qualitatively the three possible regions of operation of a spiral inductor [19]. Region I comprises the useful band of operation of a spiral inductor. Inside this region, the effect of parasitic capacitance is small and inductance value remains relatively constant and the inductor can be used as inductive component. Region II is the transition region in which reactance value becomes negative with a zero crossing, which is the self-resonance frequency of the inductor. An inductor is at self-resonance frequency when the peak electric and magnetic energies are equal. Therefore, inductance vanishes to zero at the self-resonance frequency. Above the self-resonance frequency, no net magnetic energy is available from an inductor to any external circuit. The self-resonance frequency depends on the physical inductor geometry as well as material properties. The inductor
area, the number of turns and the metal trace width values highly affect the self-resonance frequency. As the area occupied by spiral inductor increases, the self-resonance frequency decreases. Also, as the number of turns and metal trace width increase, the resonance frequency decreases. The conclusion drawn is that the specific self-resonance frequency of spiral inductor can be achieved by proper choice of geometry.

Fig. 1. Distinct operational region of a spiral inductor.

II. B. Equivalent Circuit Model

The self-resonance frequency of inductor depends on the inductance of spiral inductor and the parasitics capacitance that exists among the spiral inductor turns and between the inductor structure and ground plane. Therefore, self-resonance frequency of the inductor resonators can be optimized by changing the spiral inductor geometry and material properties. Fig. 2 shows the spiral inductor resonator on laminate substrate. The spiral inductor is connected to the ground at one end using substrate via to realize a short-circuited transmission line with less than length \( \lambda/4 \), where \( \lambda \) is the wavelength at operating frequency of interest.
The equivalent circuit model for the spiral inductor is shown in Fig. 3 (a). In this circuit $R_s$ models the metal trace resistance, $L_s$ presents the inductance and $C_p$ models parasitic capacitance. Parasitic capacitance ($C_{p1}, C_{p2}$) at the input and output that can be modeled as approximately half the total parasitic capacitance $C_p$. The series resistance for the spiral inductors can be derived from two-port Y-parameters using the following equation

$$R_s = \text{Re} \left( \frac{1}{-Y_{12}} \right)$$  \hspace{1cm} (1)

The first order estimation of $L_s$ can be determined by summing the self-inductances of each individual segment and the negative and positive mutual inductance induced by any parallel segment as described by Greenhouse [20] and based on the work of Grover [21]. Self inductance ($L$) is calculated using the following equation and is given by [21].

$$L = 2l \left[ \ln \frac{2l}{(w+t)} + 0.50049 + \frac{(w+t)}{3l} \right]$$  \hspace{1cm} (2)
In (2), \( l \) is the length (cm), \( w \) is the width (cm), and \( t \) is the metal thickness (cm), while inductance is given in nH. The mutual inductance among the different segments of the spiral inductor is predicted by

\[
M = 2lP_M
\]  

(3)

where \( M \) is the inductance in nH, \( l \) is the length (cm), and \( P_M \) is the mutual inductance parameter, which can be computed with

\[
P_M = \ln \left[ \frac{l}{GMD} + \sqrt{1 + \left( \frac{l}{GMD} \right)^2} \right] - \sqrt{1 + \left( \frac{GMD}{l} \right)^2} + \frac{GMD}{l}
\]  

(4)

In (4), GMD denotes the geometric mean distance between the metal traces. The inductance for the spiral inductors can also be derived from two-port Y-parameters using the following equation

\[
L_s = \frac{\text{Im}(-1/Y_{12})}{2\pi f}
\]  

(5)

\( C_P \) presents the shunt capacitance between ground and the trace of spiral inductor. The parallel plate approximation is valid for wide lines with large inter-winding spacing, which is the case for embedded inductance on laminate or LTCC substrate. The capacitance per unit length can be computed using equation (6)

\[
\frac{C}{l} = \varepsilon \varepsilon_0 \frac{w}{h}
\]  

(6)

In (6), \( l \) is the length, \( w \) is the width, and \( h \) is the dielectric substrate thickness. The shunt capacitance for the spiral inductors can also be predicted from two-port Y-parameters using the following equation
A parallel type of resonance can be achieved using a short-circuited transmission line [22]. The inductor resonator with shorted termination is modeled as LC resonator, as shown in Fig. 3 (b).

The inductance and capacitance in the equivalent lumped element model equal to 14 nH and 0.28 pF, respectively. The spiral inductor shows inductive nature up to 2.54 GHz where the self-resonance frequency is reached. At this resonance frequency, the spiral inductor structure can be used as a resonator. Above the self-resonance frequency, no net magnetic energy is available from an inductor to any external circuit. The spiral inductor shows capacitive nature above self-resonance frequency. The mutual inductance and parasitic capacitance among the inductor turns
enhance the total inductance and capacitance of the resonator, which help to reduce the physical dimension of the resonator.

III. SPIRAL RESONATORS FILTER DESIGN

The layout and structure of the new compact inductor resonator filter are shown in Fig. 4. This edge-coupled inductor filter was designed on two-metal layers low-cost laminate substrate using typical package laminate substrate layers, materials and design rules, as shown in Table I. All filter resonators were designed on the metal layer. The filter design consists of three spiral inductance resonators. Each resonator was designed and optimized to achieve proper resonance frequency.

Fig. 4. Cross-section and top view for new inductor resonator filter.
TABLE I. Typical Dimensions of Laminate-based substrate.

<table>
<thead>
<tr>
<th>Structure parameter</th>
<th>Dimensions</th>
<th>Material</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dielectric 1 (Molding)</td>
<td>900 µm</td>
<td>(\varepsilon_r = 4.3) (\tan \delta = 0.002)</td>
</tr>
<tr>
<td>Metal 1</td>
<td>27 µm</td>
<td>Copper</td>
</tr>
<tr>
<td>Dielectric 2</td>
<td>200 µm</td>
<td>(\varepsilon_r = 4.2,) (\tan \delta = 0.009)</td>
</tr>
<tr>
<td>Metal 2</td>
<td>27 µm</td>
<td>Copper</td>
</tr>
</tbody>
</table>

The lumped elements of the high frequency inductance equivalent circuit were extracted from the measured s-parameters with ADS optimization. The measurement data were analyzed with MathCAD and ADS for the initial lumped element model values for L, C, and R. DC resistance was extracted from the calibrated simulation tool Ansoft Quick 3D. The measured data were optimized for the lumped element model parameters using ADS – L, C and K. K is the skin effect coefficient for the AC resistance. Models were developed using optimized lumped element model parameters. Each optimized parameter was fitted to the third degree polynomials. For example, inductance and resistance for single layer inductor can be expressed as the following analytical equations (8) – (11). Equation (12) shows the relationship between \(R_{Total}\), \(R_{DC}\), and K. N is the number of turn of inductor and \(f_{re}\) is the frequency in GHz. Analytical parameter equations were created as functions of number of turn.

\[
L = -0.028N^3 + 1.0091N^2 - 1.2221N + 1.5209 \tag{8}
\]
\[
R_{DC} = 0.0002N^3 + 0.0025N^2 + 0.0153N - 0.0044 \tag{9}
\]
\[
K = 4.5714N - 4 \tag{10}
\]
\[
C = 0.0022N^3 - 0.0048N^2 + 0.0303N + 0.0199 \tag{11}
\]
\[
R_{Total} = R_{DC} \cdot (1 + K \cdot \sqrt{f_{re}}) \tag{12}
\]
The coupled structures result from identical spiral inductor resonators, which are separated by spacing. It is obvious that any coupling in those coupling structures is that of the proximity coupling through fringe fields. The nature and the extent of the fringe fields determine the nature and the strength of the coupling. The electric coupling and magnetic coupling can be obtained if the two coupled spiral inductor resonators are proximately placed. The electric and magnetic fringe fields at the coupled sides may have comparative distributions so the both the electric and the magnetic coupling occur. Design of this filter was based on the electric and magnetic mixed coupling structure. Fig. 5 shows the filter equivalent circuit modeled as a LC resonator with electric and magnetic coupling elements, respectively. The inductance and capacitance of resonators can be calculated as previously explained. The coupling of the parallel inductor segments is presented by edge coupling capacitance and mutual inductance value. The metal segments of the inductor resonator structure were assumed to be perfect electric conductors for simplicity.

![Fig. 5. Lumped-element equivalent circuit model for inductor resonator filter.](image)

According to the network theory [23], an alternative form of the equivalent circuit in Fig 5 can be obtained and are shown in Fig. 6 (a) and (b), respectively. In Fig 6 (a), electrical coupling by edge coupling capacitance can be shown that the electric coupling between the two resonant
The magnetic coupling between the two resonant inductors is represented by a \( \pi \) network. The electric and magnetic field distributions of two coupled inductor resonators are comparative so that neither the electric coupling nor the magnetic coupling can be ignored. By inserting an electric wall and magnetic wall into the symmetry plane of the mixed coupling equivalent circuit, we obtain

\[
\begin{align*}
    f_e &= \frac{1}{2\pi \sqrt{(C - C_c)(L - L_m)}} \\
    f_m &= \frac{1}{2\pi \sqrt{(C + C_c)(L + L_m)}}
\end{align*}
\]

As can be seen that both the electric and magnetic couplings have the same effect on the resonant frequency shifting. More detailed information about mixed coupling equivalent circuit can be found in [24].
Fig. 6. An alternative form of the equivalent circuit with (a) a \( \pi \) network and (b) a \( T \) network to represent the coupling.
The filter using inductor resonators has the following dimensions: area of 3660 µm × 3120 µm, trace line width of 60 µm, and spacing between lines of 60 µm. The newly designed inductor resonator filter was simulated using 3-D full wave electromagnetic software and results are shown in Fig. 7. The insertion loss in the passband and return loss of the new inductor resonator filter are respectively equal to 1 dB and 20 dB at center frequency. The filter is centered at 2.4 GHz while provided a 3-dB bandwidth of 95 MHz around the center frequency.

Fig. 7. Return loss and Insertion loss of bandpass filter using inductor resonators.
IV. SPIRAL RESONATORS FILTER SYNTHESIS

The spiral resonators filter design optimization is one of the major aspects in filter design. The center frequency and the bandwidth of the filter should be adjusted to meet the specification goals. In this section, the optimization of center frequency and bandwidth is depicted with filter geometry. The following results did not consider the losses in conductor and substrate to only emphasize on the effect of geometry and configuration on center frequency and bandwidth.

IV.A. Filter center frequency synthesis

The center frequency in the filter passband was adjusted by changing the number of resonators turns. Tuning can be performed by changing the number of resonator turns, which affect the self-resonance frequency of each inductor seen as resonators element for the filter. Fig. 8 shows the change in the resonance center frequency as function of number of inductor resonator turns. The return loss and insertion loss for both the equivalent circuit model and full-wave simulation are shown in Fig. 8 (a) and (b), respectively.
Fig. 8. (a) Return losses and (b) Insertion losses due to center frequency variations versus number of turns.
The corresponding element component values for the equivalent circuit model can be found in Table II. In the equivalent circuits, coupling effects among resonators are modeled as mutual inductance and coupling capacitance, as previously explained. The resonance frequency of the filter increases with the decrease in the number of turns of the spiral inductance resonators. Inductor resonators, with number of turns equal to 1.7, 1.5, and 1.3 turns, lead to the center frequency of 2.24 GHz, 2.4 GHz, and 2.56 GHz, respectively. When the turn numbers of inductor resonator are increased, the values of series inductance (L₁, L₂, L₃), shunts capacitance (C₁, C₂, C₃) of resonator, and mutual inductance (M₁₂, M₂₃) between resonators are increased, as reported in Table II. This can be explained by the increase in the inductance and parasitic capacitance values, which were computed by equation (2) and (3), respectively, which results in decrease of the resonance frequency. Corresponding self-resonance frequency can be calculated using equation (15).

\[
\text{SRF} = \frac{1}{2\pi \sqrt{LC}} \tag{15}
\]

However, the coupling capacitances (C₁₂, C₂₃) between the resonators remained approximately the same, since the distance among the resonators was not changed. This leads to the variation of self-resonance frequency in each resonator, without significant change in the bandwidth of the filter. Therefore, in order to tune the center frequency of the spiral inductor resonator filters, the number of turns for each resonator needs to be optimized.
TABLE II. Lumped-element Circuit Parameters for Various Number of Resonator Turns.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>1.7 turns</th>
<th>1.5 turns</th>
<th>1.3 turns</th>
</tr>
</thead>
<tbody>
<tr>
<td>L₁ (nH)</td>
<td>15.5</td>
<td>14</td>
<td>12.5</td>
</tr>
<tr>
<td>L₂ (nH)</td>
<td>15.5</td>
<td>14</td>
<td>12.5</td>
</tr>
<tr>
<td>L₃ (nH)</td>
<td>15.5</td>
<td>14</td>
<td>12.5</td>
</tr>
<tr>
<td>C₁ (pF)</td>
<td>0.3</td>
<td>0.28</td>
<td>0.26</td>
</tr>
<tr>
<td>C₂ (pF)</td>
<td>0.3</td>
<td>0.28</td>
<td>0.26</td>
</tr>
<tr>
<td>C₃ (pF)</td>
<td>0.3</td>
<td>0.28</td>
<td>0.26</td>
</tr>
<tr>
<td>C₁₂ (pF)</td>
<td>0.1</td>
<td>0.1</td>
<td>0.1</td>
</tr>
<tr>
<td>C₂₃ (pF)</td>
<td>0.1</td>
<td>0.1</td>
<td>0.1</td>
</tr>
<tr>
<td>M₁₂ (nH)</td>
<td>6.6</td>
<td>6</td>
<td>5.4</td>
</tr>
<tr>
<td>M₂₃ (nH)</td>
<td>6.6</td>
<td>6</td>
<td>5.4</td>
</tr>
</tbody>
</table>

IV.B. Filter bandwidth synthesis

The bandwidth of the proposed inductor resonator filter can be optimized by changing the distance between the resonators. The bandwidth variations versus the distance between the filter spiral resonators are shown in Fig. 9. The bandwidth of inductor resonator filters equals to 95 MHz, 75 MHz, and 55 MHz at the resonator distance of 60 µm, 90 µm, and 120 µm, respectively. It is clear that the bandwidth increases as the distance between the resonators decreases. It can be seen that as the coupling spacing decreases the two resonant peaks, which were equation (13) and
The simulated results give an insight into the characteristics of couplings and indicate that the coupling depend on the spacing. The return loss and insertion loss for both the equivalent circuit model and full-wave simulation are shown in Fig. 9 (a) and Fig. 9 (b), respectively.

Fig. 9. (a) Return loss and (b) Insertion loss for bandwidth variations versus resonators distance.
The corresponding element component values for the equivalent circuit model can be found in Table III. In the equivalent circuits, coupling effect between resonators was modeled as mutual inductance and coupling capacitance, as previously explained. As the spaces among the resonators are increased, electromagnetic coupling effects, which are mutual inductance ($M_{12}$, $M_{23}$) and coupling capacitance ($C_{12}$, $C_{23}$), are decreased. This is due to the weaker electromagnetic coupling among resonators. However, the series inductance ($L_1$, $L_2$, $L_3$) and shunt capacitances ($C_1$, $C_2$, $C_3$) of resonators are almost unchanged, since the geometry of each resonator is not changed. The target performance of the filter can be achieved by optimizing the filter structure. The center frequency and bandwidth of the filter were adjusted using the resonator number of turns and distance between resonators, respectively.
TABLE III. Lumped-element Circuit Parameters for Various Distances among Resonators.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>60 µm</th>
<th>90 µm</th>
<th>120 µm</th>
</tr>
</thead>
<tbody>
<tr>
<td>L₁ (nH)</td>
<td>14</td>
<td>14</td>
<td>14</td>
</tr>
<tr>
<td>L₂ (nH)</td>
<td>14</td>
<td>14</td>
<td>14</td>
</tr>
<tr>
<td>L₃ (nH)</td>
<td>14</td>
<td>14</td>
<td>14</td>
</tr>
<tr>
<td>C₁ (pF)</td>
<td>0.28</td>
<td>0.28</td>
<td>0.28</td>
</tr>
<tr>
<td>C₂ (pF)</td>
<td>0.28</td>
<td>0.28</td>
<td>0.28</td>
</tr>
<tr>
<td>C₃ (pF)</td>
<td>0.28</td>
<td>0.28</td>
<td>0.28</td>
</tr>
<tr>
<td>C₁₂ (pF)</td>
<td>0.1</td>
<td>0.065</td>
<td>0.032</td>
</tr>
<tr>
<td>C₂₃ (pF)</td>
<td>0.1</td>
<td>0.065</td>
<td>0.032</td>
</tr>
<tr>
<td>M₁₂ (nH)</td>
<td>6</td>
<td>5.2</td>
<td>4</td>
</tr>
<tr>
<td>M₂₃ (nH)</td>
<td>6</td>
<td>5.2</td>
<td>4</td>
</tr>
</tbody>
</table>

V. ARBITRARY SHAPE AND MULTILAYER SPIRAL RESONATORS FILTER

The presented one-layer inductor resonator filter used edge-coupled resonators to achieve electromagnetic coupling among resonators. However, with multilevel technology, broadside coupling can be used to enhance the coupling among the resonators, and decrease the size of the inductor resonator filter. The broadside coupling can be achieved by overlapped metal area using multilayer structure. Also, the broadside coupling can be optimized with proper overlap structure. As overlap area is increased, the coupling effect is also increased. In addition, arbitrary
arrangement for the resonators can be done due to the coupling mechanism used in this filter design. In the next section, a procedure is described for designing arbitrary shape and multilayer spiral resonators filter.

V. A. L-shape spiral resonator filter

The cross-sectional view of typical package laminate substrate layers and the top view of two-layer L-shape spiral resonators filter are shown in Fig. 10.

![Fig. 10. Cross-section and top view for L-shape bandpass filter using multilayer.](image)

The electromagnetic coupling between resonators in L-shape multilayer filter is enhanced due to the broadside coupling compared to the edge coupled one-layer filter. The multilayer bandpass filter using inductor type resonators has the following dimensions, resonator area 1500 µm x 1500 µm with metal trace width and space of 60 µm. Each dielectric layer has the following material characteristics: dielectric 1 layer height of 200 µm, \(\varepsilon_r\) of 4.2, dielectric layer 2 height of 50 µm, \(\varepsilon_r\) of 3.5 and molding layer height of 900 µm and \(\varepsilon_r\) of 4.3. The dimensions and material of multilayer inductor resonator filter were based on typical package substrate characteristics. The newly
developed inductor resonator bandpass filters can be used to effectively use the available space on package substrate using an arbitrary shape bandpass filter. The filter layout can be optimized based on the available real state in L, T, and I configurations. For example, L-shape bandpass filter is optimum if placed near the corner of the package substrate. T and I shape bandpass filter can be used to use narrow space among other structures. The return and insertion loss of multilayer L-shape spiral resonator filter are shown in Fig. 11. The L-shape arrangement of inductor type resonators can produce an efficient bandpass filter. In summary, the arrangement of inductor type resonator filters is flexible and can be optimized for the available real estate, on single or multi metal layers, to meet specific center frequency and bandwidth.

![S-parameters for multilayer L-shape bandpass filter.](image-url)

Fig. 11. S-parameters for multilayer L-shape bandpass filter.
V. B. Multilayer spiral resonator filter

Multilayer structure can be used to significantly reduce the footprint of the filter and enhance the electromagnetic coupling effects among the resonators. The center frequency and bandwidth variations with different multilayer structure configurations were simulated using HFSS. The return and insertion loss for the multilayer L-shaped spiral inductor resonator filters are shown in Fig. 12. Three different configurations were simulated. The resonator dimensions and aspect ratios are the same for all three configurations. However, the overlap between the resonators segments located on layer 1 and layer 2 was changed. The spiral inductor resonator segments overlap changes from full overlap, 60 \( \mu \text{m} \), to partial overlap, 30 \( \mu \text{m} \), and finally with no overlap. The 60 \( \mu \text{m} \) overlapped inductor resonator filter case has larger bandwidth, 200 MHz, compared to the partial, 175 MHz, and no overlap, 150 MHz, cases. However, the center frequency of the full overlap case has the lowest center frequency, equals to 2.42 GHz, while the center frequency for the partial and no overlap cases equals to 2.5 and 2.7 GHz, respectively. The center frequency of multilayer filter with different overlap area is changed due to the variation in the mutual inductance and coupling capacitor between the resonators. The mutual inductance enhances the self-inductance (\( L_{\text{effective}} = L + 2M \)) of the resonators because spiral inductor resonators have the same direction winding direction. The coupling capacitance between resonators enhances the parasitic capacitance in resonators. Based on equation (15), the center frequency of bandpass filter is decreased.
Fig. 12 (a) Return loss and (b) Insertion loss for 60 µm, 30 µm, 0 µm overlap distance of coupled multilayer filters.
The coupling effects among resonators can be modeled by cascading an electrical line length and the mutual coupling. These mutual inductance and loading effects increase electrical line lengths of the resonators and shift the resonance frequency downward. These results in decreasing the center frequency of the spiral inductor resonator filter. On the other hand, stronger electromagnetic coupling among the resonators results in larger bandwidth of the filter, as expected. Therefore, the center frequency and bandwidth of the arbitrary shape spiral inductor resonator filters can be optimized using proper multilayer configurations and electromagnetic coupling.
VI. MEASURED RESULTS

VI. A. Experimental Results

Fig. 13 shows the top view and cross-sectional view of a fabricated inductor resonator bandpass filter on four layers laminate substrate, respectively.

The ground layer is 200 µm apart from the resonators filter layer. The filter is connected to ground layer using via. Probing pads were included for 2-port microwave measurement. The bandpass filter was built on typical package laminate substrates dimension and materials, which
were available at the time of the design. The fabricated filter was measured using HP 8510C Network Analyzer and Universal Test Fixture. Standard short, open, thru were fabricated on the laminate substrate. A standard Thru-Reflect-Line (TRL)-method [26], using the load from the Universal Test Fixture calibration kit, was performed. Fig. 14 (a) and (b) compare the simulated and measured results of the inductor resonator filter characteristics, insertion loss and return loss. The measured and simulated center frequency equals to 2.495 GHz and 2.485 GHz, with return loss of 8.062 dB and 8.314 dB, respectively. The difference between the measured and simulated data, 10 MHz for the center frequency and 0.3 dB for the return loss at the center frequency, is due to the difference in the drawn and built dimensions of the inductor resonator filter. The measured and simulated insertion losses equal to 3.9dB and 4.3dB, respectively. The relatively high insertion loss of the measured filter is the result of weak electromagnetic edge-coupled resonators coupling, that are 60 µm apart, and the losses associated with metal conductivity and high dielectric loss, which was not considered in the initial design. The measured bandpass filter was fabricated based on the available laminate substrate technology at the time of the design. Therefore, we can increase filter performance with suitable material and design.
Fig. 14. (a) Return loss and (b) Insertion loss for measured and simulated inductor resonator bandpass filter.
Fig. 15 shows the fabricated multilayer L-shape inductor resonator filter on laminate substrate. With multilayer technology, broadside coupling can be used to enhance the coupling among the resonators.

![Image of fabricated multilayer inductor resonator filter](image)

This multilayer technology can be used to reduce the size of the inductor resonator filter. In addition, arbitrary arrangement for the resonators can be achieved due to enhanced coupling mechanism. The newly developed multilayer L-shape bandpass filter consists of area of 8.5 mm², metal trace width of 180 µm, metal thickness of 27 µm, and metal space of 60 µm. The measured insertion and return losses are shown in Fig. 16. The insertion loss in the passband and return loss of the measured L-shape inductor resonator filter equal to 1.85 dB and 13.7 dB, respectively. The center frequency of the passband equals to 2.48 GHz. These results show enhanced coupling effects increase the performance of the spiral inductor bandpass filter. This newly developed
spiral inductor resonators can be used to implement compact filter even on relatively low dielectric constant material.

![Graph showing S-parameters of measured L-shape inductor resonator bandpass filter.](image)

Fig. 16. S-parameters of measured L-shape inductor resonator bandpass filter.

VI. B. Process Variations and Proximity Effects

The manufacturing process variation affects the inductance value, quality factor and self-resonance frequency of the embedded spiral inductor in package substrate. The spiral inductor in package substrate with 100 μm width and space are used as a device under test for 3-D fullwave simulation. As shown in TABLE IV, each simulation is conducted when the metal width and space is changed by ±10% of the reference value. Self-resonance frequency is one of the most important parameters of spiral inductor resonator filter for center frequency of bandpass filter. Shifted self-resonance frequency of resonators can make the shift of a center frequency of bandpass filter. Process variation effect analysis is done on 10.3nH inductor to identify the range...
of inductance variation according to the physical parameters of the embedded inductor and summarized in TABLE IV. For all process variation parameters, the inductance varies within 5%.

**TABLE IV. Process Variation Effect on Embedded Inductor**

<table>
<thead>
<tr>
<th>Process Variation Effect</th>
<th>L (nH)</th>
<th>Q (1GHz)</th>
<th>SRF (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Values</td>
<td>10.289</td>
<td>56.919</td>
<td>3.505</td>
</tr>
<tr>
<td>Line Width +10µm</td>
<td>Value</td>
<td>10.215</td>
<td>54.351</td>
</tr>
<tr>
<td>% Difference</td>
<td>-0.73%</td>
<td>-4.51%</td>
<td>-0.59%</td>
</tr>
<tr>
<td>Line Width –10µm</td>
<td>Value</td>
<td>10.426</td>
<td>57.317</td>
</tr>
<tr>
<td>% Difference</td>
<td>1.32%</td>
<td>0.70%</td>
<td>0.30%</td>
</tr>
<tr>
<td>Layer Thickness +8µm</td>
<td>Value</td>
<td>9.728</td>
<td>54.277</td>
</tr>
<tr>
<td>% Difference</td>
<td>-5.46%</td>
<td>-4.64%</td>
<td>2.32%</td>
</tr>
<tr>
<td>Layer Thickness –8µm</td>
<td>Value</td>
<td>10.851</td>
<td>56.969</td>
</tr>
<tr>
<td>% Difference</td>
<td>5.46%</td>
<td>0.09%</td>
<td>-2.05%</td>
</tr>
<tr>
<td>Layer Alignment +25µm</td>
<td>Value</td>
<td>10.025</td>
<td>56.175</td>
</tr>
<tr>
<td>% Difference</td>
<td>-2.57%</td>
<td>-1.31%</td>
<td>0.9%</td>
</tr>
<tr>
<td>Layer Alignment –25µm</td>
<td>Value</td>
<td>10.165</td>
<td>53.707</td>
</tr>
<tr>
<td>% Difference</td>
<td>-1.21%</td>
<td>-5.64%</td>
<td>0.80%</td>
</tr>
</tbody>
</table>

The effect of physical separation on the electromagnetic crosstalk between neighboring structures must be considered to avoid spurious parasitics. All of the filters in system have very close to them a surrounding metal track. Research results have revealed certain dependence between the self-resonance frequency of the inductor and the distance between the surrounding metal track and the spiral inductor. As the surrounding metal tracks are moved closer to the spiral inductor, the resonance frequency tends to decrease. There is a certain distance from the inductor structure that this distance does not decrease the resonance frequency. Thus, in order to avoid any undesirable performance variation of the filter, extreme attention should be paid to a certain keep out area around the filter. The identical spiral inductor resonator filters have been simulated to exam how the isolation
between filter and surrounding metal track varies as the distance between them changes and shown in Fig. 17. For all distance variation parameters, the center frequencies are changed within 2%. The minimum distance of 60 µm was considered to match package layout design rule.

Fig. 17. Center frequency variation of spiral inductor bandpass filter with different distances between spiral inductor resonator filter and surrounding metal trace.
VII. CONCLUSION

A novel inductor resonator filters are designed. The design can achieve compact filter design even on low cost relatively low dielectric constant material. The design is based on the self-resonance frequency of the spiral inductors and electromagnetic coupling effects among resonators. The measured results have good agreement with the simulation results. The center frequency and bandwidth of the arbitrary shape spiral inductor resonator filters can be optimized using proper multilayer configurations and effective electromagnetic coupling. The compactness of newly developed bandpass filter makes the design of bandpass filters attractive for further development and applications in System-In-a-Package.
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REFERENCES


