

## Design a Hairpin Bandpass Filter for 5 GHz Unlicensed WiMax Band

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**Abstract**— This paper presents a hairpin bandpass filter design for 5 GHz unlicensed WiMax band. The filter is designed at 5.788 GHz centre frequency, with a bandwidth of 125 MHz bandwidth that is about 0.02% of its centre frequency, 0.1 dB ripples and consists of 3 elements. It is then simulated using Computer Simulation Technology (CST), then be fabricated on Roger 4350B that have a dielectric constant,  $\epsilon_r$  of 3.48 and substrate thickness,  $h$  of 1.524 mm. The filter is then being measured using Vector Network Analyzer (VNA) and the measurement results show a good agreement with the simulation results.

**Keywords**—Hairpin bandpass filter, Computer Simulation Technology (CST)

### I. INTRODUCTION

Filter is a device that allows the signal within the specific frequencies to pass through to the receiving device. It often being used in front of the detectors to eliminate or reduce noise due to light at wavelengths is difference from the desired wavelength. In this paper, the bandpass filter is design for 5GHz unlicensed WiMax for that reason. The bandwidth for this design is 125MHz with the operating frequency at 5.787GHz with the insertion loss of not exceeding 3 dB at the operating frequency.

Table 1. Hairpin bandpass filter design specifications

Filter Specifications	Values
Centre frequency, $f_c$	5.788 GHz
Upper cut-off frequency, $f_u$	5.850 GHz
Lower cut-off frequency, $f_l$	5.725 GHz
Bandwidth	125 MHz
Filter order, $n$	3 <sup>rd</sup> order

Figure 1 below showed an example of a bandpass filter response.

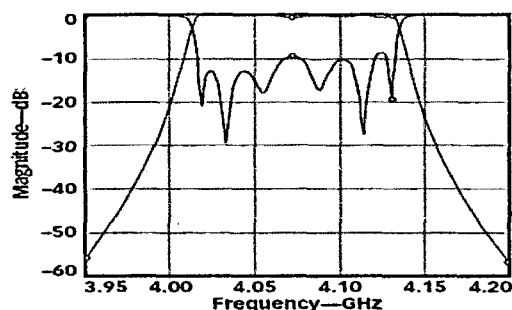


Figure 1. Bandpass filter response

### II. WIMAX OVERVIEW

WiMAX stands for Worldwide Interoperability for Microwave Access is a form of broadband wireless access which is based on the IEEE 802.16 standard for Wireless Metropolitan Area Network (WMAN). It has a theoretical range of 30 miles compared to WiFi with only 30 yards along with a stronger connection. In a typical 20MHz channel bandwidth deployment scenario, WiMAX Forum certified products will support downlink data rates of 65Mbps at close range to 16Mbps at distances of 9km to 10km, which is enough bandwidth and transmission range to deliver high-speed simultaneous access to voice, data, and video services to hundreds of businesses or thousands of residences[1].

The extended range of WiMAX's is driving a significant market opportunity and it's proving useful in delivering broadband services to rural areas where it's cost-prohibitive to install landline infrastructure. The difference between WiMax and WLAN specification are shown in Table 2.

Table 2. Comparison between WiMax and WLAN [1]

	WLAN (802.11)	WiMAX (802.16)
Bandwidth	Fixed; 20 MHz/52 subcarriers	Variable; 1 to 28 MHz/256 subcarriers
Guard interval	Fixed at 1/4*symbol time	Variable; ranges from 1/32 to 1/4*symbol time
Spectral efficiency	2.7 Mbits/s/Hz	3.1 to 3.8 Mbits/s/Hz
EVM requirements	-25 dB	-31 dB
Rx noise figure	10 dB maximum	7 dB maximum
Duplexing	TDD	TDD, FDD, HFDD
Spectrum	Unlicensed	Licensed & unlicensed
Tx dynamic range	Tx power fixed	50-dB range

In WiMax there are three duplexing techniques used which is Time Division Duplexing (TDD), Frequency Division Multiplexing (FDD) and Half Frequency Division Duplexing (HFDD). For FDD system, extensive external filtering is needed to prevent the transmit power from leaking into and interfering with the receiver. With extra filtering and dual radios, the board space is concern. With higher throughput, full FDD mode is expected to use at the base stations while subscriber station will use lower cost H-FDD or TDD.

### III. HAIRPIN FILTER THEORY

Microstrip bandpass filters are widely used at microwaves frequency with special attention at X-band. In this band, with an appropriate dielectric substrate, it is possible to perform an analysis based on quasi-transverse-electromagnetic (TEM) modes of propagation [2]. One of the popular structure is edge couple microstrip filter because there is no need to do the short circuits and easy to design and build. The main disadvantages is their large size and the solution for that is by folding the resonator the edge coupled filter and it become hairpin filter as shown in Figure 2.

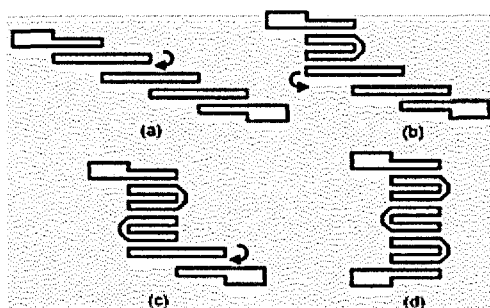


Figure 2. Four-step folding procedure to obtain a hairpin-line filter

The resonator of the hairpin filter is 180 degrees so that the length from the centre to either end of the resonator is 90 degrees [3]. Certain  $\theta$  degree is "slid" out of the coupled segment of the resonator name as sliding factor. The length of this sliding factor should be between 1 to 3 times of the resonator width to make sure the insertion loss is still in the range of specification. If sliding factor value is decreased, it will lower the value of the insertion loss since the signal power is more concentrated on the coupling structure instead of being attenuated along the transmission line path. If it is further decrease, it will introduce resonator self coupling that will lead to wider bandwidth.

The length of tapping distance or the stub length will be one of the factors that control the position of the center frequency,  $f_c$ . If the stub length is greater than it used to be, the center frequency will shift to a smaller value and if it is lower than the desired value, the center frequency will shift to a higher value.

The tapping position of the filter or the stub's location controls the return loss. As return loss increased, insertion will decrease. Other parameters such as bandwidth and center frequency are almost not affected. Figure 3 show the hairpin filter for  $n = 3$ .

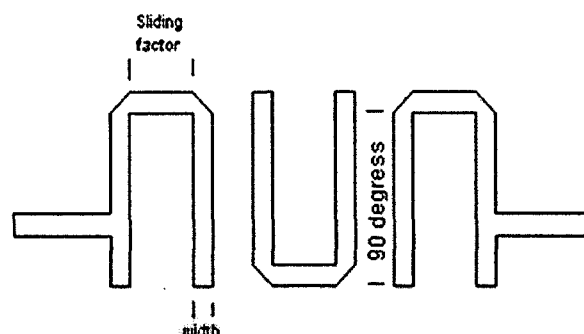


Figure 3. Hairpin Filter

Microstrip hairpin filter have several advantages that make it popular because it does not require critical grounding thus making manufacturing easier [4]. It also has a compact structure but the disadvantages of this compactness are a wider bandwidth and poor skirt rates due to unavoidable coupling [5].

### IV. METHODOLOGY

#### A. Introduction

There are several methods in designing filters. This design is using insertion loss method. The design start with a low pass filter that is selected by using normalized table to achieved design target. The target for this filter is the cut-off frequency at 5.79GHz, 0.02% bandwidth and 3-th order with 0.1 dB ripples.

The filter is then fabricated on Roger 4350B. The physical parameters of Roger 4350B is shown in the Table 3 below, obtained from [6].

Table 3. Physical parameter of Roger 4350B

Physical parameter	Values
Strip thickness, $t$	0.07 mm
Substrate thickness, $h$	1.524 mm
Dielectric constant	3.48

#### B. Design Procedure

From table 9.5 of [7], the prototype filter for  $n=3$  can be obtained. For this design, the prototype parameters are as follows:

$$\begin{aligned} \omega_c &= 1 \\ g_0 &= g_4 = 1 \\ g_1 &= g_3 = 1.0316 \\ g_2 &= 1.1474 \end{aligned}$$

Step 1:

By using the value above, filter is then converted using the formula below to obtain the impedance transformation.

$$C_{act} = \frac{g_p}{Z_0 \omega_c} \quad (1)$$

$$L_{act} = \frac{Z_0 g_p}{\omega_c} \quad (2)$$

Where  $Z_0 = 50\Omega$

Resulting circuit is shown in Figure 4.

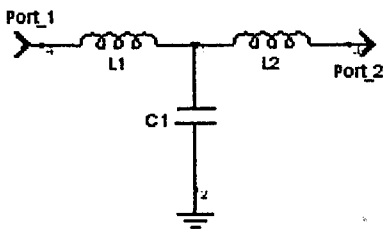


Figure 4. Low pass prototype circuit

Step 2:

The lowpass filter is then transform into bandpass filter and the value of inductor and capacitor are then calculated using the equation below.

$$L_i = L / (\omega_u + \omega_L) \quad (3)$$

$$C_i = (\omega_U + \omega_L) / \omega_0^2 C \quad (4)$$

Where  $i=1$  to  $n$

$$\omega_u = 1.05(2\pi f_c) \quad (5)$$

$$\omega_L = 0.95(2\pi f_c) \quad (6)$$

$$\omega_0 = \frac{(\omega_U + \omega_L)}{2} \quad (7)$$

Figure 5 below showed the bandpass filter circuit after the transformation from lowpass circuit.

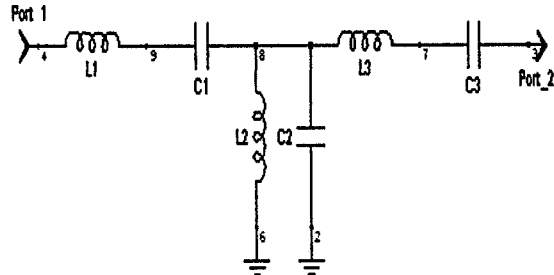


Figure 5. Bandpass filter after transformation

Step 3:

Using the J-inverter technique in [8], the admittances for the structure are obtained. The results are as shown in Table 4.

Table 4. Filters' Admittances

First coupling structure	0.1745
Intermediate coupling structure	0.0289
Final coupling structure	0.1745

From the admittance value, even and odd characteristic impedance,  $Z_{0e}$  and  $Z_{0o}$  is obtained. By using both characteristic impedances value, characteristic impedance,  $Z_0$  is calculated.

Finally, the coupling factor is calculated using the formula below.

$$C_i = 20 \log_{10} \left| \frac{Z_{0e} - Z_{0o}}{Z_{0e} + Z_{0o}} \right| \quad (8)$$

Where  $i=0$  to  $n$

Table 5 below summarized the value of  $Z_{0e}$ ,  $Z_{0o}$ ,  $Z_0$  and coupling factor.

Table 5. Calculated value for characteristic impedance and coupling factor

N	$Z_{0e}(\Omega)$	$Z_{0o}(\Omega)$	$Z_0(\Omega)$	C (dB)
0	60.248	42.798	50.779	-15.427
1	51.487	48.597	50.021	-30.789
2	51.487	48.597	50.021	-30.789
3	60.248	42.798	50.779	-15.427

Step 4:

All of the value obtained above is inserted into LineCalc to obtain the width and the length of the resonator.

Step 5:

The filter is then simulated using CST and it is being optimized to achieve the best response. The layout of the filter is realized using Autocad 2008. Additional  $\lambda/4$  length is spared in the layout for fabrication to make sure that the signal will not radiate to the aluminum plate that act as the hardware ground. The filter is then fabricated on Roger 4350B. The final dimension of the filter layout and the fabricated filter is shown in Figure 6 and Figure 7. Filter's dimension is shown in Table 6.

Table 6. Filters' dimension

Parameter	Value
Width, mm	0.81
90 degree length, mm	6.54
Sliding factor, mm	2.60
Tapping distance, mm	4.00

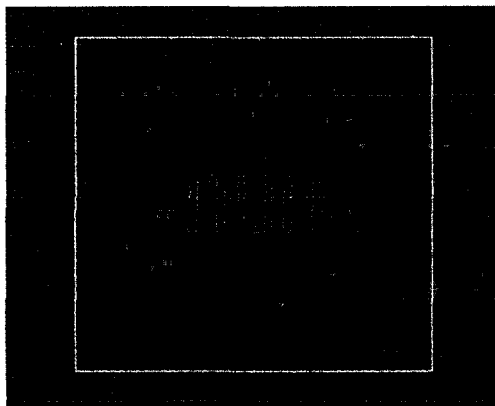


Figure 6. Filter layout from AutoCad 2008

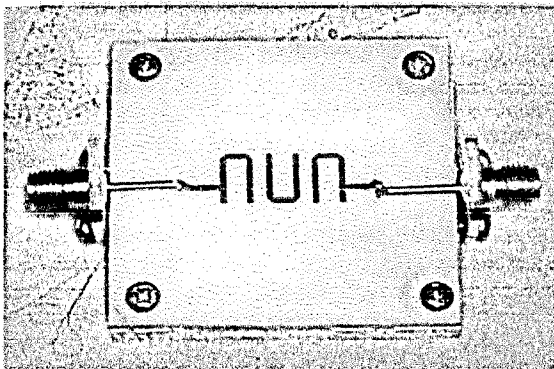


Figure 7. Fabricated filte

V. RESULTS AND DISCUSSIONS

A. Simulation Results

Figure 7 and 8 below showed the simulated result of S11 and S21 from CST in the range of 5.5 GHz to 6.1 GHz frequency. The return loss value, S11 obtained from Figure 8 is -17.74 dB and insertion loss value, S21 from Figure 9 is -1.207 dB at the centre frequency, 5.788 GHz.

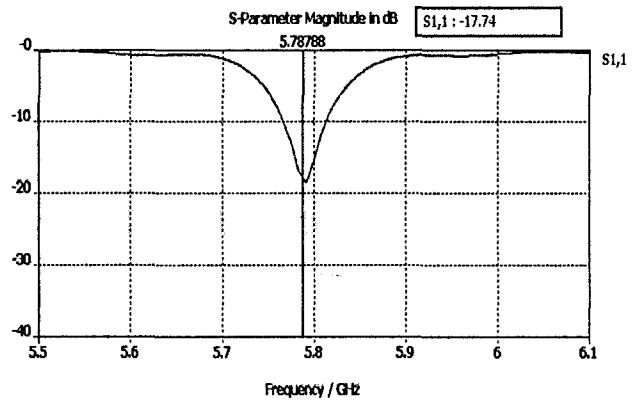


Figure 8. Return Loss, S11

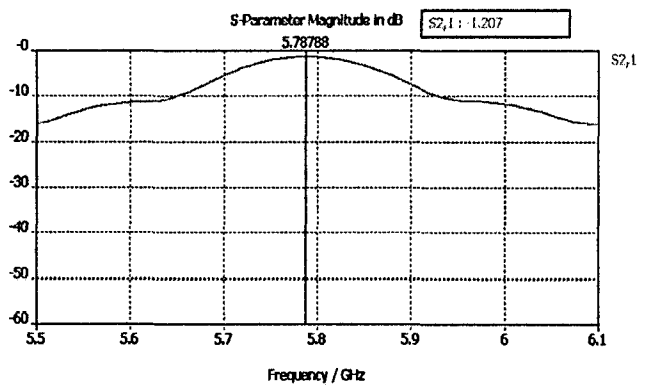


Figure 9. Insertion Loss, S21

B. Measurement Results

For this paper, the filter is measured using Wiltron 562 VNA. This VNA need to be warmed up for one hour before any measurement is done. This is to make sure that all of the memories of previous measurements are totally cleared. Then, the VNA is calibrated. When the calibration is finished, the filter is ready to measure. The measurement process should be handled carefully since both connector of the filter and VNA are fragile. The results are then transferred to Microsoft Excel.

The frequency range for this measurement is 5.5GHz to 5.9GHz. The results from the measurement are shown in Figure 10 and Figure 11.

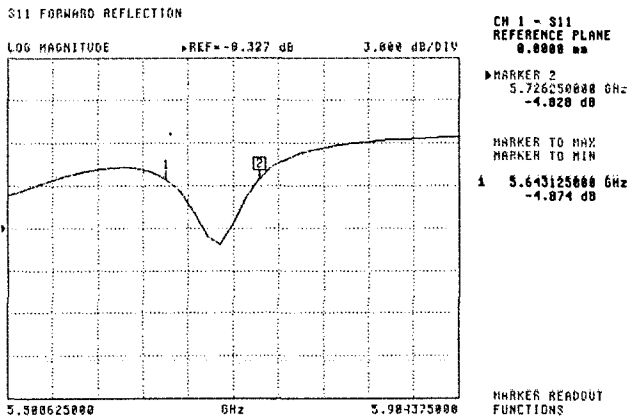


Figure 10. Return Loss, S11

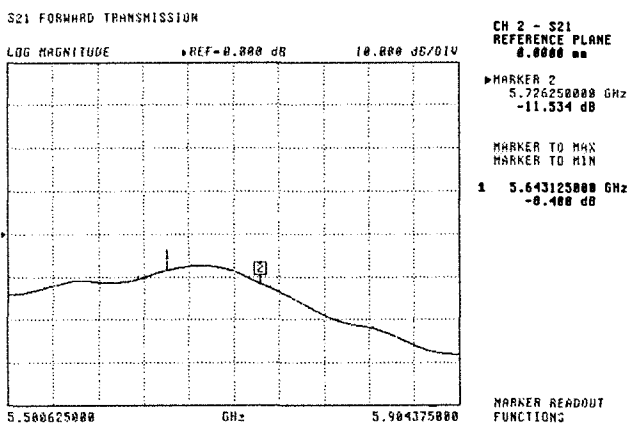
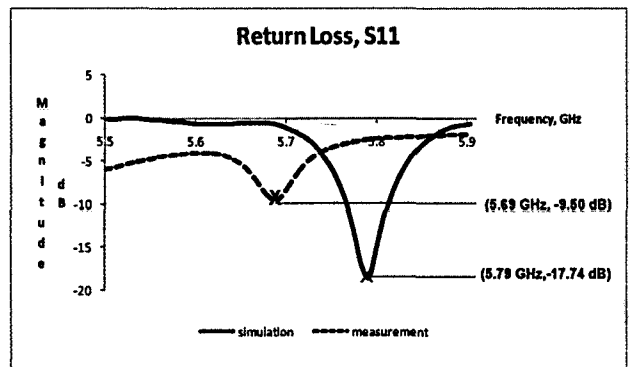
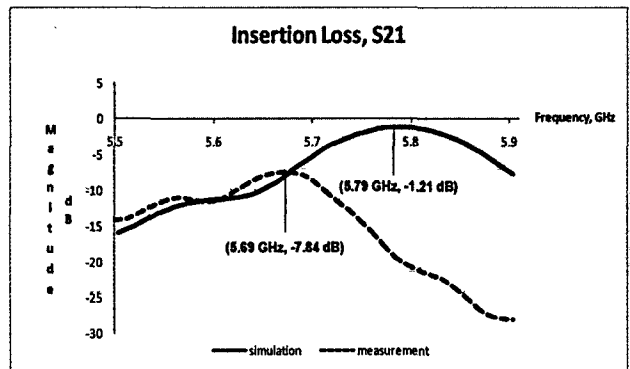


Figure 11. Insertion Loss, S21

From both graphs, the return loss value, S11 is -9.498 dB and insertion loss value, S21 is -7.471 dB at 5.7 GHz frequency.

From the measurement results, the shape of both response are as predicted in the simulation but the operating frequency is shifted to 5.7GHz, about 80MHz from original operating frequency. This is probably because of the tapping distance is longer than the simulation due to the lead solder that being used to connect the microstrip and the connector.

For a better comparison between measurement and simulation results, both responses are plotted on the same graphs as in figure below and tabulated in Table 7. Figure 12 show the comparison of S21 and Figure 13 showed the comparison of S11.



For graph the graph, it clearly show that the operating frequency from the measurement value is shifted about 80 MHz. The return loss value, S21 is also different. It decreases from -1.2129 dB to -7.836 dB. For insertion loss, S11 simulation value is -16.985 dB and measurement value is -9.498 dB.

Table 7. Comparison between simulation values and measurement values

Parameter	Simulation	Measurement
Centre frequency, $f_c$ (GHz)	5.7840	5.6906
Insertion Loss, S11 (dB)	-17.740	-9.498
Return Loss, S21 (dB)	-1.21129	-7.836

The difference in this value may caused by several factors. The matching and fabrication of the filter might be the cause of this response. The additional  $\lambda/4$  between the filter input and output to the microstrip maybe induced some losses as the signal travel through the 50Ω transmission line. Too much lead solder can cause skin effect, where most of the signal tends to ravel on part of the skin component. When that is happening, signal tends to radiate to outer space. This will cause higher attenuation and it decreases the signal

strength. The accuracy of the signal will be affected. The improper connection between connector and VNA will increase the loss of the filter, thus return loss higher.

C. Improved Circuit

Based on the results, the filter obtained is very lossy. Thus, to improve the filter responses, another filter is design while maintaining its order and bandwidth. This filter is the new and improved design using optimization technique in CST. The position of tapped point is adjusted to obtain a better return loss and the length of 50 Ω transmission line is shorten. The results for both S11 and S21 response are shown in Figure 14 and Figure 15.

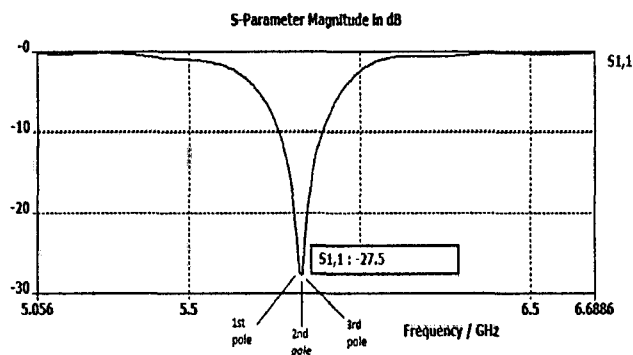


Figure 14. Return Loss, S11 for new design

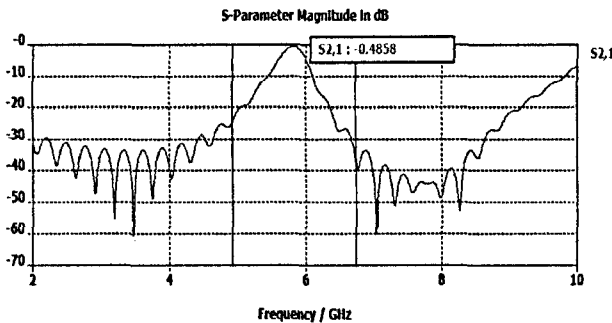


Figure 15. Insertion Loss, S21 for new design

The design cannot be proceed with fabrication due to time constraint and unavailability of VNA. The comparisons for all those responses are tabulated in Table 8.

Table 8. Comparison of all circuit's responses

Parameter	Simulation	Measurement	New Design
Centre frequency, $f_c$ (GHz)	5.7840	5.6906	5.788
Insertion Loss, S11 (dB)	-1.21129	-7.836	-0.4858
Return Loss, S21 (dB)	-17.740	-9.498	-27.61

VI. CONCLUSION

In this paper, a step by step procedure in designing a hairpin bandpass filter is done. The filter is then simulated with CST and fabricated on Roger 4350B. The results obtained from the measurement showed a good agreement with the simulation but 1.3% error in obtaining the operating frequency. The shapes of the responses are similar as the simulation results but differences in value due to some losses. A new design is created to prove that this filter can always being improved from time to time.

VII. FUTURE RECOMMENDATION

In the future, the filter should be fabricated carefully and make sure the 50Ω transmission line used is as short as possible to reduce the conductor losses. In this design, the additional  $\lambda/4$  length is spared to avoid short circuit. The connector is shorter than  $\lambda/4$  and it need to be solder to make sure it is in contact with the microstrip line. It increases the length of 50 Ω transmission line length. To avoid this situation, make sure to prepare directly the 50 Ω transmission line in the simulation design. By doing this, the results obtained might be more accurate since we have already included the 50 Ω transmission line effect in the simulation.

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