Dual-Mode Microstrip Bandpass Square Open Loop Filters

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Abstract

A coupled line dual mode resonator has been proposed in this thesis. This resonator is an enhancement of the commonly known square open-loop resonator. The models for electric, magnetic and mixed couplings have also been developed. These resonators make use of the space inside the loop resonator to achieve a more compact design and enhanced performance. The newly proposed configuration for mixed coupling is fabricated and measurement results are found to have lower insertion loss, wider bandwidth, higher coupling coefficients and a smaller size than the commonly-known four-pole square open-loop resonator.

Further development is then done on the mixed coupling configuration of the coupled line resonator. The meander loop concept and the new coupling scheme are incorporated into the design to achieve a filter with better matching and rejection level. Cascaded networks are also designed and fabricated to achieve a configuration with better than -60dB rejection level.

In this thesis, miniaturized designs based on half the width of 50ohm conductor lines are also investigated. Two of these are based on the coupled line resonator configuration proposed. One configuration is designed for narrow band purpose using the meander loop concept on a single loop. An alternative coupling scheme for the feed lines is implemented on this narrow band miniaturized design to achieve a filter that can shift the first harmonics.
Cascaded networks are also designed for the three miniaturized designs. For the coupled line wideband case, the cascaded networks show predictable responses and are able to achieve configurations with highly selectable and wideband performance. For the narrow band case, the cascaded network is able to improve the rejection band level performance.

Altogether, there are ten pieces of hardware fabricated. There is a general shift in centre frequency for the measured results. However, the shift is within the tolerated range of a few per cent.
Acknowledgement

I would like to thank Dr. Ooi Ban Leong for his valuable advice and guidance on this project. In addition, I would also like to express my gratitude to those graduate students in the Microwave Laboratory and the virtual laboratories who have also shared with me their valuable experience.

Last but not least, the Professional Officers like Hui So Chi and Guo Lin in the MMIC Laboratory, and Madam Lee and Mr. Sing from the Microwave Laboratory have also assisted me greatly in the hardware implementation. I appreciate their effort in helping me to complete my Master research work.
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<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
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<tbody>
<tr>
<td>EBG</td>
<td>Electronic Band Gap</td>
</tr>
<tr>
<td>SIR</td>
<td>Step Impedance Resonator</td>
</tr>
<tr>
<td>CAD</td>
<td>Computer Aided Design</td>
</tr>
<tr>
<td>HTS</td>
<td>High-temperature superconductors</td>
</tr>
<tr>
<td>MEMS</td>
<td>Microelectromechanical systems</td>
</tr>
<tr>
<td>MMIC</td>
<td>Monolithic microwave integrated circuits</td>
</tr>
<tr>
<td>LTCC</td>
<td>Low-temperature cofired ceramics</td>
</tr>
<tr>
<td>$\varepsilon$</td>
<td>Permittivity of the dielectric</td>
</tr>
<tr>
<td>$h$</td>
<td>Distance between the trace and the ground</td>
</tr>
<tr>
<td>$\lambda_g$</td>
<td>Group wavelength</td>
</tr>
<tr>
<td>$f_e$</td>
<td>Even mode resonant frequency</td>
</tr>
<tr>
<td>$f_o$</td>
<td>Odd mode resonant frequency</td>
</tr>
<tr>
<td>$k_E$</td>
<td>Electric Coupling coefficient</td>
</tr>
<tr>
<td>$k_M$</td>
<td>Magnetic Coupling coefficient</td>
</tr>
<tr>
<td>$k_B$</td>
<td>Mixed Coupling coefficient</td>
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1 Introduction

1.1 Motivation and purpose

Modern microwave communication systems, especially in the satellite and mobile communications, require high performance, narrowband bandpass filters having low insertion loss and high selectivity. The microstrip ring resonator has widely been used to fulfill these requirements as it is well known for its compact size, low cost and easy fabrication.

Very often, the ring resonator is being implemented as a one-wavelength-type Step Impedance Resonator (SIR). It is well-known that there are two orthogonal resonance modes within a one-wavelength ring resonator [1]. The common practice of implementing the dual mode is by introducing a small patch at the corner of the square ring resonator [2]. This is to serve as a perturbation to introduce the dual mode resonant frequencies. The feed lines are located orthogonal to each other. An example of this design is shown in Figure 1-1 (a) and the dimensions of the configuration are given in Table 1-1. Figure 1-2 shows the full wave analysis using Agilent’s momentum software. The simulation is done on a RT/Duroid 6010 substrate with a thickness, \( h = 25 \text{mil} \) and relative dielectric constant \( \varepsilon_r = 10.2 \). Two peaks corresponding to the transmission zeros (S21 is maximum) are observed at 4.51GHz and 4.62GHz. Two attenuation poles, represented by the minimum points on the graph, are also observed. They are at 4.25GHz and 5.19GHz.
Another common configuration that is often used in microwave bandpass design is an open loop resonator [3]. This type of filters has often been implemented in the form of hairpin structures [4]-[5]. Extensive research has been done on this configuration to investigate the design method and the couplings of the two open end of the hairpin structure. An example of the square open loop resonator using orthogonal feed is shown in Figure 1-1 (b). In this example, the gap-opening is at the edge of the corner opposite the two feed lines. The gap-opening is the perturbation in this case. The dimensions of this design are shown in Table 1-1 and the frequency response S21 is shown in Figure 1-3. Again, the simulation is done on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r =10.2$. Two sharp peaks corresponding to the two transmission zeros (S21 is maximum) are observed at 4.48GHz and 4.62GHz.

From these designs, it can be seen that by merely altering the square loop resonator, many of its parameters like the transmission zeros, attenuation poles and resonant frequencies can be changed. This provides the interest to investigate this type of filters further. In addition, it is relatively economical, easy and accurate to implement these planar microstrip structures. All these add to the motivation to improve on the working performance of the existing designs and to further shrink down their size for modern communication applications.
Figure 1-1 Configurations for the conventional dual mode resonators

<table>
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<tr>
<th>Item</th>
<th>$l_1$</th>
<th>$l_2$</th>
<th>$l_3$</th>
<th>$l_4$</th>
<th>$w$</th>
<th>$w_1$</th>
<th>$g$</th>
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<td>286</td>
<td>283</td>
<td>50</td>
<td>50</td>
<td>23</td>
<td>20</td>
<td>5</td>
</tr>
<tr>
<td>Dimensions for Figure 1-1(b) (mils)</td>
<td>263.5</td>
<td>283</td>
<td>50</td>
<td>50</td>
<td>23</td>
<td>N. A.</td>
<td>5</td>
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Table 1-1 Dimensions for Figure 1-1
Figure 1-2 Dual mode resonator [2] with small patch and its frequency response

Figure 1-3 Open loop dual mode resonator and its frequency response
1.2 Scope of work

The fundamental element of the filter design presented in this thesis is based on a dual mode square open loop resonator with direct-connected orthogonal feed lines. The direct connection between the feed lines and the square loop allows for little mismatch and radiation losses between them. As such, investigations will be done on features like the effects of positioning the gap-opening, the number of gaps and its relationships with the dual mode features. The coupling of the filter design will also be touched on. The filter element under investigation and the dimensions are shown in Figure 1-4 and Table 1-2. This design is implemented on a RT/Duroid 6010 substrate with a thickness, \( h = 25 \text{mil} \) and relative dielectric constant \( \varepsilon_r = 10.2 \). The full wave simulation response using Agilent’s momentum software is presented in Figure 1-5. It can be seen from the results that in addition to the sharp transmission zeros (S21 is maximum) observed at the peaks at 4.41GHz and 4.60GHz, there are also two sharp attenuation poles represented by the minimum points at 4.37GHz and 4.65GHz. The attenuation poles are very close to the transmission zeros, meaning that the filter has a very high selectivity. This response is highly desirable to the demand of modern communication network.

In this thesis, investigations and analysis will be done on this square open loop resonator with two gap-openings. In addition, motivations from the latest research done by other researchers will also be implemented on this design to evolve into new configurations to further improve the performance of the resonator.
Further to the development of this new square open loop resonator, work will also be devoted to looking into increasing the bandwidth of the filter for wideband applications and decreasing the bandwidth for applications that require extremely precise narrowband bandpass filters.

Figure 1-4 Open loop configuration with two gap-openings.

<table>
<thead>
<tr>
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<td>230</td>
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<td>10</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 1-2 Dimensions for Figure 1-4
1.3 List of Contributions arising from the present work

As a result of the investigations and designs of the work arising from the present work presented in this thesis, there are four publications contributed to some conferences and journals. They are listed below:


2 Mathematical Analysis on the square resonator with two openings

2.1 Introduction

Figure 1-4 presents a square loop with two gap-openings. The conventional square open loop resonators usually only have one gap-opening [2]. Their characteristics and different combinations have been extensively studied. The configurations to implement a two-pole and four-pole resonators are given in Figure 2-1 (a) and (b). It needs two loops to implement a two-pole design and more loops have to be added to get a highly selective response. The filter design presented in Figure 1-4 with the square loop having two gap-openings can introduce two poles with just one loop and the design has high selectivity. This leads to the motivation to investigate the effects of the additional gap-opening.

In this chapter, the mathematical analysis on the square open loop resonator with two gap-openings is presented. It is of design interest to find out how to get the transmission zeros (maximum peak on S21 graph) and attenuation poles (minimum dip on S21 graph). Therefore, the equivalent circuit analysis on the square open-loop resonator is presented and used to calculate these parameters.
2.2 Effects of the gap-openings

The design shown in the previous chapter with two gap-openings is reproduced in Figure 2-2 to show its dimensions, which are also given in Table 2-1. Figure 2-3 shows the full wave analysis using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r =10.2$. Comparison is done between open loop resonators that have one gap-opening and two gap-openings.
Figure 2-2 Configuration to investigate effects of one and two gap-openings

<table>
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<td>Dimensions for one</td>
<td>313</td>
<td>283</td>
<td>230</td>
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<tr>
<td>opening (mils)</td>
<td></td>
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<tr>
<td>Dimensions for two</td>
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<td>230</td>
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Table 2-1 Dimensions for Figure 2-2
As can be seen from the simulation results of Figure 2-3, in addition to the peak at around 4.589GHz, the additional gap-opening generates one more transmission zeros for the resonator at 4.396GHz. This gives the intuition that the gap-opening is responsible for generating the peak and is similar to the effect of introducing an additional perturbation or discontinuity to the resonator except that the filter designed is of a different type.

Figure 2-3 Effect of introducing one more gap-opening to the square loop resonator
2.3 Positions of the gap-opening

In addition to studying the effects of the additional gap-opening, it is also very important to consider the position of the gap-openings. Figure 2-4 shows the configuration for this investigation. The length $l$ is lengthened (to the right) or shortened (to the left). Towards the right is taken as the positive direction and towards the left as the negative direction. Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, $h=25\text{mil}$ and relative dielectric constant $\varepsilon_r=10.2$.

![Figure 2-4 Configuration to investigate the positions of the gap-opening](image)

<table>
<thead>
<tr>
<th>Item</th>
<th>$l_1$</th>
<th>$l_2$</th>
<th>$l_3$</th>
<th>$l_4$</th>
<th>$l$</th>
<th>$g$</th>
<th>$w$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions (mils)</td>
<td>336</td>
<td>283</td>
<td>240- $l$</td>
<td>50</td>
<td>variable</td>
<td>10</td>
<td>23</td>
</tr>
</tbody>
</table>

Table 2-2 Dimensions for Figure 2-4
From Figure 2-5, one transmission zero, represented by the peak can be observed at 4.59GHz. As the gap-opening moves further to the right, in the positive direction, the attenuation poles moved further and further apart. Therefore, in order to obtain a filter performance with sharper cut off, it is best to place the gap-opening at the corner of the square loop.

![Figure 2-5 Investigation on the positions of the gap-opening](image)

Figure 2-6 shows the configuration for the investigation on two gap-openings. The dimensions are shown in Table 2-3. Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r = 10.2$. For the gap-opening on the upper part, the $l$ is taken to be positive towards the left hand side and a lengthening of positive dimension of $l$ is to the left. This is a diagonal reflection to
the lower portion. In Figure 2-7, it can be seen that with \( l = 10\text{mil} \) and \( l = -10\text{mil} \), the shape of the graphs and the distance between the two resonant frequencies are very close. However, there is a shift in the resonant frequencies. \( l = 10\text{mil} \) have resonant frequencies at 4.43GHz and 4.60GHz. For \( l = -10\text{mil} \), the resonant frequencies are at 4.53GHz and 4.73GHz. When \( l = 0\text{mil} \), the two resonant frequencies at 4.4GHz and 4.6GHz are highly selective with the attenuation poles immediate beside the transmission zeros.

Figure 2-6 Configurations to investigate resonators with two gap-openings

<table>
<thead>
<tr>
<th>Item</th>
<th>( l_1 )</th>
<th>( l_2 )</th>
<th>( l_3 )</th>
<th>( l_4 )</th>
<th>( l )</th>
<th>( g )</th>
<th>( w )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>for Figure 2-6</td>
<td>(a) (mils)</td>
<td>303- ( l )</td>
<td>283</td>
<td>240- ( l )</td>
<td>50</td>
<td>variable</td>
<td>10</td>
</tr>
<tr>
<td>Dimensions</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>for Figure 2-6</td>
<td>(b) (mils)</td>
<td>313</td>
<td>283</td>
<td>230</td>
<td>50</td>
<td>variable</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 2-3 Dimensions for Figure 2-6
Another extreme case that is worth noticing is that when the two gap-openings are on the middle of the arm and directly opposite each other, i.e. $l = \lambda_g/8$, represented by $l_1$, $l_2$, $l_4$ and $l_5$ in Figure 2-8. The dimensions are given in Table 2-4. Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, $h=25$ mil and relative dielectric constant $\varepsilon_r=10.2$. The frequency response is shown in Figure 2-9. There is a great loss in energy at the transmission zeros and the two transmission zeros have merged to one at the fundamental frequencies, around 4.8GHz. The loss in energy at the fundamental mode is due to the fact that the field is no longer “confined” within the
loop. The merging of the transmission zeros is due to the loss of “orthogonal nature” of the configuration. In latter sections, the orthogonal nature of a design to generate dual mode will be discussed.

Figure 2-8 Configuration with the two gap-openings directly opposite

<table>
<thead>
<tr>
<th>Item</th>
<th>$l_1$</th>
<th>$l_2$</th>
<th>$l_3$</th>
<th>$l_4$</th>
<th>$l_5$</th>
<th>$l_6$</th>
<th>$g$</th>
<th>$w$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions (mils)</td>
<td>187</td>
<td>137</td>
<td>283</td>
<td>114</td>
<td>114</td>
<td>50</td>
<td>12</td>
<td>23</td>
</tr>
</tbody>
</table>

Table 2-4 Dimensions for Figure 2-8
2.4 Equivalent circuit analysis on the square open-loop resonator

With the investigation performed in the previous section, there is a need to understand those behaviors. One way of analyzing the structures is actually to view it as two “thongs” coupling with each other. The equivalent circuit of this configuration is shown in Figure 2-10.
From Figure 2-10, it can be seen that the filter design can be separated into several critical components for modeling. The design equations for these can be referenced from earlier works performed by other researchers [7]-[8]. The equations of some important elements are being reproduced in the following sub-sections.
2.4.1 Modeling the gap

The capacitances [10] making up the π network are made up of $C_p$ and $C_g$, and they are respectively expressed as

$$C_p = 0.5C_e$$  \hspace{1cm}  (2-1)$$

and

$$C_g = 0.5C_o - 0.25C_o,$$  \hspace{1cm}  (2-2)$$

where,

$$\frac{C_o}{W}(pF/m) = \left(\frac{\varepsilon_r}{9.6}\right)^{0.8}\left(\frac{s}{W}\right)^{m_o}\exp(k_o),$$  \hspace{1cm}  (2-3)$$

$$\frac{C_e}{W}(pF/m) = 12\left(\frac{\varepsilon_r}{9.6}\right)^{0.9}\left(\frac{s}{W}\right)^{m_e}\exp(k_e),$$  \hspace{1cm}  (2-4)$$

with $m_o$, $k_o$ and $m_e$, $k_e$ being given by

$$m_o = \frac{W}{h}\left[0.619\log\left(\frac{W}{h}\right) - 0.3853\right], \text{ for } 0.1 \leq \frac{s}{W} \leq 1.0,$$  \hspace{1cm}  (2-5)$$

$$k_o = 4.26 - 1.53\log\left(\frac{W}{h}\right).$$

$$m_e = 0.8675$$  \hspace{1cm}  (2-6)$$

$$k_e = 2.043\left(\frac{W}{h}\right)^{0.12}, \text{ for } 0.1 \leq \frac{s}{W} \leq 0.3,$$  \hspace{1cm}  (2-7)$$

$$m_e = \frac{1.565}{(W/h)^{0.16}} - 1,$$  \hspace{1cm}  (2-7)$$

$$k_e = 1.97 - \frac{0.03}{(W/h)}, \text{ for } 0.3 \leq \frac{s}{W} \leq 1.0.$$
2.4.2 Modeling the bend

The equations for modeling the bend, as referenced from [10] are given below:

The C and L making up the T-network in this equivalent circuit [10] are respectively given as:

\[
C_W (pF/m) = \frac{(14\varepsilon_r + 12.5)(W/h) - (1.83\varepsilon_r - 2.25) + 0.02\varepsilon_r \text{ for } (W/h) < 1}{\sqrt{(W/h)}} + \frac{0.02\varepsilon_r \text{ for } (W/h) \geq 1}{(W/h)},
\]

and

\[
L_W (nH/m) = 100 \left\{ 4\sqrt{\frac{W}{h}} - 4.21 \right\}.
\]

2.4.3 Modeling the open end

The open end is modeled by \( C_t \) (as referenced from [10]) and is expressed as

\[
C_t = \frac{(\Delta l)\sqrt{\varepsilon_{re}}}{cZ_c}, \text{ where}
\]

\( c \) is the speed of light,

\( Z_c \) is the characteristic impedance and

\[
\frac{\Delta l}{h} = \frac{\xi_1\xi_2\xi_3\xi_5}{\xi_4},
\]

with \( \xi_1, \xi_2, \xi_3, \xi_4 \) and \( \xi_5 \) being given by:

\[
\xi_1 = (0.434907)(\frac{\varepsilon_{re}^{0.81} + 0.26(W/h)^{0.8544} + 0.236}{\varepsilon_{re}^{0.81} - 0.189(W/h)^{0.8544} + 0.87}),
\]
\[ \xi_2 = 1 + \frac{(W / h)^{0.371}}{2.35 \varepsilon_r + 1}, \]  
\[ \xi_3 = 1 + \frac{0.5274 \tan^{-1} \left[ 0.084(W / h)^{1.9413 / \xi_i} \right]}{\varepsilon_r^{0.9236}}, \]  
\[ \xi_4 = 1 + 0.037 \tan^{-1} \left[ 0.067(W / h)^{1.456} \left( 6 - 5 \exp \left[ 0.036(1 - \varepsilon_r) \right] \right) \right], \]  
\[ \xi_5 = 1 - 0.218 \exp \left( -7.5 \frac{W}{h} \right). \]

### 2.4.4 Comparisons between equivalent circuit and momentum

With these desired lumped circuit-elements, the equivalent circuit in Figure 2-10 can be evaluated. Figure 2-11 shows a comparison between the calculated frequency response of the equivalent circuit and the simulated results using momentum method on the microstrip line model. The calculated results match quite closely with that using momentum. Therefore, the equivalent circuit can be used to analyze the configuration of the square open loop resonator with two gap-openings. The graph also shows the difference between modeling the configuration with and without the open end capacitance, \( C_t \). The transmission zeros (S21 is maximum) are more prominent with \( C_t \) and there is a smaller shift in resonant frequencies.
All the designs studied in this thesis are made on a RT/Duroid 6010 substrate with a thickness, $h=25$ mil and relative dielectric constant $\varepsilon_r=10.2$. With this substrate, the width of a $50\,\Omega$ line is about 23 mil. The length of one arm, $l$ is chosen to be $\lambda_g/4$. The feed line is chosen to have a length of $\lambda_g/4$. The frequency performance of the design shown in Figure 1-5 has a feed line length of 50 mil. It is represented by $l_4$ in that figure. Simulation results in Figure 2-12 shows that the feed line length does not affect much on the performance of the filter. Despite the increased conductor loss, the longer feed lines allow for easier hardware implementation.
2.5 Tuning the two Transmission Zeros

In the filter design specifications, besides the resonant frequencies, the transmission zero is also an important parameter. By evaluating the ABCD parameters of the equivalent circuit, the two transmission zeros can be found [6]. They are defined as the frequency response when $S_{21}$ is at a maximum. The equivalent circuit for the square single open loop resonator with two gap-openings is reproduced in Figure 2-13 below. It can be divided into the upper and lower sections. Each section comprises of $l_6$, the π section due to the gap, the T section due to the bend, the Ct due to the open end, and $l_1$ and $l_2$. 

Figure 2-12 Filter performance with short and long feed lines
2.5.1 ABCD parameters

The following equations are used to calculate the ABCD parameters of the equivalent circuit so as to tune the transmission zeros.

The ABCD matrix for the upper portion is given by:

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}_{\text{upper}} = M_1 M_2 M_3 M_4 M_5, \tag{2-17}
\]

and that of the lower portion is given by:

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}_{\text{lower}} = M_5 M_4 M_3 M_2 M_1, \tag{2-18}
\]

where the \( M_i \) correspond to the ABCD matrix of the different parts shown in Figure 2-13 and are given as

\[
M_1 = \begin{bmatrix}
\cos \beta l_1 & j z_o \sin \beta l_1 \\
-j y_o \sin \beta l_1 & \cos \beta l_1
\end{bmatrix}. \tag{2-19}
\]
where \( \beta \) is the propagation constant,

\[ z_o \] is the characteristic impedance of the resonator and

\[ y_o = 1/z_o. \]

Using the derived ABCD matrices for the upper and lower sections, the Y parameters for the overall circuit case can be obtained. They are given below as:

\[
Y_{upper} = \begin{bmatrix}
\frac{D_{upper}}{B_{upper}} & \frac{B_{upper} C_{upper} - A_{upper} D_{upper}}{B_{upper}} \\
\frac{B_{upper}}{B_{upper}} & \frac{B_{upper}}{B_{upper}} \\
\frac{-1}{B_{upper}} & \frac{A_{upper}}{B_{upper}} \\
\frac{B_{upper}}{B_{upper}} & \frac{B_{upper}}{B_{upper}}
\end{bmatrix}
\]
\[
Y_{\text{lower}} = \begin{bmatrix}
D_{\text{lower}} & B_{\text{lower}} C_{\text{lower}} - A_{\text{lower}} D_{\text{lower}} \\
B_{\text{lower}} & -1 \\
-1 & B_{\text{lower}} \\
B_{\text{lower}} & A_{\text{lower}}
\end{bmatrix}
\]

\[Y = Y_{\text{upper}} + Y_{\text{lower}}\]

The transmission zeros can be found by letting \( S_{21} = 0 \text{ dB} \). Therefore, The \( S_{21} \), in terms of the y-parameters, is evaluated as

\[S_{21} = \frac{-2Y_{21} Y_o}{(Y_{11} + Y_o)(Y_{22} + Y_o) - Y_{12} Y_{21}}\]

and the transmission zeros are given as

\[Y_{21} = Y_{21,\text{upper}} + Y_{21,\text{lower}} = \frac{-1}{B_{\text{upper}}} + \frac{-1}{B_{\text{lower}}} = 0\]

### 2.5.2 Calculation results by applying the equations

From the above equations and those given in section 2.4, the lumped circuit-elements, \( M_2, M_3 \) and \( M_4 \) can be found and they are tabulated in Table 2-6. The centre frequency used here is 5Hz. This frequency is chosen because the designs shown in previous sections have resonant frequencies between 4 to 5GHz.
<table>
<thead>
<tr>
<th>Item</th>
<th>Values</th>
<th>Item</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_t$</td>
<td>0.1136pF</td>
<td>$M_2$</td>
<td>$\begin{bmatrix} 0.555 &amp; -j1493 \ j0.002 &amp; 0.555 \end{bmatrix}$</td>
</tr>
<tr>
<td>$C_p$</td>
<td>0.0596pF</td>
<td>$M_3$</td>
<td>$\begin{bmatrix} 1.002 &amp; -j0.742 \ j5.68\times10^{-4} &amp; 1.002 \end{bmatrix}$</td>
</tr>
<tr>
<td>$C_g$</td>
<td>0.134pF</td>
<td>$M_4$</td>
<td>$\begin{bmatrix} 0.217 &amp; j3.267 \ j0.292 &amp; 0.217 \end{bmatrix}$</td>
</tr>
<tr>
<td>$L$</td>
<td>0.5363nF</td>
<td>$M_2 \ M_3 \ M_4$</td>
<td>$\begin{bmatrix} 437.13 &amp; -j646.38 \ j0.1629 &amp; 0.2347 \end{bmatrix}$</td>
</tr>
<tr>
<td>$C$</td>
<td>58.4pF</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 2-6 Computed values for the ABCD matrix

Figure 2-14 Design reference for the equivalent circuit
The first step in designing the resonator is to let $l_1 = l_2$, as shown in Figure 2-14. The results shown in Figure 2-11 are obtained by setting $l_1$ and $l_2$ to be equal to $\lambda_g/4$. Using $l_1$ and $l_2$ to be equal to $\lambda_g/4$, $Y_{21}$ is found to be approximately 0dB. This is the parameter to calculate the transmission zeros ($S_{21}$ is maximum) of the design. CAD software like Agilent's momentum or ADS can be used to optimize the circuit to get the required transmission zeros. It should be noted that the gap-openings cannot be greater than 50Ω line width. Therefore, the range for optimization is set.

2.6 Tuning the attenuation poles

As mentioned the introductory chapter, one attractive feature of the one-wavelength resonator is that it allows dual mode resonator circuits to be designed easily. The most common configurations of these designs are depicted in Figure 1-2 and Figure 1-3. There are some general conditions reported in [11] to realize a dual mode filter using one-wavelength ring resonator. They are

1) Input and output have to be spatially separated at 90° intervals.

2) There has to be a discontinuity to generate a reflected wave against the incident wave within the resonator.

3) There should be symmetry within the circuit geometry.
In the design shown in Figure 1-5, the perturbations are actually the gap-openings. There exists symmetry along the diagonal of the circuit. However, it is obvious that the input and output ports are not orthogonal to each other.

### 2.6.1 Concept of traveling waves

The concept of traveling wave [12] will help to understand why the circuit in Figure 1-5 needs the input and output to be at $180^\circ$ to achieve the dual mode coupling. Figure 2-15 shows the way the waves travel. The thicker lines represent the clockwise-traveling wave from the input. It travels $90^\circ$, represented by length $l$, to reach a discontinuity, namely the gap-opening. It is being reflected and continues to travel counter-clockwise, passes the input and reaches another discontinuity. However, this discontinuity also serves as a coupling session to couple some energy to the output port. The total path traveled by the wave is $360^\circ$. Similarly, counter-clockwise-traveling wave from the input, represented by the thinner line, will first reach a discontinuity and being reflected. This wave will then be reaching the output port by coupling. These two possible paths allow the combination of two orthogonal resonant modes.
Figure 2-15 Directions of wave travel to generate the dual mode

An illustration is given below to verify this concept. The configuration is shown in Figure 2-16. It has the input and output arranged orthogonally. The dimensions are shown in Table 2-7.
Figure 2-16 Configuration to study the orthogonal feed of the square loop resonator

<table>
<thead>
<tr>
<th>Item</th>
<th>$l_1$</th>
<th>$l_2$</th>
<th>$l_3$</th>
<th>$l_4$</th>
<th>$l_5$</th>
<th>$l_6$</th>
<th>$g$</th>
<th>$w$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions</td>
<td>253</td>
<td>260</td>
<td>253</td>
<td>260</td>
<td>50</td>
<td>50</td>
<td>10</td>
<td>23</td>
</tr>
</tbody>
</table>

Table 2-7 Dimensions for Figure 2-16

Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, $h=25$ mil and relative dielectric constant $\varepsilon_r=10.2$. By using the traveling wave concept, it can be observed that counter-clockwise-traveling wave reaches the gap-opening and is being reflected. In this path, it has traveled $180^\circ$. It then passes the input port and travels another $90^\circ$ to reach the second gap-opening just before the output port. The total
path traveled by the wave is $270^\circ$ and this is out of phase. This path is represented by the thinner line in Figure 2-17. The second possible path, represented by the thicker lines, is the clockwise-traveling wave. It travels $90^\circ$ to reach the gap-opening just before the output. Then, it is reflected back, passes through the input port and reaches the second discontinuity. Part of the energy is coupled forward to the second arm and travels another $180^\circ$ to the output port. Because of these two coupling processes, there is also a greater amount of energy loss. The loss is about -2dB with the simulation being executed on a lossless conductor. This is obvious in the smaller magnitude of the S21 as compared to that shown in Figure 1-5. The comparison is being shown in Figure 2-17.

Another important point obtained from the illustration shown in Figure 2-17 is that there is no dual mode in this configuration. There is only one resonant frequency at around 4.5GHz as compared to the two distinct resonant frequencies at 4.4GHz and 4.6GHz. This can be understood from the aforementioned two traveling paths of the wave. The counter-clockwise wave travels $270^\circ$, thus the field at the output due to this wave becomes zero. There is only the field due the clockwise traveling wave, resulting in single fundamental mode performance.
2.6.2 Even and odd mode analysis

This analysis method is often employed to simplify the circuit under investigation. In the circuit presented in Figure 1-5, there is also a “line of symmetry”. It is the diagonal line of symmetry that divides the configuration into two halves for odd and even mode analysis. This is shown in Figure 2-18.
The ABCD parameter matrix is given below:

\[
\begin{bmatrix}
V_1 \\
I_1 \\
\end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\
I_2 \\
\end{bmatrix}.
\]  \hspace{1cm} (2-29)

Under \textbf{even mode} condition, equal potentials are applied to each end of the circuit. There is an open circuit along the line of symmetry.

\[V_2 = V_1,\]  \hspace{1cm} (2-30)
\[ I_2 = -I_1. \]  \hfill 2-31

Now,
\[
\begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\
I_2
\end{bmatrix},
\]
\hfill 2-32
\[ V_1 = AV_2 + BI_2, \]  \hfill 2-33
\[ I_1 = CV_2 + DI_2, \]  \hfill 2-34
\[ V_1 = AV_1 - BI_1, \]  \hfill 2-35
\[ Y_e = \frac{A - 1}{B}. \]  \hfill 2-36

Under **odd mode** condition, opposite potentials are applied to each end of the circuit. There is a short circuit along the line of symmetry.
\[ V_2 = -V_1, \]  \hfill 2-37
\[ I_2 = I_1. \]  \hfill 2-38

Now, the odd mode admittance:
\[
\begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\
I_2
\end{bmatrix},
\]
\hfill 2-39
\[ V_1 = AV_2 + BI_2, \]  \hfill 2-40
\[ I_1 = CV_2 + DI_2, \]  \hfill 2-41
\[ V_1 = -AV_1 + BI_1, \quad 2-42 \]
\[ Y_o = \frac{1 + A}{B}. \quad 2-43 \]

From equations 2-36 and 2-43, the ABCD parameters of the overall circuit can be calculated as

\[ A = \frac{Y_e + Y_o}{Y_o - Y_e} = D, \quad 2-44 \]
\[ B = \frac{2}{Y_o - Y_e}, \quad 2-45 \]

And from reciprocity and symmetry,

\[ A^2 - BC = 1. \quad 2-46 \]

Hence,

\[ C = \frac{2Y_eY_o}{Y_o - Y_e}. \quad 2-47 \]

Therefore,

\[ [A \ B] = \begin{bmatrix} \frac{Y_e + Y_o}{Y_o - Y_e} & 2 \\ \frac{Y_o - Y_e}{Y_o + Y_o} & \frac{Y_o + Y_o}{Y_o + Y_o} \end{bmatrix}. \quad 2-48 \]

Using these even and odd mode circuits, it presents another way of obtaining the ABCD parameters through equivalent circuits similar to Figure 2-13. Figure 2-19 shows the equivalent circuit for even mode analysis. The open end can be modeled.
by Ct. L(arm) represents the inductance of the open-end transmission line, whereas the L(feed) represents that of the feed line.

![Figure 2-19 Equivalent circuit for even mode analysis](image)

The impedance looking into the circuit as indicated in Figure 2-19 can be calculated as

\[
Z_{\text{evenarm}} = j\omega L_{\text{arm}} = Z_o \frac{1}{j\omega C_t} + jZ_o \tan(\beta l),
\]

\[
Y_{\text{evenarm}} = \frac{1}{Z_{\text{evenarm}}},
\]

\[
Y_{\text{evencircuit}} = \left[Z_{\text{evenarm}} + [j\omega C_t]^{-1}\right]^{-1} + \left[j\omega L + [j\omega C + [j\omega L + Z_{\text{evenarm}} + [j\omega C_t]^{-1}]^{-1}]^{-1}\right],
\]

\[
Z_{\text{evencircuit}} = \frac{1}{Y_{\text{evencircuit}}},
\]

The overall impedance can then be found to be
The values of $C_t$, $C$ and $L$ can be found by the modeling equations discussed in section 2.4.

Similarly, the equivalent circuit for odd mode analysis is shown in Figure 2-20.

![Figure 2-20 Equivalent circuit for odd mode analysis](image)

The circuit is simpler than that for even mode analysis because of the absence of the capacitance to account for the open end. Again, the impedance of the circuit can be calculated as:

$$Y_{\text{odd circuit}} = Y_{\text{odd arm}} + \left[ j \omega L + \left[ j \omega C + \left[ Z_{\text{odd arm}} + j \omega L \right]^{-1} \right]^{-1} \right],$$  \hspace{1cm} 2-55

$$Z_{\text{odd circuit}} = \frac{1}{Y_{\text{odd circuit}}},$$  \hspace{1cm} 2-56

where
\[ Z_{\text{oddarm}} = jZ_o \tan \beta l, \]  

2-57

\[ Z_{\text{oddoverall}} = Z_o \frac{Z_{\text{odd circuit}} + jZ_o \tan(\beta l_{\text{feedline}})}{Z_o + jZ_{\text{odd circuit}} \tan(\beta l_{\text{feedline}})}, \]  

2-58

\[ Y_o = \frac{1}{Z_{\text{oddoverall}}}. \]  

2-59

With these \( Y_e \) and \( Y_o \), the ABCD parameters can be found. By setting \( l = \frac{\lambda_e}{4} \), the calculated values for the above parameters are:

\[ Y_e = j0.0065, \]

\[ Y_o = j0.0065, \]

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = \begin{bmatrix}
8.5854 \times 10^4 & -j1.5331 \times 10^2 \\
5.6002 \times 10^3 & 8.5854 \times 10^4
\end{bmatrix}.
\]

The even and odd mode analysis is used to investigate the attenuation poles of the dual mode resonator [12]. In order to get the attenuation poles, the \( S_{21} \) is set to a minimum. Using \( l = \frac{\lambda_e}{4} \) for all the 4 transmission arms, \(|S_{21}|\) is found to be 0.0065. This shows that using \( l = \frac{\lambda_e}{4} \) as the starting point for optimization is justified. The above even and odd equivalent models can be used with CAD software to fine-tune the attenuation poles. The lengths of the arms connected and adjacent to each other can be set to be a small deviation of each other.
3 Dual mode resonator using coupled lines

3.1 Introduction

The design configuration shown in Figure 1-5 has distinct dual mode resonant frequencies and sharp attenuation poles. Several ways have to be found to properly couple these two resonance modes to allow a bandpass performance. Extensive works have been done on the square open-loop resonators shown in Figure 2-1 to investigate the cross-coupling effects [13].

From Figure 1-5, it can be seen that the configuration can generate two resonant modes instead of using two loops like the square open-loop resonator with one gap-opening. This gives the motivation to improve the design on the configuration with two gap-openings and to further utilize the space within the resonator. This suits the needs of modern communication applications where space is a valuable asset.

In this chapter, a coupled line design making use of the space inside the square will be presented. Coupled lines have conventionally been implemented to systematically design bandpass filter. Therefore, the use of coupled lines to be incorporated into the configuration shown in Figure 1-5 in the previous section is worth investigating. In addition, the cross coupling effects of the square open loop resonators with one gap-opening have been analyzed using the concept of electric, magnetic and mixed couplings. The configurations for these couplings are shown in Figure 3-1. Configurations for the coupled line case will also be designed and
analyzed in a similar manner. These coupled line designs are fabricated and the measured results are compared against the simulation results.

Figure 3-1 Coupling effects for the single square loop resonator
3.2 Coupled lines Loop Resonator

In an attempt to further utilize the space within the square loop resonator, the circuit configuration shown in Figure 3-2 is proposed. Its dimensions are given in Table 3-1. In this design, the single loop from the previous chapter has another square loop with two gap-openings connected to it. These two loops are connected by two small “stubs”, as included in the length $l_5$. This circuit achieves the mixed coupling.

![Figure 3-2 New square open loop resonator configuration for bandpass response](image)

<table>
<thead>
<tr>
<th>Item</th>
<th>$l_1$</th>
<th>$l_2$</th>
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<th>$l_4$</th>
<th>$l_5$</th>
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<th>$w$</th>
<th>$g_1$</th>
<th>$g_2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions</td>
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<td>283</td>
<td>88.5</td>
<td>157</td>
<td>66</td>
<td>88.5</td>
<td>50</td>
<td>20</td>
<td>23</td>
<td>20</td>
<td>10</td>
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<tr>
<td>(mils)</td>
<td></td>
<td></td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 3-1 Dimensions for Figure 3-2
For the electric coupling configuration, it is presented in Figure 3-3 and the dimensions for this configuration are given in Table 3-2. This configuration implements the electric coupling in a similar manner as in Figure 3-1 (a). The electric cross coupling is achieved between the upper and lower portion of the coupled line resonator. Again, the inner and outer loops are connected. The feeding lines, as represented by $l_5$, are disconnected from the resonator in order to obtain the resonant frequencies of the resonator structure only.

![Figure 3-3 Configuration for the electric coupling](image)

<table>
<thead>
<tr>
<th>Items</th>
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<th>$l_2$</th>
<th>$l_3$</th>
<th>$l_4$</th>
<th>$g$</th>
<th>$s$</th>
<th>$w$</th>
</tr>
</thead>
<tbody>
<tr>
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<td>130</td>
<td>184</td>
<td>269</td>
<td>10</td>
<td>20</td>
<td>23</td>
</tr>
</tbody>
</table>

Table 3-2 Dimensions for Figure 3-3
Figure 3-4 shows the magnetic coupling configuration and the dimensions are given in Table 3-3. The magnetic coupling is achieved between the arms of the inner and outer loops. This is similar to that in Figure 3-1b. The inner and outer loops are again connected together by small “stubs”. The feeding lines are also disconnected to the resonator structure in order to get the resonant frequencies due to the resonator structure alone.

<table>
<thead>
<tr>
<th>Items</th>
<th>$l_1$</th>
<th>$l_2$</th>
<th>$l_3$</th>
<th>$l_4$</th>
<th>$l_5$</th>
<th>s</th>
<th>w</th>
<th>$g_1$</th>
<th>$g_2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions (mil)</td>
<td>253</td>
<td>283</td>
<td>197</td>
<td>167</td>
<td>50</td>
<td>20</td>
<td>23</td>
<td>10</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 3-3 Dimensions for Figure 3-4
3.3 Closed Coupled Inner Loop

Choosing the type of the coupled line is the first step in arriving at the configuration shown in Figure 3-2. The coupled line section can form different two-port networks by terminating two of the four ports in either open or short circuits [14]. The simplest is to implement a closed inner loop to the original configuration. This is presented in Figure 3-6 and the dimensions used in the configurations are presented in Table 3-4. The s/h ratio is chosen to be 0.5 for K=6.0, where s is the spacing between the inner and outer loop and h is the height of the substrate as shown in Figure 3-5. The even and odd modes characteristic impedances are 50ohms. This can be obtained from either ADS or reference [15].

The value of gap-opening, g can also be altered. Generally, the bigger the gap, the more losses at the resonant frequencies will be resulted.

Figure 3-5 s/h representation
Figure 3-6 Configuration for a closed inner loop

<table>
<thead>
<tr>
<th>Items</th>
<th>$l_1$</th>
<th>$l_2$</th>
<th>$l_3$</th>
<th>$l_4$</th>
<th>$w$</th>
<th>$s$</th>
<th>$g$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions (mils)</td>
<td>303</td>
<td>283</td>
<td>197</td>
<td>200</td>
<td>23</td>
<td>20</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 3-4 Dimensions for Figure 3-6
Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, \( h=25 \text{mil} \) and relative dielectric constant \( \varepsilon_r=10.2 \). Figure 3-7 shows the frequency response comparison with Figure 1-5, it can be observed that the insertion loss at the pass band is smaller, -20dB, as compared to -22dB for the single loop case. The cutoff for the attenuation is also sharper with the closed coupled line. This confirms the speculation earlier on that coupled line helps to improve the pass band response. However, this performance is not satisfactory enough, as the insertion loss in the frequency range between the two resonant modes is still high. This may be improved by converting the closed inner loop to open loop.

![Figure 3-7 Frequency response with and without the closed inner coupled loop](image-url)
3.4 Inner loop with a gap-opening

When the inner loop is opened, the configuration is given in Figure 3-8 and its dimensions are given in Table 3-5. The gap-opening is located on the top portion of the inner loop. The size of the gap-opening can be adjusted. All the other parameters are unaltered.

![Configuration with the inner loop having one gap-opening](image)

**Table 3-5 Dimensions for Figure 3-8**

<table>
<thead>
<tr>
<th>Items</th>
<th>( l_1 )</th>
<th>( l_2 )</th>
<th>( l_3 )</th>
<th>( l_4 )</th>
<th>( l_5 )</th>
<th>( w )</th>
<th>( s )</th>
<th>( g_1 )</th>
<th>( g_2 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions (mils)</td>
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<td>283</td>
<td>197</td>
<td>93</td>
<td>98.5</td>
<td>23</td>
<td>20</td>
<td>8.5</td>
<td>10</td>
</tr>
</tbody>
</table>
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r=10.2$. Figure 3-9 shows the frequency response comparison with the configuration in Figure 3-6. The lower frequency resonant mode has greater insertion loss than before. This can be explained by the fact that some field energy is being directed into the inner portion of the inner loop, leaving less energy to be coupled to the output terminal. The fact that this only affects one resonant frequency suggests that the position of the gap-opening plays a part in this. Therefore, the next investigation will be on the position of the gap-opening. From Figure 3-9, it can also be deduced that the inner gap has a greater influence on the lower frequency resonant mode.

Figure 3-9 Frequency response when a gap-opening is created in the inner loop
The configuration used in the investigation on the effect of the position of the gap-opening is shown in Figure 3-10. The dimensions are being described in Table 3-6. The configuration is similar to the one used in Figure 3-8. The length $l$ is varied by moving it along the horizontal side on the inner square loop as described in Table 3-7.

![Figure 3-10 Configuration to study the position of the gap-opening](image)

<table>
<thead>
<tr>
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<th>$s$</th>
<th>$g_1$</th>
<th>$g_2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions</td>
<td>303</td>
<td>283</td>
<td>197</td>
<td>93</td>
<td>168.5</td>
<td>$l$</td>
<td>23</td>
<td>20</td>
<td>8.5</td>
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<td>(mils)</td>
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<td></td>
<td></td>
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</tr>
</tbody>
</table>

Table 3-6 Dimensions for Figure 3-10
<table>
<thead>
<tr>
<th>Item</th>
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<tbody>
<tr>
<td>( l = 70 ) mil</td>
<td>gap is at the middle of the inner horizontal loop</td>
</tr>
<tr>
<td>( l = 0 ) mil</td>
<td>gap is on the left hand edge of the inner horizontal arm</td>
</tr>
<tr>
<td>( l = 40 ) mil</td>
<td>gap is on the left hand side of the inner horizontal arm</td>
</tr>
<tr>
<td>( l = 100 ) mil</td>
<td>gap is on the right hand side of the inner horizontal arm</td>
</tr>
<tr>
<td>( l = 145.5 ) mil</td>
<td>gap is on the right hand edge of the inner horizontal arm</td>
</tr>
<tr>
<td>( l = 249 ) mil</td>
<td>gap is on the middle of the vertical side of the inner horizontal arm</td>
</tr>
</tbody>
</table>

Table 3-7 Representation of the variations of the gap

Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, \( h=25 \)mil and relative dielectric constant \( \varepsilon_r=10.2 \). The frequency responses are shown in Figure 3-11. The two resonant frequencies have the lowest energy loss (when \( S_{21}\sim0\)dB) when \( l = 0 \).

The investigation shows that by centering the gap-opening on either the horizontal or vertical arm, similar performance can be obtained. By positioning the gap on the left hand side of the horizontal arm, the insertion loss on the lower resonant frequency is compensated. The result is best when the gap is on the left hand edge of the inner horizontal arm.
Figure 3-11 Effects of varying the position of the gap

This behavior can be understood using the coupled line theory. By positioning the gaps at the inner and outer loops further, the coupled line length is closer to $90^\circ$, as shown in Figure 3-12. The dual mode is again present as in the single loop case.
3.5 Inner loop with two gaps

The outer loop has been implemented by including two gap-openings to introduce a new type of perturbation for the dual mode resonator. In this section, the possibility to introduce two gap-openings on the inner loop will also be studied. By introducing another gap on the inner loop, there are more perturbations on the path of the wave travels. The following configurations are being studied.

![Configuration Diagram](image)

**Figure 3-13 Configuration 1 in Figure 3-17**

<table>
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<tr>
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</thead>
<tbody>
<tr>
<td>Dimensions (mils)</td>
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<td>283</td>
<td>93</td>
<td>197</td>
<td>93</td>
<td>23</td>
<td>20</td>
<td>10</td>
</tr>
</tbody>
</table>

**Table 3-8 Dimensions for Figure 3-13**
In Figure 3-13, the two gap-openings are located directly opposite to each other. Figure 3-14 shows the configuration with inner loop closed. The dimensions are also the same as in Figure 3-6. This is for comparison between inner loops with and without two gap-openings. Figure 3-15 and Figure 3-16 show two more possible positions of the two gap-openings in the inner loop.

![Figure 3-14 Configuration 2 in Figure 3-17 (same as in Figure 3-6)](image)

<table>
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<tbody>
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Table 3-9 Dimensions for Figure 3-14
Figure 3-15 Configuration 3 in Figure 3-17

<table>
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<th>$s$</th>
<th>$g$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions (mils)</td>
<td>303</td>
<td>283</td>
<td>197</td>
<td>167</td>
<td>23</td>
<td>20</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 3-10 Dimensions for Figure 3-15
Figure 3-16 Configuration 4 in Figure 3-17

<table>
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<th>$w$</th>
<th>$s$</th>
<th>$g$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions (mils)</td>
<td>303</td>
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<td>197</td>
<td>93</td>
<td>23</td>
<td>20</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 3-11 Dimensions for Figure 3-16
Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, \( h=25 \text{mil} \) and relative dielectric constant \( \varepsilon_r=10.2 \). Figure 3-17 shows the investigations on the above configurations for two inner gaps and the comparison with a single inner gap.

Figure 3-17 Effects of having two gaps on the inner loop and their various positions

Compared with the closed inner loop, the inner loop with two gaps separates the two resonant modes further apart. This means that the inner loop with two gaps provides the potential to design bandpass filter with wider bandwidth. The arrangement of the pairs of the gaps on the inner and outer loop as shown in configuration 3 in Figure 3-17 gives the widest bandwidth. These results agree with the traveling wave analysis discussed earlier on in section 3.4.
3.6 Connecting the inner and outer loops

Simulation results in the previous section shows that the inner loop can improve the insertion loss on the range of frequencies between the two resonant modes. However, this is still not satisfactory enough for a bandpass filter performance. One possible solution is to connect the inner and outer loops together. Figure 3-18 to Figure 3-22 show some configurations for investigations.

![Diagram of Configurations](image)

**Figure 3-18 Configuration 1 in Figure 3-23**

<table>
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<th>$w_2$</th>
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</table>

**Table 3-12 Dimensions for Figure 3-18**
Figure 3-19 Configuration 2 in Figure 3-23

<table>
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<th>$w$</th>
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Table 3-13 Dimensions for Figure 3-19
Figure 3-20 Configuration 3 in Figure 3-23

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Table 3-14 Dimensions for Figure 3-20
Figure 3-21 Configuration 4 in Figure 3-23 (same as Figure 3-2)

<table>
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<th>$s$</th>
<th>$w$</th>
<th>$g_1$</th>
<th>$g_2$</th>
</tr>
</thead>
<tbody>
<tr>
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<td>88.5</td>
<td>157</td>
<td>66</td>
<td>88.5</td>
<td>50</td>
<td>20</td>
<td>23</td>
<td>20</td>
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</tbody>
</table>

Table 3-15 Dimensions for Figure 3-21
Figure 3-22 Configuration 5 in Figure 3-23 (same as Figure 3-15)

<table>
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<th>$l_4$</th>
<th>$w$</th>
<th>$s$</th>
<th>$g$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions (mils)</td>
<td>303</td>
<td>283</td>
<td>197</td>
<td>167</td>
<td>23</td>
<td>20</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 3-16 Dimensions for Figure 3-22

Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, $h=25$ mil and relative dielectric constant $\varepsilon_r=10.2$. Figure 3-23 and Figure 3-24 give the frequency responses. There is a pass band filter response for configurations 1 to 4, showing that connecting the inner and outer loops does provide the cross coupling required for bandpass filter response.
Figure 3-23 S21 response for studying the positions to connect the inner and outer loops

Figure 3-24 S11 response for studying the positions to connect the inner and outer loops
3.7 Investigating the coupling effect of the coupled structure

This chapter started off with the intention to investigate the possibility of using coupled line to improve the performance of the bandpass filter. In the previous subsections of this chapter, this has been proven. The next section will be on the coupling effects of the filter configuration designed in the previous sections.

3.7.1 Cross coupling effects on the working design

Having determined the design to be used for the connected coupled line resonator, it is of interest to find out the resonant frequencies of the filter configuration of Figure 3-2. This is shown in Figure 3-25. These resonant frequencies studies are done by tapping at the original feed lines positions and also by tapping at the sides of the filter configuration. With the feed lines not directly connected to the resonator structure, the resonant frequencies obtained are purely due to the configuration. When the feed lines are connected directly to the resonator structures, there are other loading effects and the resonant frequencies obtained are not purely due to the configuration alone. It can be seen that the overall configuration of Figure 3-2 has two resonant frequencies. The centre of these resonant frequencies is 3.7 GHz. The feeding positions have a slight effect on the resonant frequencies. It “spreads” the two resonant frequencies apart when tapping by the sides at the centre (red graph).
3.7.2 Cross coupling on half of the design

A closer look at the effect of coupling positions on the resonant frequencies is shown in Figure 3-26. The investigation is on half of the resonator in Figure 3-2. This half of the resonator has a resonant frequency of 3.7GHz, which tallies with that in Figure 3-25. All the three types of tapping shown do not affect the resonant frequency. By tapping at the corners and at the centre, the attenuation poles are a reflection of each other. By attaching the feed lines directly on the resonator, the energy has spread out more evenly to the neighboring frequencies of the resonant frequency. This also explains why the configuration in Figure 3-2 manages to exhibit a bandpass filter performance.
Figure 3-26 Investigating the effects of coupling positions on the resonant frequencies

### 3.8 Electric, magnetic and mixed couplings

Having done the full wave analysis using Agilent’s momentum software in the previous sections, there is a better insight into the general behavior of the resonator developed so far. It may be time to understand the configuration by looking into the equivalent circuit.

A well-known type of coupling analysis on the square open loop resonators has been done by J. S. Hong and M. J. Lancaster [13]. They have categorized three general types of coupling for the square open loop resonators, namely, electric, magnetic and mixed couplings. Using this concept, the respective coupling
coefficients can be found from the electric and magnetic resonant frequencies. Coupling coefficient tables can be generated for different types of spacing, dielectric constants, width of the microstrip lines etc. Through the value of coupling coefficient, filters with a specified Q can be designed. In the proposed filter configuration here, the couplings present are extremely complicated. There are much more variables that can be modeled and studied. Therefore, only some prominent parameters will be studied here.

The three types of couplings mentioned above can also be obtained from the coupled resonator developed in the previous sections. The filter configuration in Figure 3-2 contained several types of couplings in a single design. This makes it worth studying about.

From Figure 3-26, it can be seen that by connecting the feeding lines directly on the resonator structure merely spread energy around the resonant frequency. The resonant frequency itself does not shift position. Recalling that the resonant frequencies are obtained when S21 is a maximum and the S21 can be determined by the Y parameters, ABCD parameters and so on. It is therefore reasonable to analyze the configuration in Figure 3-2 by detaching the feeding line from the resonator design. The configurations and dimensions investigated are given in Figure 3-27 and Table 3-17.
Figure 3-27 Configurations for direct and indirect feeding

Table 3-17 Dimensions for Figure 3-27

<table>
<thead>
<tr>
<th>Item</th>
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<th>$l_2$</th>
<th>$l_3$</th>
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<th>$l_5$</th>
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<th>$s$</th>
<th>$w$</th>
<th>$g_1$</th>
<th>$g_2$</th>
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</thead>
<tbody>
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<td>88.5</td>
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<td>20</td>
<td>23</td>
<td>20</td>
<td>10</td>
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</tr>
<tr>
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<td>283</td>
<td>88.5</td>
<td>157</td>
<td>66</td>
<td>88.5</td>
<td>50</td>
<td>20</td>
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<td>20</td>
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</tr>
</tbody>
</table>
Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, $h=25$mil and relative dielectric constant $\varepsilon_r=10.2$. The wideband full wave responses of these two circuits are shown in Figure 3-28. From the graph, it can clearly be seen that the difference between the direct feed and indirect feed in this design is just that the direct feed brings up the $S_{21}$ graph at the pass band. The resonant frequencies remain unchanged here.

Figure 3-28 Wideband response for comparing the effects of direct and indirect feeding
3.8.1 Types of couplings found in the new filter configuration

Figure 3-29 shows the types of coupling that can be found in the new configuration. Both internal and external couplings exist. In addition, electric and magnetic couplings can also be found in the design. Therefore, this filter configuration is a mixed coupling configuration.

Figure 3-29 Types of couplings found in the newly proposed filter configuration
3.8.2 Electric coupling

Figure 3-30 Configurations to study the electric couplings
The coupled line designs for electric coupling are shown in Figure 3-30. Table 3-18 tabulates the dimensions used for these designs. These configurations are mainly designed based on the attempt to achieve electric coupling between the upper and lower portions of the loop. Each half of the inner and outer loops is grouped together as coupled lines.

Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r=10.2$. Figure 3-31 shows the full wave simulation response. It can be seen that the configuration in (a) exhibits an additional pair of resonant frequencies very close to the fundamental ones. The fundamental resonant frequencies are both at around 4.75GHz and 4.92GHz.

<table>
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<th>$s$</th>
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<tbody>
<tr>
<td>Dimensions for (a) (mil)</td>
<td>88.5</td>
<td>130</td>
<td>184</td>
<td>269</td>
<td>10</td>
<td>20</td>
<td>23</td>
<td>N. A.</td>
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<tr>
<td>Dimensions for (b) (mil)</td>
<td>88.5</td>
<td>130</td>
<td>184</td>
<td>269</td>
<td>10</td>
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<td>N. A.</td>
</tr>
<tr>
<td>Dimensions for (c) (mil)</td>
<td>88.5</td>
<td>130</td>
<td>184</td>
<td>269</td>
<td>10</td>
<td>20</td>
<td>23</td>
<td>23</td>
</tr>
</tbody>
</table>

Table 3-18 Dimensions for Figure 3-30
The wideband response for the above two configurations are presented in Figure 3-32. By looking at these wideband responses, it may be deduced that the configuration in (a) has stronger coupling because a second pair of resonant frequencies is located very near to the first. The figure also shows that if the connecting line between the inner and outer loops is placed at the corners as in configuration (c), the coupling is made weaker because the second pair of resonant frequencies is located further.
3.8.3 Magnetic coupling

The configuration to investigate magnetic coupling for the coupled line resonator is not easy to arrive at. Two possible configurations are shown in Figure 3-33 and the corresponding dimensions are given in Table 3-19. These arrangements are similar to the configurations used for investigating electric coupling in the previous section. One is a coupled resonator with inner and outer loops connected and the other without connections.
Figure 3-33 Configurations to investigate the magnetic coupling for coupled line resonator

<table>
<thead>
<tr>
<th>Items</th>
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<th>$l_6$</th>
<th>$l_7$</th>
<th>$w_1$</th>
<th>$w_2$</th>
<th>$s$</th>
<th>$g$</th>
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</thead>
<tbody>
<tr>
<td>Dimensions for (a) (mils)</td>
<td>130</td>
<td>283</td>
<td>87</td>
<td>184</td>
<td>87</td>
<td>130</td>
<td>50</td>
<td>23</td>
<td>N. A.</td>
<td>20</td>
<td>10</td>
</tr>
<tr>
<td>Dimensions for (b) (mils)</td>
<td>130</td>
<td>283</td>
<td>87</td>
<td>184</td>
<td>87</td>
<td>130</td>
<td>50</td>
<td>23</td>
<td>N. A.</td>
<td>20</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 3-19 Dimensions for Figure 3-33
Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, \( h = 25 \text{mil} \) and relative dielectric constant \( \varepsilon_r = 10.2 \). The pair of resonant frequencies exhibited by the two configurations is close but does not coincide with each other, just like the case with electric coupling. For structure (a) in Figure 3-34, the resonant frequencies are at 3.5GHz and 4.0GHz. The resonant frequencies for structure (b) are at 4.7GHz and 5.0GHz.

![S21 graph](image)

Figure 3-34 Frequency response to investigate magnetic coupling of the coupled resonator
It is understood that the magnetic coupling between the two loops is due to the close proximity of the conductor arms. Figure 3-35 reproduces the configuration shown in Figure 3-4. This configuration maximizes the coupled line length between the inner and outer loops. This resembles the case shown in Figure 3-1 for the single line square loops.

![Figure 3-35 Configuration to study the magnetic coupling for coupled lines resonator](image)

<table>
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<tr>
<th>Items</th>
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<th>$l_4$</th>
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<td>20</td>
<td>23</td>
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</tbody>
</table>

Table 3-20 Dimensions for Figure 3-35
Again, full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, $h=25\text{mil}$ and relative dielectric constant $\varepsilon_r=10.2$. The frequency response of the configuration is observed by tapping adjacent to the “arms”, as shown by the blue graph. It can be seen that the pair of resonant frequencies exhibited by this type of configuration and that configuration with connected inner and outer loops as shown in Figure 3-34 is very close. The fundamental resonant frequencies are both at around 3.6GHz and 4GHz. The resonant frequencies for the first harmonics are different for the case connected at the corner (configuration shown in Figure 3-33) and that connected at the side (configuration in Figure 3-35).

Figure 3-36 Further investigation into the configurations for magnetic coupling
3.8.4 Mixed coupling

For the mixed coupling, many configurations have already been studied in previous sections. The configuration shown in Figure 3-2 is being discussed here to look into the resonant frequencies by tapping at different parts of the resonator. Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant \( \varepsilon_r=10.2 \). As shown in the frequency response in Figure 3-29, there are both electric and magnetic couplings found in this configuration. The graph in black and in blue in Figure 3-37 tapped at the part where there are magnetic and electric couplings respectively. This is to explore the field at different parts of the resonator.

![S21](image)

**Figure 3-37** Investigating the mixed coupling of the configuration shown in Figure 3-2
3.9 Equivalent circuits for the three types of couplings

It is not easy to represent the new filter configuration using equivalent circuits and that the axis of symmetry is also not two dimensional. One possible way of finding the equivalent circuits for the three types of coupled structures developed is to use the left-handed and right-handed transmission lines method. Another way is by representing the capacitive coupling using J-inverter and the inductive coupling using K-inverter. Both these methods need to take into account of the close internal couplings within the neighboring lines.

For a quick analysis here, the equivalent circuits presented are treated as if the coupled line structure can be simplified to be represented by just one lumped-C and lumped-L. This can be understood by emphasizing that it is the cross coupling that is of interest. The coupled line itself contributes to the internal electric and magnetic couplings represented by the internal C and L respectively. By doing so, there will also be a line of symmetry to apply electric and magnetic walls to find out the resonant frequencies. The important parameter that is of interest is the coupling coefficients for the three types of couplings. These parameters are obtained from the measured S21 responses.
3.9.1 Equivalent circuit for electric coupling

![Equivalent Circuit for Electric Coupling](image)

Figure 3-38 Equivalent circuit for electric coupling

The resonant frequencies by applying electric and magnetic wall at the line of symmetry are given by [13]:

\[ f_e = \frac{1}{2\pi \sqrt{L_{lumped} (C_{lumped} + C_{mutual})}} \]
\[ f_m = \frac{1}{2\pi \sqrt{L_{lumped} (C_{lumped} - C_{mutual})}} \]

From the measured results, the electric coefficient, \( k_E \) is given by:

\[ k_E = \frac{f_m^2 - f_e^2}{f_m^2 + f_e^2} = \frac{4.577^2 - 4.465^2}{4.577^2 + 4.465^2} = 0.0248 \]
3.9.2 Magnetic coupling

The resonant frequencies by applying electric and magnetic walls on the above circuit are given by [13]:

\[ f_e = \frac{1}{2\pi \sqrt{(L_{lumped} - L_{mutual})/C_{lumped}}} \]  \hspace{0.5cm} \text{(3-4)}

\[ f_m = \frac{1}{2\pi \sqrt{(L_{lumped} + L_{mutual})/C_{lumped}}} \]  \hspace{0.5cm} \text{(3-5)}

Using the measured results, the magnetic coupling coefficient, \( k_M \) is given by:

\[ k_M = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2} = \frac{3.724^2 - 3.382^2}{3.724^2 + 3.382^2} = 0.096 \]  \hspace{0.5cm} \text{(3-6)}
3.9.3 Mixed Coupling

![Mixed Coupling Diagram](image)

Figure 3-40 Equivalent circuit for mixed coupling

\[
f_e = \frac{1}{2\pi \sqrt{(L_{\text{lumped}} - L_{\text{mutual}})(C_{\text{lumped}} - C_{\text{mutual}})}}, \tag{3-7}
\]

\[
f_m = \frac{1}{2\pi \sqrt{(L_{\text{lumped}} + L_{\text{mutual}})(C_{\text{lumped}} + C_{\text{mutual}})}}, \tag{3-8}
\]

Using the measured results, the mixed coupling, \( k_B \), is given by:

\[
k_B = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2} = \frac{3.592^2 - 3.269^2}{3.592^2 + 3.269^2} = 0.0939 = k'_M + k'_E, \tag{3-9}
\]

where \( k'_M, k'_E \) are the magnetic and electric couplings found in the mixed coupling structure.
3.10 Measurement results

The configurations shown in Figure 3-2 to Figure 3-4 and Figure 3-27 (a) to study the electric, magnetic and mixed couplings are fabricated. The designed configurations are first made into laser masks. This type of masks is chosen because it is very accurate. The design parameters are sometimes very small, as small as 10mil, high accuracy is desirable. The laser masks are then fabricated using the photo-etching process in the school laboratory. This fabrication process is not complicated and not expensive, making the study of these microstrip resonators economical.

From the measured and simulated results shown in Figure 3-41 to Figure 3-44, there are shifts in centre frequencies between the simulated and measured results. They are 6.1%, 5.2%, 4.3% and 5% for the electric, magnetic, mixed couplings and the direct feed line for mixed coupling respectively. However, these shifts are within acceptable range and it can therefore be concluded that the designs for the three types of couplings are valid.
Figure 3-41 Simulated and measured results for electric coupling configuration

Figure 3-42 Simulated and measured results for magnetic coupling configuration
Figure 3-43 Simulated and measured results for the mixed coupling configuration

Figure 3-44 Measured and simulated results for the mixed coupling with direct feed lines
The measured results and simulated results showed in Figure 3-44 are the filter responses of the newly developed dual mode coupled line resonator achieved in this chapter. It can be seen from the graph that the insertion loss across the pass band is just 0.7766dB. There are sharp cutoffs at the pass band edges. The achieved 3dB bandwidth of this resonator is 454MHz with a fractional bandwidth of 13.2%.

The photographs for the three pieces of hardware are shown in Figure 3-45 to Figure 3-47. They are all fabricated on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r = 10.2$. In order to have more secured connections with the Agilent Network analyzer, all the four pieces of fabricated hardware are mounted on aluminum base.
Figure 3-46 Hardware for the magnetic coupling configuration

Figure 3-47 Hardware for the mixed coupling configuration
Figure 3-48 Hardware for the mixed coupling case with direct feed lines
3.11 Overall performance

Having investigated all the features of this new filter design, the summarized measured performance of this resonator is shown in Table 3-21 and the comparisons with a four-pole square loop resonator shown in [13] is shown in Table 3-22. As the resonator is doubly loaded and it is symmetrical, the loaded Q is given by the equation in the table [16].

<table>
<thead>
<tr>
<th>Property</th>
<th>Measured values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Midband insertion loss, IL</td>
<td>0.7766</td>
</tr>
<tr>
<td>Frequency range</td>
<td>3.209GHz – 3.663GHz</td>
</tr>
<tr>
<td>3dB Bandwidth</td>
<td>454MHz</td>
</tr>
<tr>
<td>Fractional bandwidth</td>
<td>13.2%</td>
</tr>
<tr>
<td>Coupling coefficient, k</td>
<td>0.132</td>
</tr>
<tr>
<td>( Q_{\text{loaded}} ) [16]</td>
<td>( \frac{1}{2} \left( \frac{f_o}{\Delta f_{3dB}} \right) = \frac{3.428}{0.454} = 7.55 )</td>
</tr>
</tbody>
</table>

Table 3-21 Summary of the performance of the newly proposed coupled line resonator

<table>
<thead>
<tr>
<th>Property</th>
<th>Four-pole square open-loop</th>
<th>Newly proposed coupled line</th>
</tr>
</thead>
<tbody>
<tr>
<td>IL</td>
<td>2.1769dB</td>
<td>0.7766dB</td>
</tr>
<tr>
<td>Fractional bandwidth</td>
<td>4%</td>
<td>13.2%</td>
</tr>
<tr>
<td>Coupling coefficient, k</td>
<td>0.023</td>
<td>0.132</td>
</tr>
<tr>
<td>Size</td>
<td>( 0.37\lambda_g \times 0.37 \lambda_g )</td>
<td>( 0.21\lambda_g \times 0.21\lambda_g )</td>
</tr>
</tbody>
</table>

Table 3-22 Comparisons between a four-pole square loop resonator and the newly proposed coupled line resonator
From the comparison with the four-pole square loop resonator, it can be seen that the newly proposed coupled line resonator has lower insertion loss across the pass band, 0.7766dB as compared to 2.1769dB for the single line square loop resonator. The fractional bandwidth of the design is also broadened from 4% to 13.2%. The goal of miniaturizing the design has also been achieved.

Figure 3-49 below shows the current plot for the newly proposed coupled line resonator. The warmer color (near the red side of the spectrum) represents higher magnitude of current. The cooler color (near the blue side of the spectrum) represents lower magnitude of current. From the current density distribution, the internal and external electric and magnetic couplings can be deduced. It can therefore be confirmed that this is a mixed coupling coupled line resonator.

Figure 3-49 Current plot for the newly proposed coupled line resonator
4 Meander Square Coupled Line Open Loop Resonator

4.1 Introduction

In the previous chapter, the novel square coupled line open loop resonator is designed. The design has a high coupling coefficient, reasonably large fractional bandwidth and reasonable loaded Q. However, the rejection band is not satisfactory enough. It is about -25dB. Therefore, further improvement has to be sorted. In addition, as size is a major concern for modern wireless communications applications, it is necessary to find ways to further miniaturize the design. From [17], it is known that by employing a meander loop structure, the performance of the filter can be improved. One such configuration [10] is shown in Figure 4-1. This configuration is the meander loop version to that shown in Figure 2-1 (a).

In addition, from [18], higher order filter design can be implemented from lower order sections to improve the filter performance. One such configuration is shown in Figure 4-2.

Therefore, in this chapter, these features will be incorporated into the newly proposed filter configuration in the previous chapter.
Figure 4-1 Meander loop configuration for single loop resonator

Figure 4-2 New coupling scheme for two pole filters from [18]
4.2 Layout of the new meander coupled line resonator

The layout for the new meander coupled line resonator is shown in Figure 4-3 and the corresponding dimensions for the design are shown in Table 4-1. This is again investigated on a RT/Duroid 6010 substrate with a thickness, $h=25\text{mil}$ and relative dielectric constant $\varepsilon_r=10.2$.

![Figure 4-3 Layout of the new meander loop resonator](image)

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<tr>
<td>Dimensions (mils)</td>
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<td>134</td>
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<td>200</td>
<td>110</td>
<td>40</td>
<td>23</td>
<td>20</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 4-1 Dimensions for Figure 4-3
4.3 Implementing an alternate coupling scheme

The design in Figure 4-3 starts off by referencing the novel coupling schemes from [18]. The paper presents some ideas on how to implement higher order filter characteristics from lower order sections and at the same time retain the coupling coefficients. Some common higher order characteristics include better rejection band performance. Therefore, this layout in Figure 4-2 is tried on the design from the precious chapter. It is meant to be for stripline filters and the lengths of the upper and lower loop are set to $\lambda$ and $\lambda/2$. In the case for the coupled line resonator in this thesis, the configuration is different and so the lengths for the upper and lower loop are also not the same as that mentioned in [18]. The main idea used here is the coupling scheme at the input and output ports of the resonator.

In this new coupling scheme at the input and output ports, the feed lines are also connected directly to the resonator. Figure 4-4 shows the resulting configuration used to investigate the new coupling scheme. Table 4-2 tabulates its dimensions. The length $l$ in the configuration is made to vary to study its effects. The full wave response is compared to the design in Figure 4-5.
Figure 4-4 New coupling scheme implemented on the coupled line resonator

Figure 4-5 Coupled line resonator developed in previous chapter

<table>
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<tr>
<th>Item</th>
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<td>93/113</td>
<td>21.25</td>
<td>166.5/</td>
<td>NA</td>
<td>20/40</td>
<td>20</td>
<td>23</td>
<td>10</td>
<td>NA</td>
<td>NA</td>
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<tr>
<td>Dimensions for Figure 4-4 (mils)</td>
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<td>157</td>
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<td>20</td>
<td>23</td>
<td>NA</td>
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<td>10</td>
</tr>
</tbody>
</table>

Table 4-2 Dimensions for Figure 4-4 and Figure 4-5
Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, $h=25\text{mil}$ and relative dielectric constant $\varepsilon_r=10.2$. The S21 and S11 are shown in Figure 4-6 and Figure 4-7. At first, as little changes to the dimensions of the original designs are made. The responses are shown by the graph in red. The rejection band has better response, but the insertion loss is made worse. It is then suspect that the coupling at the input and output ports are not enough. In order to improve these couplings, the space between the inner and outer loops is increased at the feed-point sides. The new response is presented by the graph in blue. In this, both the rejection band and the insertion loss are improved but the bandwidth has been reduced.

![Figure 4-6 The S21 of the designs used to implement the new coupling schemes](image-url)
4.4 Performance of the design with new coupling scheme

Some calculated parameters for comparing the performance of the above design with new coupling scheme is shown in Table 4-3.

<table>
<thead>
<tr>
<th>Property</th>
<th>Simulated values</th>
</tr>
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<tbody>
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<td>Frequency range</td>
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<td>3dB Bandwidth</td>
<td>383MHz</td>
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<tr>
<td>Fractional bandwidth</td>
<td>10.85%</td>
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<tr>
<td>Coupling coefficient, k</td>
<td>0.106</td>
</tr>
<tr>
<td>$Q_{\text{loaded}}$ [16]</td>
<td>$Q_{\text{loaded}} = \frac{1}{2} + \frac{1}{(\Delta f)_{3\text{dB}}} = \frac{f_o}{Q_e} = \frac{3.530}{0.383} = 9.22$</td>
</tr>
</tbody>
</table>

Table 4-3 Performance of the filter with new coupling scheme
4.5 Implementing the meander coupled line resonator

The design presented in the previous session has an improved performance, but the S11 is not low enough. A closer look at the design configuration shown in Figure 4-4 will notice that the outer loop from the original design has “disappeared” as a result of the implementation of the new coupling scheme. As such, it is necessary to implement this back to the new design. One such attempt is shown in Figure 4-3. Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r=10.2$. The comparisons of the S21 and S11 of the design with and without incorporating the outer loop are shown in Figure 4-8 and Figure 4-9. From the graphs, it can be observed that with the same performance for the rejection band, the matching is much better. However, there suffers a reduction in bandwidth.

![Figure 4-8 S21 of the modified design for implementing the new coupling scheme](image)

In order to increase the bandwidth, further modifications to the design in Figure 4-3 are made. These configurations are shown in Figure 4-10 to Figure 4-13. Two alternative designs are being proposed here. Both have modifications done on the inner space of the inner loop. Configuration 1 shown in Figure 4-10 has “two arms” attached to the inner loop. This also creates extra gap-openings on the coupled structures. The gap-openings are on alternate sides of each other, just like those hairpin structures. Configuration 2 in Figure 4-11 opens two gap-openings on alternate sides of the meander loop resonator. Configuration 3 and 4 are designs shown earlier and are included here for comparison purpose.
Figure 4-10 Configuration 1 for Figure 4-14

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<td>44.5</td>
<td>1133</td>
<td>21.25</td>
<td>146.5</td>
<td>104</td>
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Table 4-4 Dimensions for Figure 4-10
Table 4-5 Dimensions for Figure 4-11

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Figure 4-12 Configuration 3 for Figure 4-14

Figure 4-13 Configuration 4 for Figure 4-14

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<th>$s$</th>
<th>$w$</th>
<th>$g$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions for Figure 4-12 (mils)</td>
<td>253</td>
<td>110</td>
<td>134</td>
<td>110</td>
<td>200</td>
<td>110</td>
<td>40</td>
<td>20</td>
<td>23</td>
<td>10</td>
</tr>
<tr>
<td>Dimensions for Figure 4-13 (mils)</td>
<td>253</td>
<td>110.25</td>
<td>44.5</td>
<td>113</td>
<td>21.25</td>
<td>146.5</td>
<td>40</td>
<td>20</td>
<td>23</td>
<td>10</td>
</tr>
</tbody>
</table>

Table 4-6 Dimensions for Figure 4-13 and Figure 4-15
Figure 4-14 S21 response for the alternative designs to increase the bandwidth of the filter

Figure 4-15 The S11 for the alternate designs to increase the bandwidth of the filter
Full wave analysis is done using Agilent’s momentum software. The simulation is executed on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r=10.2$. Figure 4-14 and Figure 4-15 show the full wave simulation S21 and S11 frequency responses. Although the alternative designs are able to improve the bandwidth, that is done at the expense of the matching. As the original intention of lowering the S21 at the rejection band has achieved without these alternative designs, the filter configuration of Figure 4-3 is fabricated instead.

4.6 Measurement results

The designed configurations are made into laser masks. Laser mask is chosen because of its high dimension accuracy. The laser masks are then fabricated using the photo-etching process in the school laboratory. This fabrication process is not complicated making the choice for the study of this microstrip resonators economical.

Measurement results and the photograph of the hardware of the newly developed coupled line meander loop resonator are shown in Figure 4-16 and Figure 4-17. The measured results for this design has a rejection band better than -25dB and S11 is close to -20dB. The measured 3dB bandwidth is 202MHz with a fractional bandwidth of 5.98%. The insertion loss is 1.2565dB in this case, 0.4799dB more than the coupled line resonator presented in the last chapter.
From the measured and simulated results, there is a shift in centre frequencies of 4.85% between the simulated and measured results. This shift is within acceptable range.

Losses can also be observed from the measured results because the simulation results do not take into account the conductor loss. In the simulated conditions, the ground is assumed to be infinite and the effects of the connectors at the input and output ports are also not included. There is also the tolerance of the fabrication process to be taken into consideration. Accounting for all those, the measured and simulated results agree well with each other.

Figure 4-16 Simulated and measured response of the new meander loop coupled line resonator
4.7 Overall performance

Table 4-7 summarizes the measured performance of the coupled line meander loop resonator. From the comparisons, the insertion loss of this new design has increased from 0.7766dB to 1.2565dB and the bandwidth has decreased from 13.2% to 5.98%. However, this design has improved matching at the pass band. The narrow band feature can also be employed in narrow band applications.
<table>
<thead>
<tr>
<th>Property</th>
<th>Measured values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Midband insertion loss, IL</td>
<td>1.2565dB</td>
</tr>
<tr>
<td>Frequency range</td>
<td>3.28GHz – 3.482GHz</td>
</tr>
<tr>
<td>3dB Bandwidth</td>
<td>202MHz</td>
</tr>
<tr>
<td>Fractional bandwidth</td>
<td>5.98%</td>
</tr>
<tr>
<td>Coupling coefficient, k</td>
<td>0.0597</td>
</tr>
</tbody>
</table>

\[ Q_{\text{loaded}} = \frac{1}{2} + \frac{1}{Q_e} \left( \frac{f_o}{(\Delta f)_{3dB}} \right) = \frac{3.38}{0.202} = 16.73 \]

Table 4-7 Performance of the meander coupled line resonator

The comparisons of performance for the three configurations designed are shown in Table 4-8. Although the design with just the coupling scheme implemented is not fabricated, due to the good agreement between the simulated and the measured results, it can be deduced that the design with just the new coupling scheme implemented has filter performance in between the coupled line resonator and the meander resonator.

<table>
<thead>
<tr>
<th>Property</th>
<th>Coupled line resonator</th>
<th>New coupling scheme (simulated)</th>
<th>Meander resonator</th>
</tr>
</thead>
<tbody>
<tr>
<td>IL</td>
<td>0.7766dB</td>
<td></td>
<td>1.2565dB</td>
</tr>
<tr>
<td>Freq. range</td>
<td>3.209-3.663 (GHz)</td>
<td>3.344-3.727 (GHz)</td>
<td>3.28– 3.482 (GHz)</td>
</tr>
<tr>
<td>3dB BW</td>
<td>454 MHz</td>
<td>383 MHz</td>
<td>202 MHz</td>
</tr>
<tr>
<td>Fract. BW</td>
<td>13.2%</td>
<td>10.85%</td>
<td>5.98%</td>
</tr>
<tr>
<td>k</td>
<td>0.132</td>
<td>0.106</td>
<td>0.0597</td>
</tr>
<tr>
<td>(Q_{\text{loaded}} [16])</td>
<td>7.55</td>
<td>9.22</td>
<td>16.73</td>
</tr>
</tbody>
</table>

Table 4-8 Performance comparisons for the three configurations designed
The current plot of the meander loop resonator is shown in Figure 4-18. The warmer color (near the red side of the spectrum) represents higher magnitude of current. The cooler color (near the blue side of the spectrum) represents lower magnitude of current. The current plot shows that higher magnitude of current is found at the upper left hand corner and lower right hand corner of the configuration. This resonator is also of mixed coupling type. The electric and magnetic couplings are stronger in this meander loop resonator and more distinguishable than the previous chapter.
5 Coupling of several meander resonators

5.1 Introduction

With the completion of the filter configurations in the previous sections, it is now of interest to implement cross coupling among several of these new filters configurations. For the conventional square open loop resonator shown in Figure 5-1, the electric (blue), magnetic (red) and mixed (black) couplings between any two adjacent resonators are calculated in a matrix and each has its cross coupling coefficients. There have also been various schemes to implement the cross couplings between the resonators [18] to [21]. With the cascaded network, it is possible to design filter with much better rejection band performance and filters with multiple poles.

The coupled line meander loop configuration presented in the last chapter has exhibited some cross coupling effects just like those with several of the square open loop resonators. Coupling schemes like that for the single loop square open loop resonator maybe possible to develop. However, as the coupling effects for this proposed coupled line meander loop resonator are rather complex, CAD tool, the Agilent momentum software, will be used to do the investigation to find out the cascaded network. The purpose of this investigation is to design a cascaded network with improved rejection bands performance for high selective narrow band applications.
5.2 Layout of the cascaded meander loop network

The configuration of the network is shown in Figure 5-2 and the dimensions of the design are shown in Table 5-1. This is again fabricated on a RT/Duroid 6010 substrate with a thickness, \( h = 25 \text{mil} \) and relative dielectric constant \( \varepsilon_r = 10.2 \). Only two units of the meander loop resonators are cascaded. As there are many cross coupling involved in the design, only CAD simulations using Agilent’s momentum is used in the study of the design.

The cascaded design is similar to the electric coupling network for a two-pole single square open loop case. The orientation of the design is flipped for the second resonator.
Table 5-1 Dimensions for Figure 5-2

<table>
<thead>
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<th>$s$</th>
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<tbody>
<tr>
<td>Dimensions (mils)</td>
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<td>134</td>
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<td>200</td>
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<td>50</td>
<td>23</td>
<td>15</td>
<td>10</td>
<td>15</td>
</tr>
</tbody>
</table>

5.3 Exploring the feeding positions

One feature of the coupling theory for the single square open loop resonator is the positions of the feed lines attached to the resonators. The different feeding positions investigated are presented in (b) to (c) in Figure 5-3 in addition to the configuration (a) shown in Figure 5-2. Their dimensions are tabulated in Table 5-2. The feeding positions are at the different combinations at the top, center and bottom sides of the resonators. The second resonator is a mirror image of the first resonator.
Figure 5-3 Configurations (b), (c) and (d) in Figure 5-4
Table 5-2 Dimensions for Figure 5-3

<table>
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<th>( l_3 )</th>
<th>( l_4 )</th>
<th>( l_5 )</th>
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<th>( s )</th>
<th>( g )</th>
<th>( d )</th>
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</thead>
<tbody>
<tr>
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<td>134</td>
<td>110</td>
<td>200</td>
<td>110</td>
<td>63</td>
<td>50</td>
<td>23</td>
<td>15</td>
<td>10</td>
<td>15</td>
</tr>
</tbody>
</table>

Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, \( h=25 \text{mil} \) and relative dielectric constant \( \varepsilon_r =10.2 \). The frequency responses are shown in Figure 5-4. Configurations (a) and (d) both have attenuations poles (minimum points) located close to 2.9 GHz and 4.5GHz. For configuration (c), the attenuations poles have disappeared. The S21 response for configuration (b) shows the sharpest cutoff at the pass band but the higher frequency rejection band performance for this configuration is not as good as configurations (a), (c) and (d).

For the S11 response in Figure 5-5, configurations (b) to (d) have two poles observed at 3.46GHz and 3.56GHz. The difference is in the insertion loss at the pass band. Configuration (b) is able to achieve the -10dB insertion loss throughout the pass band and is therefore the most desirable among the three. For configuration (a), two poles are observed at 3.44GHz and 3.46GHz. The insertion loss obtained across this narrow band is approximately -25dB.

Taking into consideration of conductor loss in actual hardware, configuration (a) is chosen as the desired feeding configuration.
Figure 5-4 S21 response for the different feeding positions

Figure 5-5 S11 response for the different feeding positions
5.4 Exploring the distance between the resonators

Another feature of the coupling theory used in the single square open loop resonator is the distance between the resonators. This parameter is found to affect the coupling coefficient, k of the design. The chosen configuration and dimensions for investigation from the previous section are reproduced in here. The parameter “d” is made to vary with values 5, 10, 15 and 20mil.

![Figure 5-6 Configuration to investigate the optimized distance between the resonators](image)

<table>
<thead>
<tr>
<th>Items</th>
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<th>$l_6$</th>
<th>$l_7$</th>
<th>$l_8$</th>
<th>w</th>
<th>s</th>
<th>g</th>
<th>d</th>
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<td>110</td>
<td>200</td>
<td>110</td>
<td>63</td>
<td>50</td>
<td>23</td>
<td>15</td>
<td>10</td>
<td>variable</td>
</tr>
</tbody>
</table>

Table 5-3 Dimensions for Figure 5-6
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r = 10.2$. From the S21 response in Figure 5-7, when the two resonators are closer, the coupling is stronger. This is reflected by the higher band attenuation pole being closer to the pass band edge. When $d=5$mil, this attenuation pole is at 4.4GHz whereas when $d=20$mil, the attenuation pole is at 4.55GHz.

For the S11 response shown in Figure 5-8, the matching for $d=5$mil is not as good as $d=10$mil. This shows that too close a distance will affect the insertion loss. However, when $d=20$mil, the two distinctive poles observed between 3.4GHz and 3.5GHz have almost disappeared. The optimized distance between the two resonators is when $d=15$mil.

![Figure 5-7 S21 response for investigating the distance between the resonators](image-url)
5.5 Exploring the orientation of the two resonators

As mentioned in sections 5.2 and 5.3 the second resonator is a mirror image of the first resonator. This orientation is like the electric coupling between two single square loops. In this section, full wave simulations using Agilent’s momentum software is used to do the comparison between this mirrored resonators and the replicated resonators shown in Figure 5-9. The dimensions are shown in Table 5-4.
Figure 5-9 Different orientations of the two resonators

<table>
<thead>
<tr>
<th>Items</th>
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<th>$l_3$</th>
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<th>$l_7$</th>
<th>$l_8$</th>
<th>$w$</th>
<th>$s$</th>
<th>$g$</th>
<th>$d$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions for both (a) and (b) (mils)</td>
<td>253</td>
<td>110</td>
<td>134</td>
<td>110</td>
<td>200</td>
<td>110</td>
<td>63</td>
<td>50</td>
<td>23</td>
<td>15</td>
<td>10</td>
<td>15</td>
</tr>
</tbody>
</table>

Table 5-4 Dimensions for Figure 5-9
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, $h=25\text{mil}$ and relative dielectric constant $\varepsilon_r =10.2$. The $S_{21}$ response is shown in Figure 5-10. For configuration (b), the attenuation poles ($S_{21}=\text{minimum}$) have disappeared and there is a greater insertion loss. In configuration (a), the attenuation poles are present at 3.15GHz and 4.5GHz. However, in both configurations, the rejection band can achieve a $S_{21}$ close to -60dB. The $S_{11}$ response shown in Figure 5-11 clearly shows that the matching in (b) is not good.

Figure 5-10 $S_{21}$ response for configurations to investigate the orientation of the two resonators
5.6 Measurement results

The configuration in Figure 5-2 is being fabricated using the laser mask and photo-etching technique. Measurement results and the photograph of the hardware of the newly developed coupled line meander loop resonator are shown in Figure 5-12 and Figure 5-13. The hardware is mounted on aluminum base to provide better grounding and secured connections with the Agilent Network analyzer.

The measured results for this design have a rejection band better than -60dB and S11 across the pass band is also better than -10dB. The measured -10dB bandwidth
is 203MHz with a fractional bandwidth of 5.99%. This measured result is better than the simulated results. However, the insertion loss is 3.2636dB, more than the measured results from earlier designs.

Losses can also be observed from the measured results because the simulation results do not take into account the conductor loss. In this cascaded network, much cross coupling effects take place and this is expected to take much more effect. In the simulated conditions, the ground is assumed to be infinite and the effects of the connectors at the input and output ports are also not included. Accounting also for the tolerance of the fabrication process, the measured and simulated results agree well with each other.
Figure 5-13 Hardware of the cascaded network of meander loop coupled line resonator

5.7 Overall Performance

The summarized measured performance is shown in Table 5-5. The measured insertion loss for this configuration is higher than earlier designs. This is because the conductor loss is more for the cascaded network. Table 5-6 and Table 5-7 show the comparisons between the simulated performances of the fabricated resonators proposed. The cascaded network has similar performance as the single meander loop, but much better rejection band (-60dB) as compared to -25dB.
### Table 5-5 Measured performance of the cascaded network of meander loop resonators

<table>
<thead>
<tr>
<th>Property</th>
<th>Measured values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Midband insertion loss, IL</td>
<td>3.2636dB</td>
</tr>
<tr>
<td>Frequency range (-10dB)</td>
<td>3.291GHz – 3.494GHz</td>
</tr>
<tr>
<td>-10dB Bandwidth</td>
<td>203MHz</td>
</tr>
<tr>
<td>Fractional bandwidth</td>
<td>5.99%</td>
</tr>
<tr>
<td>Coupling coefficient, k</td>
<td>0.0598</td>
</tr>
</tbody>
</table>

\[
Q_{\text{loaded}}[16] = \frac{1}{2} + \frac{1}{Q_c} = \frac{f_o}{(\Delta f)_{3dB}} = \frac{3.39}{0.203} = 16.70
\]

### Table 5-6 Comparison of the cascaded network with the previous resonators designed

<table>
<thead>
<tr>
<th>Property</th>
<th>Coupled line resonator</th>
<th>New coupling scheme</th>
<th>Meander resonator</th>
<th>Cascaded two units of meander resonator</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fract. BW</td>
<td>11.46%</td>
<td>10.85%</td>
<td>7.43%</td>
<td>4.29%</td>
</tr>
<tr>
<td>k</td>
<td>0.077</td>
<td>0.106</td>
<td>0.043</td>
<td>0.0429</td>
</tr>
</tbody>
</table>

### Table 5-7 Comparisons between the measured results of the designed resonators

<table>
<thead>
<tr>
<th>Property</th>
<th>Coupled line resonator</th>
<th>Meander resonator</th>
<th>Cascaded unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>IL</td>
<td>0.7766dB</td>
<td>1.2565dB</td>
<td>3.2636dB</td>
</tr>
<tr>
<td>Fract. BW</td>
<td>13.2%</td>
<td>5.98%</td>
<td>5.99%</td>
</tr>
<tr>
<td>k</td>
<td>0.132</td>
<td>0.0597</td>
<td>0.0598</td>
</tr>
<tr>
<td>(Q_{\text{loaded}}[16])</td>
<td>7.55</td>
<td>16.73</td>
<td>16.70</td>
</tr>
</tbody>
</table>
The current plot of the cascaded network is shown in Figure 5-14. The warmer color (near the red side of the spectrum) represents higher magnitude of current. The cooler color (near the blue side of the spectrum) represents lower magnitude of current. The current plot shows that higher magnitude of current is found at the bottom right corner of the first resonator and lower left corner of the second resonator. This allows the two to couple with each other.

Figure 5-14 Current plot for the cascaded network
6 Design Considerations

6.1 Introduction

In the previous chapters, the step-by-step evolution of the final meander coupled line resonator has been shown. Having this pattern developed, this chapter will give a guideline to design this type of filter.

6.2 Design Parameters

Table 6-1 summarizes the relationships between the critical parameters in filter designs and the design considerations when designing the coupled line resonators. Some of these are the starting points of the designs above, some are pointers observed when doing the optimization for the designs presented above.

At the present stage, the designs rely much on the optimization tools in CAD software. The design equations like those in [13] are possible to develop. They need a database of the performances for many of these types of resonators and some critical parameters like the coupling coefficients and Q to be computed. However, with the pointers below, it is still relatively easy to design the meander loop resonator proposed.
<table>
<thead>
<tr>
<th>Item</th>
<th>Descriptions</th>
</tr>
</thead>
<tbody>
<tr>
<td>Centre frequency</td>
<td>This is set by choosing the $\lambda_g/4$ line length for each of the arm in the resonator.</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>This can be chosen from the specific configuration from Table 5-7.</td>
</tr>
<tr>
<td>Rejection level</td>
<td>This can be determined from the type of resonator.</td>
</tr>
<tr>
<td>Width of the gap-opening</td>
<td>This affects the insertion loss. Smaller gap-opening, better insertion loss. But the gap-opening has to be big enough for fabrication and for providing the capacitance for perturbations.</td>
</tr>
<tr>
<td>Coupled line length for implementing the new coupling scheme at the input and output ports</td>
<td>This affects the matching of the meander loop resonator. Usually its required length is longer than the design before implementing the new coupling scheme.</td>
</tr>
<tr>
<td>Spacing between the inner and outer loops</td>
<td>This can refer to [15] or ADS for the desired $K$ and odd and even mode characteristic impedances.</td>
</tr>
</tbody>
</table>

Table 6-1 Overview of the design features of the various coupled line resonators
7 Miniaturized meander loop filter

7.1 Introduction

Having developed the dual mode resonator using the configurations shown in the previous chapters, it may be of interest to try to use the configuration designed in miniaturized applications.

Miniaturized structures have received extensive attentions from many researchers in the past few years and have been incorporated into antennas and filter designs. Many of these miniaturized structures are under the Electronic Band Gap, EBG category. They are actually periodic structures that forbid the propagation of energy in a frequency band and are well-known for its harmonic suppression properties. In addition, the slow-wave characteristics exhibited by periodic structures can also reduce the circuit components’ size.

Many literatures have been published for using EBG structure on the ground plane [22], [23], [24], few are on using them on the transmission line itself [25]-[26]. In this chapter, it is the intention to use the pattern designed in the previous chapter on the transmission line to achieve a bandpass filter with some performance characteristics like the EBG structures. Three units of the miniaturized units are cascaded and fabricated.
7.2 Layout of the miniaturized design

Figure 7-1 shows the layout of the miniaturized unit. This is similar to the meander loop resonator in chapter 4 and 5. The width of all the conductors is half of the original 50 ohm line width. It is 10mil instead of 23mil. The width of the middle coupling arm, represented by $2w_2$, is 20mil. The distance “d” between the cascaded networks is now 10mil instead of the previous 15mil in chapter 5. The two resonators are also not mirror images of each other.

![Figure 7-1 Layout of the miniaturized design](image)

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<td>10</td>
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</tbody>
</table>

Table 7-1 Dimensions for Figure 7-1
7.3 Effects of the width of the conductor

The pattern developed in Figure 4-3 has many “elements” inside the outer square loop. If a filter of higher frequency is to be designed using the same substrate, this will post a limitation because these “elements” just does not have enough space inside the outer loop. In this section, the configuration shown in Figure 7-2 and the dimensions tabulated in Table 7-2 are used to study the effects of the width of the conductor. The design is implemented on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r = 10.2$.

The length, $l$, is remained unchanged in this study because it affects the coupling of the input and output ports. The gap, $g$, is also unchanged because it determines the loss of the resonators.

Full wave analysis using Agilent’s momentum software is performed. Figure 7-3 shows the results of reducing the width from 20mil to 10mil in a step of 2. The designs have centre frequency 3.7GHz. It shows that decreasing the width of the conductor lines will widen the bandwidth of the filter and there is a slight “stretch” in the frequency response.
Figure 7-2 Configuration to study the effects of the width of the conductor

<table>
<thead>
<tr>
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<td>200</td>
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<td>23</td>
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<td>10</td>
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<tr>
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<td>247</td>
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<td>102.5</td>
<td>185</td>
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Table 7-2 Dimensions for Figure 7-2
7.4 Width of line connecting to feeding line

The idea is implemented on a filter with intentional centre frequency at 7 GHz. Two designs have been investigated. One is with all the transmission lines halved, the other is with the centre line remaining the same width. This is done in consideration that the centre line together with the two transmission lines above and below form the new coupling scheme. Therefore, it maybe necessary to retain the width for matching purposes so that the source and load feeding lines can be attached at those points.
Figure 7-4 Configuration to study to the line width connecting to the feed line

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Table 7-3 Dimensions for Figure 7-4
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, $h=25\text{mil}$ and relative dielectric constant $\varepsilon_r=10.2$. The resonators are being tapped (not directly connected to the structure). The gap between the feed lines and the resonator is 10mil. From the response shown in Figure 7-5, it can be seen that the two resonant frequencies are now further apart and that the attenuation poles at the sides of the resonant frequencies have “disappeared”. With the two resonant frequencies further apart to each other means that it is possible to design a filter with wider bandwidth. Configuration (a) has resonant frequencies at 5.8GHz and 6.95GHz. Configuration (b) has resonant frequencies at 5.95GHz and 7.15GHz. 

Figure 7-5 Frequency response of the filter design conductor lines width halved
7.5 Matching problem on the miniaturized unit

Matching is a problem in the above design because the width of the transmission lines is halved. Originally, with all the transmission lines having the width same as a 50Ω transmission line, i.e. 23mil, the feed lines can be connected to the filter configuration without worrying about this effect. With the conductor line width decreased, there will be a “step” section between the feed lines and the resonator structures and this will affect the matching of the resonator. In addition, it will also affect the predicted performance discussed so far from previous chapters. The following three feeding positions using the configuration in Figure 7-4 (a) illustrate these effects. The feed line has a length of 50mil and width of 23mil.

Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r=10.2$.

Figure 7-6 shows the proposed design with the same feeding positions as in the case the square loop resonator without implementing the new coupling scheme and without the meander lines. The S11 in the pass band is not as low as desired.

Figure 7-7 shows the design with feed lines at opposite positions as in Figure 7-6. The matching has improved a lot here. This is because the feed line is directly connected to a gap on the outer loop. This can be explained by the fact that the current can only flow to the part where there is a conductor. This means that the
impedance seen by the source is larger. As a result, when using the direct feed method, the S11 in the pass band will improve.

Figure 7-8 shows the feeding positions same as the meander lines designs. Again, the S11 in the pass band is not low enough, indicating a poor matching.

![Figure 7-6 Frequency response of the miniaturized structure with first type of feeding](image)
Figure 7-7 Frequency response of the miniaturized structure with a second type of feeding

Figure 7-8 Frequency response for the miniaturized design with a third type of feeding
7.6 Using the feed lines to improve matching

From the previous section, it can be seen that the matching is not good for the new miniaturized structure. The miniaturized structure exhibits low impedance as a result of the decreased conductor width. Therefore, when the feed line is connectedly directly to the miniaturized structure, it is better to connect it using a $\lambda_g/4$ transmission line. In this case, if the miniaturized structure is designed to have 50$\Omega$ at the feed points, this 50$\Omega$ will not be altered at the source and load points. Figure 7-9 shows the improved matching after using $\lambda_g/4$ lines as the feeding lines.

Figure 7-9 Frequency response of the miniaturized structure with $\lambda/4$ feeding lines
7.7 Cascaded miniaturized structure

As the miniaturized structures presented here intend to implement some of the features of the EBG structures such as periodicity, they should be able to be cascaded and provide additional poles at the pass band [25]. In addition, when the units are cascaded, there is electrical connection between them and there is no need to “flip” the units because it is not the coupling effects that are taking place.

The layouts of the different numbers of cascaded networks are shown in Figure 7-10. The dimensions for a single unit, the distance between the units and the feed line length are shown in Figure 7-11, Figure 7-12 and Table 7-4.

Figure 7-10 Three cascaded networks for the miniaturized designs
Figure 7-11 Cascaded network configuration

Figure 7-12 Feed lines attached to the cascaded miniaturized units

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Table 7-4 Dimensions for Figure 7-11 and Figure 7-12
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, $h=25$ mil and relative dielectric constant $\varepsilon_r=10.2$. The S21 response is shown in Figure 7-13. As the number of cascaded units increased, the filter becomes more selective. The rejection band is better than -45 dB and there are distinctive attenuation poles (S21 is minimum) at 5.6 GHz and 8.3 GHz for the 3-units network. Figure 7-14 shows the S11 response. The 1-unit network has 1 pole (possibly 2 poles, but because of other design issues, the two poles are merged to one). The 2-unit network has 4 poles and 3-unit network has 6 poles. This suggests that each additional unit has increased the number of poles in the pass band by two.

![S21 response](image)

**Figure 7-13** The S21 responses of the cascaded miniaturized networks
7.8 Measurement results

The 3-unit configuration is fabricated using the laser mask and photo-etching technique. Measurement results and the photograph of the hardware of the newly developed coupled line meander loop resonator are shown in Figure 7-15 and Figure 7-16. The hardware is mounted on aluminum base to provide better grounding and secured connections with the Agilent Network analyzer.

The simulated results are done using ie3d, with simulation conditions under open space. As the width of the conductor is small and the lines are much closer to each other, these simulation conditions such as open and enclosed space play an important part in the accuracy of the simulations. The measured results for this
design has a rejection band better than -40dB and S11 across the pass band is also better than -10dB.

The measured 3dB bandwidth is 1.6GHz with a fractional bandwidth of 23.6%. The insertion loss is 1.815dB, relatively acceptable. The simulation results do not take into account the conductor loss. Accounting for all those, the measured and simulated results agree well with each other.

Figure 7-15 Simulated and measured response for the miniaturized design
7.9 Overall Performance

The summarized performance of the measured results for the 3-unit cascaded network is tabulated in Table 7-5. The 3dB bandwidth achieved for this miniaturized network is 1.6GHz and the fractional bandwidth is 23.6%. This is a very wideband response and can be used in wideband applications in communications.

As this is a wideband filter, its measured performance is not compared with the filter designs in previous chapter.
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**Table 7-5 Performance of the 3-unit cascaded miniaturized network**

The current plot of the cascaded network is shown in Figure 7-17. The warmer color (near the red side of the spectrum) represents higher magnitude of current. The cooler color (near the blue side of the spectrum) represents lower magnitude of current. The current plot shows that higher magnitude of current is found at the connecting portions between the units. This allows the current to flow to the next unit. Again this is a mixed coupling configuration as can be deduced from the current distribution.

![Current plot for the miniaturized cascaded network](image_url)
8 Miniaturized structures to suppress spurious harmonics

8.1 Introduction

The miniaturized structure that results from the dual mode meander square open loop resonator seems to be a rather complicated one compared to the EBG structures that are commonly known. This is a disadvantage. Therefore, this chapter will further investigate on the miniaturized structure from the previous chapter to simplify it and look at the ability of this new design to suppress or delay spurious harmonics like what EBG structures do. It is well known that EBG structures exhibit the ability to suppress spurious harmonic responses [27]-[28].

8.2 Layout of the simplified miniaturized structure

Figure 8-1 and Figure 8-2 show the new simplified miniaturized structure based on the design in previous chapter. The centre two “loops” that are used to have a good match are taken away from the miniaturized design. The feed line length \( l \) is \( \lambda_g/4 \). The distance between the two units is 10mil.
Table 8-1 Dimensions for Figure 8-1 and Figure 8-2
8.3 Exploring the pattern inside the loop

In order to investigate the effects of different patterns to be included inside the loop, the following configurations in Figure 8-3 are explored. The dimensions are given in Figure 8-4 and Table 8-2. The simulation is based on a RT/Duroid 6010 substrate with a thickness, h=25mil and relative dielectric constant $\varepsilon_r =10.2$. Configuration (a) is that from the previous chapter. It has two arms attached inside. Configuration (b) has two circles connected to the loop. Configuration (c) has nothing inside the loop.

(a)

(b)

(c)

Figure 8-3 Configurations for the different patterns inside the loop
Figure 8-4 Dimension reference for the different configurations
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Table 8-2 Dimensions for Figure 8-4
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, \( h=25 \)mil and relative dielectric constant \( \varepsilon_r =10.2 \). Figure 8-5 and Figure 8-6 show the S21 and S11 response of the three configurations. The first harmonic of configuration (b) and (c) have shifted to around 12.1GHz instead of around 10.1GHz in configuration (a). This shows that both the configurations in (b) and (c) are able to “migrate” the first harmonics, making the filter more selectable.

The S11 response of configuration (b) and (c) are also very similar. The return loss corresponds to the S21 response. In all three configurations, the insertion loss across the pass band is better than -10dB.

Figure 8-5 S21 response for investigation of the different patterns inside the loop
8.4 Investigating the position of the circles inside the loop

In Figure 8-3 (b), the circles are connected to the inner loops. Current can pass through the circles. In this section, investigation will be done on the effects of connecting and disconnecting the circles from the inner loops.
Figure 8-7 Layout to investigate positions of circles in inner loop

Figure 8-8 Dimension reference for the Figure 8-7
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Table 8-3 Dimensions for Figure 8-8

Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, $h=25$ mil and relative dielectric constant $\varepsilon_r = 10.2$. Figure 8-9 and Figure 8-10 show the frequency response of the two cases presented. It can be seen that the frequency responses for S21 and S11 are the exactly the same whether the circles are connected directly to the inner loop or not.

In both configurations presented here, the center of the loop is kept “clear”. This gives the insight that when the filter design is miniaturized like the configurations shown in this chapter and in the previous chapter, the center of the loop being “clear” will give a migration of the first harmonics.
Figure 8-9 S21 response for investigating the positions of the circles

Figure 8-10 S11 response for investigations of the positions of the circles
8.5 Measured results

The configuration in Figure 8-2 is fabricated using the laser mask and photo-etching technique. Measurement results and the photograph of the hardware of the newly developed coupled line meander loop resonator are shown in Figure 8-11 to Figure 8-13. The hardware is mounted on aluminum base to provide better grounding and secured connections with the Agilent Network analyzer.

The measured results for this design has a rejection band better than -40dB. The measured 3dB bandwidth is 1.575GHz with a fractional bandwidth of 23.6%. This is a wideband response. The insertion loss is an acceptable value of 1.3104dB. Losses exist because the simulation results do not take into account the conductor loss. In addition, in the simulated conditions, the ground is assumed to be infinite and the effects of the connectors at the input and output ports are also not included. Accounting also for the tolerance of the fabrication process, the measured and simulated results agree well with each other.
Figure 8-11 Simulated and measured results for S21 of the simplified miniaturized configuration

Figure 8-12 Simulated and measured results for S11 for the simplified miniaturized configuration
8.6 Overall Performance

Table 8-4 shows the comparison in performance of the common filter parameters between the miniaturized design in Chapter 7 and the simplified one in this chapter. The performance between the two is very similar. This shows that the interposed loops in the miniaturized structure do not play as big a role as in the coupled line configurations.

The current plot of the cascaded network is shown in Figure 8-14. The warmer color (near the red side of the spectrum) represents higher magnitude of current. The
cooler color (near the blue side of the spectrum) represents lower magnitude of current. The current plot shows that higher magnitude of current is found at the connecting portions between the units. This allows the current to flow to the next unit. The current distribution is very similar to the configuration fabricated from Chapter 7.

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Table 8-4 Performance comparison between the newly proposed miniaturized structures

Figure 8-14 Current plot for the simplified miniaturized structure
9 Miniaturized meander loop resonator

9.1 Introduction

The miniaturized designs from the previous two chapters give a rather wideband response. In this chapter, miniaturized structure developed from the meander loop resonator concept with a narrowband response will be presented. This narrowband filter also implements the mode coupling control method [29] at the feed lines to achieve harmonic suppression. Figure 9-1 is an illustration of this. The feed lines at the input and output ports are extended along the arm of the resonators, forming half a loop by itself. Research has demonstrated that two distinct modes with closely spaced resonant frequencies are generated in ring resonators coupled to a microstrip line [30]. An optimized length of the open-end microstrip line at the input and output ports can “migrate” the second resonant [29].

Figure 9-1 Illustration of mode coupling control in microstrip filters
9.2 Layout of the new narrow band miniaturized single loop resonator

The layout and dimensions of the proposed miniaturized narrowband design are shown in Figure 9-2 and Table 9-1. The filter is designed to have a centre frequency of 7GHz. This filter design makes use of the decreased conductor line width. The conductor width is 10mil instead of the $\lambda_g/4$ conductor width of 23mil. This is like a single meander loop design. There are also two gap-openings on the loop. At the input and output ports, something like the optimized length of the open-end microstrip line in Figure 9-1 are implemented.

![Figure 9-2 New miniaturized configuration from the single loop resonator](image)

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Table 9-1 Dimensions for Figure 9-2
9.3 Defining the Resonance Frequencies

In the earlier section, it has been discussed that the meander loop [17] can improve the response of the bandpass filter. As the coupled line resonators designed in previous chapters gives rather wideband response, the resonant frequencies are affected by many coupling effects within the resonator. For a narrowband filter and with only a single square loop, the resonant frequencies will be easier to define.

Figure 9-3 shows the starting point in designing the new miniaturized resonator. The two filters of conductor width 10mil and 23mil respectively are studied. The arm of the square loop has a length of $\lambda_g/4$. Detailed dimensions are furnished in Table 9-2.

![Figure 9-3 Configuration to study effects of the conductor width with resonant frequencies](image)

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</tbody>
</table>

Table 9-2 Dimensions for Figure 9-3
A full wave analysis is executed by using the Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, $h=25\text{mil}$ and relative dielectric constant $\varepsilon_r=10.2$. It can be observed from Figure 9-4 that the resonance frequencies of the thinner line are at 6.7GHz and 7.0GHz whereas for the 50ohm line resonator, the resonant frequencies are at 6.9GHz and 7.3GHz. It can therefore be deduced that the thinner conductor line has the effects of miniaturizing the resonator.

![Figure 9-4 Effects of conductor width and resonant frequencies](image)

Figure 9-4 Effects of conductor width and resonant frequencies
9.4 Feeding positions of the Dual Mode Resonator

The frequency response from Figure 9-4 shows that the resonator does not have attenuation poles at the edges of the pass band. This can be understood from the odd and even mode analysis from section 2.6.2. The equations to calculate the even and odd modes impedances are reproduced below in equations 9-1 to 9-6. The resonator configuration in Chapter 2 has the feed lines attached to the opposite corners of the resonator. From 9-2 and 9-5, in order for \( Y_e \) and \( Y_o \) = 0, i.e. the condition for attenuation poles to occur, \( l_{feedline} \) cannot be 0. In the configuration in Figure 9-3 represents those with \( l_{feedline} \) =0, therefore there are no attenuation poles.

\[
Y_{\text{even circuit}} = \left[ Z_{\text{even arm}} + \left( j \omega C_j \right)^{-1} \right]^{-1} + \left[ j \omega L + \left[ j \omega C + \left( j \omega L + Z_{\text{even arm}} + \left( j \omega C_j \right)^{-1} \right)^{-1} \right]^{-1} \right]^{-1} \tag{9-1}
\]

\[
Z_{\text{even overall}} = Z_o \frac{Z_{\text{even circuit}} + jZ_o \tan(\beta l_{\text{feedline}})}{Z_o + jZ_{\text{even circuit}} \tan(\beta l_{\text{feedline}})} \tag{9-2}
\]

\[
Y_e = \frac{1}{Z_{\text{even overall}}} \tag{9-3}
\]

\[
Y_{\text{odd circuit}} = Y_{\text{odd arm}} + \left[ j \omega L + \left[ j \omega C + \left( Z_{\text{odd arm}} + j \omega L \right)^{-1} \right]^{-1} \right]^{-1} \tag{9-4}
\]

\[
Z_{\text{odd overall}} = Z_o \frac{Z_{\text{odd circuit}} + jZ_o \tan(\beta l_{\text{feedline}})}{Z_o + jZ_{\text{odd circuit}} \tan(\beta l_{\text{feedline}})} \tag{9-5}
\]

\[
Y_o = \frac{1}{Z_{\text{odd overall}}} \tag{9-6}
\]
The configurations to prove that directly connecting the feed lines to the resonator can create the attenuation poles ($S_{21} = \text{minimum}$) is shown in Figure 9-5. The dimensions are given in Table 9-3. The feed line length is $159\text{mil}, \lambda_g/4$.

![Figure 9-5 Configurations to investigate the effects of feed lines positions](image)

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<tr>
<td>Dimensions for (a) (mils)</td>
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<td>Dimensions for (b) (mils)</td>
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<td>N. A.</td>
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</tbody>
</table>

Table 9-3 Dimensions for Figure 9-5
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, $h=25$ mil and relative dielectric constant $\varepsilon_r=10.2$. It can be seen that the resonant frequencies do not shift much as a result of the introduction of the feed lines ($\lambda_g/4$). Therefore, another more simplified way of setting the frequencies for the attenuation poles is to first analyze the square loop resonator with two openings and then add in the feed lines.

A point to note for this type of filter is that the fractional bandwidth of the resonator is around 5%.

![Figure 9-6 Effects of directly and indirectly connecting the feed lines](image)
9.5 Meander Loop Resonator

As discussed, meander loop will help to miniaturize the design. Figure 9-7 shows the configuration with this concept implemented on a single square loop resonator with two gap-openings. Each loop has an area of 30 mil x 30 mil and it is attached to the outer side of the original configuration in Figure 9-5 (b). Dimensions are given in Table 9-4.

![Figure 9-7 Configurations to study meander single loop resonators](image)

<table>
<thead>
<tr>
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<tr>
<td>Dimensions for (b) (mils)</td>
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</tr>
</tbody>
</table>

Table 9-4 Dimensions for Figure 9-7
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, \( h = 25 \text{mil} \) and relative dielectric constant \( \varepsilon_r = 10.2 \). Figure 9-8 shows the frequency response comparison for the meander loops configurations. The resonator with 2 meander loops has the resonant frequencies shifted from 6.2GHz for 1 meander loop to 5.4GHz. This indicates that with more meander loops, the resonator can be further miniaturized.

![Figure 9-8 Meander loops implemented on the single loop resonators](image)

**9.6 Suppression of Spurious Harmonics**

Figure 9-9 presents a third type of meander loop design. It is designed based on the same principle used in [31] to “migrate” the coupling mode between the excitation lines and the microstrip resonator. One of the meander loops is “opened-up” to
serve as the open-end microstrip line connected to the excitation-coupled lines of the resonator. It is being compared with the configurations in Figure 9-7 (a) and (b).

Detailed dimensions are given in Table 9-5.

![Figure 9-9 Meander configuration to suppress spurious harmonics](image)

<table>
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</table>

Table 9-5 Dimensions for Figure 9-9

Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, $h=25$ mil and relative dielectric constant $\varepsilon_r = 10.2$. Figure 9-10 shows the wideband response of the filter designs in Figure 9-7 (a) and (b) and Figure 9-9, annotated as (c) in Figure 9-10. The centre frequency is around 7GHz. It can be seen that in addition to giving a better match across the pass band, the design with two meander loops also produces an additional attenuation pole at around the first harmonics.
Figure 9-10 Harmonic migration with the new coupling control

Figure 9-11 S11 for the three configurations
Optimization can be done by tuning the “opened-up” lines. The configuration and the dimensions for this investigation are given in Figure 9-12 and Table 9-6.

![Figure 9-12 Configuration to study the length optimization of the meander loop resonator](image)

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</tbody>
</table>

Table 9-6 Dimensions for Figure 9-12
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, $h$=25mil and relative dielectric constant $\varepsilon_r$=10.2. Figure 9-13 shows the response by first lengthening the vertical line directly attached to the feed line by 20 mils. This has improved the rejection band. Then, the design is made “symmetrical” by lengthening all the open-ends by 20 mils. The response is represented by the thicker line in Figure 9-13. An additional attenuation pole is introduced at 14.5GHz as a result of the optimization and the rejection band has also improved.

Figure 9-13 Optimization of the coupled excitation lines
9.7 Measured results

The configuration in Figure 9-2 is fabricated using the laser mask and photo-etching technique. Measurement results and the photograph of the hardware of the newly developed coupled line meander loop resonator are shown in Figure 9-14 to Figure 9-16. The measured results are similar to the simulated ones. The good high band rejection can be seen clearly from Figure 9-14 and the S11 is better than -10dB across the pass band. The hardware is mounted on aluminum base to provide better grounding and secured connections with the Agilent Network analyzer.

The measured result for this design has a rejection band around -25dB. The measured 3dB bandwidth is 262MHz with a fractional bandwidth of 3.94%. This is a narrowband response. There are two attenuation poles (S21 is minimum) at around 6GHz and 7.5GHz. The insertion loss is 2.1841dB. Losses are found in the fabricated hardware only because the simulation results do not take into account the conductor loss. In addition, in the simulated conditions, the ground is assumed to be infinite and the effects of the connectors at the input and output ports are also not included. There is also the tolerance of the fabrication process to be taken into consideration. This is especially crucial because of the small size of the configuration. Accounting for all those, the measured and simulated results agree well with each other.
Figure 9-14 Simulated and measured results for the miniaturized narrowband resonator

Figure 9-15 Simulated and measured S11 of the miniaturized narrowband resonator
9.8 Overall Performance

Table 9-7 summarizes the measured performance of the newly proposed narrowband single loop resonator and compares it against the simulated results. This design has a very narrowband response and can be used for very narrowband application in communications. The design of this filter demonstrates that for a miniaturized design, both the wideband and narrowband applications can be achieved. For wideband applications, coupled line configuration can be used and for narrowband configuration, single loop can be used.
<table>
<thead>
<tr>
<th>Property</th>
<th>Simulated results</th>
<th>Measured results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Midband insertion loss, IL</td>
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<td>2.1841dB</td>
</tr>
<tr>
<td>Frequency range</td>
<td>6.725GHz – 7.050GHz</td>
<td>6.525GHz – 6.787GHz</td>
</tr>
<tr>
<td>3dB Bandwidth</td>
<td>325MHz</td>
<td>262MHz</td>
</tr>
<tr>
<td>Fractional bandwidth</td>
<td>4.72%</td>
<td>3.94%</td>
</tr>
<tr>
<td>Coupling coefficient, k</td>
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<td>0.0393</td>
</tr>
<tr>
<td>$Q_{\text{loaded}}$ [16]</td>
<td>21.2</td>
<td>25.4</td>
</tr>
</tbody>
</table>

Table 9-7 Summary of the performance of the narrowband miniaturized configuration

The current plot of the configuration in Figure 9-2 is shown in Figure 9-17. The warmer color (near the red side of the spectrum) represents higher magnitude of current. The cooler color (near the blue side of the spectrum) represents lower magnitude of current. The current plot shows that higher magnitude of current is found at the input port of the design. Both electric and magnetic couplings can be observed from the current distribution.

Figure 9-17 Current plot of the narrowband meander loop design
10 Cascaded miniaturized narrow band units

10.1 Introduction

The miniaturized design for narrowband can be cascaded to form filters with even better higher band rejection. However, this narrowband filter seems to be very difficult to match. This is because the configuration is smaller and the cross coupling between the conductor lines are stronger. In this chapter, a new way of connecting the individual unit will be explored. It is found that this new way of connecting the periodic unit can provide a better rejection band.

10.2 Layout of the cascaded network

Two ways are proposed here to connect the narrowband miniaturized unit in Chapter 9. Figure 10-1 (a) shows the cascaded network with the same type of connections between the two units as in the miniaturized wideband case. As with the case shown for the meander loop coupled line resonator, the connection length between the two units is made as small as possible. It is chosen to be 10mil in (a). In another alternative method of cascading, the connection length is made to be zero, meaning the two units are touching each other. If the two units are touching, the frequency response will be affected significantly for the meander loop coupled line resonator. Investigation will be performed for this non-coupled line case. The dimensions are given in Figure 10-2 and Table 10-1.
The two units forming the cascading configurations are mirror images of each other.

(a)

Figure 10-1 Configurations (a) and (b) to study the cascading networks
Figure 10-2 Dimensions reference for Figure 10-1

<table>
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<tr>
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</table>

Table 10-1 Dimensions for Figure 10-2
Full wave analysis is done using Agilent’s momentum software. The simulation is based on a RT/Duroid 6010 substrate with a thickness, \( h = 25\text{mil} \) and relative dielectric constant \( \varepsilon_r = 10.2 \). The comparisons for the S21 and S11 of the two cascaded networks are shown in Figure 10-3 and Figure 10-4. It can be observed that not only is the S11 at the pass band not as good for the case with longer connection, it also introduces spurious harmonics. This has contradicted the good response of the single unit. The cascaded network with connection length equal to zero shows a better response for the higher frequency range. The attenuation pole for the lower frequency side is still present.

Figure 10-3 Comparison of the S21 responses for the two cascaded networks
Figure 10-4 Comparisons of S11 responses of the two cascaded networks

Figure 10-5 Wideband performance for the 2 cascaded networks
Figure 10-5 shows the wideband response for the two proposed type of cascaded networks. The originally migrated harmonics response has appeared again. This is due to the close proximity of the two units. By applying EBG patterns on the ground plane of the configuration will remove this spurious harmonics.

### 10.3 Measurement results

The configuration with zero connection length between the two units has been fabricated using the laser mask and photo-etching technique. Measurement results and the photograph of the hardware of the miniaturized cascaded narrowband resonator are shown in Figure 10-6 to Figure 10-8. The hardware is mounted on aluminum base to provide better grounding and secured connections with the Agilent Network analyzer.

The measured result for this design has a -10dB bandwidth of 125MHz with a fractional bandwidth of 1.92%. This is a narrowband response. The side band rejection is better than -45dB. The insertion loss is 5.3799dB. Losses are due to conductor loss. There is also the tolerance of the fabrication process to be taken into consideration. This is especially crucial because of the small size of the configuration. Many filters used commercially are followed by amplifiers. Therefore, the loss is not the main concern. Accounting for all those, the measured and simulated results agree well with each other.
Figure 10-6 Simulated and measured S21 responses of the cascaded network

Figure 10-7 Simulated and measured S11 responses of the cascaded network
10.4 Overall Performance

Table 10-2 shows the performance comparisons between the simulated and measured results, the two agree quite well except for the loss. The table also shows the comparisons with the non-cascaded configuration. The loss for the single unit is lower, with 2.1841dB while that for the cascaded configuration, the loss is 5.3799dB. The bandwidth is narrower for the cascaded one. Fractional bandwidth of the single unit is 3.94% and the cascaded unit is 1.92%.
<table>
<thead>
<tr>
<th>Property</th>
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<th>Measured results of cascaded unit</th>
<th>Measured results of single unit</th>
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</thead>
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<td>5.4dB</td>
<td>2.2dB</td>
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<td>Frequency range</td>
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<td>6.463GHz-6.588GHz</td>
<td>6.513GHz-6.813GHz</td>
</tr>
<tr>
<td>Bandwidth (-10dB)</td>
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<td>125MHz</td>
<td>262MHz</td>
</tr>
<tr>
<td>Fractional bandwidth</td>
<td>1.09%</td>
<td>1.92%</td>
<td>3.94%</td>
</tr>
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<td>Coupling coefficient, k</td>
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<td>$Q_{\text{loaded}}$ [16]</td>
<td>92.2</td>
<td>52.2</td>
<td>25.4</td>
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</tbody>
</table>

**Table 10-2** Comparisons between simulated and measured performance for the cascaded units
The current plot for the configuration in Figure 10-1 is shown in Figure 10-9. The warmer color (near the red side of the spectrum) represents higher magnitude of current. The cooler color (near the blue side of the spectrum) represents lower magnitude of current. The current plot shows that higher magnitude of current is found at the input and output ports of the design. Mixed coupling as can be deduced from the current distribution is found in this design.

![Figure 10-9 Current plot for the cascaded network](image-url)
11 Conclusion

In this Master thesis, research has been done on the investigation and design of microstrip bandpass filters using square open loop resonators. Based on the theory of existing square open loop resonator with one gap-opening, mathematical analysis is performed on square loop with two gap-openings. This new configuration can be designed using the equivalent circuits for modeling gaps, bends and open end. The ABDC and Y parameters of this equivalent circuit is calculated and found that the transmission zeros, that is, when S21 frequency response of the filter is maximum, obtained from this mathematical analysis tallies with the full wave simulation results using Agilent’s momentum software. In addition, by the odd and even analysis, the attenuation poles, meaning when the S21 frequency response is minimum, can also be calculated and is found to be close to the simulated results using full wave analysis.

Starting from the square open loop resonator with two gap-openings, coupled line resonator is proposed. This dual mode coupled line square loop resonator consists of inner and outer square loops joined together. It has two gap-openings on each of the inner and outer loops, and can reduce the size of the single loop resonator. Similar to the single loop case, electric, magnetic and mixed coupling configurations of the coupled line resonators have been proposed. The mixed coupling configuration is fabricated and measured. It is found to have low insertion loss of 0.7766dB. In
addition, its fractional bandwidth is 13.2%. There is also a size reduction of 27.7% compared to the four-pole square loop resonator.

The coupled line microstrip resonator is further improved on using the meander loop concept. It has been published that meander loop configuration can reduce the size of square loop resonators. It is therefore implemented on the coupled line resonator. In addition, a coupling scheme that is known to be able to implement higher order filter characteristics from lower order sections is applied to the coupled line microstrip resonator. The fabricated resonator is found to have a fractional bandwidth of 5.98% and insertion loss of 1.2565dB.

Two units of the coupled line meander-loop resonators are cascaded together. Investigations on their orientations show that they should be mirror images of each other. This is like coupling two single square loop resonators for the electric coupling case. This cascaded unit is fabricated and has a measured 3dB bandwidth of 5.99%. The overall performance of the cascaded configuration is very similar to the single unit configuration. However, by cascading two units, a measured rejection band of better than -60dB is achieved. This makes the filter highly selective and desirable.

As it is always the demand of modern communication to reduce the size of circuits, smaller filters are highly desirable. Therefore, from the proposed coupled line meander loop resonators, miniaturized structures with the width of conductor line
reduced to half that of the 50ohm line width is designed. It is found that this miniaturized design can increase the bandwidth of the filter tremendously. In addition, in designing the cascading of the units to get better rejection band, the crossing coupling effect do not have to take into account. Three units of the miniaturized units are cascaded by directly connecting them side by side. The fabricated hardware has a bandwidth of 23.6% and a rejection band better than -40dB.

Another miniaturized cascaded configuration is also designed by removing the inner interposed loops from the inner loops. This filter is found to be able to push back the first harmonics. The measured bandwidth of this miniaturized filter is also 23.6%. The main difference between this design and the one with the interposed inner loops is the delay in first harmonics.

Using the design concept of conductor line width half that of the 50ohm, a single loop narrowband bandpass filter with meandered configuration is designed. This meander loop resonator also has two gap-openings in the loop. In addition, one of the meander loops is “opened-up” to serve as the open-end microstrip line connected to the excitation-coupled lines of the resonator. It is known that this can “migrate” the second resonant frequencies of the square loop resonator and give a good narrowband response. The configuration of this narrowband filter is fabricated and has a measured bandwidth of 3.94%. This narrowband filter can be used in many communications applications requiring high selectivity.
The cascaded network consisting of two units of the narrowband meander loop resonator is also designed. They are cascaded side by side and it is found that when the distance between the two units is zero, meaning they are touching each other, the response is the best. However, the high frequencies response is not as good as the single unit. Nevertheless, the cascaded network is able to achieve a better rejection band performance of better than -40dB as compared to about -20dB for the single unit.

As a result of this research on coupled line square loop resonator and reduced conductor line width filters, ten filters designed have been fabricated and their performance measured. All the measurement results agree well with the simulated results.
12 Future Works

The work done in this thesis has devoted much attention to studying the internal coupling effects of the newly proposed dual mode coupled line resonator. Mathematical models and design equations can be synthesized to design the filter with specific rejection band levels just like the square open-loop resonator. In addition, mathematical coupling schemes can be applied to the single resonator and be used to design more highly selective filters. The cascading method shown in Chapter 5 is the very first step of developing even more completed cascading filter network.

Mathematical models using left-handed and right-handed transmission line models can also be applied to the proposed coupled line resonator to generate equations for defining the transmission and attenuation zeros of the band pass filter.

In addition, the coupled line square open loop resonator can also be explored to be implemented on other types of filters. Striplines and coplanar microstrip filters are some examples of filters that can use this coupled line square open loop structure. As microstrip lines are a type of transmission lines, other types of transmission lines like coaxial lines and waveguides can also be applied by this coupled line resonator concept. It can also be applied on other types of materials, such as High-temperature superconductors (HTS), ferroelectrics, micromachining or microelectromechanical systems (MEMS), hybrid or monolithic microwave integrated circuits (MMIC),
active filters, photonic bandgap (PBG) materials/structures, and low-temperature cofired ceramics (LTCC).

HTS microstrip filters use HTS thin films instead of conventional conductor films on the surface. It can lead to significant improvement of microstrip filter performance with regard to the pass band insertion loss and selectivity. This is particularly important for narrow-band filters, which play an important role in many applications. Recent research has already been using hairpin band pass filters and single meander loop structures on high power HTS applications. Similar research can be explored for the coupled line resonators proposed in this thesis.

Ferroelectric tunable filters are fast, small, lightweight, and, because they work on electric fields, have low power consumption. The range of tuning is quite large and devices are relatively simple in nature. Ferroelectric materials can be incorporated into the coupled line resonator by depositing conducting films on both sides of the surfaces of bulk the ferroelectric substrates. Because of the high dielectric constant of ferroelectric substrate, the sizes of the resonators can be very small and using HTS thin films can also help to reduce conductor losses.

Micromachine filters can be implemented by suspending the microstrip coupled line loop resonator on thin dielectric membranes to eliminate dielectric loss and dispersion problems, giving pure TEM mode of propagation and conductor-loss limited performance.
MIC circuits are hybrid microwave circuit in which a number of discrete active and passive components attached externally to an etched circuit on a common substrate. Generally, MIC circuits have advantages of having low cost, high yield and good reliability.

Miniaturization is one of the concerns of modern demand of communication circuits. The designs presented in this thesis can be explored to combine with EBG structures to further improve the performance of the filters. EBG patterns can be applied on the ground of the coupled line filter in this thesis.

Further works can also be on implementing the coupled line square loop resonator and the miniaturized filters at the feed lines of Antennas and at the input and output ports of microwave amplifiers to form active microwave filters.
REFERENCES


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