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Butterworth Filter Design at RF and X-band Using Lumped and Step Impedance Techniques

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Abstract— **Low pass filters are widely used in telecommunications for a variety of commercial and military applications. The design and simulation technology differs according to the cut-off frequency. In this work we design, simulate and compare various types of passive Butterworth filters, regarding their order and the type of topology used, for RF frequencies using lumped components with cut-off frequency around 3 KHz. Then, low frequencies are transferred to microwave frequencies and a low pass even order Butterworth microwave filter, using step impedance microstrip lines with cut-off frequency around 3 GHz is analyzed, designed and simulated.**

Key Words—**Butterworth Filters, Lumped Elements, Microstrip Lines, Step Impedance.**

I. Introduction

Low pass filters are widely used in telecommunications for a variety of commercial and military applications. In telecommunications transmitters they are used to block produced harmonic emissions and potentially interfere with other communications [1] - [3].

Passive filters using inductors and capacitors have been studied excessively in literature and they are most effective in RF frequencies between 30 KHz and 300 MHz. Below 30 KHz, active filters using operational amplifiers in Shallen-Key topology are usually more cost effective and above 300 MHz striplines and microstrip are generally used. Generally, at frequencies below 1 GHz, filters are usually implemented using lumped elements such as resistors, inductors and capacitors.

There are a number of standard mathematical approaches to design a normalized Low Pass Prototype that approximates an ideal low-pass filter response. Among the well known methods are: Maximally flat or Butterworth function, Equal ripple or Chebyshev approach and Elliptic function. The Bessel approximation has the slowest cutoff rate, but this is a trade-off with its favorable linear phase response, which reduces phase distortion. The Chebyshev approximation provides rapid cutoff beyond the cut-off frequency but the designer compromise this with its low ripple in the pass band. A Butterworth approximation has a characteristic between the two.

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II. Passive Low Pass Filters in RF Frequencies

A. Butterwoth Low Pass Prototype

There are a number of standard approaches to design a normalized Low Pass Prototype of Figure 1 that approximate an ideal low-pass filter response with cutoff frequency of unity. [1], [4]-[6]. Among the well known methods is the maximally flat or Butterworth function.

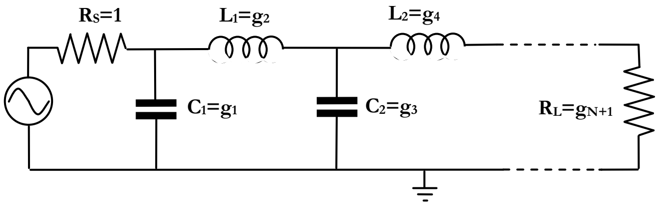

Fig. 1. Π-topology for medium to high impedance loads for the Nth order Butterworth Low Pass filter

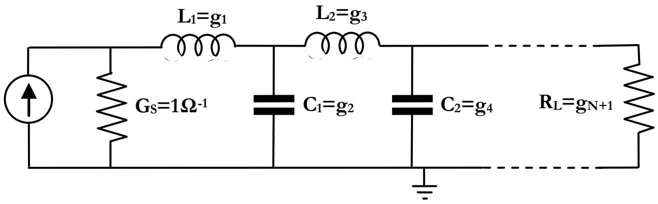

Fig. 2. T-topology for low impedance loads for the Nth order Butterworth Low Pass filter

A filter response is defined by its insertion loss (IL in dBs), or power loss ratio (P_{LR}) which is the reverse ratio of the transducer power gain G_T. The G_T is defined as the ratio of the Power delivered to the Load (P_L) to the Power Available from the source (P_{AVS}). [7]-[9]. The basic idea is to approximate the ideal Power Loss Ratio ($1/|H(\omega)|^2$, where $H(\omega)$ is the amplitude response) of a passive filter using Butterworth polynomials function as it is shown in eq.(1)

$$P_{LR} = \frac{P_{AVS}}{P_L} = 1 + \left(\frac{\omega}{\omega_C}\right)^{2N} \tag{1}$$

where ω is the circular frequency (ω=2πf) in rad/sec and ω_C is the cut-off frequency.

Table I gives the factors of the polynomials for N=1 to N=5. We use Table 1, to design a low pass prototype Butterworth filter. The values of g_i correspond to inductance and capacitance in the Butterworth filter as shown in figures 1 and 2.

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TABLE I
ELEMENT VALUES FOR MAXIMALLY FLAT (BUTTERWORTH) LOW PASS PROTOTYPE

N	g1	g2	g3	g4	g5	g6
1	2.000	1.000				
2	1.414	1.414	1.000			
3	1.000	2.000	1.000	1.000		
4	0.765	1.848	1.848	0.765	1.000	
5	0.618	1.618	2.000	1.618	0.618	1.000

g0=1 and ω_C=1rad/s

The filter has a unity cut-off frequency and unity load impedance. It can be converted into a low-pass filter, with a specified cut-off frequency and a specified impedance, using frequency scaling and impedance transformation. For a new load impedance of R_0=50Ω and cut-off frequency of ω_C, the original resistance R_i , inductance L_i and capacitance C_i (i=1,2,3,…) are changed by the following formulas and a low pass filter as showed in figure 3, is designed.

$$R = R_0 R_N \tag{2}$$

$$L = R_0 \frac{L_N}{\omega_C} \tag{3}$$

$$C = \frac{C_N}{R_0 \cdot \omega_C} \tag{4}$$

B. Odd Order Π and T Butterworth Models

The first circuit design is a 3rd order Butterworth filter in Π-topology. The specifications of the filter are: R_S= R_L = 50Ω, cut-off frequency f_C= 3 KHz. For this purpose, we will firstly design the filter with ω_C=1 rad/s. Using Table 1, the schematic is as in Figure 1, where C_1=g_1=1F, L_1=g_2=2H, C_2 = g_3 = 1F. Using transformations given by equations (2) to (4) we find the values of the elements as shown in figure 3: (C_1= C_2=1.06μF, L_1=5.3mH).

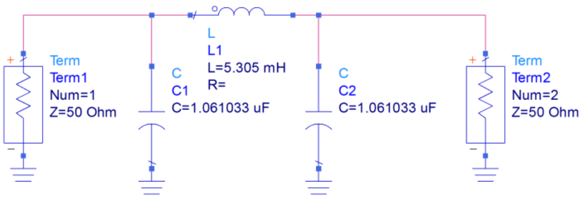

Fig.3. A low pass 3rd order passive filter at 3 KHz in Π-topology.

The second circuit design is a 3rd order Butterworth filter in T-topology. The specifications of the filter are the same. Using Table 1, the schematic is as in Figure 4, where L_1=g_1=1H, C_1=g_2=2F, L_2=g_3=1H. Using equations (2) to (3) the values of the lumped elements are: L_1=L_2=2.65mH and C_1=2.12 μF, as showed in figure 4.

Another way to find the elements in T-topology with known elements of the Π-topology is the Wye-Delta transformation. Both topologies give a frequency response as it is shown in figure 5 where it is clear that the cut-off frequency is 3 KHz and the slope is -18 dB/octave.

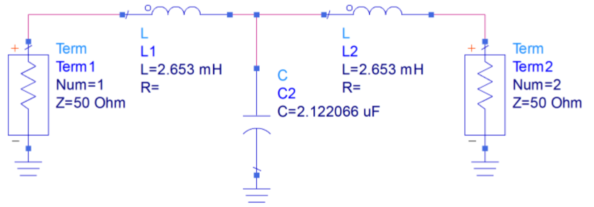

Fig.4. A low pass 3rd order passive filter at 3 KHz in star topology.

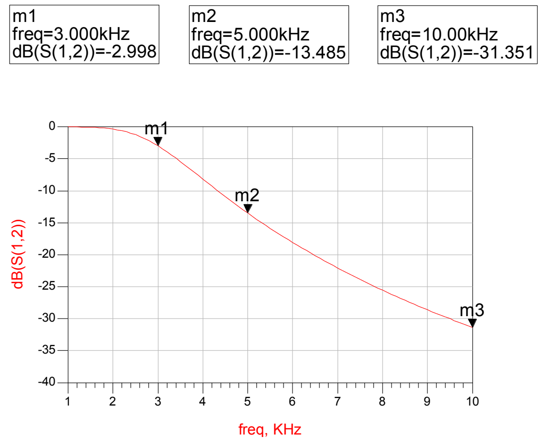

Fig.5. The Bode diagram for 3rd order low pass filter.

Following the same algorithm for a 5th order Low-Pass Filter in Π-topology, the elements parameters are L_1=L_2=4.3 mH C_1= C_3=0.66μF and C_2= 2.12 μF. The frequency response is showed in figure 6. The only difference is that the slope now is -30dB/octave.

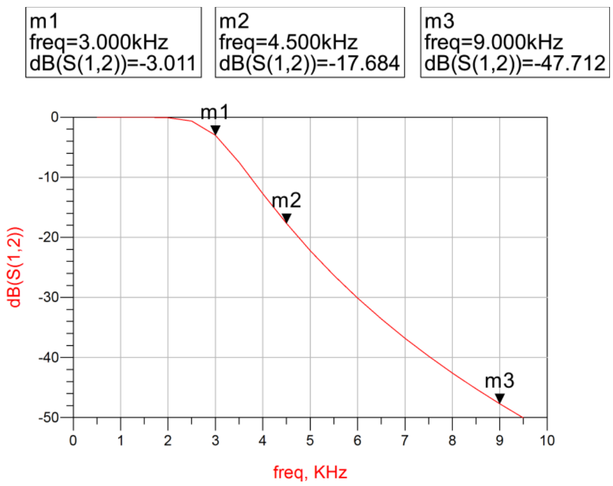

Fig.6. The Bode diagram for 5rd order low pass filter at 3 KHz.

C. Even Order Π and T Butterworth Models

In order to design a 4th order Low Pass Filter in Π-topology, having ω_C=1rad/s and R_L=1Ω, we use figure 1 and Table 1 to find the elements' values: C_1=g_1=0.765F, L_1=g_2=1.848H, C_2 =g_3=1.848F, L_2=g_4=0.765 H. We use then, equations (2) to (4) in order to find the elements for 50Ω impedance and f_C=3 KHz: C_1=g_1=0.812μF, L_1=g_2=4.9mH, C_2=g_3=1.961μF, L_2=g_4=2.03 mH.

Following the same algorithm for a 4th order Low-Pass Filter in T-topology, we find the values of the lumped elements L_1=2.03mH, C_1=1.96μF, L_2=4.9mH and C_2=812nF. The frequency response for both circuits is showed in figure 8. The only difference is that the slope now is -24dB/octave.

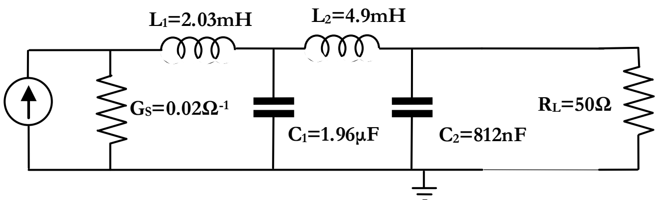

Fig.7. A low pass 4rd order passive filter at 3 KHz in T-topology.

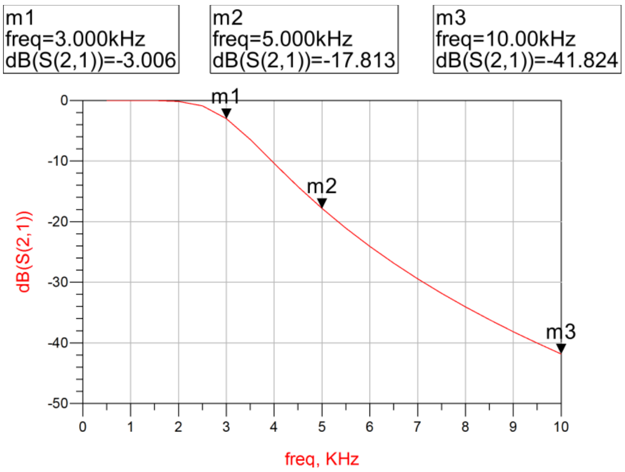

Fig.8. The Bode diagram for the 4rd order lumped filter.

III. Low Pass Filter Design in Microwave Frequencies

A relatively easy way to implement low-pass filters in microstrip or stripline is to use alternating sections of high and low characteristic impedance (Z_0) transmission lines. Such filters are usually referred to as stepped-impedance filter and are popular because they are easy to design and take up less space than similar low-pass filters using stubs. However due to the approximation involved, the performance is not as good and is limited to application where a sharp cut-off is not required [10]-[15].

A. Microstrip Line Network analysis

A transmission line of length l terminated at a load Z_L and having a characteristic impedance of Z_0 is shown in figure 9. Assume that it is lossless (α=0), which means that the propagation constant γ=a+jβ is the phase constant β of the transmission line. The electrical length of the line is βl. For EM wave propagation that is of TEM mode or quasi-TEM mode, the propagation constant can be approximated as:

$$\beta = \frac{2\pi}{\lambda} = \frac{\omega}{\upsilon_p} \cong \omega\sqrt{\mu_o \varepsilon_{eff} \varepsilon_o} = \sqrt{\varepsilon_{eff}}\,\frac{\omega}{c} = \sqrt{\varepsilon_{eff}}\,k_o \quad (5)$$

The vacuum permittivity is $\varepsilon_0=8.854\times10^{-12}$ A·s/V·m.The ε_{eff} is the effective dielectric constant of the material, ε is the complex, frequency dependent absolute permittivity of the microstrip material ($\varepsilon=\varepsilon_{eff}\cdot\varepsilon_0$). The relative permittivity ε_r is not always equal to the effective dielectric constant as we will discuss in the paragraph B. The relative permeability $\mu_0=4\pi\times10^{-7}$ V·s/A·m is the magnetic constant.

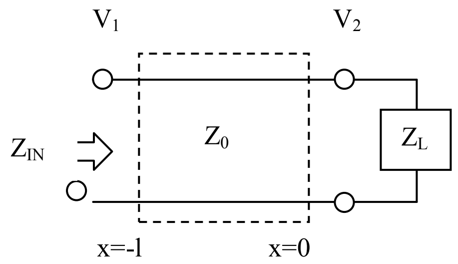

Fig. 9. A transmission line as a two port network

The input impedance of the transmission line is given by the following formula.

$$Z_{in} = Z_0 \frac{Z_L + jZ_0 \tan(\beta l)}{Z_0 + jZ_L \tan(\beta l)} \quad (5)$$

A short length of a transmission line is a reciprocal two-port network and can be described in terms of admittance parameters as

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{12} & Y_{11} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \quad (6)$$

where

$$Y_{11} = \left.\frac{I_1}{V_1}\right|_{V_2=0} = \left.\frac{I_2}{V_2}\right|_{V_1=0} \text{ and } Y_{12} = \left.\frac{I_2}{V_1}\right|_{V_2=0} = \left.\frac{I_1}{V_2}\right|_{V_1=0} \quad (7)$$

or, in terms of impedance parameters as

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{12} & Z_{11} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} \quad (8)$$

where, using equation (5) we can find:

$$Z_{11} = \left.\frac{V_1}{I_1}\right|_{I_2=0} = \left.\frac{V_2}{I_2}\right|_{I_1=0} = -jZ_0 \cot(\beta l) \quad (9a)$$

$$Z_{12} = \left.\frac{V_2}{I_1}\right|_{I_2=0} = \left.\frac{V_1}{I_2}\right|_{I_1=0} = -jZ_o \frac{1}{\sin(\beta l)} \quad (9b)$$

These representations lead naturally to the Π and T equivalent circuits shown in in figures 10 and 11. Because the network is lossless, there are three degrees of freedom (only the imaginary parts of the three matrix elements). This implies that the Π and T equivalent circuits as shown in figures 10 and 11 can be constructed from purely reactive elements as shown in figures 1 and 2 respectively.

Assuming short lines, with electrical length βl<π/4, the series element of figure 11 can be thought of as inductor (positive reactance) and the shunt element can be considered

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a capacitor (negative reactance). What is more, for such electrical lengths, sin(βl)= βl and tan(βl/2)= βl/2 (small angle approximation).

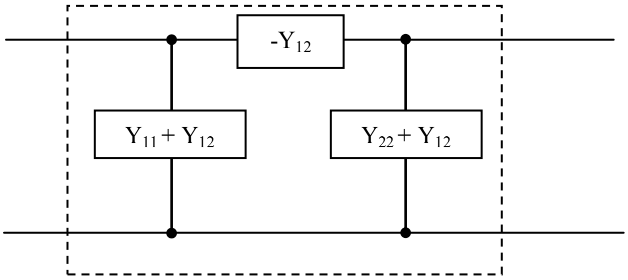

Fig. 10. The Π-equivalent circuit for the transmission line

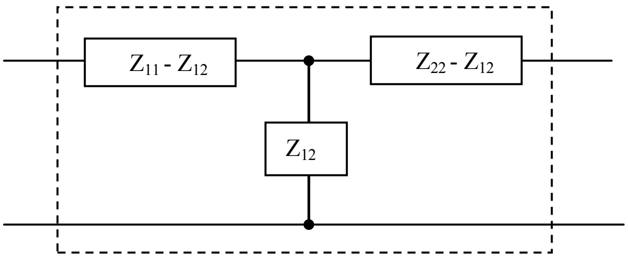

Fig. 11. The T-equivalent circuit for the transmission line

$$Z_{11} - Z_{12} = -jZ_0\left(\frac{\cos\beta l - 1}{\sin\beta l}\right) = \\ = jZ_0\left(\tan\frac{\beta l}{2}\right) = jZ_0\frac{\beta l}{2} \quad (10)$$

and

$$Z_{12} = -jZ_o\frac{1}{\sin(\beta l)} = -jZ_0\frac{1}{\beta l} \quad (11)$$

Because the ideal inductor's impedance Z is the reactance X where $jX=Z_{11}-Z_{12}$ and the ideal capacitor's admittance Y is the susceptance B where $jB=1/Z_{12}$ and assuming large impedance $Z_0=Z_H>50\Omega$, equations (10) and (11) reduce to

$$X = \omega L = Z_H\beta l \quad \text{and} \quad B = \omega C = \frac{\beta l}{Z_H} = 0 \quad (12)$$

Assuming a low impedance $Z_0=Z_L<50\Omega$, equations (10) and (11) reduce to

$$X = \omega L = Z_L\beta l = 0 \quad \text{and} \quad B = \omega C = \frac{\beta l}{Z_L} \quad (13)$$

The actual values of Z_H and Z_L are usually set to the highest and lowest characteristic impedance that can be practically fabricated. Typical values are Z_H=100 to 150Ω and Z_L=10Ω to 20 Ω. Since a typical low-pass filter consists of alternating series inductors and shunt capacitors in a ladder configuration, we could implement the filter by using alternating high and low characteristic impedance section transmission lines. Using (12) and (13), the relationships between inductance and capacitance to the transmission line length at the cutoff frequency ωc are:

$$l_{HIGH} = l_L = \frac{\omega_c L}{Z_H\beta} \quad (14a)$$

$$l_{LOW} = l_C = \frac{\omega_c C Z_L}{\beta} \quad (14b)$$

B. Microstrip Line Field Analysis

A cross section of microstrip and strip line on a printed circuit board is shown in Figure 12. For stripline the propagation mode is TEM since the conducting trace is surrounded by similar dielectric material. Hence $\varepsilon_{eff}=\varepsilon r$. For microstrip line the propagation mode is a combination of TM and TE modes. This is due to the fact that the upper dielectric of a micostrip line is usually air while the bottom dielectric is the printed circuit board dielectric [16]. A TEM mode cannot be supported as the phase velocities for electromagnetic waves in air and the board are different, resulting in mismatch at the air-dielectric boundary. However at frequencies lower than 6GHz, the axial E and H fields are small enough that we can approximate the propagation mode as TEM, hence the name quasi-TEM. For microstrip line the effective dielectric constant ε_{eff} falls within the range 1 and εr. At low frequency most of the electromagnetic field is distributed in the air, while at high frequency the electromagnetic field crowds towards the dielectric [17]-[19].

Empirical formulas are obtained from the numerical solution by the methods of curve fitting. Assuming the conductors and dielectric are lossless, and ignoring the effect the conductor thickness t, an example of the empirical formulas for ε_{eff}, Z_L (W<H) and Z_H (W>H) are given by [1], [7]:

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\cdot\frac{1}{\sqrt{1+\frac{12H}{W}}} \quad (15)$$

where, for low and high impedances' parts the ratio H/W could be found by the followin

$$\frac{W_L}{H} = \frac{8e^A}{e^{2A} - 2} \quad (16)$$

$$\frac{W_H}{H} = \frac{2}{\pi}\left\{E - 1 - \ln(2E-1) + \frac{\varepsilon_r - 1}{2\varepsilon_r}\left[\ln(E-1) + 0.39 - \frac{0.61}{\varepsilon_r}\right]\right\}$$

where

$$A = \frac{Z_L}{60}\sqrt{\frac{\varepsilon_r + 1}{2}} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1}\left(0.23 + \frac{0.11}{\varepsilon_r}\right) \quad (17a)$$

$$E = \frac{377\pi}{2Z_H\sqrt{\varepsilon_r}} \quad (17b)$$

TABLE II

ARITHMETIC VALUES FOR DIFFERENT IMPEDANCE'S OF A MICROSTRIP LINE

$Z_0(\Omega)$	ε_r	A	E	H(mm)	W(mm)	ε_{eff}
10	3.48	0.394		1.6	95.39	3.41
20	3.48	0.644		1.6	15.01	3.06
50	3.48	1.392	6.349	1.6	3.63	2.73
120	3.48		2.645	1.6	0.45	2.43
150	3.48		2.116	1.6	0.06 (imp)	2.33

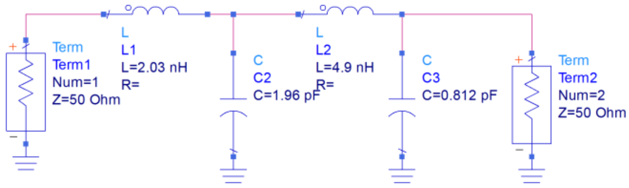

Fig. 12. Lumped elements for the 4th order Butterworth filter with cut-off frequency 3 GHz.

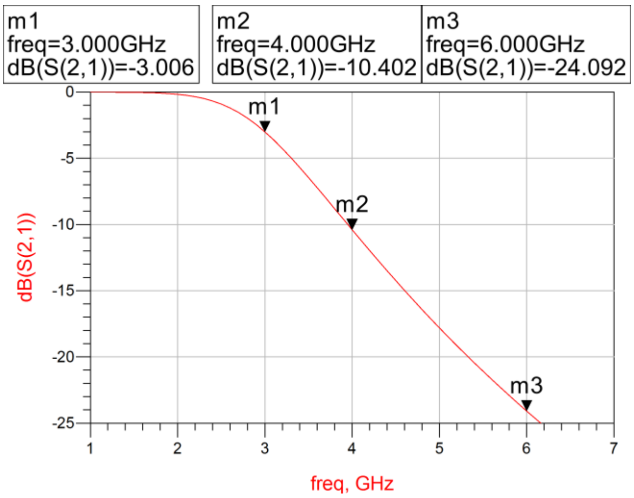

Fig. 13. The response of the circuit in figure 12.

TABLE III
DESIGN PARAMETERS FOR THE 4TH ORDER T-TOPOLOGY MICROSTRIP FILTER

g_i	L or C	Z_H or Z_L	ε_{eff}	W(mm)	l (mm)
0.765	2.03nH	120	2.43	0.45	3.256
1.848	1.96pF	20	3.06	15.01	6.723
1.848	4.9nH	120	2.43	0.45	7.858
0.765	0.81pF	20	3.06	15.01	2.778

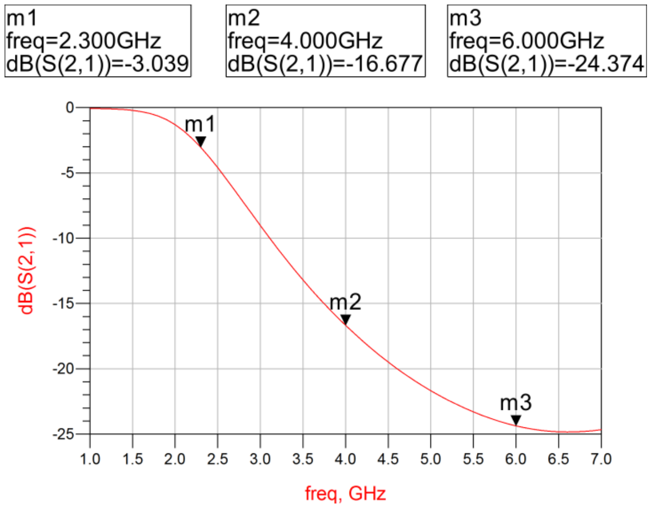

Fig. 15. The response of the circuit in figure 14.

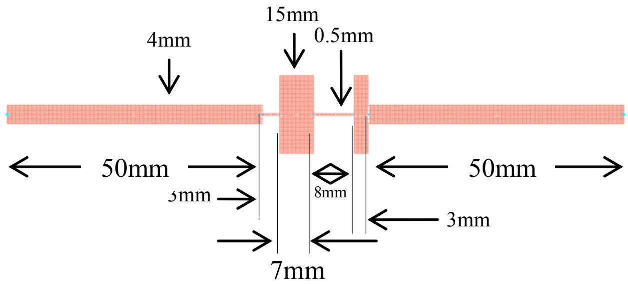

Fig. 16. Approximate dimensions of the microwave filter.

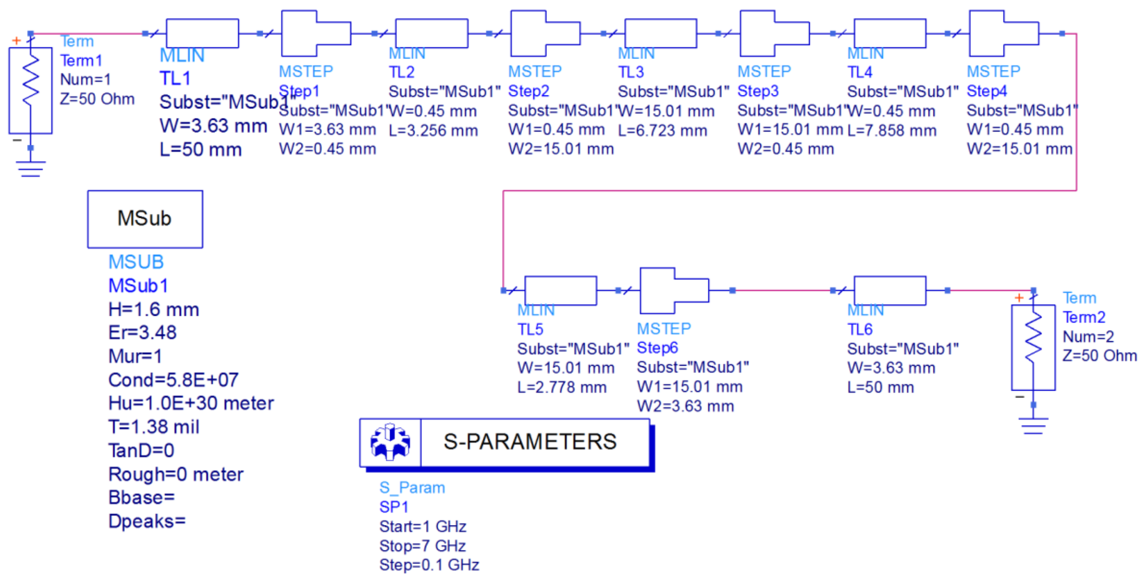

Fig. 14. The microstrip low pass filter design using step impedances.

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For the substrate, we choose a typical Rogers 4350 printed circuit board with $\varepsilon_r = 3.48$ and H = 1.6mm. For Z_L=10Ω, the results are A=0.3943, W=95mm and ε_{eff}=3.41. For Z_H=150Ω, using equations (15), (16) and (17b) the results are B=2.12, W=0.06mm and ε_{eff}=2.33. Summarizing the above, the next table is extracted.

Combining equations (5) and (14), the following equations will help in designing the 4th order microstrip filter which is shown in fig. 12. Table III summarize all the designing parameters.

$$l_L = l_{HIGH} = \frac{c \cdot L}{Z_H \cdot \sqrt{\varepsilon_{eff}}} \qquad (18a)$$

$$l_C = l_{LOW} = \frac{c \cdot C \cdot Z_L}{\sqrt{\varepsilon_{eff}}} \qquad (18b)$$

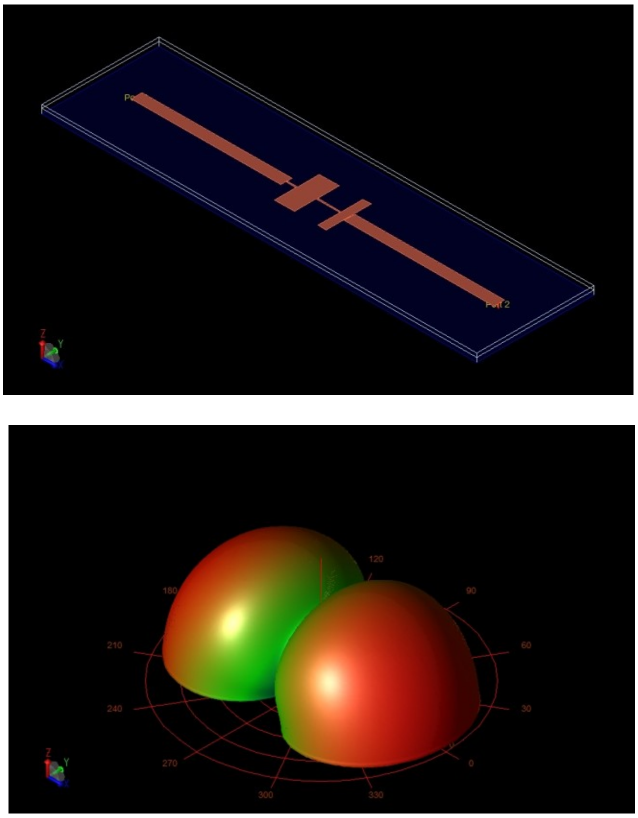

Fig. 17. The layout of the microstrip low pass filter at 3 GHz and its radiation pattern.

IV. Conclusion

The procedure followed in this paper for designing a microstrip filter is to find the element values for filters with an arbitrary number of stages and arbitrary topology. For a normalized low-pass design, where the source impedance is 1Ω and the cutoff frequency is ω_C = 1 rad/s, however, the element values for the ladder-type circuits of Figures 1 and 2 can be tabulated by using simple equations. Since a typical low-pass filter consists of alternating series inductors and shunt capacitors in a ladder configuration, we could implement the filter by using alternating high and low characteristic impedance section transmission lines. Using equations (14), we can find the transmission lines length at the cutoff frequency ω_C in relation with inductance and capacitance mentioned. Equations 15 and 16 give the ratio W/H for each transmission line. What is more, equations 18 give an easy way formula to find the dimensions of the filter, when the effective dielectric constant for the microstrip lines is given by equation (15).

The calculated amplitude response of the filter, with and without losses is presented in figures 15 and 13 respectively. The effect of loss is to increase the passband attenuation to about 5 dBs moving the cut-off frequency at 2.4 GHz. The lumped-element filter gives more attenuation at higher frequencies. This is because the stepped-impedance filter elements depart significantly from the lumped-element values at higher frequencies. The stepped- impedance filter may have other passbands at higher frequencies, but the response will not be perfectly periodic because there are non commensurable lines.

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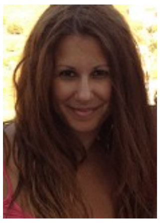

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