

KU-BAND SUBSTRATE INTEGRATED WAVEGUIDE BANDPASS FILTER DESIGN USING MECHANICAL TUNING

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By
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We certify that we have read this thesis and that in our opinion it is fully adequate,
in scope and in quality, as a thesis for the degree of Master of Science.

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ABSTRACT

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M.S. in Electrical and Electronics Engineering

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Substrate integrated waveguide (SIW) filters are widely used due to their easy fabrication techniques and relatively high quality factor. In microwave frequencies, manufacturing tolerances become comparable with wavelength and affect filter performance. Tuning resonator center frequencies and couplings between resonators allow correcting the errors that are caused by imperfections. An iris bandpass filter is designed in Ku-band using coupling matrix approach which relates mathematical coupling values to physical dimensions. SIW tuning is usually done by adding varactors to the circuit. This introduces nonlinearities which may be undesirable while also requiring external control voltages. A mechanical tuning method for SIW, similar to tuning screws that are used in waveguide filters, is analyzed for resonators and couplings. It is modified by using small surface mount resistors to achieve a more compact structure. Tuning mechanisms are inserted to designed filter and fabricated on a low loss substrate. Time domain method for tuning coupled filters is introduced as a tuning algorithm. Using mechanical tuners, SIW filter is tuned to desired filter performance.

Keywords: Substrate Integrated Waveguide, Bandpass, Tunable Filter, Ku-band.

ÖZET

MEKANİK AYARLAMA İLE KU-BANT GÖMÜLÜ DALGA KILAVUZU BANT GEÇİREN FİLTRE TASARIMI

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Elektrik-Elektronik Mühendisliği, Yüksek Lisans

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Gömülü dalga kılavuzu (GDK) yapısındaki filtreler kolay üretim teknikleri ve görece yüksek kalite faktörleri sebebi ile yaygın olarak kullanılmaktadır. Mikro-dalga frekanslarında, dalga boyu ile karşılaştırılabilir hale gelen üretim toleranslarından filtre performansı etkilenmektedir. Bu bozukluklar rezonatörlerin merkez frekansı ve rezonatörler arasındaki bağlaşım ayarlanarak düzeltilebilir. Matematiksel bağlaşım değerlerini fiziksel boyutlarla ilişkilendiren bağlaşım matrisi yöntemi ile Ku-bantta çalışan iris filtre tasarlanmıştır. GDK yapısındaki filtreler genellikle varaktör kullanılarak ayarlanır. Bu yöntem filtrenin doğrusallığını bozmakta ve fazladan kontrol gerilimi gerektirmektedir. Normal dalga kılavuzu yapılarında kullanılan ayar vidalarına benzer bir yapı rezonatörler ve bağlaşım için analiz edilmiştir. Ayar yapısını daha kompakt yapabilmek için döndürülebilir küçük yüzey monte dirençler kullanılmıştır. Bu mekanizma önceden tasarlanan filtreye eklenmiş ve az kayıplı bir alttaş kullanılarak üretimi yapılmıştır. Ayarlama algoritması olarak zaman alanında kullanılan bir yöntem sunulmuştur. Bu mekanik ayar yapıları kullanılarak GDK filtre istenilen performansa ayarlanabilmektedir.

Anahtar sözcükler: Gömülü Dalga Kılavuzu, Bant Geçiren, Ayarlanabilir Filtre, Ku-bant.

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Chapter 1

Introduction

With increasing radar and communications technologies, tighter specifications for microwave filters are needed. Most commonly used type of filters are bandpass filters which requires careful design process. Parameters like insertion loss, return loss, bandwidth and attenuation have to be optimized depending on application. However, no matter how much the designer tries to achieve the perfect filter, there are usually trade-offs and errors. Most of the errors become apparent during the manufacturing phase. There are many manufacturing parameters that affect the filter performance directly. In addition, these parameters are never exact, they have some tolerances which are usually shared by the manufacturer. A via hole may be misaligned by a margin or distance between two microstrip lines may be wrong due to etching process. Even dielectric constant can vary across the panel.

Upper Ku-Band is commonly used for radar applications. Since the frequency is relatively high, high performance filters are difficult to design and manufacture. In addition, insertion loss becomes critical due to the noise figure of receiver. For this reason resonators must have high quality factor. Another concern for this frequency band is the attenuation of SATCOM frequency which is close to 14 GHz. The filter bandwidth cannot be wider than necessary to satisfy the attenuation requirement.

In microwave frequencies metal waveguides are used extensively to construct filters, dividers, couplers, etc. With widespread availability and high accuracy of 3D electromagnetic simulation software, high performance can be achieved with relative ease. However, metal waveguides are often bulky and expensive to manufacture. In addition, it is difficult to integrate with other parts of the system, requiring special transition designs usually with a tuning mechanism to compensate for manufacturing and alignment errors.

Substrate integrated waveguide (SIW) has been proposed to overcome these difficulties while keeping the advantages of conventional waveguide [1]. SIW consists of two arrays of vias parallel to each other on a printed circuit board. These vias behave as short edges of a rectangular waveguide while top and bottom layer of PCB make up for long edges. Since there are no additional components, SIW structures can be easily realized with conventional PCB manufacturing techniques. This makes the integration of SIW and other planar circuitry on the board trivial. Many RF products such as filters, couplers, dividers, amplifier, etc are being developed using SIW technology [2–7].

SIW filters have some advantages like high quality factor, low radiation and conductor losses, etc. compared to microstrip filters [8]. However, fabrication tolerances and errors of microstrip are also present in SIW. These tolerances may cause serious problems especially in higher microwave frequencies where wavelength becomes comparable with measure of errors. The problems usually manifest themselves as mismatch in pass band, wider or narrower bandwidth, higher or lower center frequency or all at the same time. To overcome the fabrication tolerances, tuning mechanisms are used in microstrip, waveguide and SIW structures. Tuning screws are inserted to waveguide filters to control resonance frequencies of resonators and couplings between resonators [9, 10]. Varactor diodes are used in microstrip and SIW filters for the same purposes [11, 12]. Disadvantage of varactor diodes are nonlinearity of the devices and necessity of an external voltage to control the capacitance. Tuning screws on microstrip and SIW is not practical since they are both planar structures. However, there is one proposed method to mechanically tune SIW resonators and filters [13].

The purpose of this thesis is to design and tune a bandpass filter with specific design criteria using synthesis and tuning methods that are usually used in rectangular waveguide filters. The designed filter will work in Ku-Band with center frequency of 16.5 GHz. At least 1 GHz of passband with minimal ripple is required. Return loss should be lower than -15 dB between 16 and 17 GHz. In addition, attenuation at 14 GHz must be better than 30 dB.

In Chapter 2, a Ku-band SIW bandpass filter is designed using coupling matrix method in CST Microwave Studio environment.

In Chapter 3, tuning mechanism is explained and integrated to the designed filter in simulation environment.

In Chapter 4, designed filter is manufactured, measured and tuned using time-domain tuning method.

Chapter 2

SIW Design

2.1 SIW Basics

Substrate integrated waveguides have very similar properties with rectangular waveguides. In order to design a SIW device effectively, a basic understanding of rectangular waveguides is required.

In a rectangular waveguide, both TE and TM mode waves can propagate. However, dominant mode for propagation is TE_{10} having the lowest cut-off frequency [14]. The cut-off frequency for this mode is calculated by;

$$f_{c_{10}} = \frac{1}{2a\sqrt{\mu_r\epsilon_r}} \quad (2.1)$$

where μ_r and ϵ_r are medium parameters that fills the waveguide. It is important to notice that only the longer dimension of the waveguide, which is denoted by a , determines the cut-off frequency. It must be ensured that operating frequency is higher than the cut-off frequency of the waveguide. In addition, it is a good idea to design the waveguide in a way that the cut-off frequency of next propagating mode should be higher than operating frequency. This is to ensure that only one mode propagates so that no energy is lost.

In a SIW structure, distance between the via arrays will make up the longer dimension and will be denoted by a_s . Additionally, d and p parameters will represent via diameter and distance between the centers of individual vias respectively.

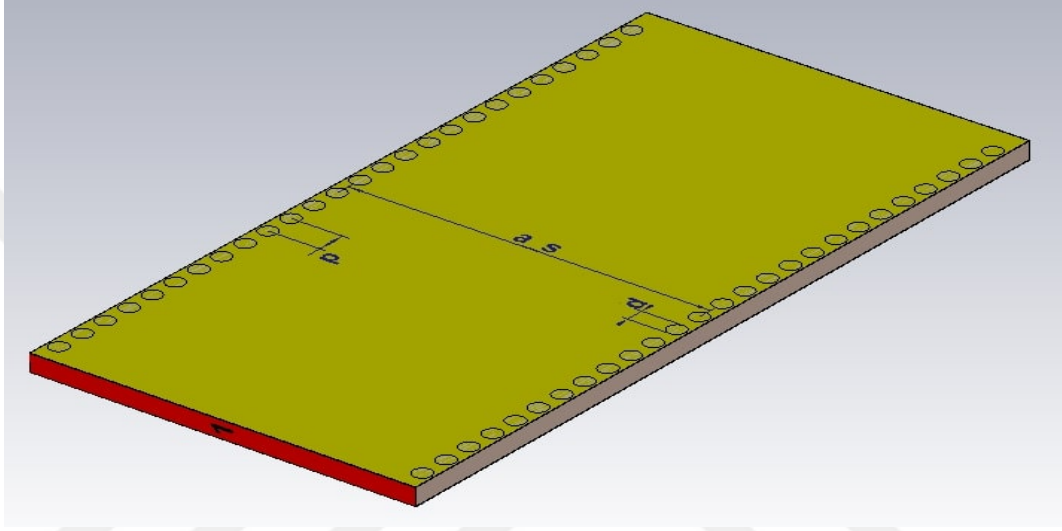


Figure 2.1: Substrate integrated waveguide

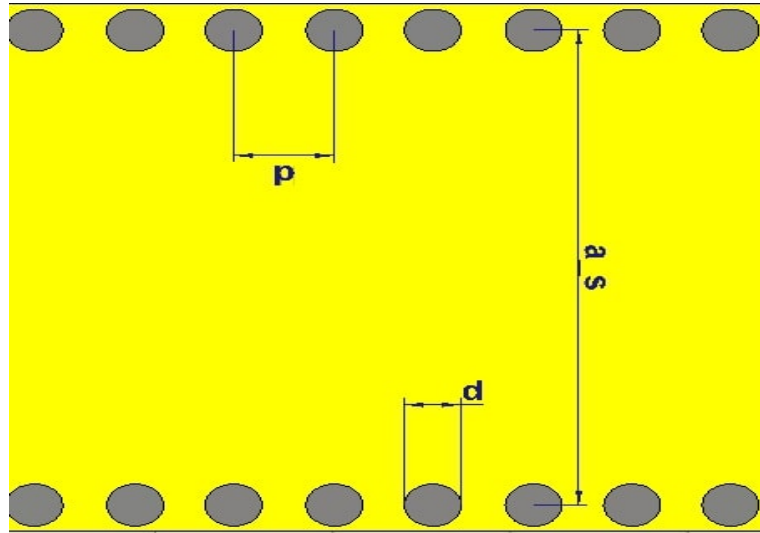


Figure 2.2: View from top of a SIW

Equation for a_s calculation is given in [15];

$$a_s = a_d + \frac{d^2}{0.95p} \quad (2.2)$$

where a_d is the width that a dielectric filled waveguide must have to have the

same cut-off frequency with an air filled waveguide and it can be calculated as

$$a_d = \frac{a}{\sqrt{\epsilon_r}} \quad (2.3)$$

A SIW is simulated for different a_s values using CST Microwave Studio and s_{21} for the dominant mode is given in Figure 2.3. Substrate is chosen as 20 mils thick Arlon CLTE-XT due to its low loss properties in microwave frequencies.

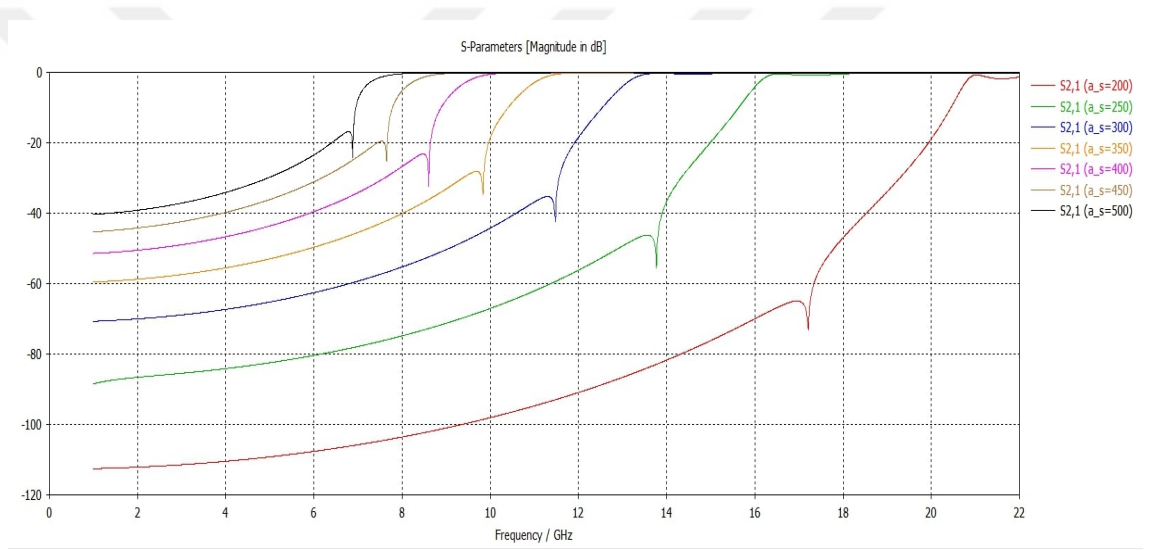


Figure 2.3: s_{21} of SIW when a_s is swept

Figure 2.3 shows that cut-off frequency decreases when a_s is increased as expected from Equation 2.1. Since the designed filter will operate in upper Ku-Band, a_s is selected as 350 mils which provides a cut-off frequency close to 11 GHz.

Parameters p and d have to be carefully selected in order to minimize leakage losses at the operating bandwidth. A rule of thumb for selection of these parameters is given in [16] as;

$$d < \frac{\lambda_g}{5} \quad (2.4)$$

$$p < 2d \quad (2.5)$$

where λ_g is wavelength of a guided wave. It can be calculated as given in [14];

$$\lambda_g = \frac{2\pi}{\sqrt{(\frac{\epsilon_r(2\pi f)^2}{c^2}) - (\frac{\pi}{a})^2}} \quad (2.6)$$

For $a_s = 350$ mils, guided wavelength becomes 520 mils. This makes the maximum value of d close to 100 mils. In order to reduce total filter size, and for a cleaner look, d is chosen as 16 mils which is the smallest available via diameter that can be easily manufactured in available facilities. After choosing a_s and d , an optimization is run to get the ideal p value that will result in minimum loss which is 26 mils. Finally, all required parameters of SIW are known.

2.2 SIW Transition

Although SIW as a transmission line has many advantages, it still have to be integrated to the other planar circuitry. There are many ways to implement a microstrip-to-SIW transition. The simplest way to do this is using a tapered microstrip line connecting to the SIW. This transition can be seen in Figure 2.4.

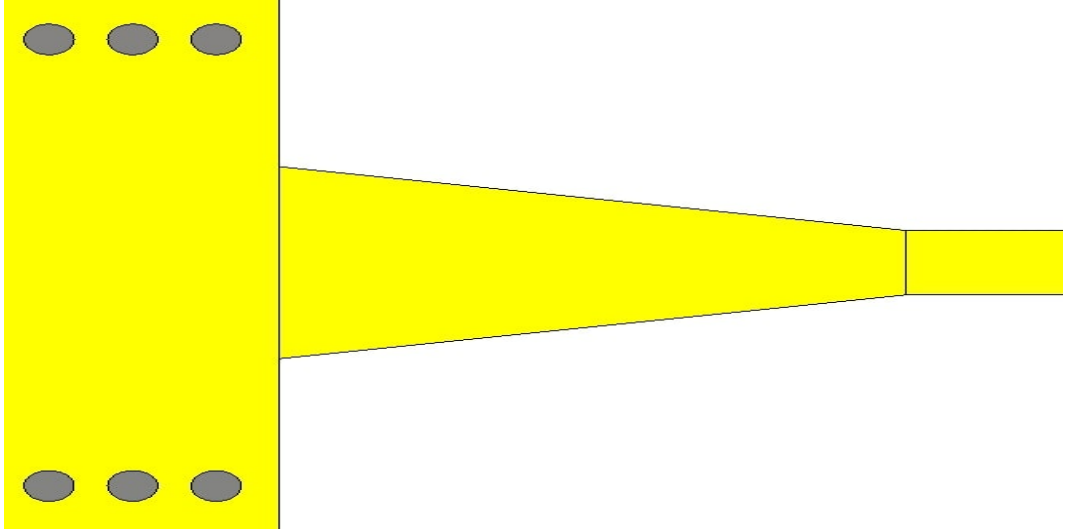


Figure 2.4: Tapered line transition

The parameters w_{ms} , w_t and l_t are width of 50Ω microstrip line, width of the tapered line and length of the taper respectively. Design of these parameters is

very straightforward and explained in great detail in [17]. After calculating the initial values for the taper parameters an optimization is run in CST to ensure the best performance since design of the transition affects the filter performance significantly. The through SIW design with tapered line at both input and output ports can be seen in Figures 2.5 and 2.6.

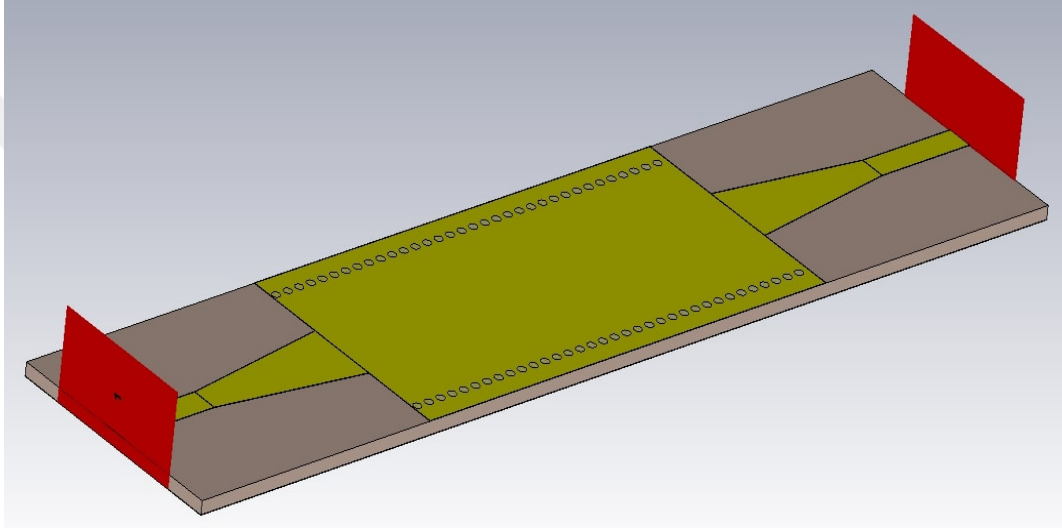


Figure 2.5: A SIW through line

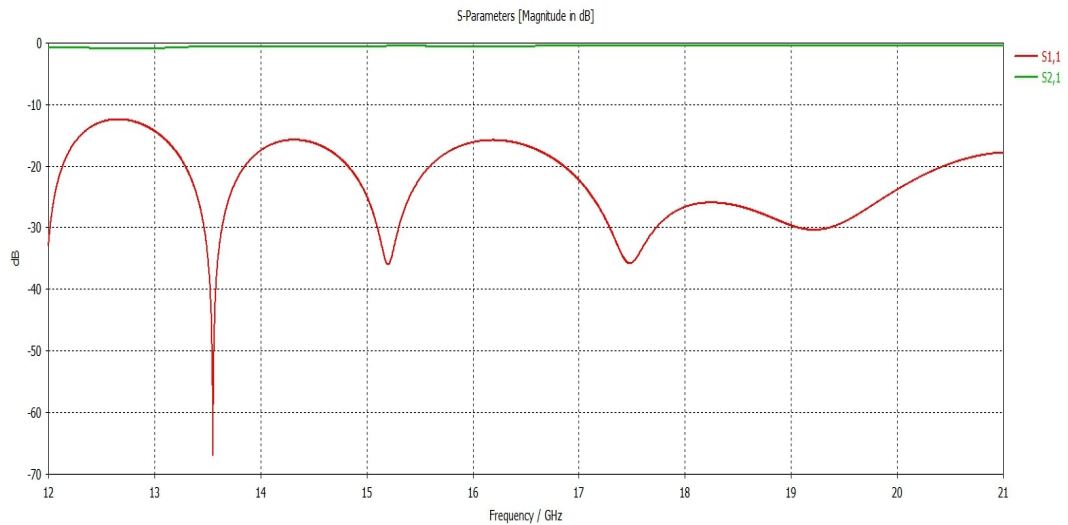


Figure 2.6: Simulation results of through line

After simulation and optimization, the SIW is manufactured and measured. The substrate is 20 mils thick Arlon CLTE-XT which has a dielectric constant of 2.94.

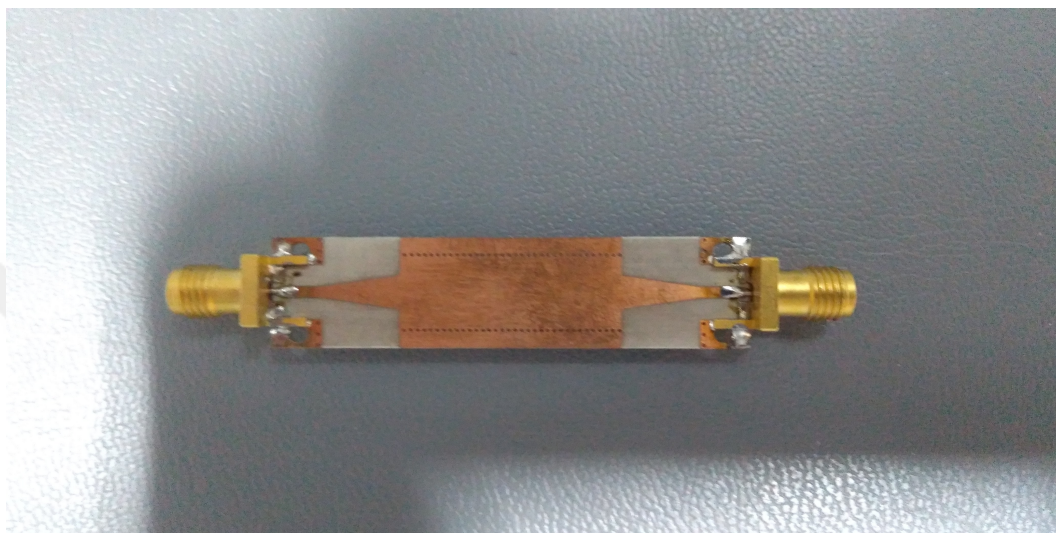


Figure 2.7: SIW through line PCB

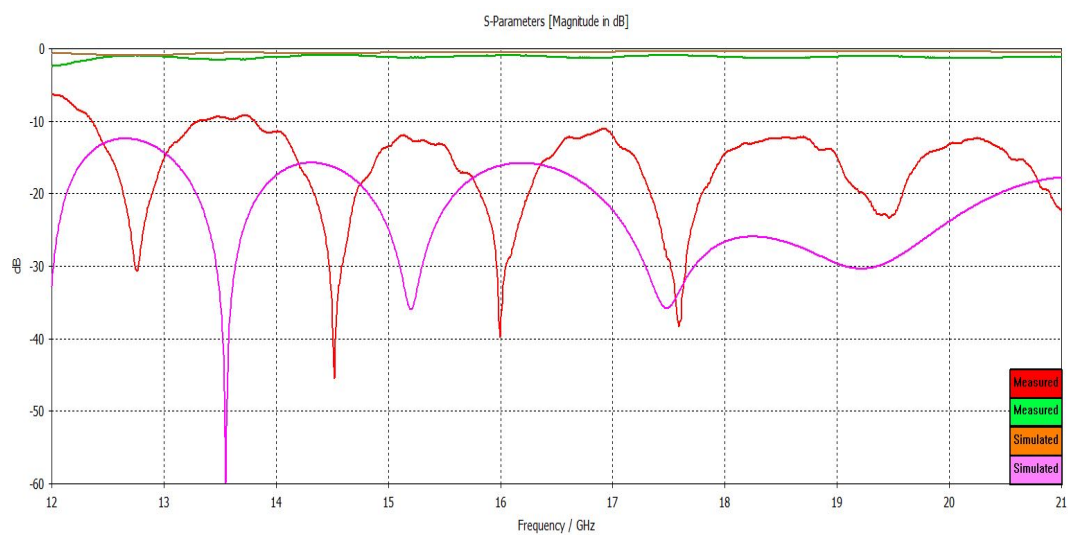


Figure 2.8: SIW through line simulated and measured results

Figure 2.8 shows that simulation and measurement results are similar though there is some additional mismatch on the PCB. This difference can be attributed to end launch SMA connectors that are used in measurement setup but not included in simulation.

2.3 SIW Filter Design

The usual method for synthesizing microwave filters is to calculate mathematical polynomials, obtain a prototype and extract element values one by one. A more practical way to design is to use coupling matrix method in which one can relate the mathematical results to physical structure relatively easily. This method was first proposed in 1971 [18] and is still widely used in microwave filter design.

2.3.1 Coupling Matrix

In simplest terms, a filter consists of some resonators coupled to each other in some manner. A coupling matrix is basically a numerical representation of these couplings. Any filter can be created if these couplings can be formed in physical shape. The necessary dimensions to obtain the required couplings can be easily extracted and optimized by using a CAD software.

A coupling matrix has $N \times N$ elements where N represents the filter order. Row and column numbers are resonator numbers. Nonzero elements show that there is a coupling between those resonators. Coupling of source and load to the filter can be also included which makes the matrix have $(N + 2) \times (N + 2)$ elements.

For simplicity, a third order iris filter is designed using this method. Only sequential resonators have nonzero couplings which makes the matrix take the form:

$$C = \begin{bmatrix} 0 & c_{S1} & 0 & 0 & 0 \\ c_{S1} & 0 & c_{12} & 0 & 0 \\ 0 & c_{12} & 0 & c_{23} & 0 \\ 0 & 0 & c_{23} & 0 & c_{3L} \\ 0 & 0 & 0 & c_{3L} & 0 \end{bmatrix} \quad (2.7)$$

where c_{12} , c_{23} denote coupling between the resonators and c_{S1} , c_{3L} are couplings between source to first resonator and third resonator to load respectively. Note that due to symmetry; $c_{S1} = c_{3L}$.

Using the code provided in [19], coupling matrix elements are extracted for an iris filter with 16.5 GHz center frequency and 2 GHz bandwidth. Required coupling values are found as; $c_{S1} = c_{3L} = 6.7546$, $c_{12} = c_{23} = 0.1478$ Now that coupling values are known, physical dimensions can be found using a 3D electromagnetic solver like CST.

2.3.2 Single Resonator

A single SIW resonator can be approximately simulated as a dielectric cavity sitting in a perfect electric conductor (PEC) environment. Using the CST eigenmode solver, resonant frequency can be found easily. Simulation setup in CST can be seen in Figure 2.9 in which background material is PEC.

w_{res} and l_{res} are width and length of the resonator. Resonant frequency depends on the volume of the cavity. Figure 2.10 shows the resonant frequency when $w_{res} = 340$ mils and l_{res} is swept.

In order to get a resonance at 16.5 GHz, resonator length should be 265 mils.

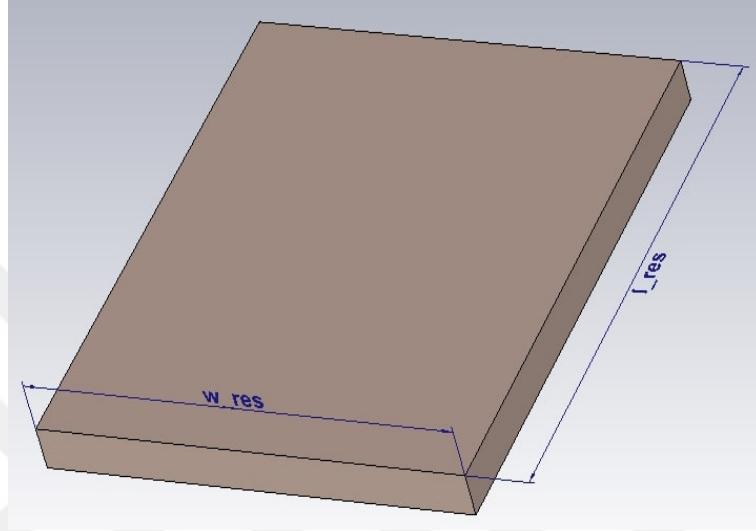


Figure 2.9: Single resonator in PEC

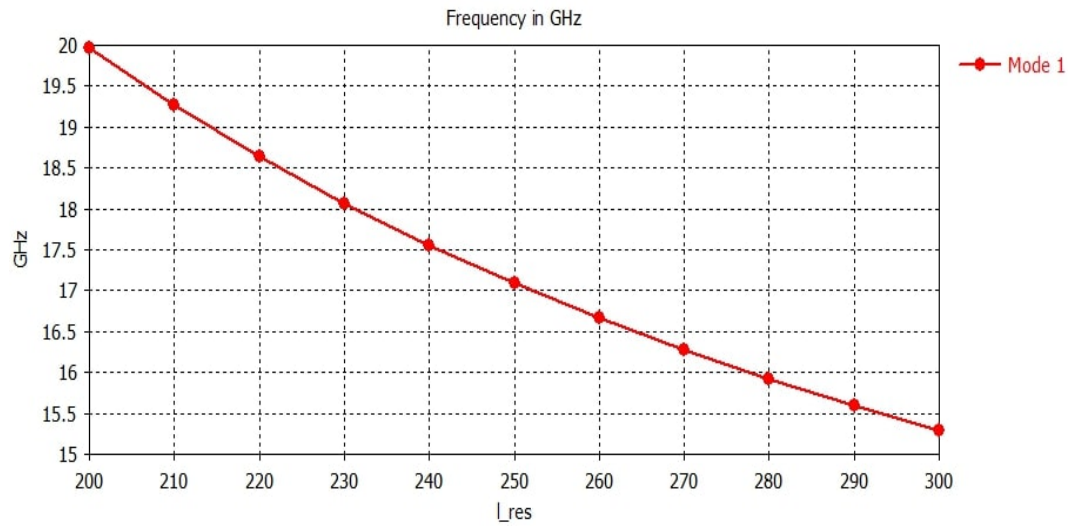


Figure 2.10: Resonant frequency of a single SIW resonator

2.3.3 Source and Load Coupling

In Figure 2.11 coupling mechanism between source and first resonator is shown. Width of the coupling aperture is w_{ext} and distance to the source port is l_{ext} . Note that this structure is also used for load and third resonator coupling. Using the external Q calculator in eigenmode solver, the parameters can be optimized for required coupling and loaded frequency.

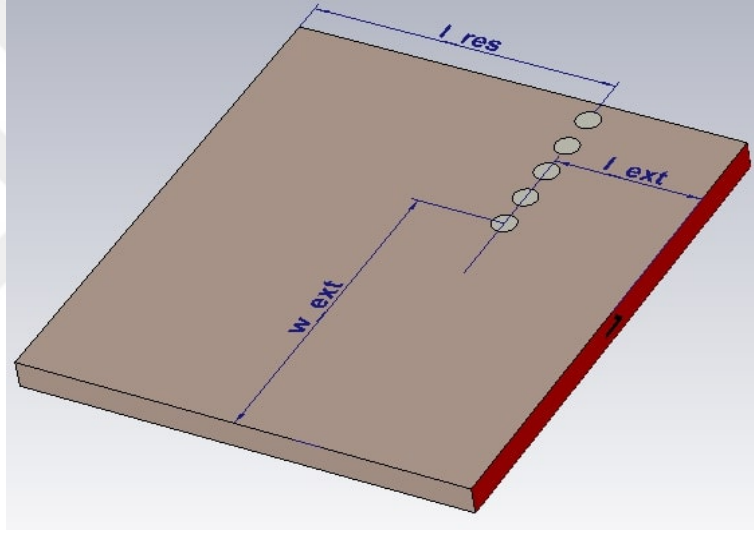


Figure 2.11: Coupling between source(load) and first(last) resonator

For $w_{ext} = 225$ mils, $l_{ext} = 100$ mils and $l_{res} = 210$ mils; external Q is 7.27 and loaded frequency is 16.23 GHz. Although these values are not exact, they provide a starting point for a shorter optimization time.

2.3.4 Coupling Coefficient

In order to couple the resonators, an aperture is used similar to regular waveguide iris filters. When two modes are coupling to each other, coupling coefficient can be calculated using the equation;

$$k = \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2} \quad (2.8)$$

where f_1 and f_2 are center frequencies of two modes that have the lowest cut-off frequency.

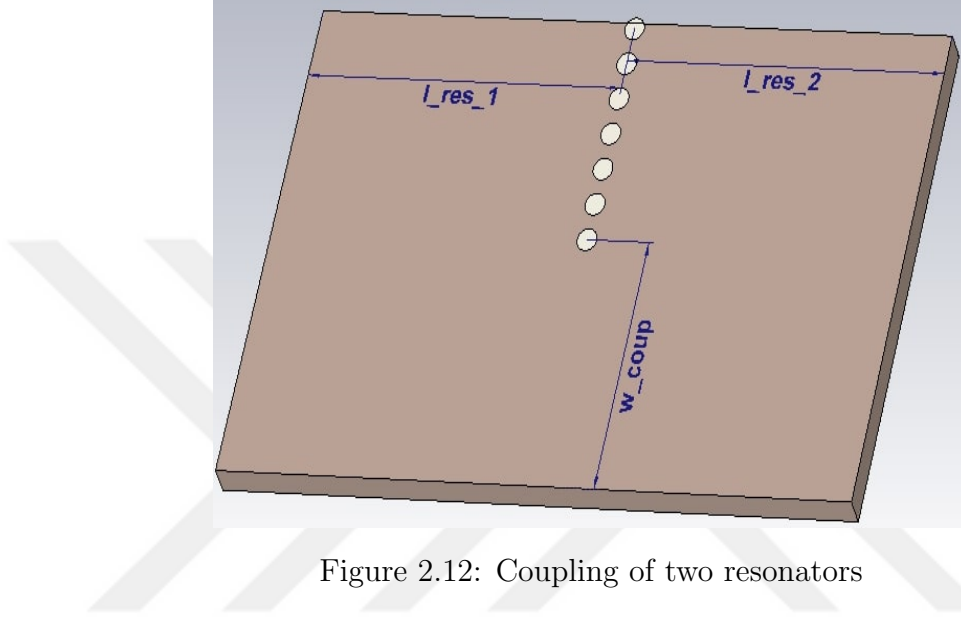


Figure 2.12: Coupling of two resonators

Center frequency of the coupling, which can be calculated as geometric mean of f_1 and f_2 , also needs to be considered. It is important to note that since resonators are coupled, effective volume of the structure is increased and resonance frequency is decreased. In order to keep the same frequency, resonators are kept shorter and coupling gap is swept.

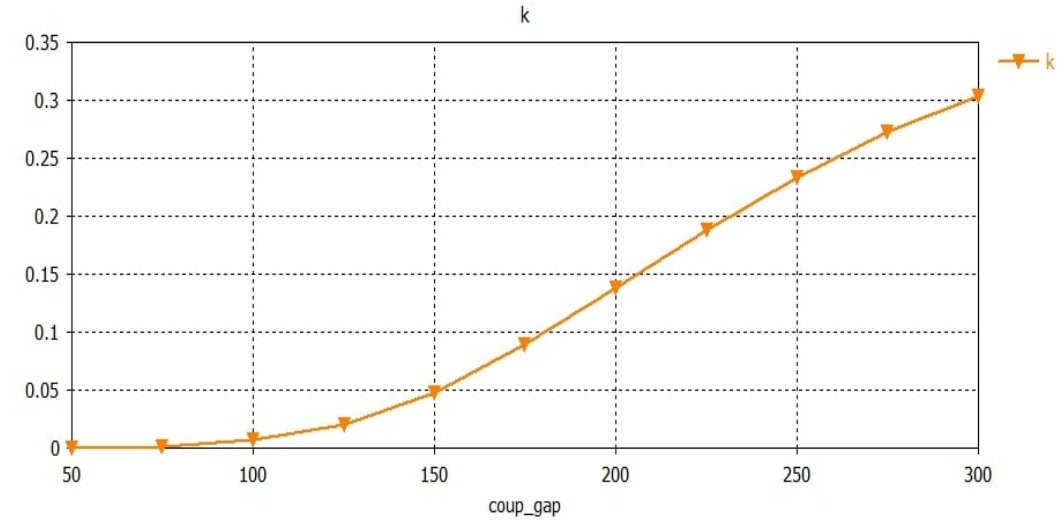


Figure 2.13: Coupling coefficient vs coupling gap

To get a 0.147 coupling coefficient width of the gap should be approximately 200 mils.

2.3.5 Ku-Band Band Pass Filter

Now that physical values are known, the filter can be designed by putting together previous simulations. Figure 2.14 shows the final topology of filter. Initial values from synthesis and final values after optimization are given in Table 2.1.

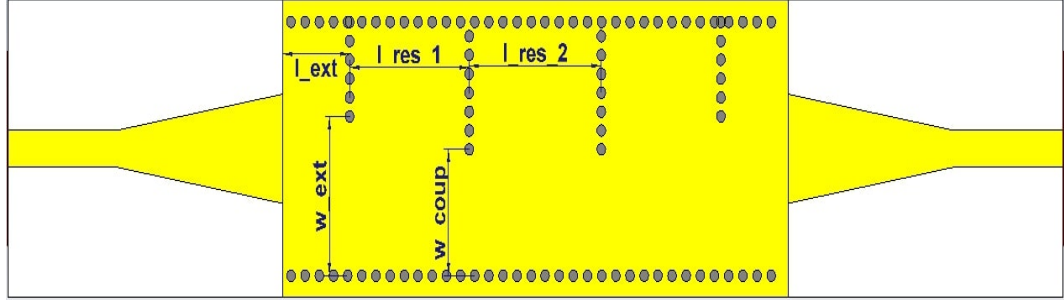


Figure 2.14: Final filter topology

Table 2.1: Filter Parameters

Parameter	Before Opt. (mils)	After Opt. (mils)
l_{ext}	100	124
w_{ext}	225	220
w_{coup}	200	174
l_{res1}	210	220
l_{res2}	240	243

Designed filter response in Figure 2.15 meet the criteria given in Chapter 1. The return loss in passband is better than -15 dB while attenuation at 14 GHz is higher than 30 dB.

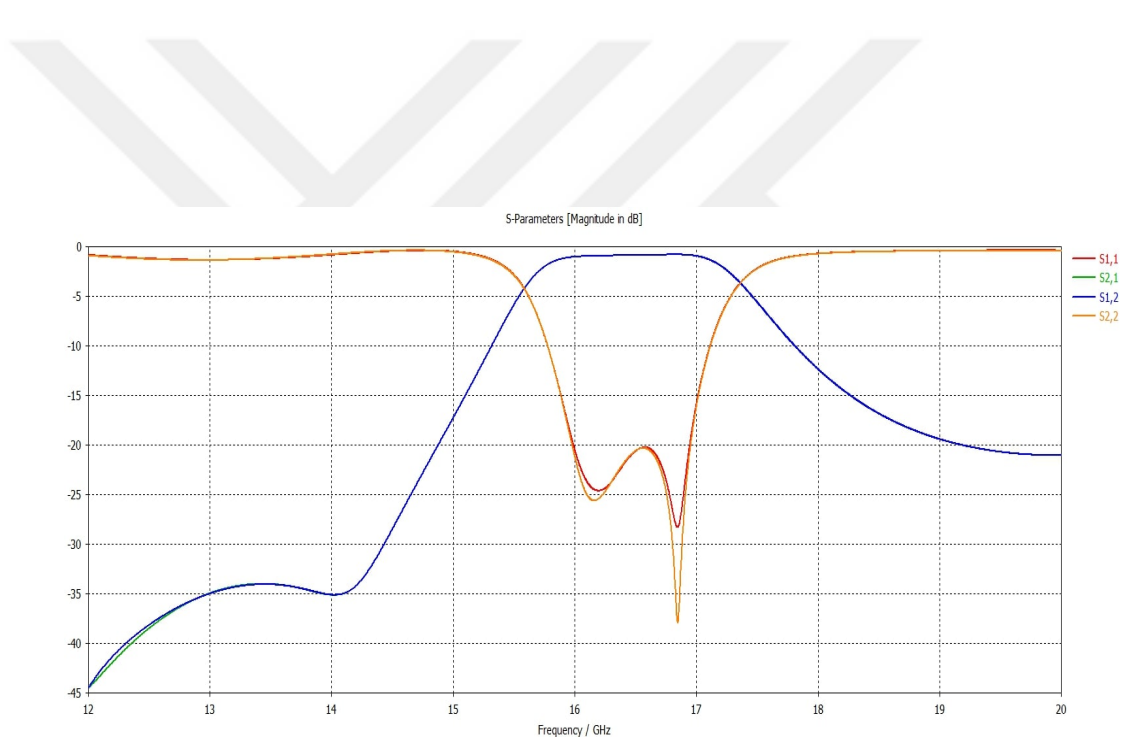


Figure 2.15: Simulated Ku-band filter response

Chapter 3

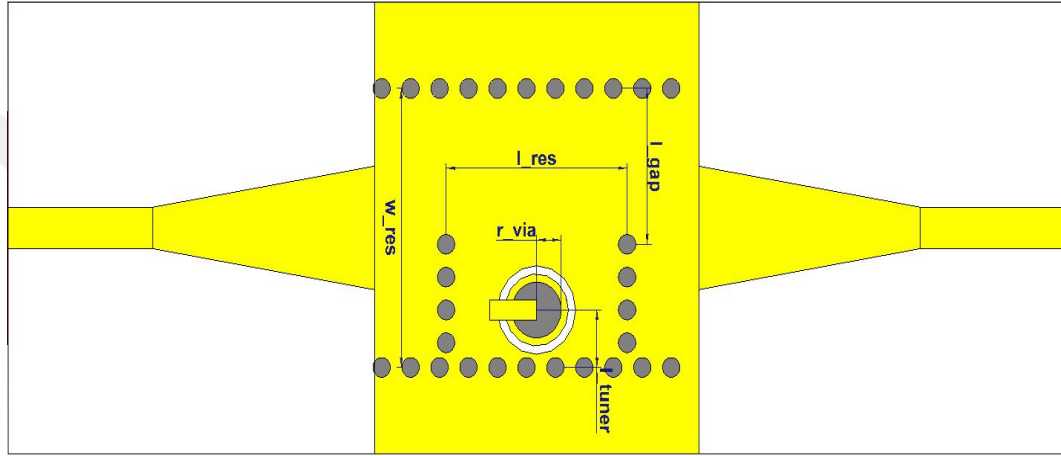
SIW Tuning

Most common method for tuning regular rectangular waveguide filters is to insert screws to resonators and coupling apertures. The screws distort the electric field in the cavity or the aperture, shifting the resonance frequency and changing the coupling. Although this method is very effective in waveguide filters, it is not applicable in SIW due to its planar structure. In [13], a mechanical way of tuning SIW resonators and filters is proposed. A via is inserted to the structure with an offset from the center of the resonator, similar to a tuning screw which compresses the electrical field and increasing the resonance frequency.

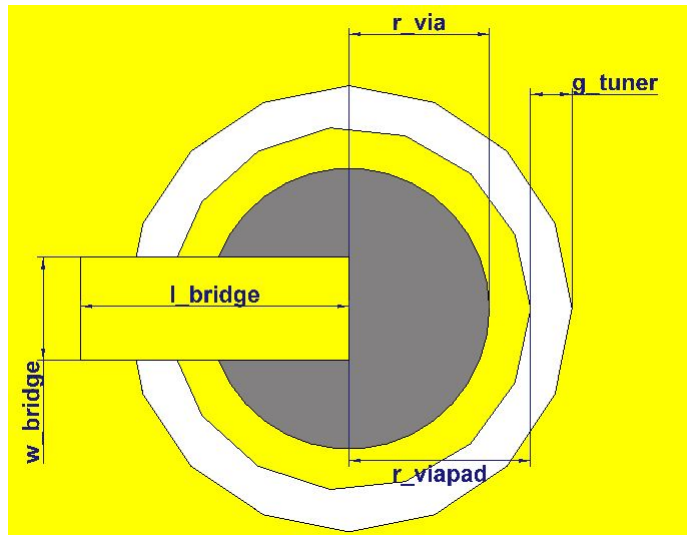
3.1 SIW Resonator Tuning

In [13], vias are inserted to resonators with an offset (l_{tuner}) from the center. Top pad of via is isolated with a gap (g_{tuner}) from the SIW top layer. A metallic contact is inserted between the via and rest of the SIW. The setup can be seen in Figure 3.1.

Center frequency of resonator can be adjusted to the desired value by rotating the metallic contact around the via. When the contact is facing towards the center, frequency is at the highest value while when contact is facing away from



(a) Tunable SIW resonator



(b) Tuning structure closeup

Figure 3.1: Tuning of a single SIW resonator

the center, frequency is at the lowest value. The change in resonance frequency with respect to contact angle can be seen in Figure 3.2.

Rotating the contact provides a tuning range more than 1 GHz. This tuning range greatly depends on the parameters like offset from the center and radius of the via hole. Other parameters are optimized to be as small as possible while ensuring the needed tuning range. Optimized parameter values are given in Table 3.1.

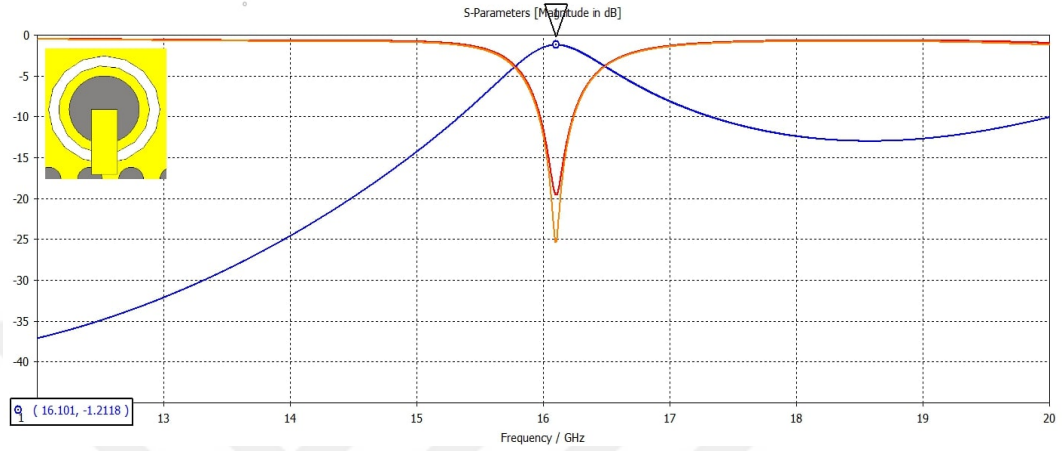
Table 3.1: Tuning Parameters

Parameter	Value (mils)
w_{res}	340
l_{res}	250
r_{via}	34
l_{tuner}	70
l_{gap}	190
w_{bridge}	25
l_{bridge}	65
r_{viapad}	44
g_{tuner}	10

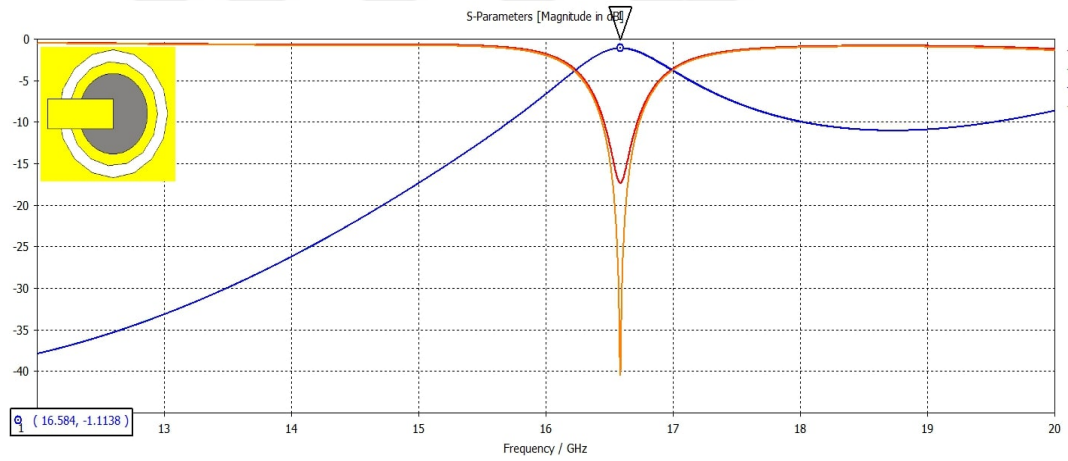
3.2 SIW Coupling Tuning

The same tuning structure can be inserted to coupling apertures of the filter as seen in Figure 3.3. When tuning bridge is facing to the outer wall of SIW, the coupling is at maximum. When it is facing inward, the coupling is at minimum. In Figure 3.4, frequencies of the two modes that are coupled to each other can be seen and coupling coefficient can be calculated using Equation 2.8.

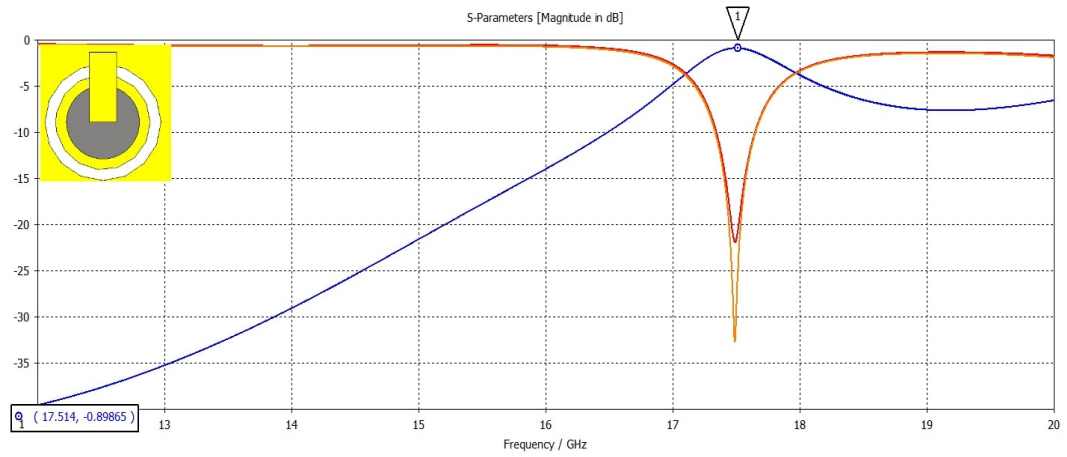
For simplicity of design, tuner parameters (Figure 3.1b) are chosen as the same values in previous section. All parameter values are given in Table 3.2.



(a) Tuned to lowest frequency



(b) Tuned to middle frequency



(c) Tuned to highest frequency

Figure 3.2: Tuning of a single SIW resonator

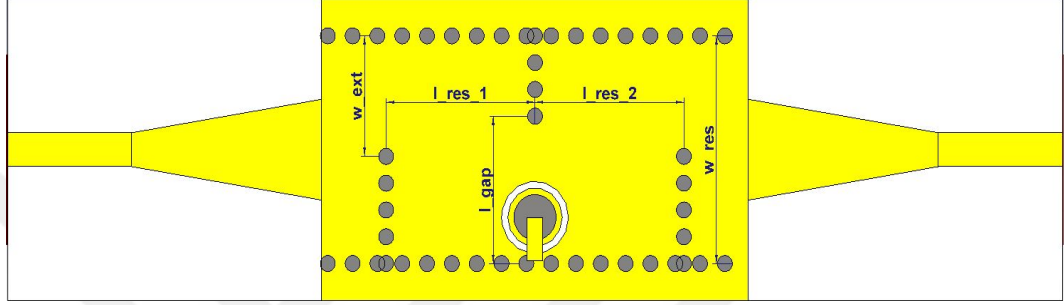
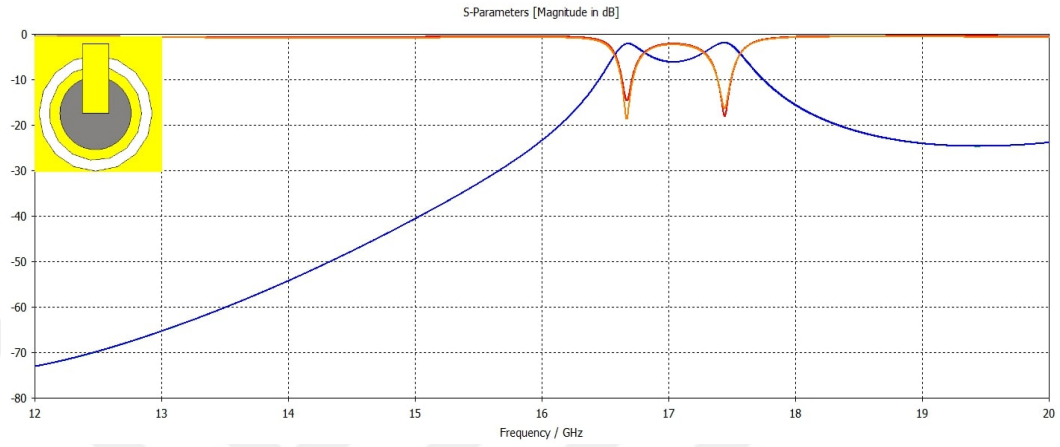


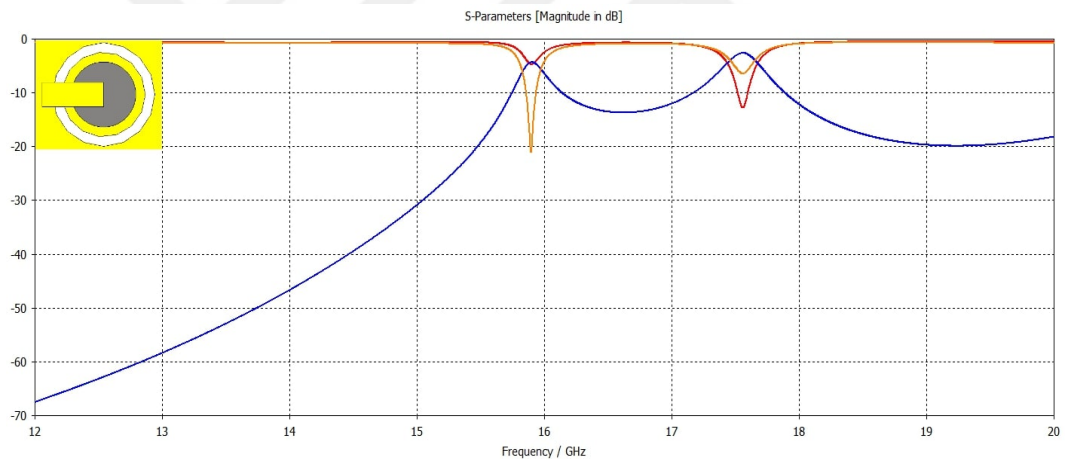
Figure 3.3: Coupling between two resonators

Table 3.2: Tuning Parameters

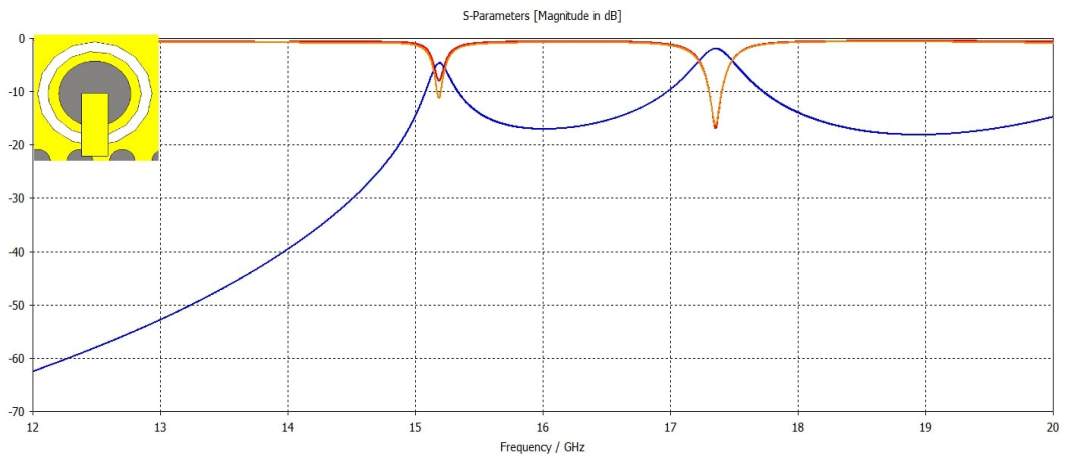
Parameter	Value(mils)
w_{ext}	180
l_{res_1}	240
l_{res_2}	240
l_{tuner}	70
l_{gap}	220
w_{res}	340
w_{bridge}	25
l_{bridge}	65
r_{viapad}	44
g_{tuner}	10



(a) Tuned to low coupling



(b) Tuned to medium coupling



(c) Tuned to high coupling

Figure 3.4: Tuning of the coupling between two resonators

3.3 SIW Filter Tuning

By combining tuning mechanisms of previous sections with the filter designed in Chapter 2.3.5, the desired tunable bandpass filter can be designed. Since the manufacturing tolerances are random, the error in filter response will be unpredictable. For this reason, the filter is designed while keeping tuning mechanisms at middle position. This allows the user to tune the filter both ways depending on the error.

Figure 3.5 shows the reference filter with tuners in middle position. Tuning mechanism dimensions are same with previous section. Other parameters are optimized to match the filter response in Figure 2.15.

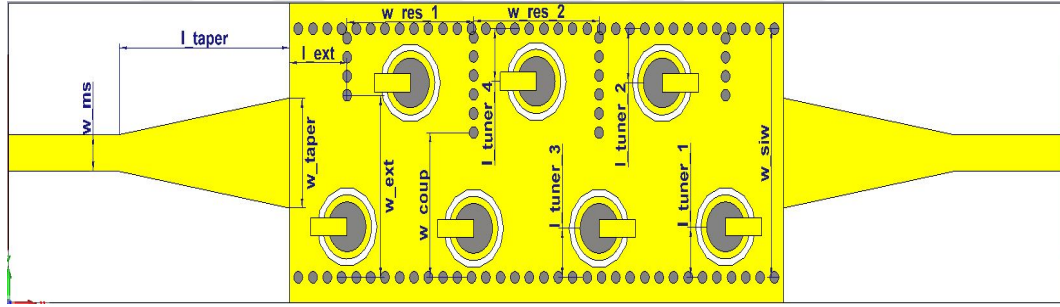


Figure 3.5: Designed reference filter with tuners in middle position

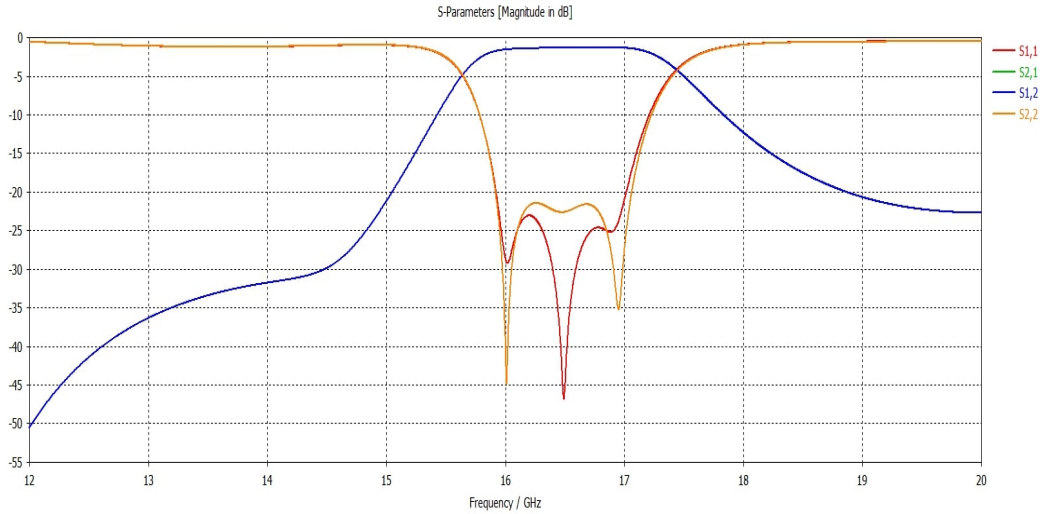


Figure 3.6: Reference filter s-parameters

Table 3.3: Reference filter parameters

Parameter	Value (mils)
w_{ms}	50.5
l_{taper}	306
w_{taper}	100
l_{ext}	104
w_{res_1}	228
w_{res_2}	226
w_{ext}	249
w_{coup}	198
w_{siw}	340
l_{tuner_1}	69
l_{tuner_2}	74
l_{tuner_3}	72
l_{tuner_4}	67

The final parameters are very close with the filter without the tuning components. So the design process explained in Chapter 2 does not need to be repeated when tuning elements are added.

It is important to note that filter is not completely symmetrical due to the tuner in the middle. This can be seen in Figure 3.6 where s_{11} and s_{22} are both below -20 dB but are unequal. However this asymmetry is ignored since it would not cause a problem in application.

Chapter 4

Measurement and Tuning

Tuning a band pass filter may be a cumbersome and time consuming process. It usually consists of changing every parameter randomly until required response is achieved. Some kind of intuition can be gained with experience but it is still an iterative progress. Since resonators are coupled to each other, tuning a single resonator effects the whole filter response. While looking at frequency response, it is impossible to understand what kind of effect each parameter has. Since manufacturing tolerances are unpredictable, all seven tuners must be considered independently.

The ideal solution for this would be a method to see how the response changes for each individual tuner. In planar filters, usually only resonator frequency can be tuned. The tuning method that will be explained in next section is commonly used in waveguide filters. However, since it is shown that both center frequency and coupling can be tuned, this method can be used with the designed SIW filter.

4.1 Time Domain Tuning

In [20], a time domain tuning method is proposed to identify which resonator or coupling should be tuned. This method uses time domain response of the

reflection coefficient of the filter. This provides an algorithm to tune the filter so the designer can know which parameter to change.

4.1.1 Time Domain Response

CST Microwave Studio has a post processing option to calculate time domain response of the reflection coefficient of a filter. First step is to use time domain transform for simulated filter. This will be the "golden response" that manufactured filters will be tuned to. This result will be imported to the network analyzer and provide a template to make the tuning process easier.

Before applying the transformation, frequency span, number of points and start-stop time must be decided. These parameters also must be consistent with the measurement parameters in network analyzer. The application note [21] also gives some rules-of-thumb regarding the choice of these parameters as;

1. Simulation frequency must be equal to the center frequency of the filter.
2. Frequency span should be 2 to 5 times the bandwidth of the filter.
3. Number of points can be chosen as 201 for a good balance between speed and resolution.
4. Start time should be set to $t = -(2/\pi BW)$
5. Stop time should be set to $t = (2N + 1)/(\pi BW)$ where N is number of resonators and BW is defined for -3 dB points in Hz.

For the designed filter, 3dB bandwidth is about 1.7 GHz. Start and stop time are calculated and entered as input to post-processing screen as seen in Figure 4.1.

Figure 4.2 shows time domain response of reflection coefficient when all resonators and coupling are properly tuned.

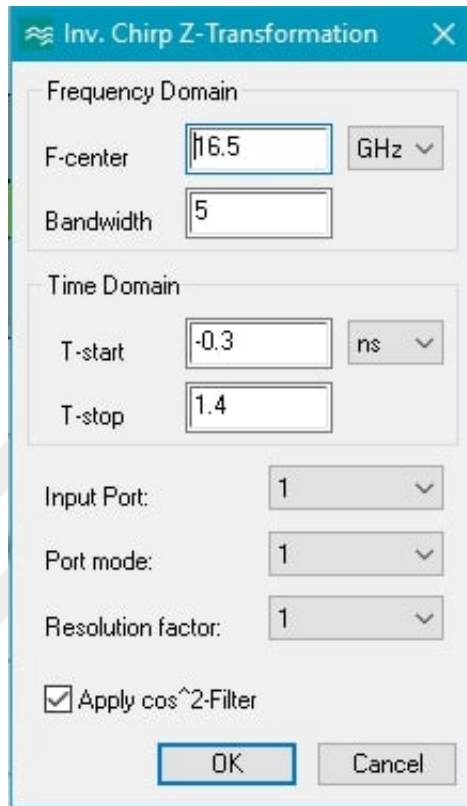


Figure 4.1: Post processing screen in CST MWS

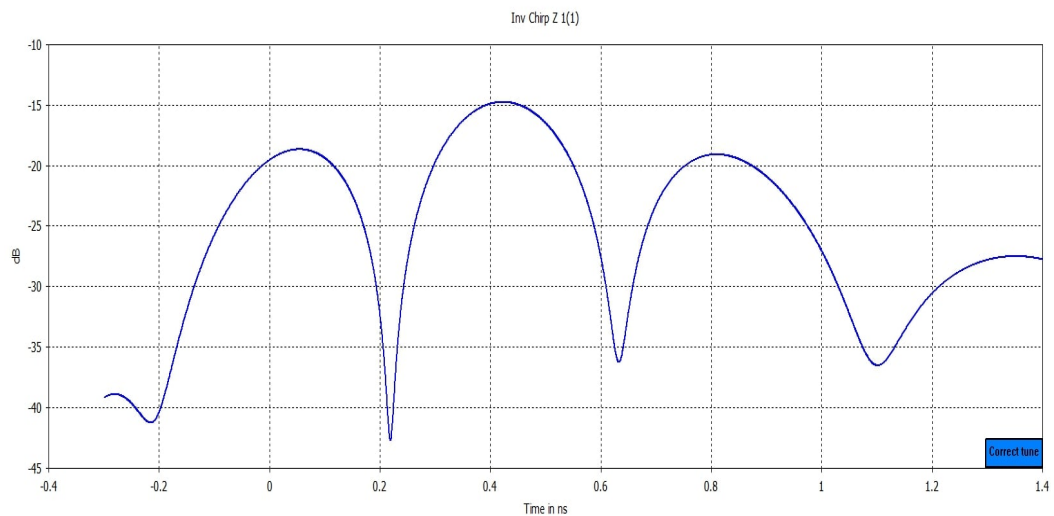


Figure 4.2: Everything correctly tuned

Note that there are three dips for three resonators. These represent the times when almost no power is reflected from the filter. It can be deduced from the dips whether any of the resonators are mistuned. Since first dip represents first resonator and second dip represents second resonator and so on, the designer can determine which of the resonators is mistuned.

4.1.2 Tuning of Resonators

In this section, the three resonators are purposefully mistuned one at a time by rotating the tuner 10 degrees and time domain response is observed. Figure 4.3 and 4.4 show time domain response when first resonator is tuned to a higher and lower frequency respectively.

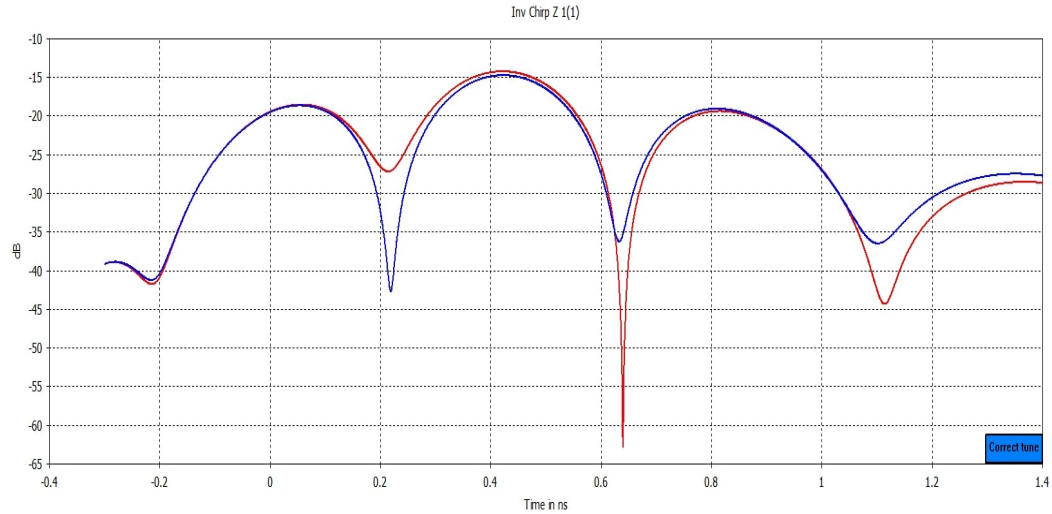


Figure 4.3: First resonator mistuned to a higher frequency

When first resonator is mistuned, regardless of the direction, the first dip disappears. This shows some of the power reflects back from the resonator. Another important point is the changes in other dips. Although only first resonator is mistuned, all of them are affected by it and appear mistuned as well. In other words, a problem at first resonator can mask other problems in following resonators. So it is a good idea to start tuning with the first resonator.

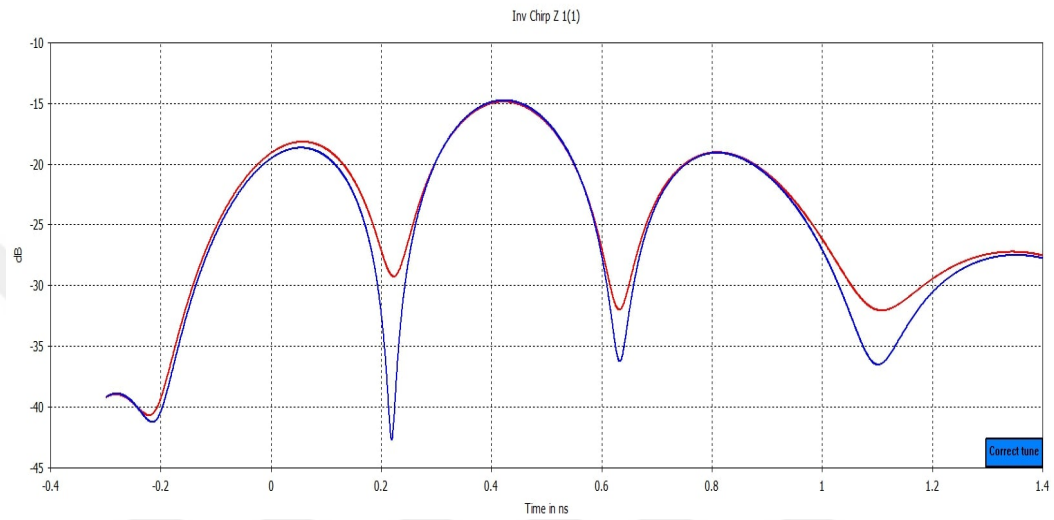


Figure 4.4: First resonator mistuned to a lower frequency

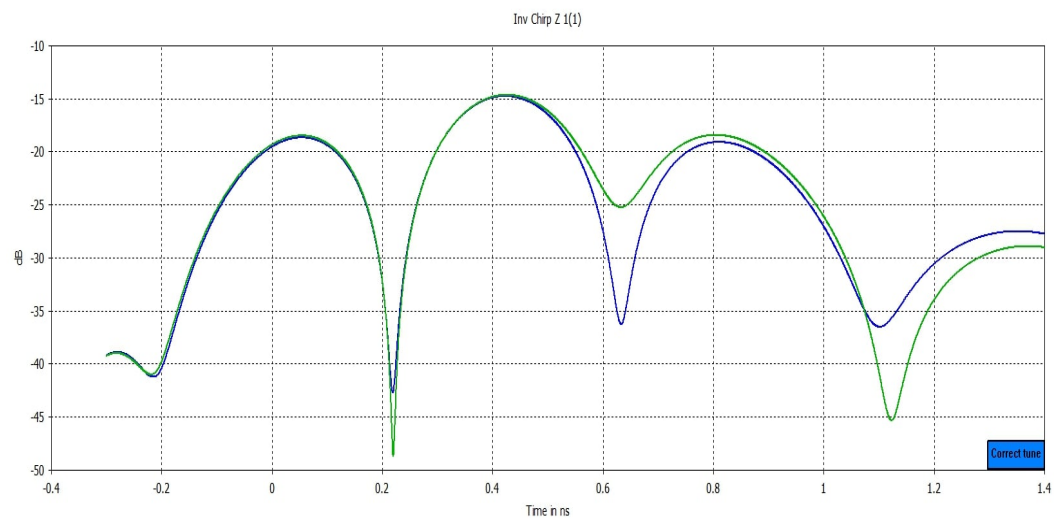


Figure 4.5: Second resonator mistuned to a higher frequency

Figure 4.5 shows reflection coefficient when only second resonator is mistuned. This time the dip for the second resonator is increased. First resonator is unaffected while third resonator also appears to be mistuned although its parameters are not changed. Since third resonator is behind the second resonator, it gives false information.

To overcome this problem, especially in higher order filters which would have much more dips, time domain transformation of reflection coefficient of second port can be used. Doing this will allow the tuner to tune outer resonators first and work towards the center. However, this is an iterative process since tuning the resonator at second port will affect the resonator in first port.

Finally, in Figure 4.6 third resonator is mistuned. As expected, there are no problems with first and second resonator. Only on the third resonator, there appears to be some reflected power.

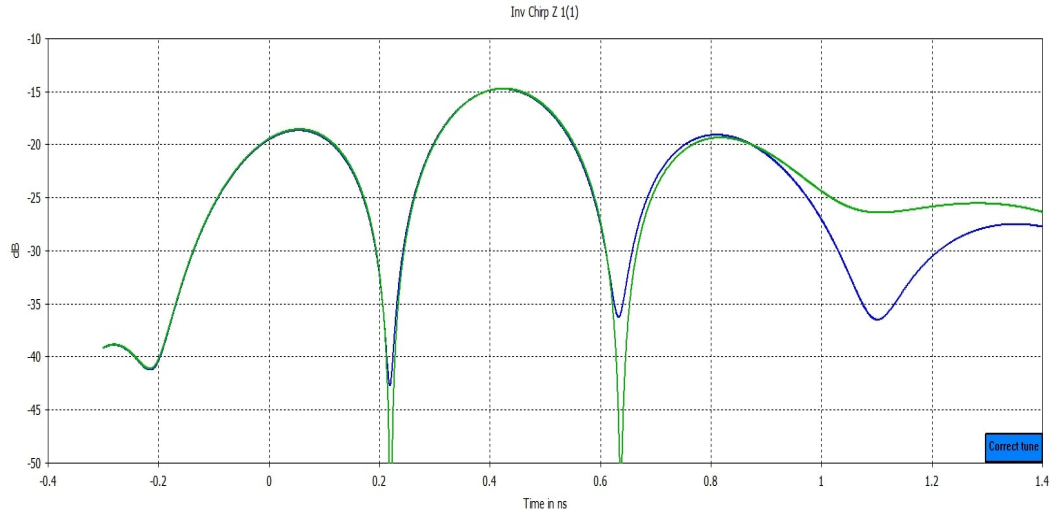


Figure 4.6: Third resonator mistuned to a higher frequency

When a resonator is not tuned correctly, the corresponding dip disappears. It is a good idea to tune the resonators before couplings. However, while tuning the couplings, effective volume of the resonators also change, hence couple of more iterations of tuning the resonators might be required.

4.1.3 Tuning of Couplings

There are four peaks for four coupling apertures. All couplings are mistuned on purpose one at a time, and changes in response is observed. Figure 4.7 and 4.8 shows when source is coupled to first resonator more and less than intended.

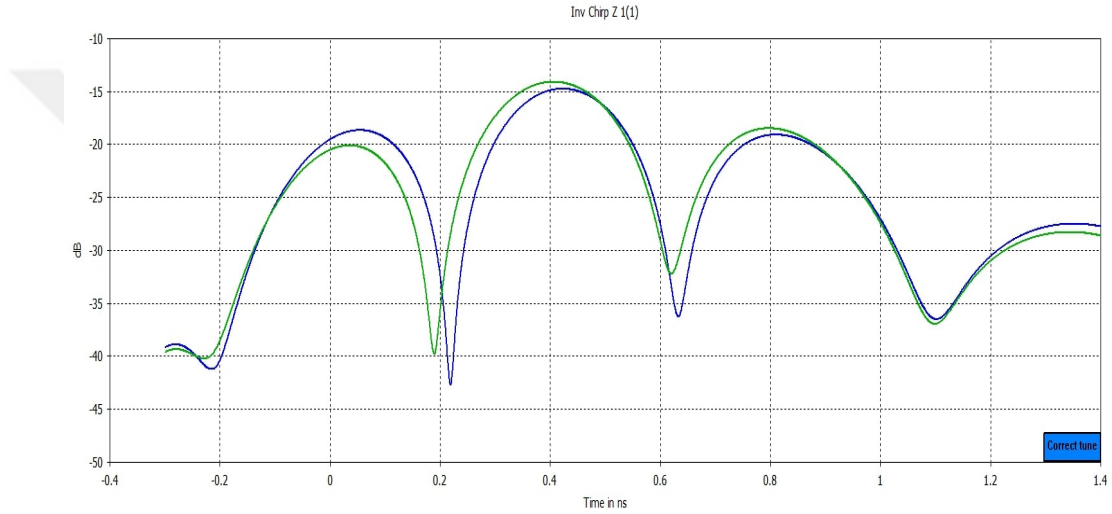


Figure 4.7: Source and first resonator overcoupled

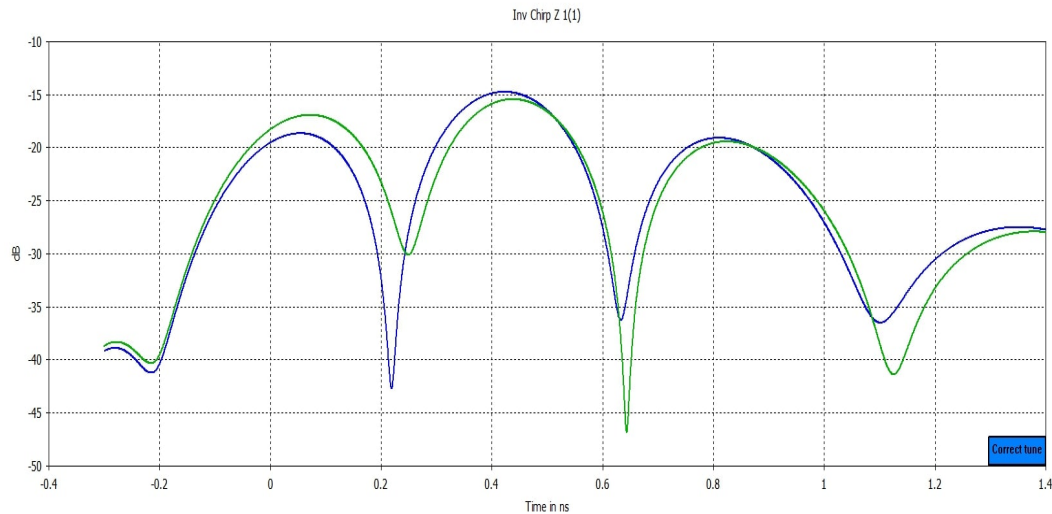


Figure 4.8: Source and first resonator undercoupled

Since the coupling is increased, more power goes through the filter and less is reflected. In Figure 4.7, it is seen that the first peak decreases as expected. The

same train of thought can also be applied for Figure 4.8 where the peak is increased. Note that other points in both cases are also affected, just like the case when only first resonator was mistuned.

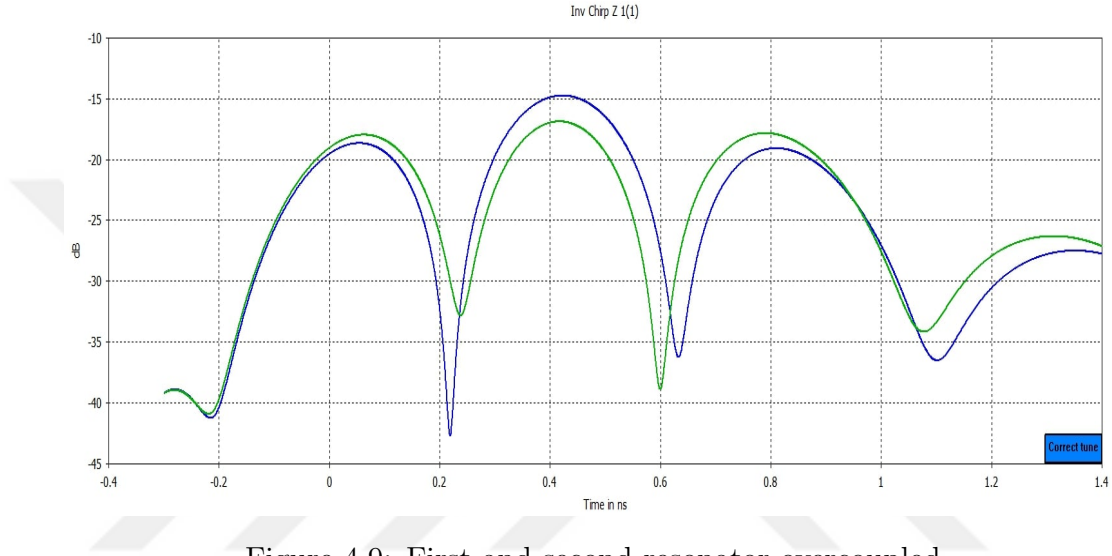


Figure 4.9: First and second resonator overcoupled

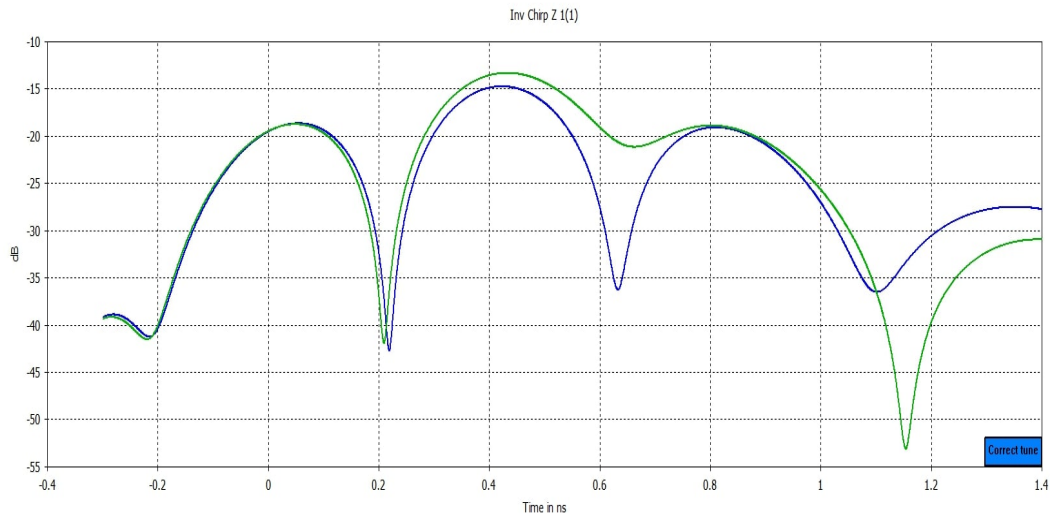


Figure 4.10: First and second resonator undercoupled

Tuning of coupling between first and resonator is shown in Figures 4.9 and 4.10. Similar to previous case, second peak in the figure rises or declines depending whether coupling is decreased or increased.

The same pattern continues for coupling between second and third resonator

in Figures 4.11 and 4.12.

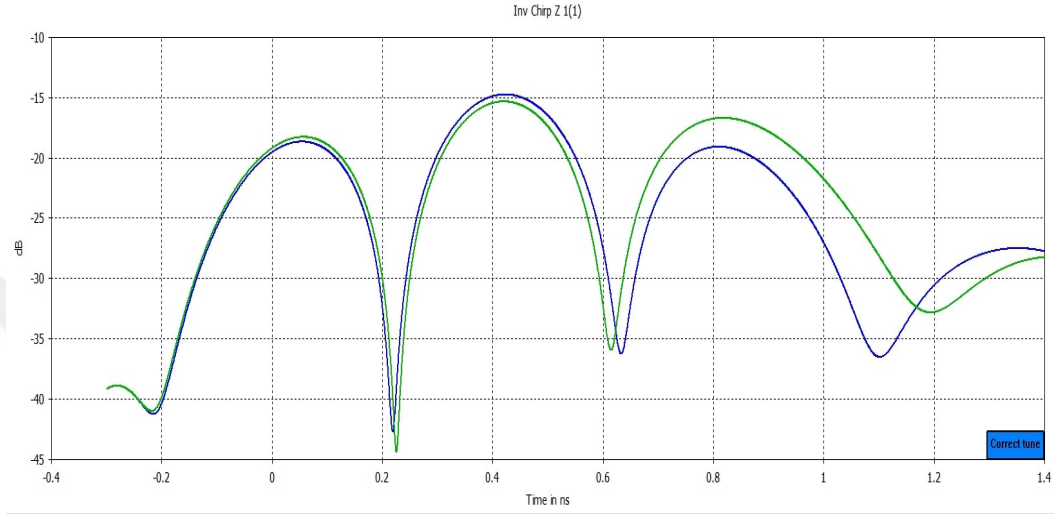


Figure 4.11: Second and third resonator undercoupled

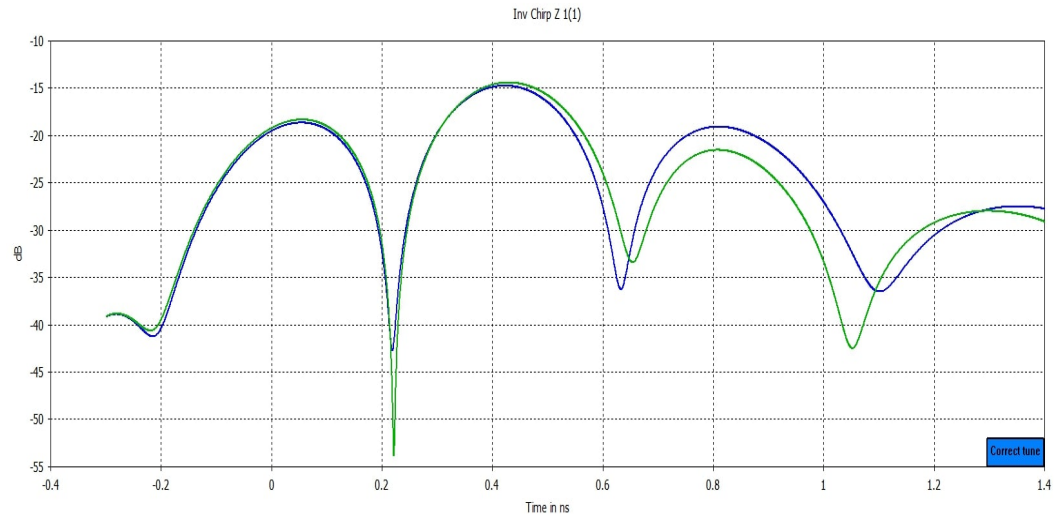


Figure 4.12: Second and third resonator overcoupled

Finally, Figures 4.14 and 4.13 shows coupling between third resonator and the load. As expected, fourth peak is affected corresponding to the fourth aperture in the filter.

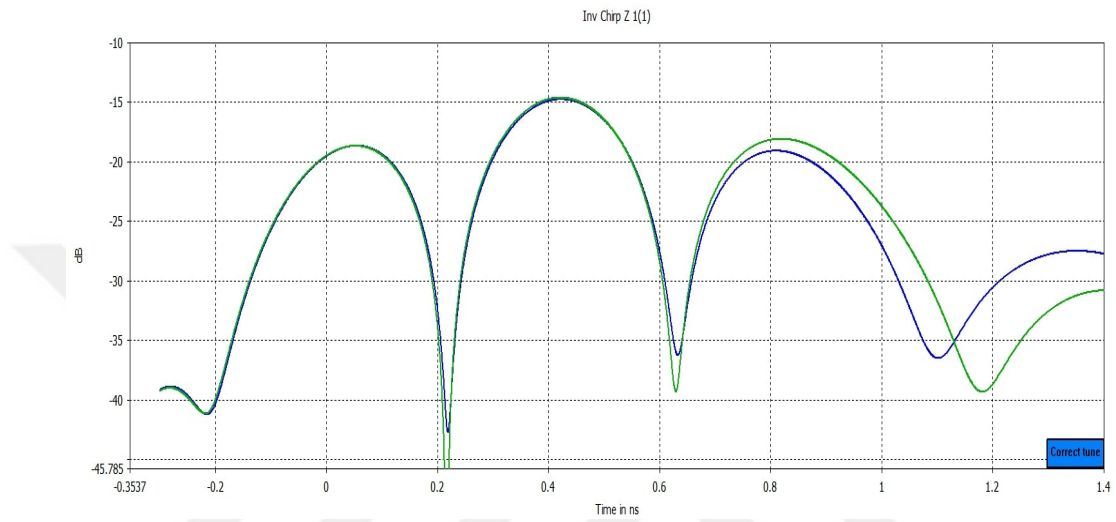


Figure 4.13: Load and third resonator overcoupled

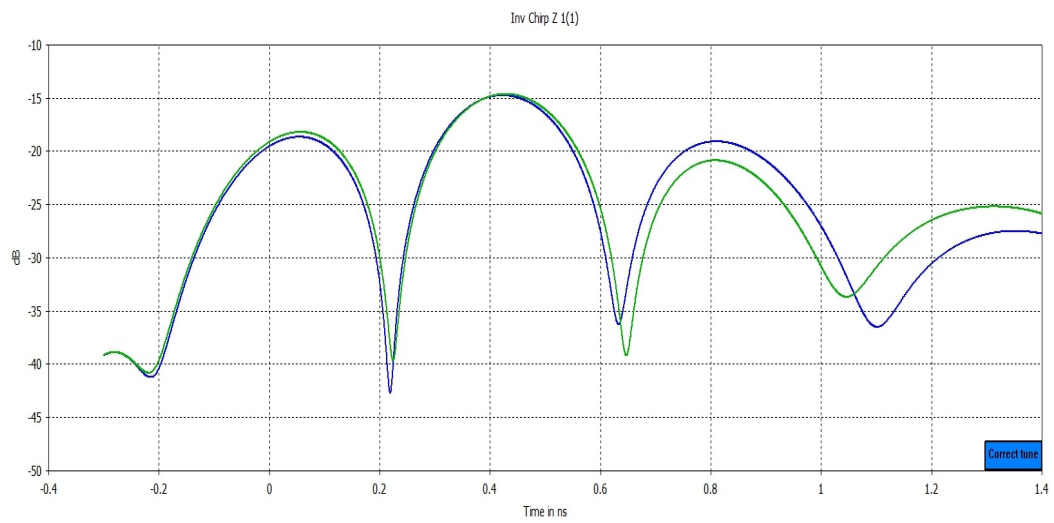


Figure 4.14: Load and third resonator undercoupled

4.2 Implementation

In [13], rotating contacts are built by a piece of plastic with a metallic strip. The plastic is screwed through the via. In a real world application, the screw needs to go inside a hole in the module. Also, a screw head creates extra height which might be undesirable in certain situations. To eliminate these problems, metallic contacts are built by using 0201 package surface-mount zero ohm resistors. Another advantage would be the frequency range in which this method can be applicable. In higher microwave frequencies, resonator size becomes relatively small compared to the screw dimensions. Using 0201 package resistors make, using a smaller tuning via possible. However, this method has the disadvantage of having parasitic effects due to lumped element package but these are very small and can be tuned out. Constructed filter and soldered tuning resistors can be seen in Figures 4.15 and 4.16.

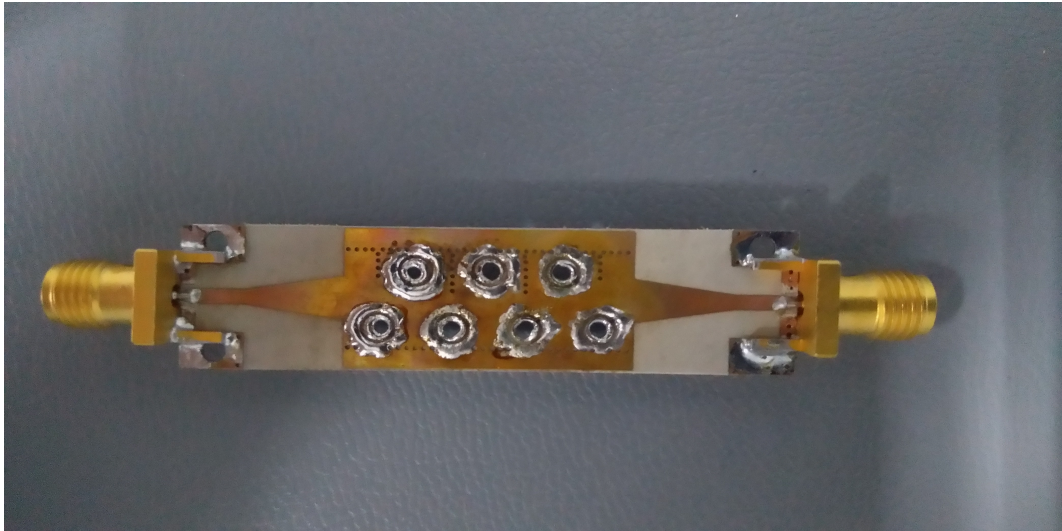


Figure 4.15: Manufactured filter with resistors soldered

4.3 Measurement

To measure the filter, Keysight PNA-X network analyzer is used. In order to use the time domain tuning method, analyzer must have the bandpass transform

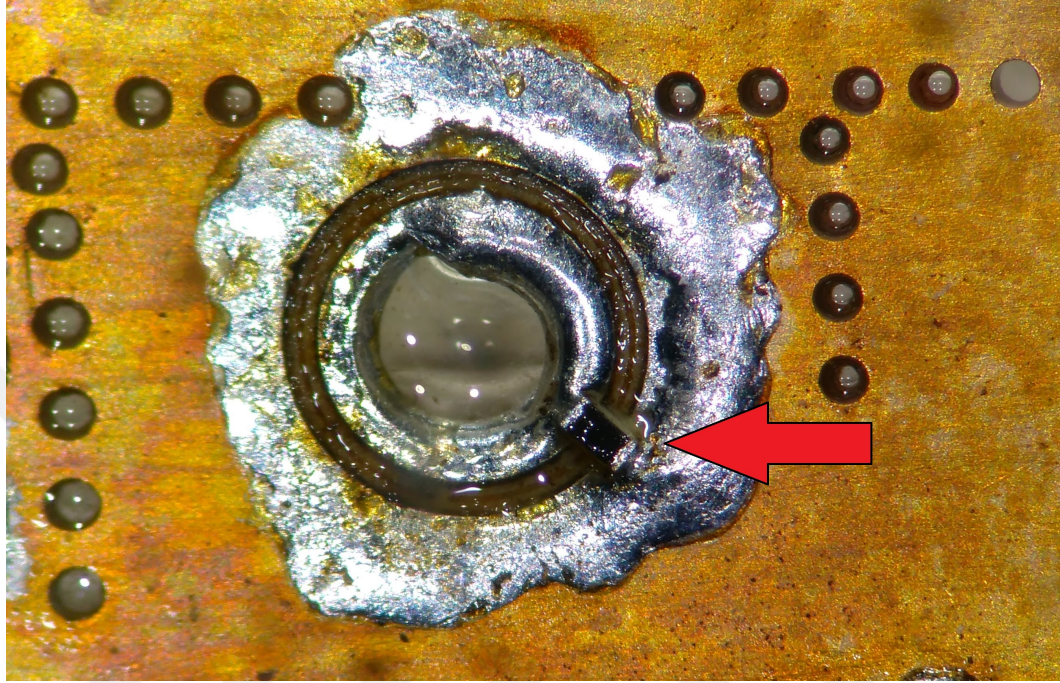


Figure 4.16: One of the tuning structures with the resistor soldered

option installed. After adjusting the settings, the filter is connected to analyzer ports. This measurement setup can be seen in Figure 4.17. Initially, tuning resistors are positioned in middle positions as in Figure 3.5.

As expected initial results are not ideal and the filter needs tuning. To start the tuning process, first s-parameter results of the simulation in Figure 3.6 are exported from CST and imported to the network analyzer. Start-stop time and number of points should be exactly same with the simulation. This will serve as the template. Figure 4.18 shows initial time domain response of the measured filter with the template on the same graph. Although time domain response is usually enough for filter tuning, it is always a good idea to monitor the frequency response at the same time.

As explained in Seciton 4.1, tuning is done by starting with the resonators while assuming couplings are correct. After finding the dip positions for resonator tuners, couplings are adjusted. This might take a couple of iterations. Time domain response of the tuned filter is given with the template in Figure 4.21.

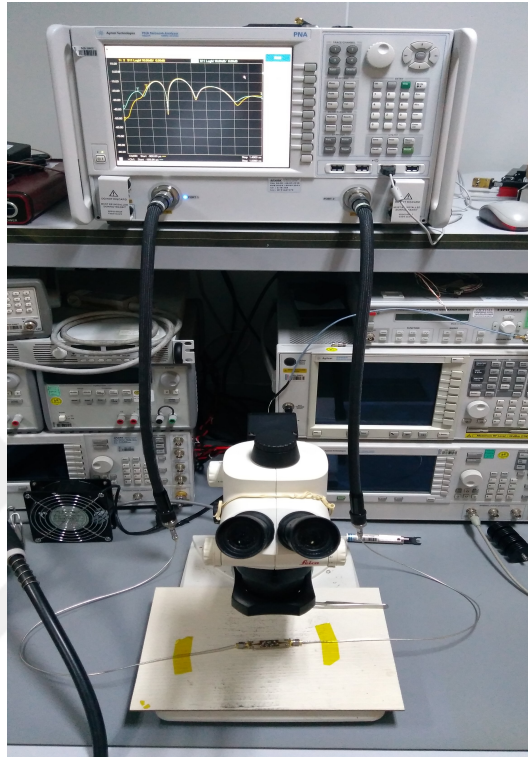


Figure 4.17: Measurement setup for tuning filter

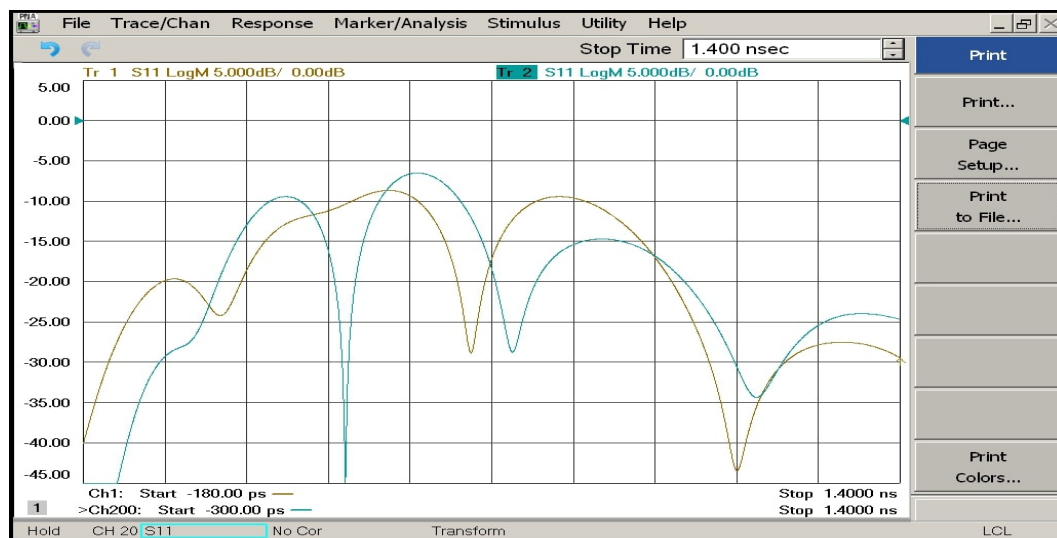


Figure 4.18: Initial time domain response for untuned filter

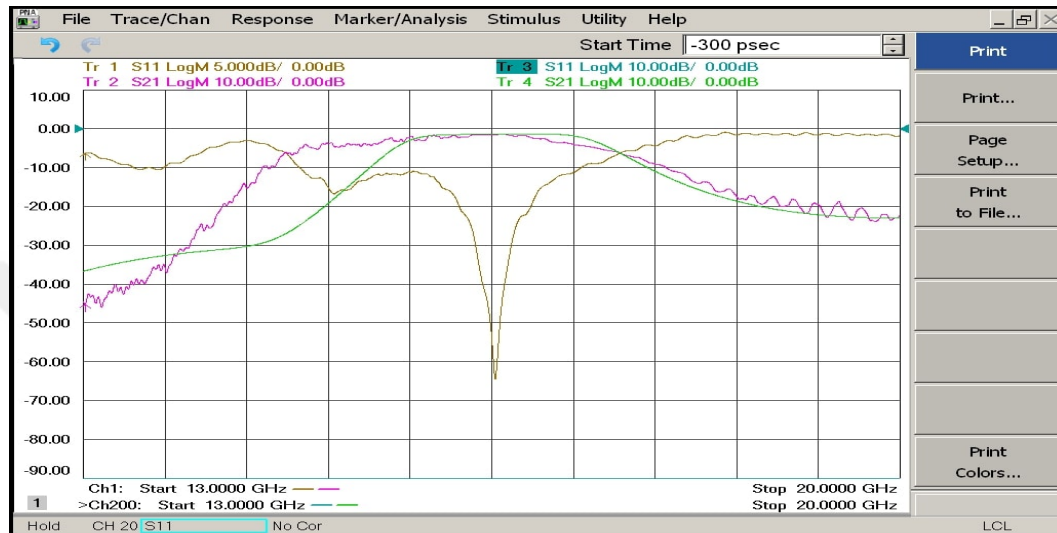


Figure 4.19: Initial frequency domain response for untuned filter

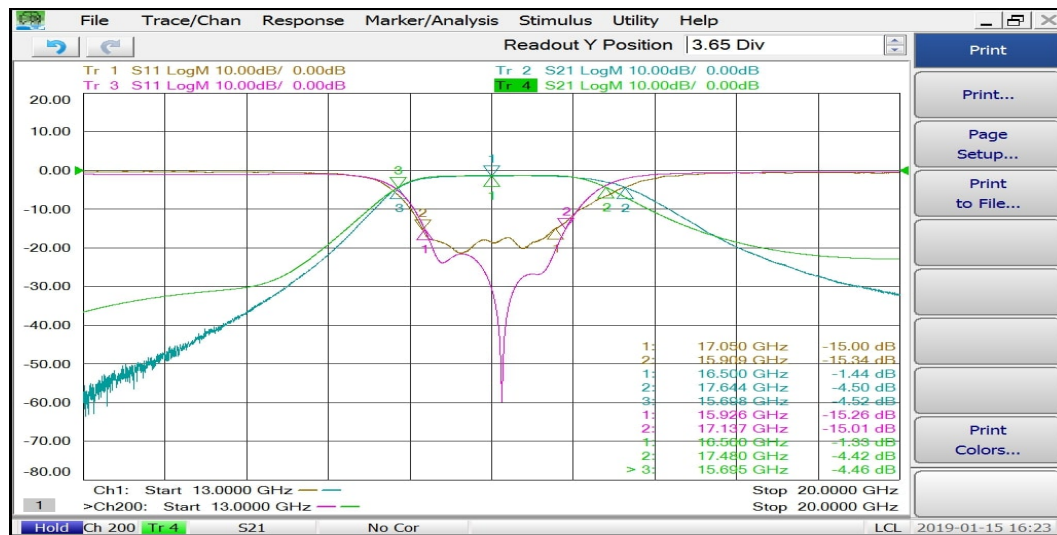


Figure 4.20: Tuned filter response

In addition, final response of tuned filter is given in Figure 4.20. Note that the bandwidth of the measured filter is slightly wider than desired. An explanation for this can be seen in Figure 4.21 where last coupling between third resonator and load is mistuned. It is possible to tune to the exact reference filter by spending more time on tuning. For this case, error is only about 100 MHz at the upper side band. Since error is very insignificant and priority is to attenuate lower side band, this error is ignored.

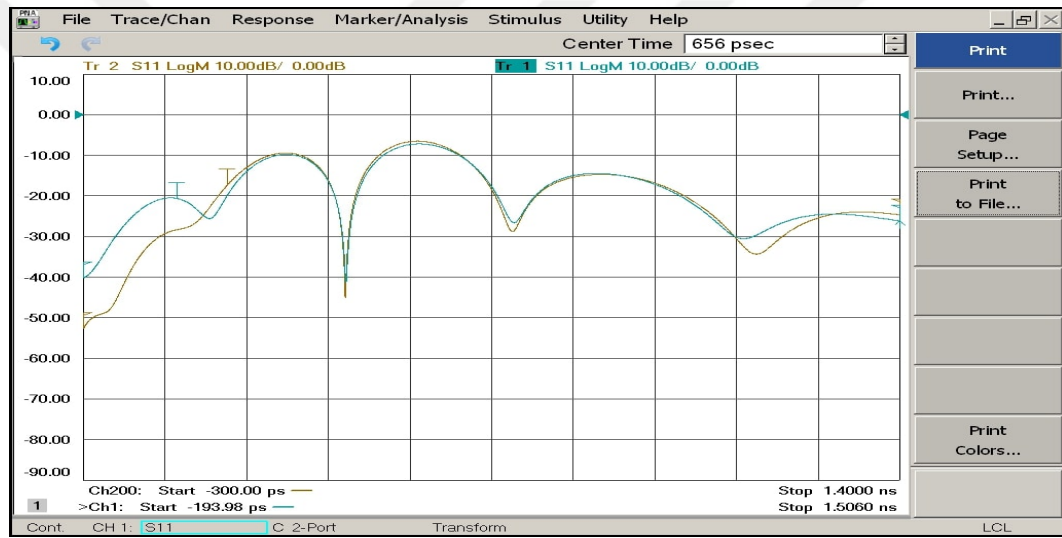


Figure 4.21: Time domain response for tuned filter

It is crucial to point out that in Figure 4.21, center time for template filter and measured filter is different. This is done deliberately due to the connectors used during measurement. These connectors are not modeled in CST simulation but introduce a delay while measuring the filter.

Chapter 5

Conclusion

In this thesis, a mechanical tuning mechanism for substrate integrated waveguide structures is analyzed and implemented for the purpose of nullifying the effects of manufacturing tolerances. SIW design equations are used for determining SIW parameters and transition to microstrip structure. A through line SIW is fabricated and measured. Coupling matrix approach is used to design a SIW bandpass filter which operates in Ku-band. Resonator and coupling dimensions are determined with eigenmode solver. The filter has better than -15 dB return loss and minimal insertion loss between 16 and 17 GHz while maintaining attenuation greater than 30 dB at 14 GHz. Application of the proposed tuning mechanism is analyzed using CST design environment. Tuning of a single resonator and coupling between two resonators are simulated and necessary parameters for tunable filter design are extracted. After designing the tunable filter, it is manufactured using common PCB production techniques. Time-domain tuning method, which is commonly used in combline waveguide filters, is introduced which enables the designer to individually tune all resonators and couplings. Using time domain transformation of reflection coefficient, the filter is tuned to the design criteria. Measurements show that tuning to the exact design parameters is possible depending on the sensitivity of tuners.

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