Designing Microstrip Bandpass Filter at 3.2 GHz

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Abstract: Bandpass filters play a significant role in wireless communication systems. Transmitted and received signals have to be filtered at a certain center frequency with a specific bandwidth. In designing of microstrip filters, the first step is to carry out an approximated calculation based on using of concentrated components like inductors and capacitors. After getting the specifications required, we realized the filter structure with the parallel-coupled technique. Experimental verification gives comparison, how close the theoretical results and measurements look like.
Keywords: Wireless communication, WiMAX, bandpass filter, parallel-coupled microstrip, computational electromagnetics.

1. Introduction

The advances of telecommunication technology arising hand in hand with the market demands and governmental regulations push the invention and development of new applications in wireless communication. These new applications offer certain features in telecommunication services, that in turn offer three important items to the customers. The first is the coverage, meaning each customer must be supported with a minimal signal level of electromagnetic waves, the second is capacity that means the customer must have sufficient data rate for uploading and downloading of data, and the last is the quality of services (QoS) which guarantee the quality of the transmission of data from the transmitter to the receiver with no error. In order to provide additional transmission capacity, a strategy would be to open certain frequency regions for new applications or systems. WiMAX (Worldwide interoperability Microwave Access) which is believed as a key application for solving many actual problems today is an example [1].

In realization of such a system like WiMAX we need a complete new transmitter and receiver. A bandpass filter is an important component must be found in the transmitter or receiver. Bandpass filter is a passive component which is able to select signals inside a specific bandwidth at a certain center frequency and reject signals in another frequency region, especially in frequency regions, which have the potential to interfere the information signals. In designing the bandpass filter, we are faced the questions, what is the maximal loss inside the pass region, and the minimal attenuation in the reject/stop regions, and how the filter characteristics must look like in transition regions [2].

In the process to fulfill these requirements there are several strategies taken in realization of the filters, for example, the choice of waveguide technology for the filter is preferred in respect to the minimal transmission loss (insertion loss). This strategy is still actual in satellite applications. The effort to fabricate waveguide filters prevents its application in huge amounts. As alternative, microstrip filter based on printed circuit board (PCB) offers the advantages easy and cheap in mass production with the disadvantages higher insertion losses and wider transition region. In this work we would like to give a way to conceive, design and fabricate bandpass filter for the WiMAX application at the frequency 3.2 GHz with parallel-coupled microstrips as opposed to the [3] which designed filter for wireless local area network 5.75 GHz, and [4] which used the composite resonators and stepped impedance resonators for filter realization.

Received: January 9, 2010. Accepted: April 18, 2010

2. Basics of Filter

A. Transfer Function

In Radio Frequency (RF) applications, for defining transfer function we use the scattering parameter S_{21}. In many applications we use instead the magnitude of S_{21}, the quadrate of S_{21} is preferred

$$|S_{21}(j\Omega)|^2 = \frac{1}{1+\varepsilon^2 F_n^2(\Omega)} \tag{1}$$

ε is the ripple constant, $F_n(\Omega)$ filter function and Ω is frequency variable. If the transfer function is given, the insertion loss response of the filter can calculated by

$$L_A(\Omega) = 10\log\frac{1}{|S_{21}(j\Omega)|^2}\,dB \tag{2}$$

For lossless conditions, the return loss can be found by

$$L_R(\Omega) = 10\log\left[1-|S_{21}(j\Omega)|^2\right]dB \tag{3}$$

B. Butterworth Filter

Filters designed with Butterworth approach show the maximal flat characteristics in the pass region. Figure 1 shows the attenuation characteristics of lowpass Butterworth filter. In pass region, $f < f_c$, the attenuation of ideal lowpass filter is 0 dB, good approximation must have characteristics close to zero from the frequency zero Hertz to a certain so-called cut-off frequency f_c. For $f > f_c$, the ideal lowpass filter attenuates the signal completely or $L_A \rightarrow \infty$. The Butterworth approach is expected to have the attenuation factor as high as possible.

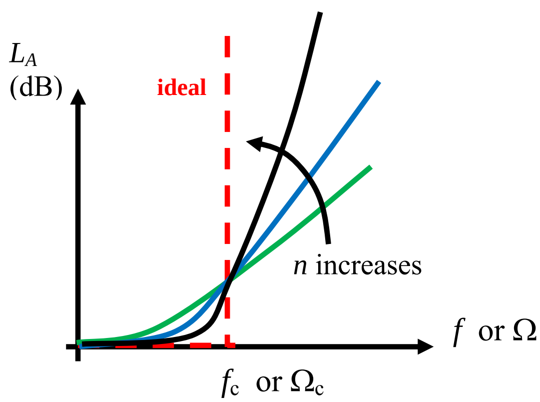


Figure 1. Characteristics of Butterworth filters.

The quadrate of the magnitude of the transfer function

$$|S_{21}(j\Omega)|^2 = \frac{1}{1+\Omega^{2n}} \tag{4}$$

Figure 2 gives the circuit implementation of the filter by means of concentrated components like inductors (L) and capacitors (C), for the even and odd filter degree (*n*).
The Butterworth approach for designing filter uses the condition attenuation of 3 dB at the frequency $\Omega=\Omega_c=1$, so that the following equations can be used for collecting the values of L and C for the circuits

$$g_0 = g_{n+1} = 1 \tag{5}$$

$$g_i = 2\sin\left(\frac{(2i-1)\pi}{2n}\right) \tag{6}$$

for i=1 to n.

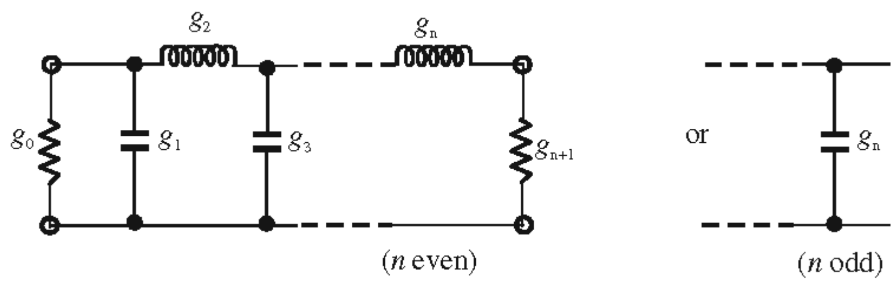


Figure 2. Realization of filter using LC components.

The value of n can be determined if an additional constraint is given, for example, the filter must have minimal attenuation factor at a certain frequency.

C. Chebyshev Filter

In practical implementation, the specification for losses in pass region can normally be higher than zero. Chebyshev approach exploits this not so strictly given specification values. It can be 0.01 dB, or 0.1 dB, or even higher values. The Chebyshev approach thereby shows certain ripples in the pass region, this can lead to better (higher) slope in the stop region. Figure 3 shows the attenuation characteristics for lowpass filter based on Chebyshev approach.

The quadrate of the magnitude of the transfer function with Chebyshev approach is given by

$$\left|S_{21}(j\Omega)\right|^2 = \frac{1}{1+\varepsilon^2 T_n^2(\Omega)} \tag{7}$$

$T_n(\Omega)$ is Chebyshev function type 1 with order n.

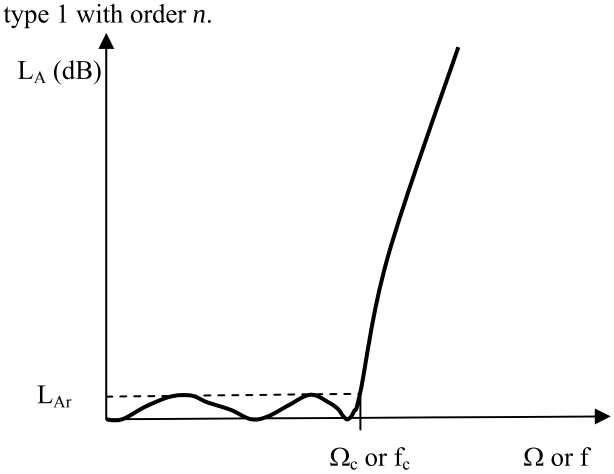


Figure 3. Attenuation characteristics for Chebyshev approach.

The component values can be calculated with the following rules

$$g_0 = 1 \tag{8}$$

$$g_1 = \frac{2}{\gamma}\sin\left(\frac{\pi}{2n}\right) \tag{9}$$

$$g_i = \frac{1}{g_{i-1}} \frac{4\sin\left(\frac{(2i-1)\pi}{2n}\right)\sin\left(\frac{(2i-3)\pi}{2n}\right)}{\gamma^2 + \sin^2\left(\frac{(i-1)\pi}{n}\right)} \sin\left(\frac{(2i-1)\pi}{2n}\right) \tag{10}$$

for i=2 to n

$$g_{n+1} = \begin{cases} 1 & \text{for odd } n \\ \coth^2\left(\frac{\beta}{4}\right) & \text{for even } n \end{cases} \tag{11}$$

where

$$\beta = \ln\left[\coth\left(\frac{L_{A,r}}{17.37}\right)\right] \text{ and } \gamma = \sinh\left(\frac{\beta}{2n}\right)$$

D. Transformation to Bandpass Filter

The previous observation was done for lowpass implementation. A transformation to bandpass is needed for getting bandpass characteristics.

In the transformation, the component L will be converted to serial combinations of L_s and C_s, whereas the component C becomes parallel combination of L_p and C_p. With the cut-off frequencies ω_1 and ω_2 as lower and upper boundary, we can calculate the center frequency and the relative frequency bandwidth as follows

$$\omega_o = \sqrt{\omega_1\omega_2} \text{ and } FBW = \frac{\omega_2 - \omega_1}{\omega_o}$$

And the values for the new components are

$$L_s = \left(\frac{1}{FBW \cdot \omega_o}\right) Z_o \cdot g \tag{12}$$

$$C_s = \left(\frac{FBW}{\omega_o}\right)\frac{1}{Z_o \cdot g} \tag{13}$$

for the serial combination, and

$$C_p = \left(\frac{1}{FBW \cdot \omega_o}\right)\frac{g}{Z_o} \tag{14}$$

$$L_p = \left(\frac{FBW}{\omega_o}\right)\frac{Z_o}{g} \tag{15}$$

for the parallel combination.

Z_o is the value of the load impedance, normally set to 50 Ω.

3. Filter Realization with Microstrip Technology

A. Microstrip Transmission Line

Microstrip transmission line is the most used planar transmission line in Radio frequency (RF) applications [5]. The planar configuration can be achieved by several ways, for example with the photolithography process or thin-film and thick film technology. As other transmission line in RF applications, microstrip can also be exploited for designing certain components, like filter, coupler, transformer or power divider.

If a microstrip transmission line, as depicted in Fig, 4, is used for transport of wave with relative low frequency, the wave type propagating in this transmission line is a quasi-TEM wave. This is the fundamental mode in the microstrip transmission line.

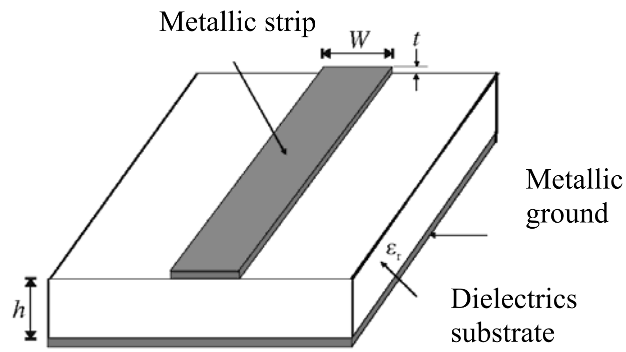


Figure 4. Microstrip transmission line.

The width of the strip *W* together with the dielectric constant and the thickness of the substrate determines the characteristic impedance Z_o of the line [5].

B. Designing Bandpass Filter

Figure 5 shows the filter structure observed in this work. This filter type is known as parallel-coupled filter. The strips are arranged parallel close to each other, so that they are coupled with certain coupling factors. We use the following equations for designing the parallel-coupled filter

$$\frac{J_{01}}{Y_o} = \sqrt{\frac{\pi FBW}{2 g_0 g_1}} \tag{16}$$

$$\frac{J_{j,j+1}}{Y_o} = \frac{\pi FBW}{2} \frac{1}{\sqrt{g_j g_{j+1}}} \tag{17}$$

for j=1 to n=1

$$\frac{J_{n,n+1}}{Y_o} = \sqrt{\frac{\pi}{2} \frac{FBW}{g_n g_{n+1}}} \tag{18}$$

g_0, g_1, .., g_n can be taken from table, FBW is the relative bandwidth as explained before, $J_{j,j+1}$ is the characteristic admittance of J inverter and Y_o is the characteristic admittance of the connecting transmission line.

With the data of characteristic admittance of the inverter, we can calculate the characteristic impedances of even-mode and odd-mode of the parallel-coupled microstrip transmission line, as follows [11, 12]

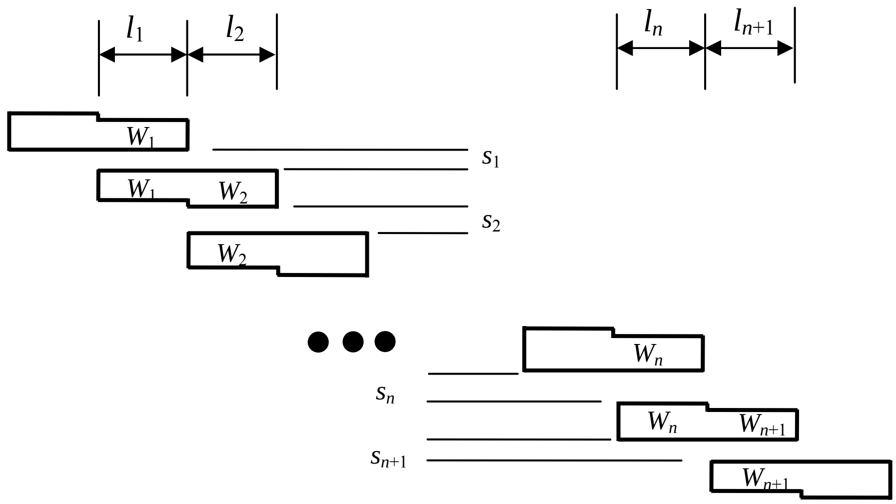


Figure 5. Parallel-coupled bandpass filter.

$$\left(Z_{0e}\right)_{j,j+1} = \frac{1}{Y_o}\left[1 + \frac{J_{j,j+1}}{Y_o} + \left(\frac{J_{j,j+1}}{Y_o}\right)^2\right] \tag{19}$$

for j = 0 to n, and

$$\left(Z_{0o}\right)_{j,j+1} = \frac{1}{Y_o}\left[1 - \frac{J_{j,j+1}}{Y_o} + \left(\frac{J_{j,j+1}}{Y_o}\right)^2\right] \tag{20}$$

for j = 0 to n.

4. Filter Calculations with Sonnet v12 and Measurements

For filter fabrication we use a PCB of type RO TMM10 provided by the Rogers Corp (www.rogerscorp.com) with the thickness 0.762 mm (0.03 inch). The substrate of type RO TMM10 has the relative permittivity of 9.2 and tangent loss of 0.0022.

In order to have the wave impedance of 50 ohms, the microstrip line designed in this PCB must have the strip width of 0.7 mm or 0.8 mm. Figure 6 shows the microstrip transmission line expected to have the wave impedance of 50 ohms as the external connector for connecting to other components.

Figure 7 visualizes the reflection factor gained by computer simulation with Sonnet [6] for strip width of 0.7 mm and 0.8 mm. We see the strip width of 0.8 mm gives much more better results compared to strip width of 0.7 mm.

Sonnet is a commercial software for electromagnetic calculation based on the method of moment [7], which is implemented in so-called multilayered structures [8]. Sonnet has been used successfully for more than a decade [9, 10].

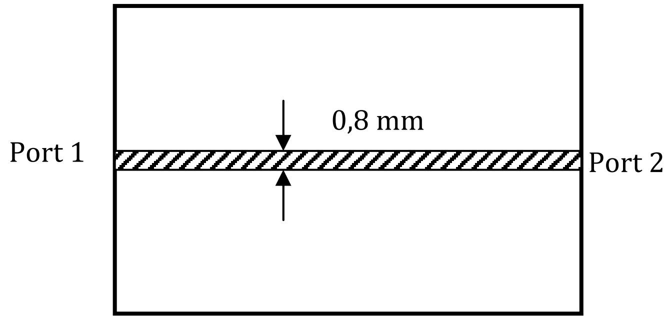


Figure 6. Connecting Microstrip line.

The design of bandpass filter will be done at the center frequency of 3.2 GHz with the bandwidth of 0.1 GHz, or FBW = 0.1/3.2 = 0.03125.

In designing the filter, the filter order of *n* =3 is used. With eq. (5) and (6) we get $g_0=1=g_4$; $g_1=1$;$g_2=2$; $g_3=1$;

$$\frac{J_{01}}{Y_o} = \sqrt{\frac{\pi FBW}{2g_0 g_1}} = 0.2216 = \frac{J_{3,4}}{Y_o} \text{ and}$$

$$\frac{J_{1,2}}{Y_o} = \frac{\pi FBW}{2}\frac{1}{\sqrt{g_1 g_2}} = 0.0347 = \frac{J_{2,3}}{Y_o}$$

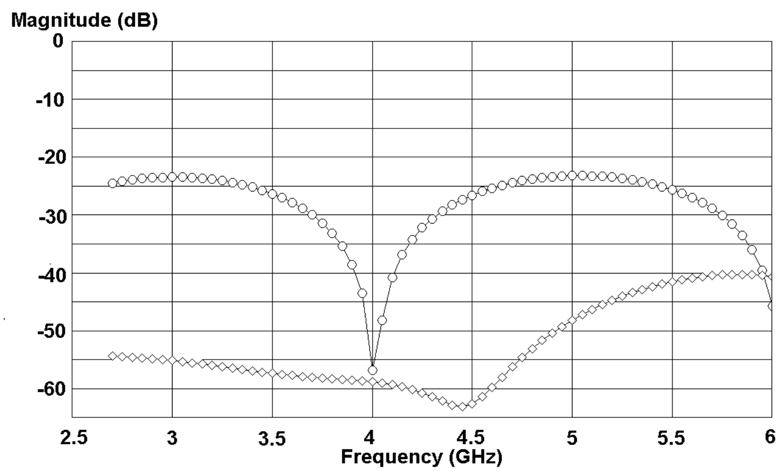


Figure 7. Reflection factor (o) 0.7 mm and (◊) 0.8mm.

Then we calculate the even-mode and odd-mode characteristic impedances of this parallel-coupled microstrip line using eq. (19) and (20), which leads to

$$(Z_{0e})_{0,1} = 63.535\Omega = (Z_{0e})_{3,4}$$

$$(Z_{0o})_{0,1} = 41.3753\Omega = (Z_{0o})_{3,4}$$

$$(Z_{0e})_{1,2} = 51.795\Omega = (Z_{0e})_{2,3}$$

$$(Z_{0o})_{1,2} = 48.3852\Omega = (Z_{0o})_{2,3}$$

With the procedure explained in [11,13,14], we can determine the width of parallel-coupled microstrip lines W and the distance between them s. A pair of parallel-coupled microstrip lines with certain width and separation distance will deliver a pair of characteristic impedances, the even mode and the odd mode ones. Figure 8 shows the graph for microstrip line with the relative permittivity of 9.2 and thickness 0.763 mm. We see, if the separation is small the even mode impedance is high, and the odd mode impedance is small. In order to achieve the impedance pair $(Z_{0e})_{0,1} = 63.535\Omega$ and $(Z_{0o})_{0,1} = 41.3753\Omega$, we built a search algorithm which match the values of W and s to approximate the values of the impedance pair above. The process is visualized in Figure 8. According to Figure 8, we get W_1=0.7 mm and s_1=0.46 mm and the effective relative permittivities 6.67 and 5.4 for even and odd mode, respectively. The length of the resonator required is l_1'=9.568 mm, which must be geometrically reduced in order to take into account the fringe effects [13] due to open end with Δl_1=0.3178 mm, so that the geometrical length of the first resonator becomes l_1 = 9.25 mm. The same data will be also used for the fourth resonator.

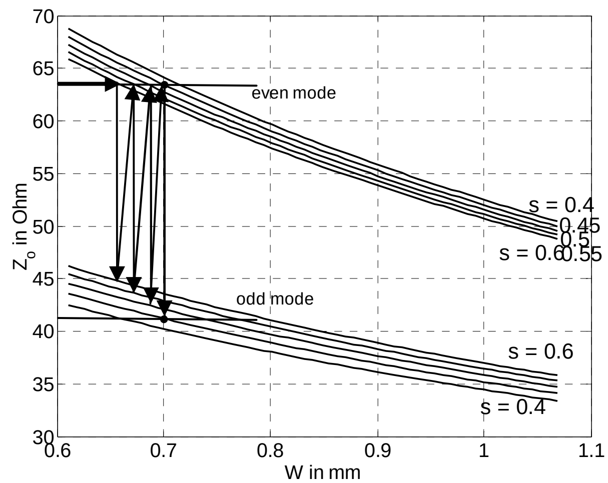


Figure 8. Impedance of coupled microstrip, small separation s

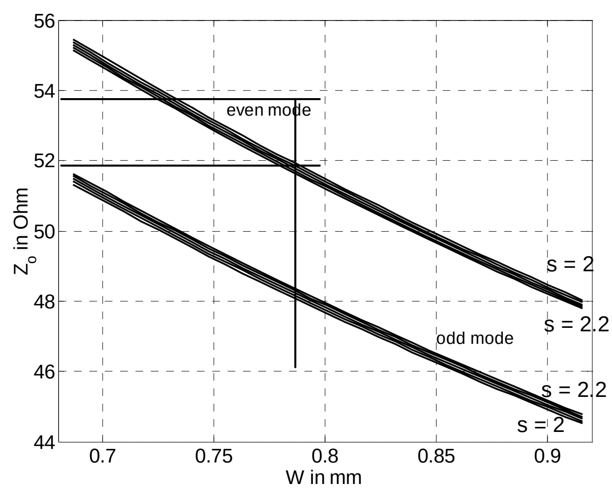


Figure 9. Impedance of coupled microstrip, large separation s

For the second and third resonator we must find the value for W_2 and s_2, which yield $(Z_{0e})_{1,2}$ = 51.795 ohms and $(Z_{0o})_{1,2}$ = 48.3852 ohms. The iterative calculation gives W_2=0.78 mm and s_2=2.2 mm as given in Figure 9, and the effective relative permittivities 6.542 and 5.8855 for even and odd mode respectively. The resonator length must be reduced from 9.41 mm to 9.09 mm.

Figure 10 shows the schematic of the bandpass filter designed. This figure gives four resonators built by four pairs of parallel-coupled microstrip.

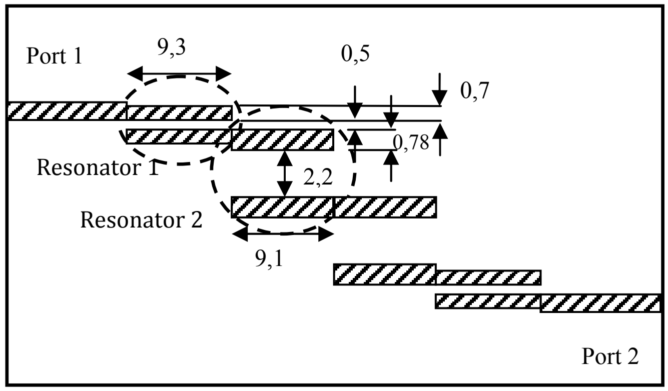


Figure 10. Schematic of bandpass filter (geometrical data in mm).

In the next part we would like to consider the effects of geometrical deviations of the strips to the reflection and transmission factors of the filter. It is very important due to the small dimensions of the filter in respect to the uncertainties in microstrip fabrication later.

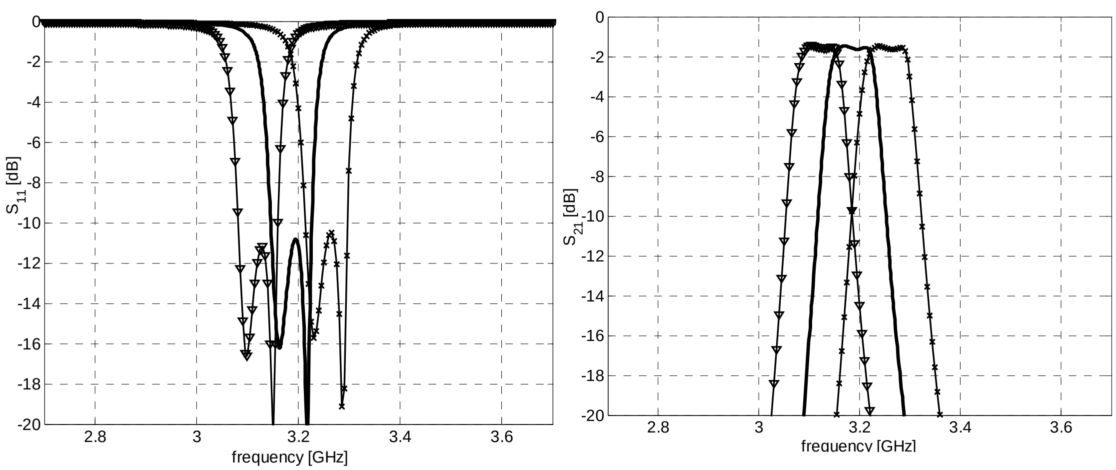


Figure 11. Comparison of effects of resonator length (*l*). Solid lines: data given in Figure10, triangles: all resonator lengths increased 0.2 mm, stars: all lengths reduced 0.2 mm

Figure 11 gives effects of changes in resonator lengths (*l*). The characteristics of the filter are shifted about 50 MHz with almost unchanged contour to lower or higher frequencies, if the resonators are 0.2 mm longer and shorter, respectively.

Similarly with the deviation of the strip widths as shown in Figure 12, however the tolerances in strip width give smaller impacts to the reflection and transmission factors.

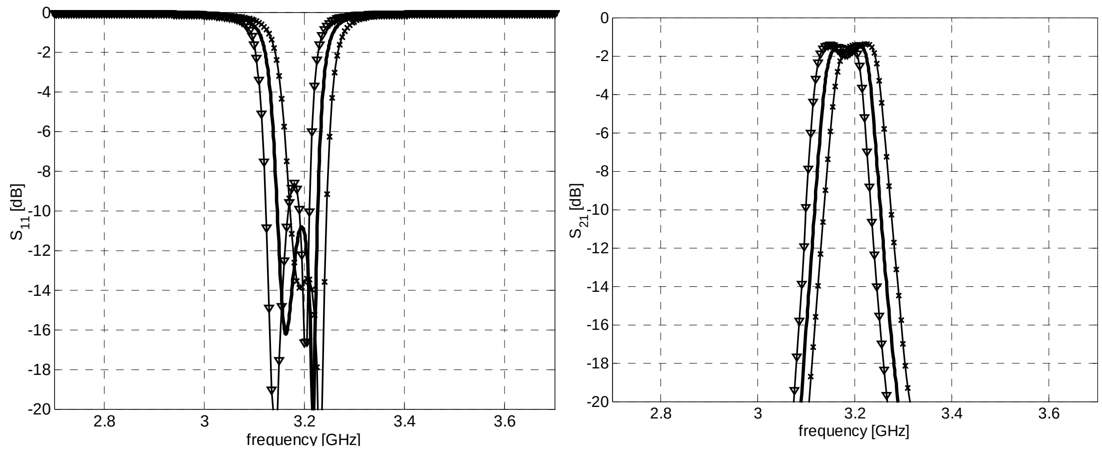


Figure 12. Comparison of effects of resonator width (W). Solid lines: data given in Figure10, triangles: width wider by 0.1 mm, stars: width narrower by 0.1 mm.

We built a bandpass filter with the data given in Figure 10, and show again the characteristics of the filter separatery in Figure 13. We see a resonant at desired frequency of 3.2 GHz with reflection factors smaller than -10dB. The insertion loss occurring in S_{21}, of about -2 dB, primarily due to the tangent loss of the substrate.

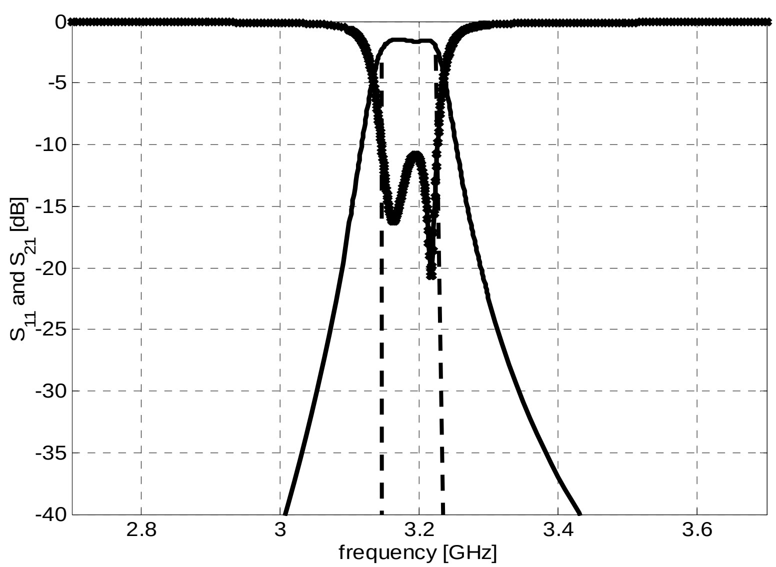


Figure 13. Reflection (stared) and transmission factor (solid) of the filter in Figure 10.

Figure 14 gives the reflection and transmission factor measured for the range 50 MHz to 6 GHz. This result ensures that the bandpass characteristics are indeed valid for a wide range of spectrum. In order to get detailed characteristics in the neighborhood of 3.2 GHz, Figure 15 gives more information needed to quantify the measurement results. With the 3 dB boundary, we get just about 50 MHz bandwidth, expected was 100 MHz, and higher loss of -7.5 dB.

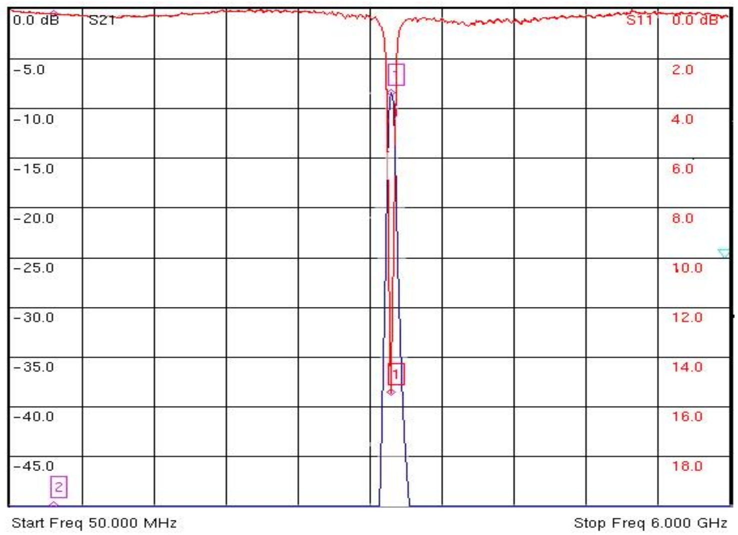
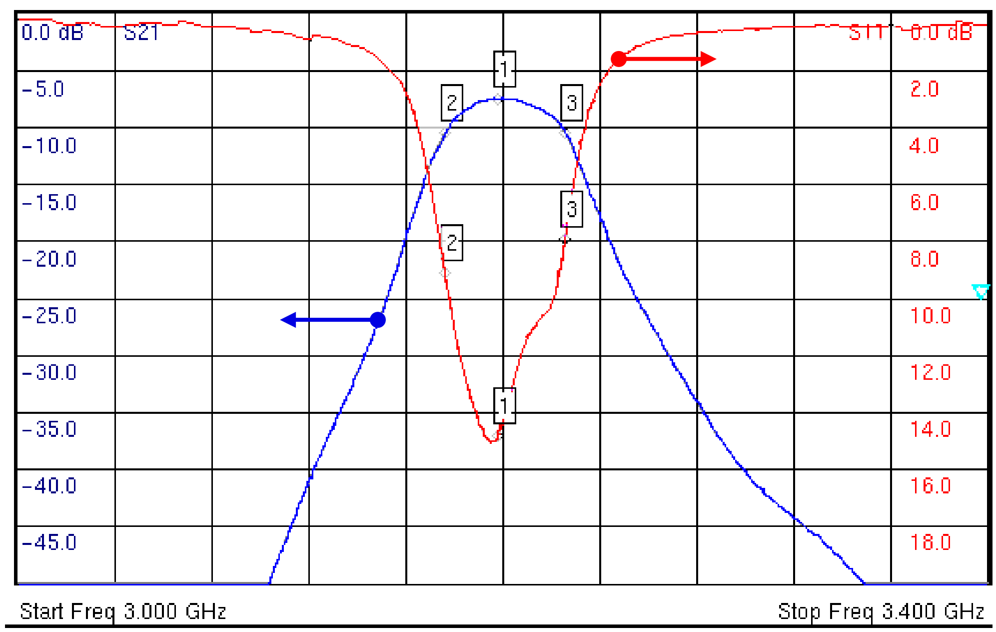


Figure 14. Measured reflection and transmission factor 50 MHz to 6 GHz.



Mkr	Ref Freq	Ref Ampl
M1	3.198091 GHz	-7.52/14.85dB
M2	3.176000 GHz	-10.46/9.14dB
M3	3.225455 GHz	-10.57/7.97dB

Figure 15. Detailed view of measured reflection and transmission factor.

The additional loss obtained in the measurement originates likely from losses of the coaxial connectors and their poor contacts to the microstrip line, which cannot be considered in the simulation with Sonnet. Figure 16 shows the measured reflection factor in complex plane in form a Smith's chart, we see, at the resonant frequency 3.2 GHz, indicated by marker 1, the point is close to the centre of the Smith's chart.

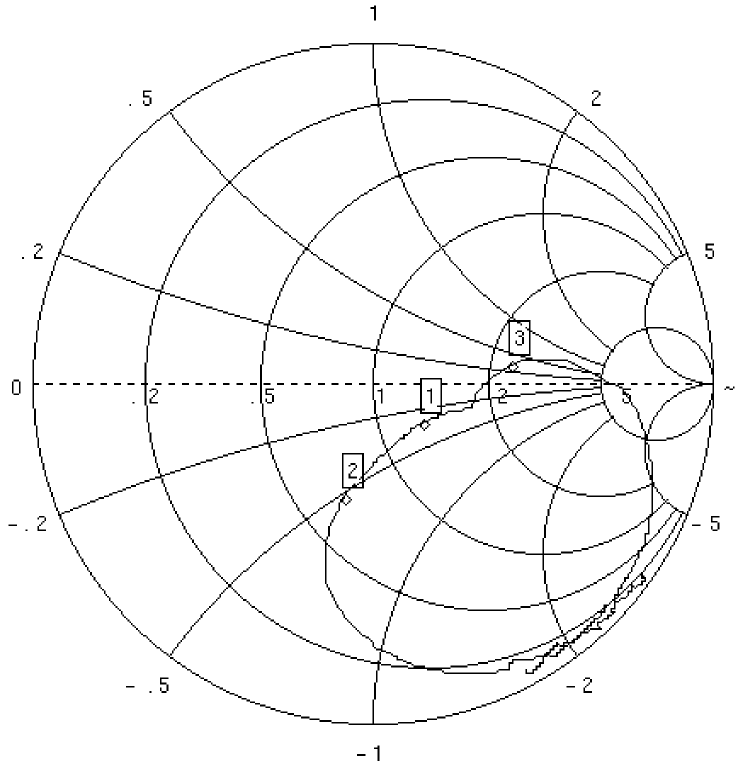


Figure 16. Measured reflection factor in Smith's chart.

5. Conclusion

Designing of bandpass filter with Butterworth approach in combination with concentrated components, i.e. inductors and capacitors and its computational verification in form of parallel-coupled microstrip lines with the program Sonnet give very good filter characteristics at the center frequency 3.2 GHz with frequency bandwidth of about 100 MHz as required at the specification stage. At the center frequency the insertion loss and reflection factor has the values about -2 dB and better than -15 dB, respectively.

The measurement gives also very good filter characteristics at the frequency 3.2 GHz, however with larger insertion loss of about -7.5 dB and smaller bandwidth of about 50 MHz. This larger loss originates likely from losses of the coaxial connectors and their poor contacts to the microstrip line.

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