

Chapter 7

RF Filters

Introduction

Filters used for RF signals are nearly always Bandpass filters using coupled resonator design techniques. The basic properties of the most common types of filters are:

LC bandpass filters.

Typical unloaded Q values are 60-150, resulting in a large % Bandwidth filter with a high insertion loss. Frequency range 455 kHz to 2 GHz.

Helical Filters

Typical unloaded Q values > 1000 . The Q value can be increased, by increasing the size of the resonator cavity. Frequency range 30 MHz – 1 GHz.

Coaxial Filters

Typical unloaded Q value > 5000 . Frequency range 80 MHz – 2 GHz. Using a Ceramic dielectric results in a small size for mobile radio use. Air dielectric coaxial filters are used for high power diplexers in Radio and TV transmitters.

Microstrip and Stripline filters

Made using lines on PCB. Typical unloaded Q value 250. Wide % bandwidth.

Transmission Line Filters

Designs based on Short or open circuited stubs. Wide Bandwidth.

Interdigital Filters

High Q and $< 10\%$ bandwidth.

Acoustic filters:

Ceramic filters

Commonly used frequencies only. 455 kHz, 10.7 MHz IF filters, 800 MHz and 1.9 GHz diplexers for cellular phones.

Crystal Filters

Very High Q values and very low bandwidths. Typical 6 kHz Bandwidth at 10.7 MHz. Used for SSB filters. High cost compared to other filters.

SAW filters

Commercial Frequencies to 1 GHz. 10.7 MHz, 70 MHz. High insertion loss, can control amplitude response and group delay independently.

Several of the above filter types are discussed in detail in these notes.

Filter Design Revision

Most RF filter designs start with a normalised lowpass filter design and that filter has the required specifications in terms of filter order required, amplitude and group delay performance. This normalised lowpass filter is then transformed into the appropriate bandpass filter. There are two starting points:

- 1 Normalised LC filter tables or equations
- 2 K and Q filter tables.

These two are related and one can be derived from the other. For some filter types, like Butterworth filters, simple equations can be used for the normalised lowpass filter prototype. For RF designs K and Q tables are however better, since they include the effect of finite Q values in the resonators used. A good book for filter tables is A. I. Zverev, Handbook of Filter Synthesis, Wiley, 1967 and more recently from Artec House. This book is still in print.

Filter Tables

n	q0	I. L.	q 1	q n	k 12	k 23	k 34	k 45
	INF.	0.000	1.4142	1.4142	0.7071			
	14.142	0.915	1.4142	1.4142	0.7071			
	7.071	1.938	1.4142	1.4142	0.7071			
2	4.714	3.098	1.4142	1.4142	0.7071			
	3.536	4.437	1.4142	1.4142	0.7071			
	2.829	6.021	1.4142	1.4142	0.7071			
	2.357	7.959	1.4142	1.4142	0.7071			
	2.020	10.459	1.4142	1.4142	0.7071			
	INF.	0.000	1.0000	1.0000	0.7071	0.7071		
	20.000	0.958	0.8041	1.4156	0.7687	0.6582		
	10.000	2.052	0.8007	1.5359	0.7388	0.6716		
	6.667	3.300	0.9097	1.6301	0.7005	0.6879		
3	5.000	4.742	0.8226	1.7115	0.6567	0.7060		
	4.000	6.443	0.8406	1.7844	0.6077	0.7256		
	3.333	8.512	0.8625	1.8497	0.5524	0.7470		
	2.857	11.157	0.8884	1.9068	0.4883	0.7706		
	INF.	0.000	0.7654	0.7654	0.8409	0.5412	0.8409	
	26.131	1.002	0.5376	1.4782	1.0927	0.5668	0.6670	
	13.066	2.162	0.5355	1.6875	1.0745	0.5546	0.6805	
4	8.710	3.489	0.5417	1.8605	1.0411	0.5373	0.6992	
	6.533	5.020	0.5521	2.0170	1.0004	0.5161	0.7207	
	5.226	6.822	0.5656	2.1621	0.9547	0.4906	0.7444	
	4.355	9.003	0.5819	2.2961	0.9051	0.4592	0.7706	
	3.733	11.772	0.6012	2.4159	0.8518	0.4192	0.7998	

	INF.	0.000	0.6180	0.6180	1.0000	0.5559	0.5559	1.0000
	32.361	1.045	0.4001	1.5527	1.4542	0.6946	0.5285	0.6750
	32.361	1.045	0.5662	0.7261	1.0947	0.5636	0.5800	0.8106
	16.180	2.263	0.3990	1.8372	1.4414	0.6886	0.5200	0.6874
	16.180	2.263	0.5777	0.7577	1.0711	0.5408	0.6160	0.7452
	10.787	3.657	0.4036	2.0825	1.4088	0.6750	0.5080	0.7066
5	10.787	3.657	0.5927	0.7869	1.0408	0.5144	0.6520	0.6860
	8.090	5.265	0.4111	2.3118	1.3670	0.6576	0.4927	0.7290
	8.090	5.265	0.6100	0.8157	1.0075	0.4844	0.6887	0.6278
	6.472	7.151	0.4206	2.5307	1.3195	0.6374	0.4732	0.7542
	6.472	7.151	0.6293	0.8449	0.9722	0.4501	0.7267	0.5681
	5.393	9.425	0.4321	2.7375	1.2675	0.6149	0.4479	0.7821
	5.393	9.425	0.6508	0.8748	0.9355	0.4103	0.7663	0.5048

Table 1. Butterworth Response, K And Q Value filter table, From A. I. Zverev. "Handbook of Filter Synthesis", Wiley, pp 341.

Butterworth Filters

Butterworth filters have a maximally flat amplitude response. That means all the derivatives of the amplitude with frequency are zero at DC. The Butterworth response is a good compromise between attenuation characteristic and group delay.

For a Butterworth Lowpass filter of n^{th} order, with a cut off frequency of f_c and a source and load resistance of R , the filter component values are given by:

$$C_i = \frac{1}{\pi f_c R} \sin \frac{(2i-1)\pi}{2n} \quad i=1,3,5,\dots \quad \text{Eqn. 1}$$

$$L_i = \frac{R}{\pi f_c} \sin \frac{(2i-1)\pi}{2n} \quad i=2,4,6,\dots \quad \text{Eqn. 2}$$

The first component ($i=1$) is a shunt capacitor, the second ($i=2$) component is a series inductor etc. The normalised q values and coupling coefficients are related to the normalised lowpass impedances for a filter, using the following equations:

$$q_0 = Z_1 \quad q_n = Z_n \quad k_{ij} = \frac{1}{\sqrt{Z_i Z_j}} \quad \text{Eqn. 3}$$

Where Z_i is the normalised impedance as obtained from LC filter tables or from equations 1 and 2. Applying equation 3 to equations 1 and 2 gives:

$$q_0 = q_n = 2 \sin \frac{(2-1)\pi}{2n} = 2 \sin \frac{\pi}{2n} \quad \text{Eqn 4}$$

$$k_{ij} = \frac{1}{2 \sqrt{\sin \frac{(2i-1)\pi}{2n} \sin \frac{(2i+1)\pi}{2n}}} \quad i=1,3,5,\dots \quad \text{Eqn. 5}$$

These equations will then allow the k and q values for $IL = 0$ and $q_0 = \infty$ as shown in table 1 to be calculated.

Figure 1 shows the resulting circuit for a 7th order Lowpass filter with a 500 MHz cut-off frequency. The frequency response is shown in figure 2. The group delay of Butterworth filters is reasonably flat but has a rise near the cut off frequency. The step response of these filters exhibits some ringing, which degrades its use for data communications.

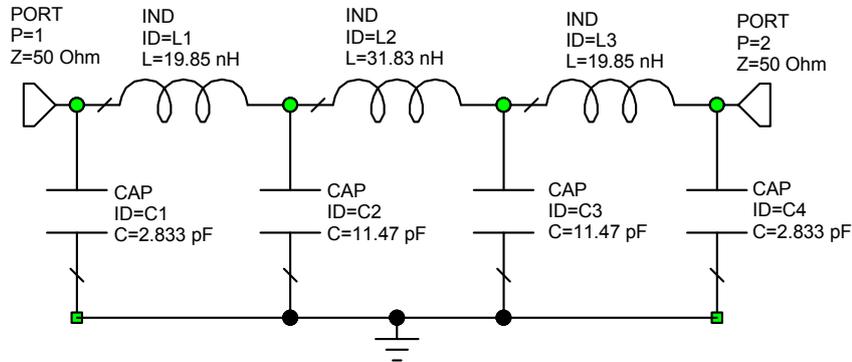


Figure 1. Butterworth Lowpass Filter.

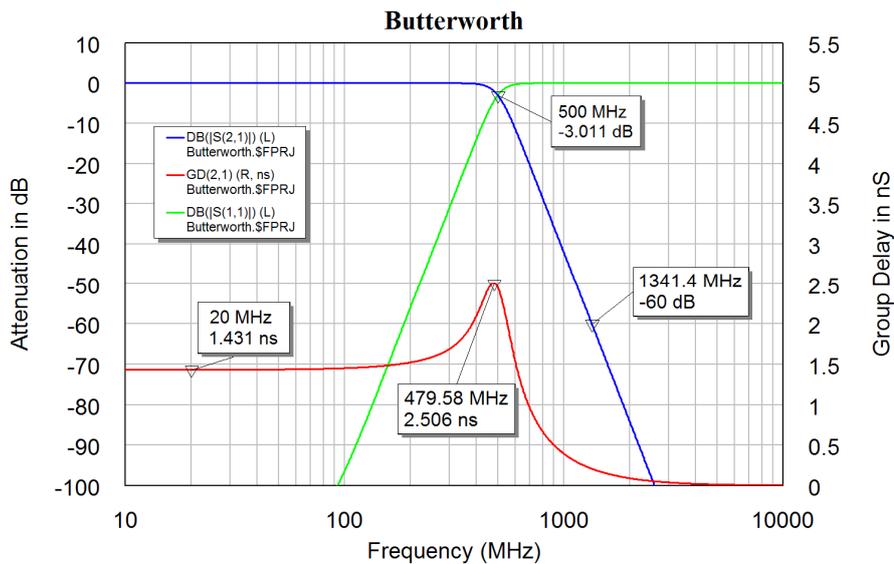


Figure 2. Butterworth Frequency Response.

Bessel filter

The Bessel filter has a maximally flat group delay and does not have any ringing or overshoot in its step response. As a result the Bessel filter type is very suitable for data communication systems.

Simple equations like equations 1 and 2 do not exist for Bessel filters. As a result the normalised filter element values have to be obtained from filter tables. Many filter tables for Bessel filters are normalised for a group delay of one second, rather than a 3 dB cut-off frequency. The Bessel filter in figure 3 has a cut-off frequency of 500 MHz so that accurate comparisons between the filter types can be made.

The Bessel filter has a less sharp roll-off of the attenuation characteristic, compared with the Butterworth filter, but has a flatter group delay as can be seen by comparing figure 2 and 4. For the Butterworth filter the 60 dB attenuation level occurs at 1.3 GHz while for the Bessel filter it occurs at 2.4 GHz. The group delay for the Bessel filter is less than that for the Butterworth filter, so that the signal appears sooner at the filter output.

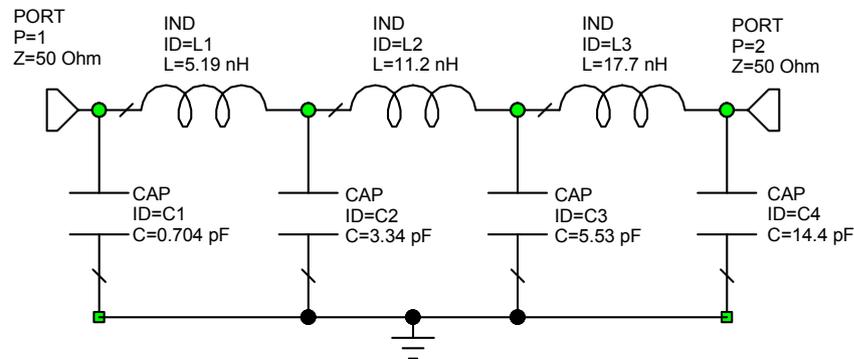


Figure 3. Bessel Lowpass Filter.

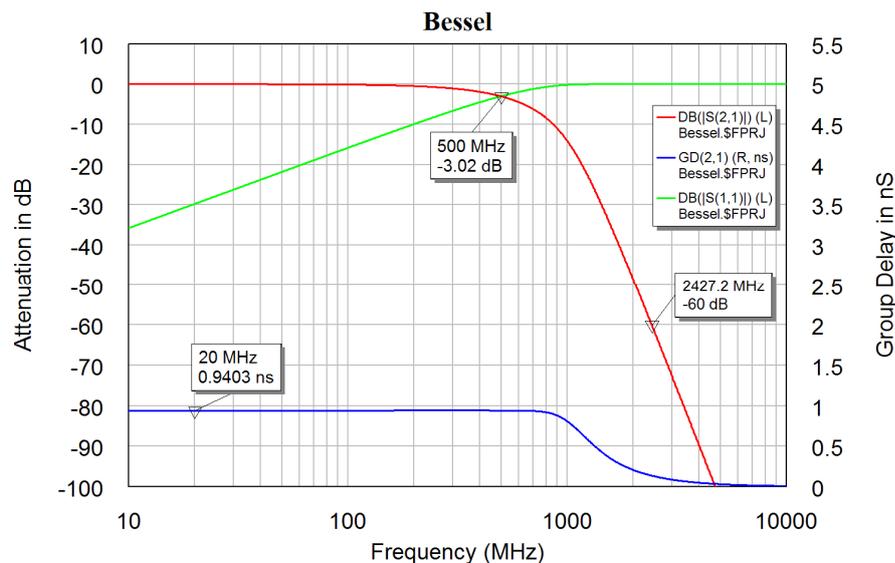


Figure 4. Bessel Lowpass Filter Frequency Response.

Chebyshev Filter

The pole positions for the Chebyshev filter lie on an ellipse, hence the Chebyshev filter is also called an elliptical filter. The Chebyshev filter has ripples in the passband and the value of the ripples is a parameter that is selected as part of the filter design. The filter of figure 5 has a 0.1 dB passband ripple as shown on figure 6. The inductors and capacitors for the Chebyshev filter are larger than those of the Bessel or Butterworth filters. The Chebyshev filter has a very sharp attenuation response in the stop band, with a -60 dB attenuation at 967 MHz, compared with 1.3 GHz for the same attenuation for the Butterworth filter. Increasing the passband ripple will increase the passband attenuation. The group delay for the Chebyshev filter is larger than that of the Bessel and Butterworth filters. As a result the Scale on figure 6 is different from figure 2 and 4.

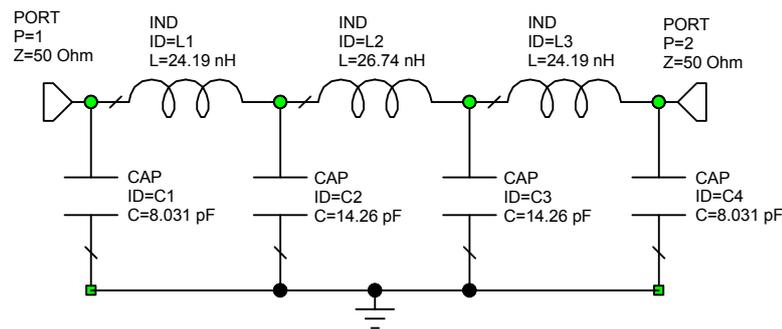


Figure 5. Chebyshev Lowpass Filter.

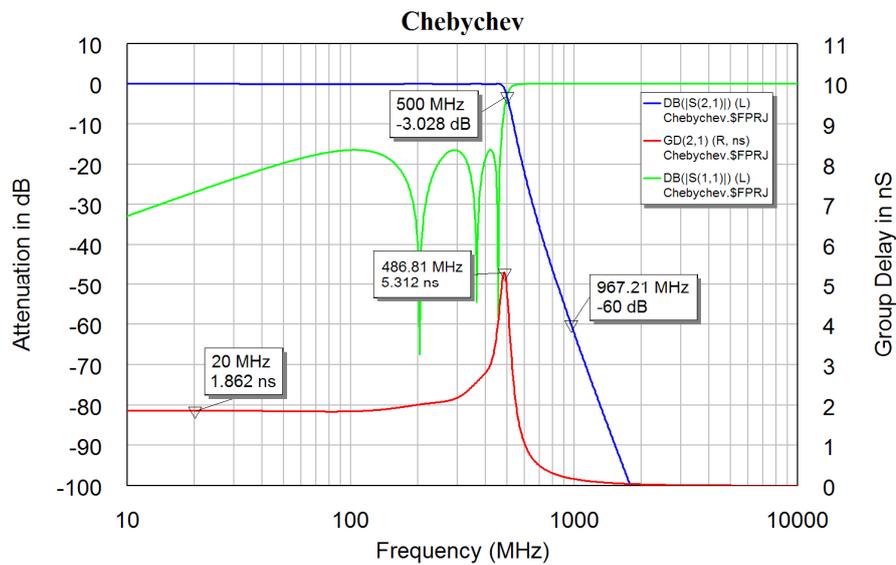


Figure 6. Chebyshev Lowpass Filter Frequency Response.

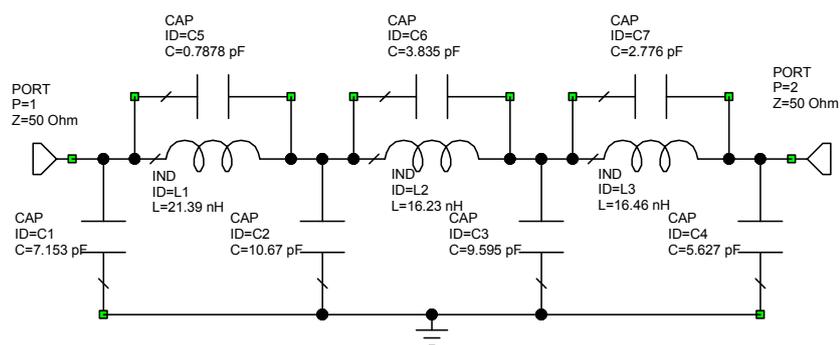


Figure 7. Cauer-Chebyshev Lowpass Filter.

A Cauer-Chebyshev filter has capacitors in parallel with the inductors to cause zeros in the stopband as shown in figure 7. The zeros are placed such that the minimum stopband attenuation values are the same as shown in figure 8. This gives an even higher stopband attenuation, with a 60 dB attenuation at 628 MHz compared with 1.3 GHz for the Butterworth filter. The Cauer-Chebyshev filter has an even bigger variation

in group delay than the Chebyshev filter and will thus have very poor transient response. The group delay of the Chebyshev and Cauer-Chebyshev filters are poor and these filters should not be used for data communications or applications where the group delay is important.

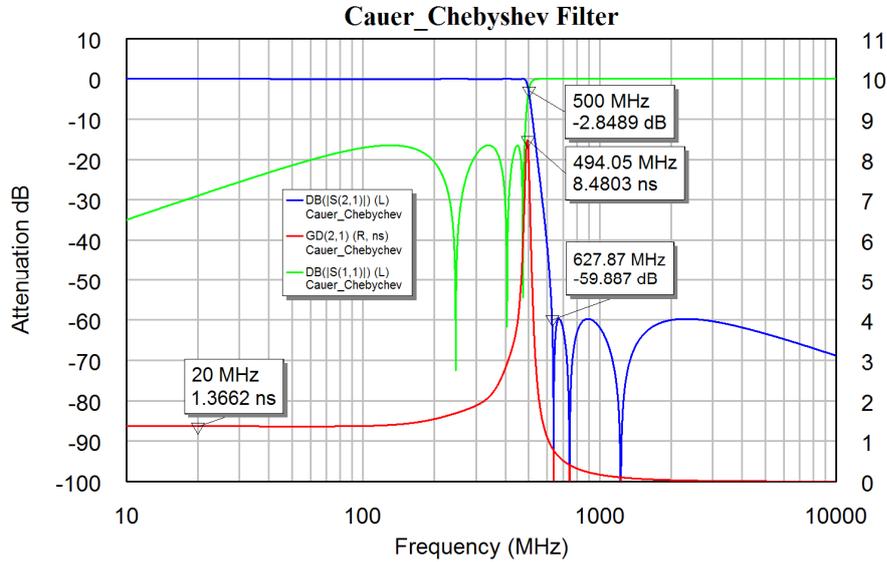


Figure 8. Cauer-Chebyshev Lowpass Filter Frequency Response.

Figure 9 shows the passband response of the four different filter types. The 0.1 dB passband ripple of the Cauer Chebyshev and Chebyshev filters can clearly be seen. The Bessel filter has a significant attenuation for frequencies above 100 MHz.

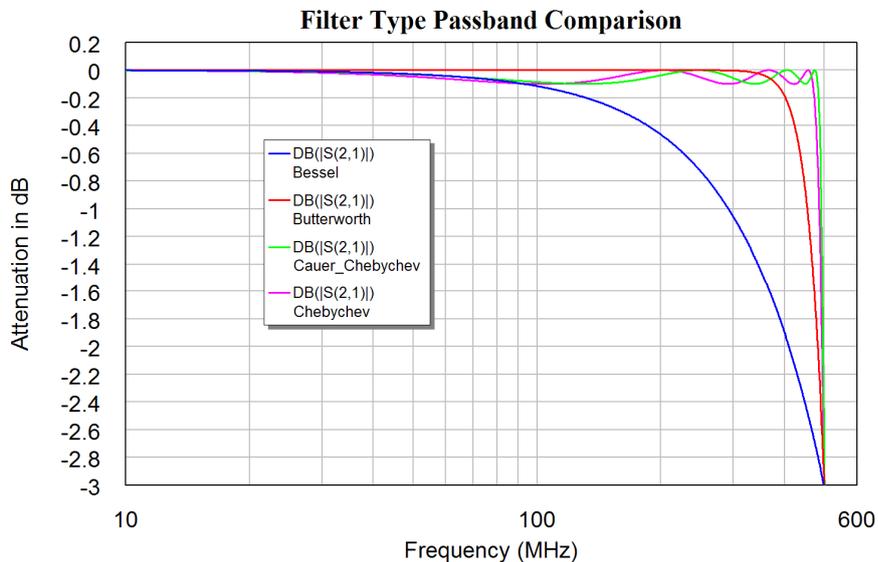


Figure 9. Passband Frequency Response of the filters of Figures 2, 4, 6 and 8.

Both the Chebyshev and Cauer-Chebyshev filters have poles on an ellipse and are elliptical filters. As a result it is better not to specify a filter as an elliptic filter since it does not fully specify the filter type.

From filter tables, which include losses due to the finite Q values of the filter elements, it can be determined that for a given Q value for the components, the Chebyshev filter has a higher passband insertion loss than a Butterworth filter and a Bessel filter has a lower insertion loss than a Butterworth filter. The same applies for Bandpass filters, so that for the same unloaded Q of the resonators used in coupled resonator RF filters, the Bessel filter will have the lowest insertion loss and the Chebyshev filter will have the highest insertion loss.

The selection of the correct filter type is critical to ensure that the filter used satisfies both the transient response requirements and the passband insertion loss and stopband attenuation requirements for the filter. In digital communication applications, the Bit Error Rate (BER) after demodulating is the critical parameter. To reject signals in an adjacent channel intuition suggests the use of a Chebyshev filter. However, computer simulations carried out at JCU for a 3rd Generation Mobile Radio system, have shown that the lowest demodulated BER when there is a large unwanted signal in an adjacent channel is obtained with a Bessel type filter response for the Channel filters.

RF Lowpass Filter design

Lowpass filters at RF and microwave frequencies can be designed and constructed using conventional LC filter design principles. Care should be taken to ensure that:

The impedance of the leads are kept as small as possible. In particular the impedance of the leads on capacitors connecting the capacitors to ground, or the length of any PCB layout track connecting the components to the ground-plane must be kept as small as possible, since it's inductance will effect the frequency response of the filter.

Components must be used that have self resonant frequencies well above the cut off frequency of the filter if a lowpass filter is used and well above any useable frequency if a high pass filter is used. Most commonly used capacitors have self resonant frequencies well below 1 GHz.

It is feasible to design conventional LC lowpass filters with cut off frequencies up to 1 GHz. When the frequency becomes too high for LC components to be used, inductors can be replaced with thin transmission lines of the length required to give the desired inductance and capacitors can be replaced by wide transmission lines.

As an example consider the design of a 7th order Butterworth lowpass filter with a cut off frequency of 2 GHz. From the equations 1 and 2 for Butterworth filters or from tables, by denormalising and using a cut off of 2 GHz and input impedance of 50 Ω , the component values shown in table 2 are obtained for the filter.

L C	Value	Line Length	Line Width
L1	1.771 nH	3.6 mm	0.5 mm
C2	1.985 pF	4.2 mm	10 mm
L3	7.170 nH	13.5 mm	0.5 mm
C4	3.183 pF	6.6 mm	10 mm
L5	7.170 nH	13.5 mm	0.5 mm
C6	1.985 pF	4.2 mm	10 mm
L7	1.771 nH	3.6 mm	0.5 mm

Table 2. LC filter values and corresponding microstrip line size.

For Rogers RO4003 substrate with $H = 0.83$ mm and $T=35$ microns, the width for a 50Ω transmission line is 1.9 mm. the thinnest line that is desirable is 0.5 mm and the thickest is 10 mm. That corresponds to impedances of 96.5Ω and 14.3Ω respectively. One needs to find out what length of transmission lines corresponds to the inductance required for the filter. In Microwave Office it is easy to compare the impedance of an inductor with one of a transmission line, as shown in figure 10.

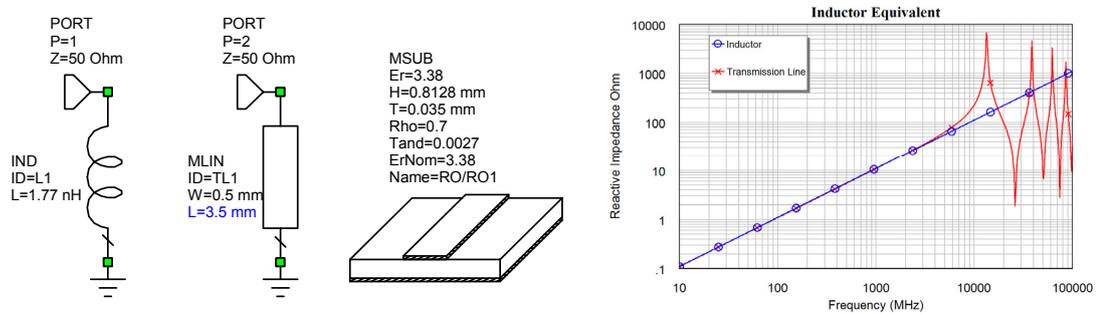


Figure 10. Replacement of Inductors using Microstrip Transmission lines

The length of the transmission line is tuned to ensure that at low frequency both have the same impedance. It can be seen that the short section of transmission line accurately represents the inductor for frequencies up to 10 GHz. This process is repeated for all the inductances. Note for larger inductances, at higher frequencies the transmission line will be a quarter wavelength resulting in an infinite impedance or half a wavelength, resulting in a very small impedance. For lowpass filters the length of the transmission line should be less than one quarter wavelength for all frequencies of interest.

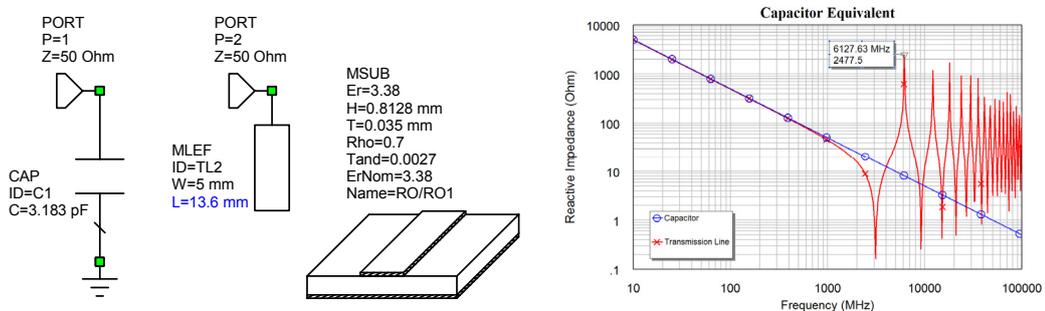


Figure 11. Replacement of capacitors using series Microstrip Transmission lines.

The Capacitors can be replaced with open circuited transmission lines as shown in figure 11. The resulting line sections have been placed in table 2 and it can be seen that some line sections are wider than they are long, which will cause some errors in the simulation as the actual fields will be different from what is assumed in the simulation of the transmission lines.

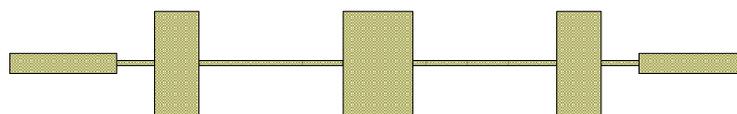


Figure 12. PCB Layout of the 2 GHz Lowpass filter.

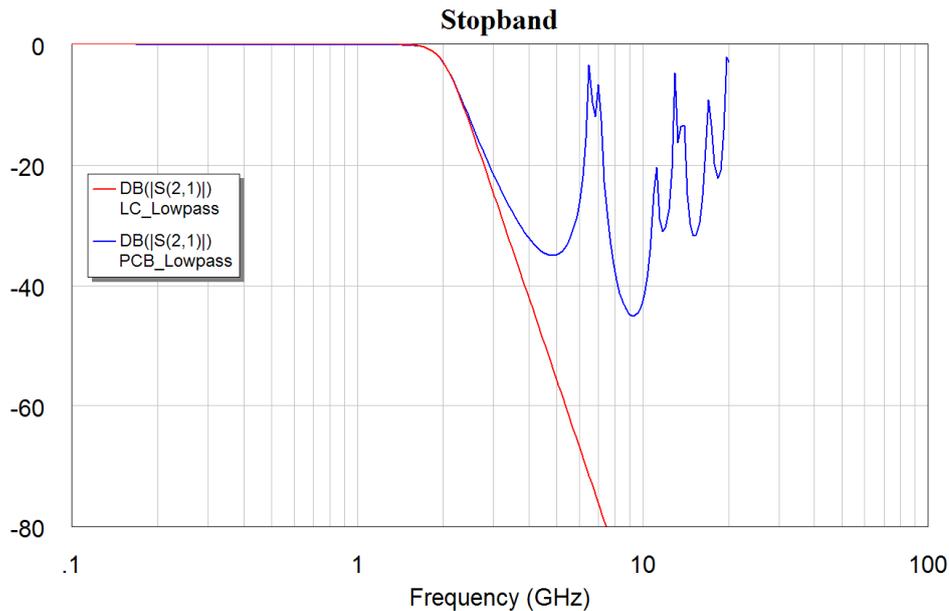


Figure 13. Frequency Response of the LC filter and the corresponding Microstrip design.

Figure 13 compares the frequency response of the ideal LC filter (red) with the PCB filter of figure 12 (blue) and it shows that the PCB filter behaves well in the passband, but that the stopband performance is not very good. That is due to the elements becoming comparable to a quarter or half wavelength long at certain frequencies. By replacing the capacitors with open circuited shunt transmission lines as shown in figure 14, it is possible to obtain a better stopband performance. In addition by making the lengths of the stubs connected at each cross-point slightly different the effect of the transmission lines being one quarter or one half wavelength is minimised. In figure 13, the transmission line has a high impedance of 2.48 k Ω at 2.48 GHz and in figure 14, by using stubs, the impedance is 1.82 k Ω at 9.2 GHz.

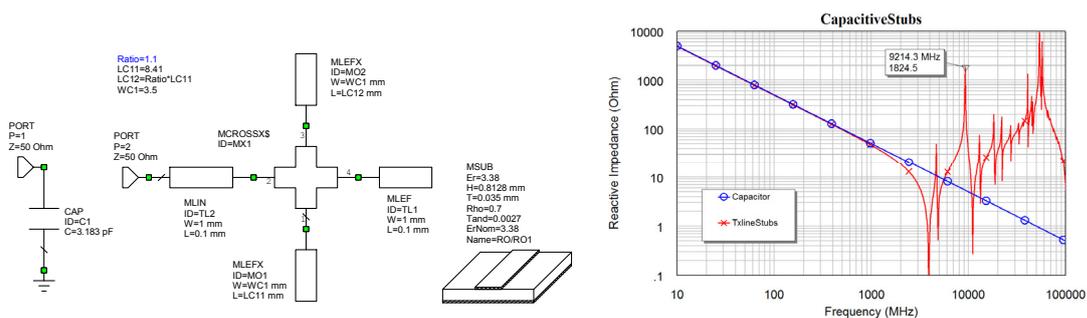


Figure 14. Replacement of capacitors using shunt Microstrip Transmission lines.

By using radial stubs as shown in figure 15, the resonance can be extended even further to give an peak impedance of 1.59 k Ω at a frequency of 12.5 GHz. The shape of the radial stub is shown in figure 15. The angle for that stub is set to 60 degrees. A wider angle stub will push the high impedance frequency even further. It is possible in a filter to vary the angles to ensure that the resonances do not coincide.

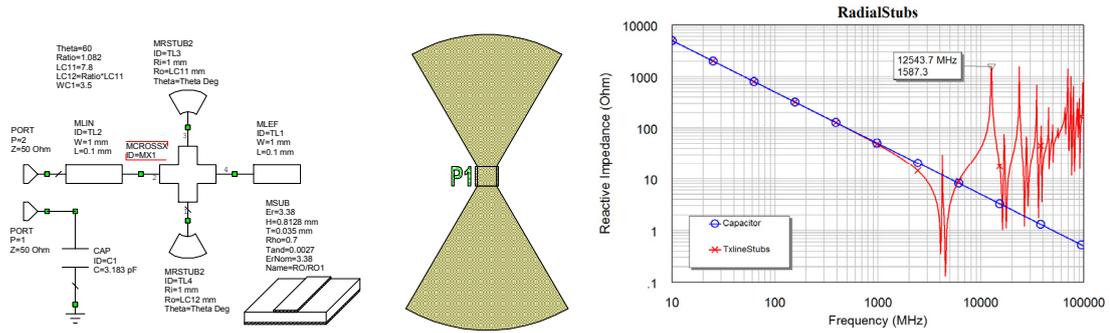


Figure 15. Replacement of capacitors using shunt Radial Stub Microstrip Transmission lines.

The transmission line replacement for all the elements in the desired lowpass filter can be calculated in a similar manner. Once the circuit schematic for the whole filter is complete, the lengths and width of the stubs and the lengths of the inductive sections are optimised to ensure the filter meets the passband and the stopband specifications. As part of this, the width to length ratio of the stubs can be changed to increase the attenuation at particular frequencies in the stopband. The resulting circuit diagram is shown in figure 16.

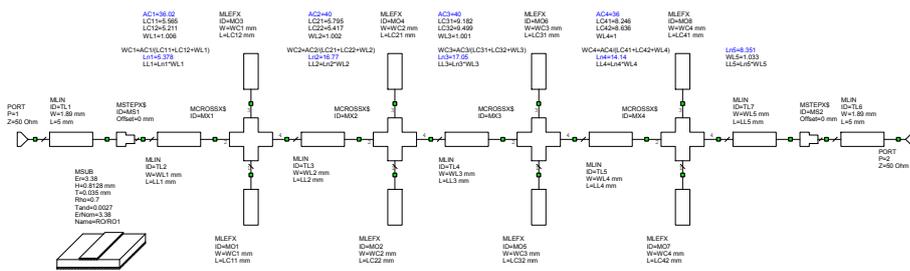


Figure 16. Circuit Schematic for the 2 GHz Lowpass Filter.

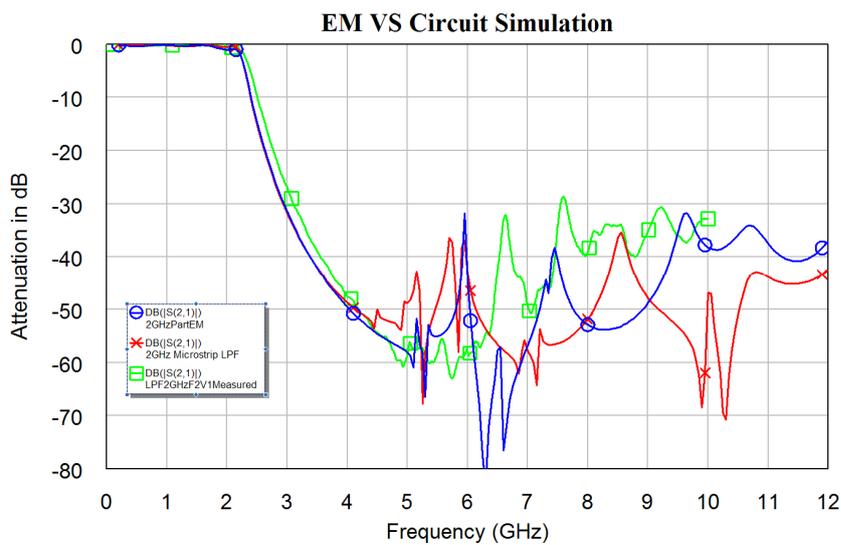


Figure 17. Frequency Response of the 2 GHz Lowpass Filter.

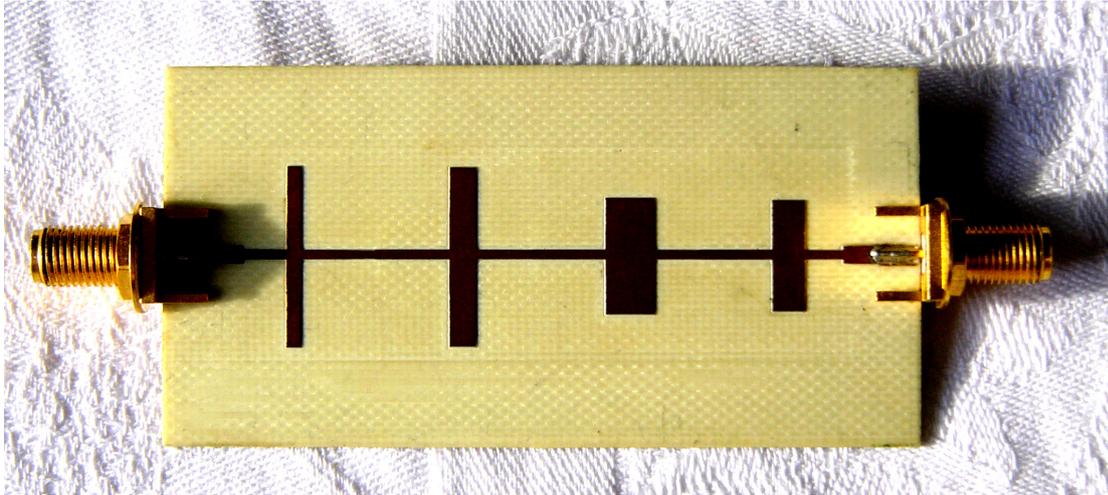


Figure 18. 2 GHz Lowpass filter hardware.

The behaviour of the resulting filter is also calculated using a full Electro-Magnetic simulation of the capacitive stub lines. Since the transmission lines replacing the inductors can be modelled accurately using a non-EM simulation, only a partial EM simulation, for modelling the capacitors, is required. The frequency response obtained from the circuit simulation, the partial EM simulation and the measured performance is shown in figure 17. It can be seen that there is a good agreement for frequencies below 4 GHz but that the stopband behaviour is still not adequate. The filter hardware realisation is shown in figure 18.

The attenuation in the stopband can be improved by adding open circuited stubs which create a transmission zero in the stopband at the frequencies where the filter without the stubs has an inadequate attenuation.

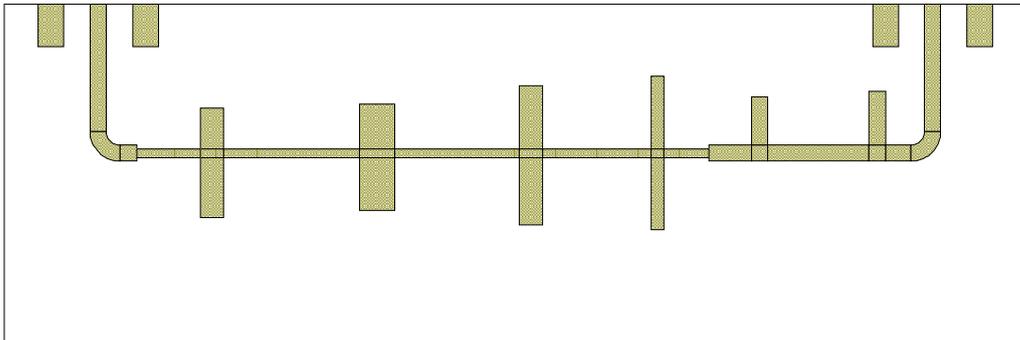


Figure 19. 2 GHz Lowpass Filter with Stubs Added.

In this application, a good stopband performance to 6 GHz was required. The stubs will affect the passband performance and the circuit elements need to be re-tuned and optimised to restore the passband performance to match that of figure 17. The resulting PCB layout is shown in figure 19.

Figure 20 shows the simulated and measured performance of the resulting filter. The simulated performance has a very good stopband performance. The measured performance has a significant amount of coupling between the input and output, which is due to electromagnetic coupling between the input and output tracks and leads

connecting the filter. The stopband performance was improved by enclosing the filter and placing microwave absorber on the ground-plane on the back of the filter and on the inside lid of the enclosure.

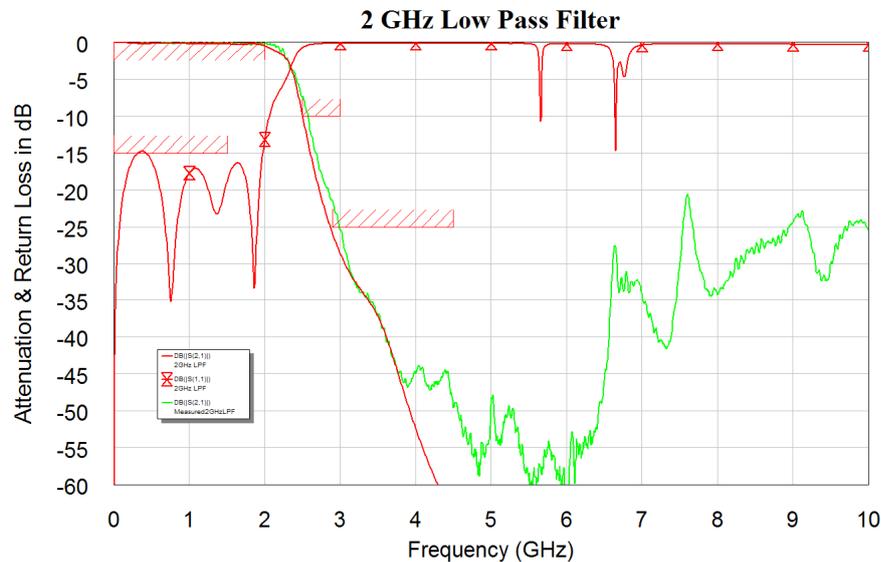


Figure 20. Simulated and Measured Performance of 2 GHz Lowpass Filter with Stubbs.

Bandpass Filter Design

LC Bandpass Filters

To illustrate the design principles, a filter with a centre frequency of 18.48 MHz and a bandwidth of 2 MHz is designed using different techniques. This particular filter was required for a weather satellite data receiver, designed at JCU, to filter the IF data signal prior to demodulation. Since the transmitted signal is a Binary Phase Shift Keyed (BPSK) modulated RF signal containing digital data, a Bessel filter results in the lowest BER after demodulation.

Low-Pass to Band-Pass Transformation

A simple way of designing bandpass LC filters is to use the low-pass to band-pass transformation, described in most filter books. The low-pass to band-pass transformation will often give impractical component values for the bandwidths required for most RF filters. As a result, the low-pass to band-pass transformation is not suitable for most bandpass filter designs.

More practical component values are obtained using coupled resonator designs. The filter resulting from the low-pass to band-pass transformation is shown in figure 21. A 2.0 mH and a 4.4 mH inductor is required. These are a very large inductors and will be difficult to make with acceptable Q values. Figure 22 shows the frequency response of this filter. The low-pass to band-pass transformation has transformed the amplitude response correctly, but the group delay is no longer flat.

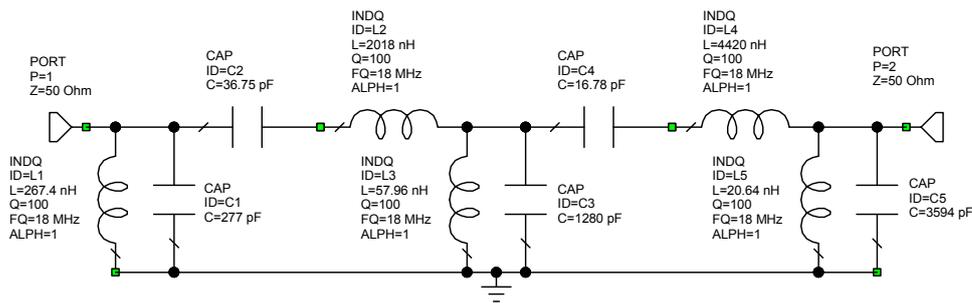


Figure 21. Filter from Lowpass to Bandpass Transformation.

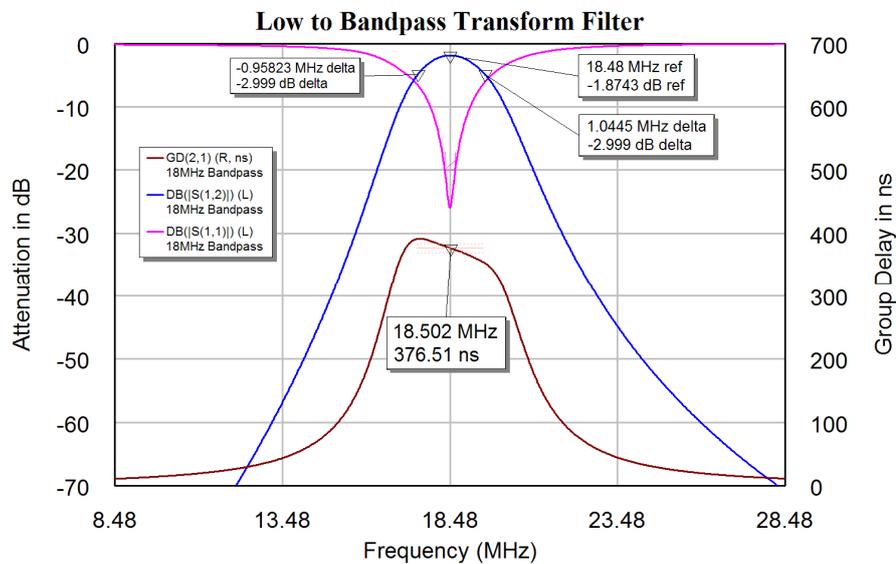


Figure 22. Frequency Response of Filter of Figure 19.

LC Coupled Resonator Filter

In coupled resonator filters, the filter consists of a series of resonators, which are coupled using impedances, coupling loops or apertures. The centre frequency of the filter is determined by the frequency of the resonator and the bandwidth of the filter is determined by the coupling between the resonators. The input impedance of the filter is determined by the coupling into the first resonator and output impedance is determined by the coupling out of the last resonator. For LC coupled resonator filters, the resonators consist of parallel LC circuits. The coupling between LC resonators can be done by connecting capacitors between the resonators, this is called capacitive coupling. The coupling can also be done by connecting inductors between the resonators, this is called inductive coupling. Capacitive coupling is the cheapest as standard capacitors can be used, while the inductors often need to be wound individually. Filter tables, equation 3 or for Butterworth filters, equations 5 give the required coupling factors (K) for the desired filter.

The loading required to achieve the desired first and last resonator Q values, can be obtained from the Q values of the filter tables, from equation 3 or for Butterworth filters, equation 4. In many cases, filters with 50 Ω input and output impedances are

required, while the desired loaded Q values of the resonators require very different impedances. A capacitive impedance transformer can be used to obtain the required impedance transformations.

Equations 3 and 4 only apply for lossless resonators. The K and Q filter tables in Zverev allow the filter to be designed to accommodate the finite Q values of the LC resonators, which result in the filter having an insertion loss. Table 1 shows the K and Q value table from Zverev for Butterworth filters. Denormalising the filter coefficients:

$$K_{ij} = k_{ij} \frac{BW}{F_c} \text{ and } Q_n = q_n \frac{F_c}{BW} \quad \text{Eqn. 6}$$

Zverev shows that the coupling capacitor is given by:

$$C_{ij} = K_{ij} \sqrt{C_i C_j} = k_{ij} \frac{BW}{F_c} \sqrt{C_i C_j} \quad \text{Eqn. 7}$$

where C_i and C_j are the resonator capacitors for the i th and j th resonator and K_{ij} is the denormalised coupling coefficient, k_{ij} is the normalised coupling coefficient from the tables, F_c is the centre frequency and BW is the filter bandwidth. Note the coupling capacitors are to be included in the total capacitance of the resonator so that for the resonator j the resonating capacitance is given by:

$$C_j = C_{res} - C_{ij} - C_{jk} \quad \text{Eqn. 8}$$

where C_{res} is the total resonating capacitance as determined from the resonant frequency and inductance for the resonator. Using coupling inductors results in:

$$L_{ij} = \frac{\sqrt{L_i L_j}}{K_{ij}} = \frac{F_c}{k_{ij} * BW} \sqrt{L_i L_j} \quad \text{Eqn. 9}$$

Where L_i and L_j are the resonator inductors. In a similar manner, the coupling inductors must be considered to be part of the inductance of the resonator, so that the total resonator inductance is made up the inductance of the resonator in parallel with the coupling inductors on either side.

$$\frac{1}{L_j} = \frac{1}{L_{res}} - \frac{1}{L_{ij}} - \frac{1}{L_{jk}} \quad \text{Eqn. 10}$$

where L_{res} is the total resonating inductance as determined from the resonant frequency and capacitance for the resonator. To facilitate the design of these filters, a visual studio .net program, shown in figure 23, has been developed for LC filters with capacitive coupling and is available for download from the AWR web site.

The design process is illustrated using the Bessel filter with a 18.48 MHz centre frequency and 2 MHz bandwidth filter, in the previous design. The filter design requires 4 resonators. Inductors with an inductance of 200 nH and an unloaded Q_0 of 100 are available. In LC resonators, the Q of the capacitors used is generally much higher than the Q of the inductors, so that by using equation 6, the resonator has a normalised q of:

$$q_0 = \frac{BW}{F_c} Q_0 = \frac{2}{18.48} 100 = 10.8. \text{ From the Bessel filter tables in Zverev, the nearest}$$

value corresponds to a q_0 of 10.048 and results in an insertion loss of 1.85 dB. Since the actual Q is slightly higher, one should obtain an actual insertion loss that is slightly less

as is evident from figures 25, 27 and 29. The corresponding K and Q values from this filter table for this value of q_0 are: $q_1 = 0.3891$, $q_4 = 0.5098$, $k_{12} = 1.7120$, $k_{23} = 0.6794$ and $k_{34} = 0.8512$. Entering those values in the LC Bandpass Filter design program gives the filter values as shown in figure 23. Entering those values in a Microwave Office schematic gives the filter shown in figure 24.

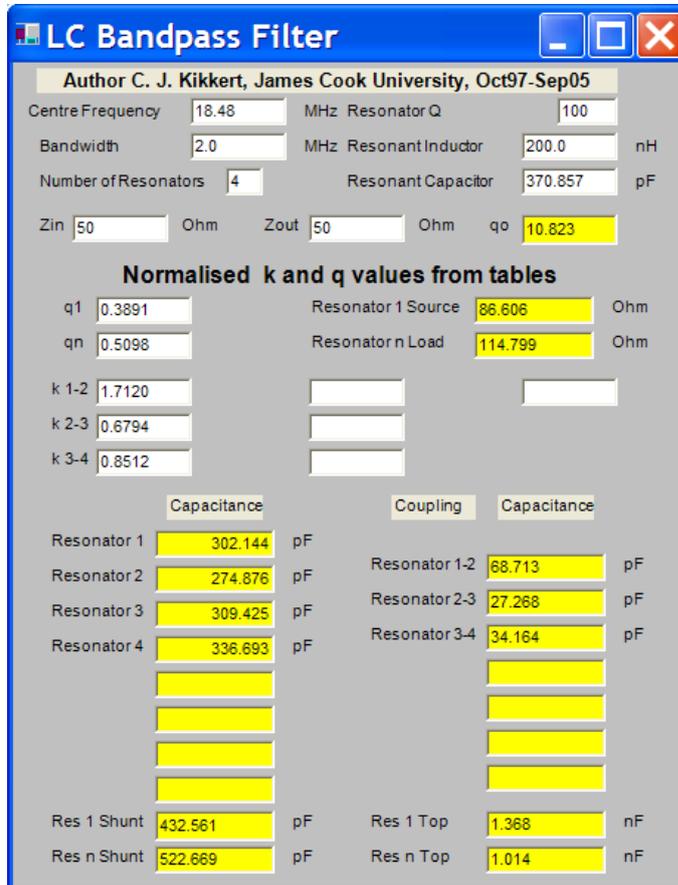


Figure 23. LC Bandpass Filter Design Program for a Bessel Filter.

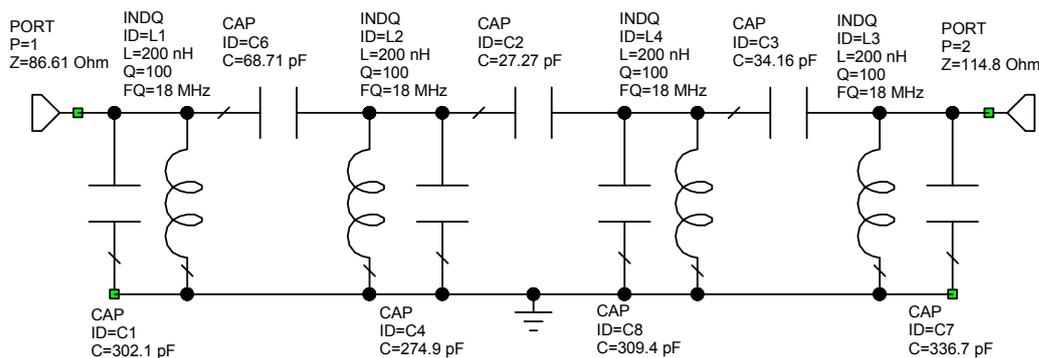


Figure 24. Capacitive Coupled Bessel LC Coupled Resonator Filter.

By comparing figure 22 and figure 25, it can be seen that capacitive coupling reduces the attenuation at high frequencies and increases the attenuation at low frequencies. In addition, the group delay is less flat than for the lowpass to bandpass transformation.

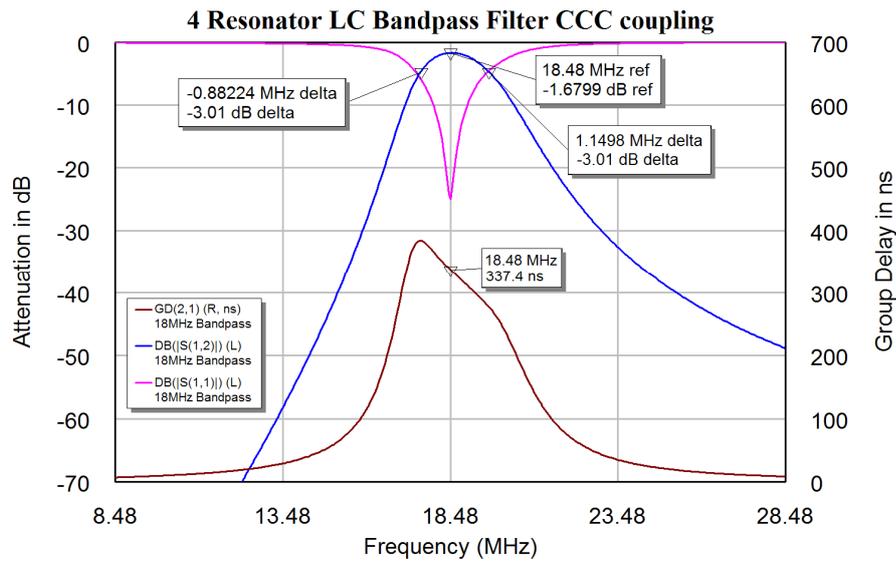


Figure 25. Frequency Response of the Filter of Figure 24.

Instead of using capacitive coupling, inductive coupling can be used. For capacitive coupling the total resonating capacitance is made up of the capacitor of a resonator plus the adjacent coupling capacitors according to equation 8, which is incorporated in the LC Bandpass filter design program of figure 23. For inductive coupling the capacitance of the resonators are simply the resonant capacitance (370.857 pF in the design example) and the inductance of the resonator is adjusted to account for the coupling inductors, resulting in the circuit of figure 26. Figure 27 shows the frequency response of the filter. The group delay for inductive coupling has an opposite slope to the one for capacitive coupling. By using a combination of capacitive and inductive coupling, as shown in figure 28, a flatter group delay is obtained, as shown in figure 29.

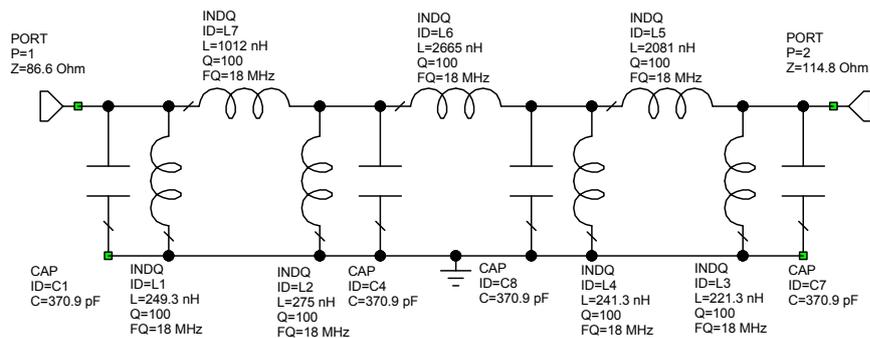


Figure 26. Inductively Coupled LC Resonator Filter

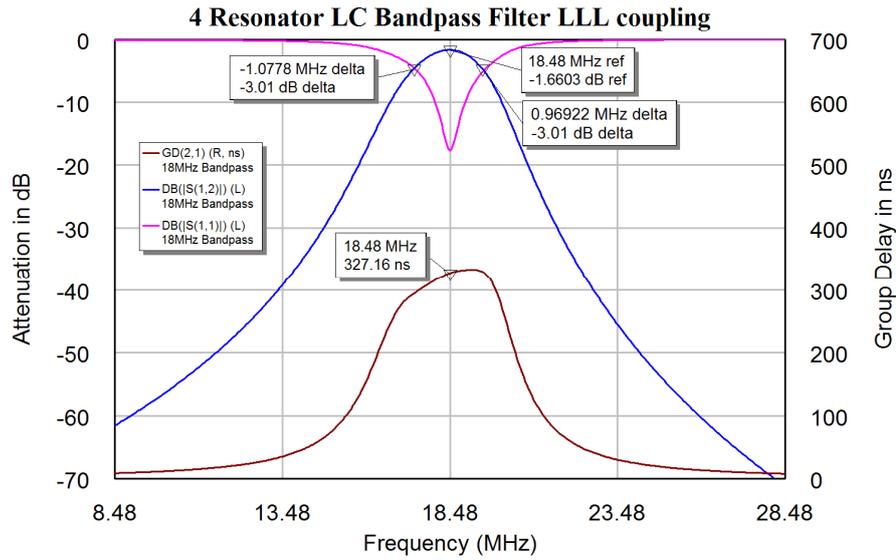


Figure 27. Frequency Response of the Filter of Figure 20.

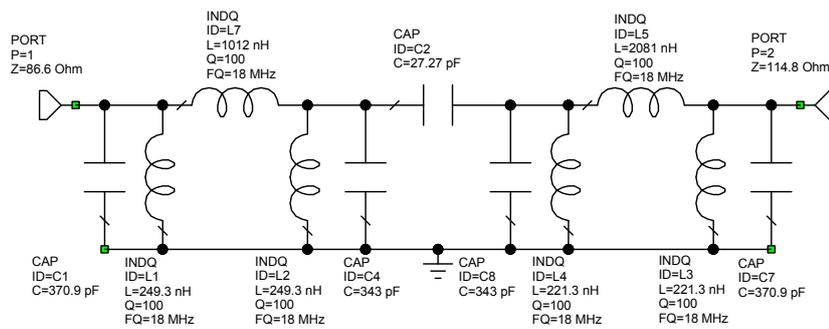


Figure 28. Filter with Both Capacitive and Inductive Coupling.

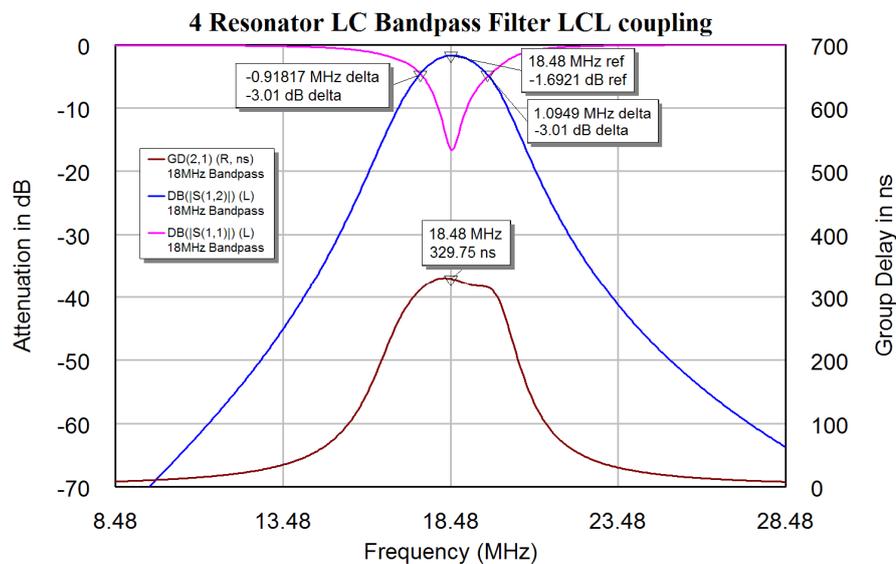


Figure 29. Frequency Response of the Filter of Figure 28.

The input and output impedances of the filter can be changed to 50 Ω, by replacing C1 and C7 with capacitive transformers as shown in figure 30. This will slightly change the frequency response and the group delay. To achieve the best input and output match and a flat group delay, the filter is optimised to meet these specifications. To achieve the lowest insertion loss at the centre frequency. S_{11} and S_{22} are optimised to be better than -30 dB at 18.48 MHz. The group delay is optimised to be as close to 325 ns as possible. Having a constant group delay results in a Bessel type filter and the filter bandwidth is determined by the group delay. The 325 ns group delay results in a 2 MHz bandwidth. The component values after optimisation are shown in figure 30. Figure 31 shows the corresponding frequency response. The resulting filter has a very flat group delay.

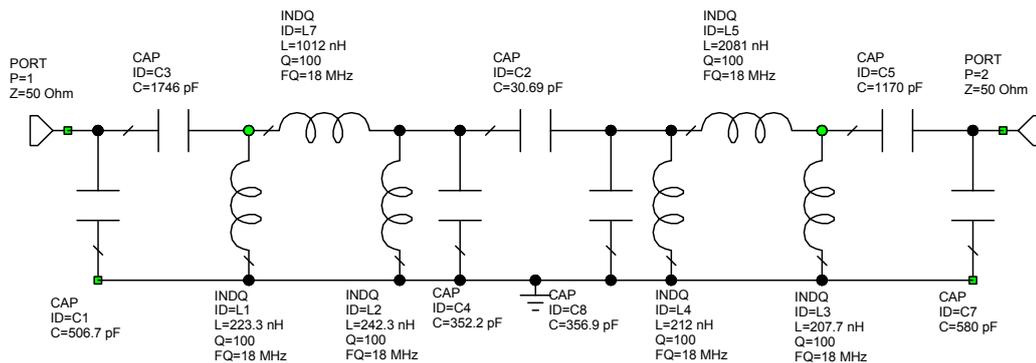


Figure 30. Optimised Coupled Resonator LC Filter.

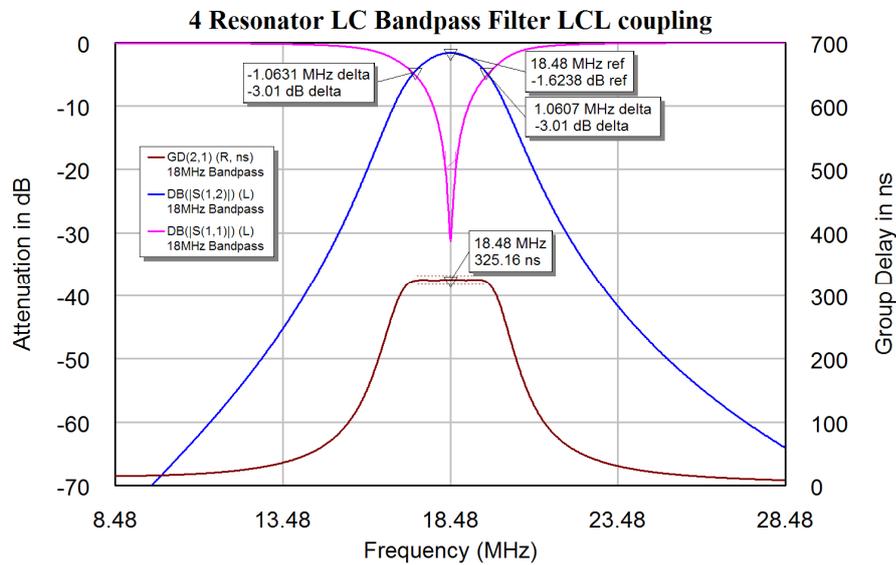


Figure 31. Frequency Response of the Filter of Figure 30.

HF Filters

At higher frequencies tuneable inductors become difficult to make and the inductors can be replaced by microstrip inductors as shown in figure 32. This is a Capacitive Coupled LC resonator filter with a 160 MHz centre frequency and a 27 MHz bandwidth. The vertical transmission lines form the inductors. The parallel plates at the end of the

transmission lines are pads for capacitors. Sufficient space is allowed to have 2 capacitors in parallel to permit the capacitors to be realised with a high precision. The first and last inductors are tapped transmission lines, to allow the input and output to be coupled to the filter at the correct impedances. Figure 33 shows the frequency response of the filter. This filter had to satisfy passband insertion loss constraints and as a result it corresponds to a Butterworth type frequency response and the group delay has peaks at the passband edges, similar to figure 2. Like the Bessel type filter of figure 25, the group delay is bigger at the lower corner frequency than the upper corner frequency. Like the other capacitive coupled filters, the attenuation at higher frequencies (220 MHz) is far less than the attenuation at lower frequencies (100 MHz). The measured frequency response of the filter is close to the calculated response shown in figure 33.

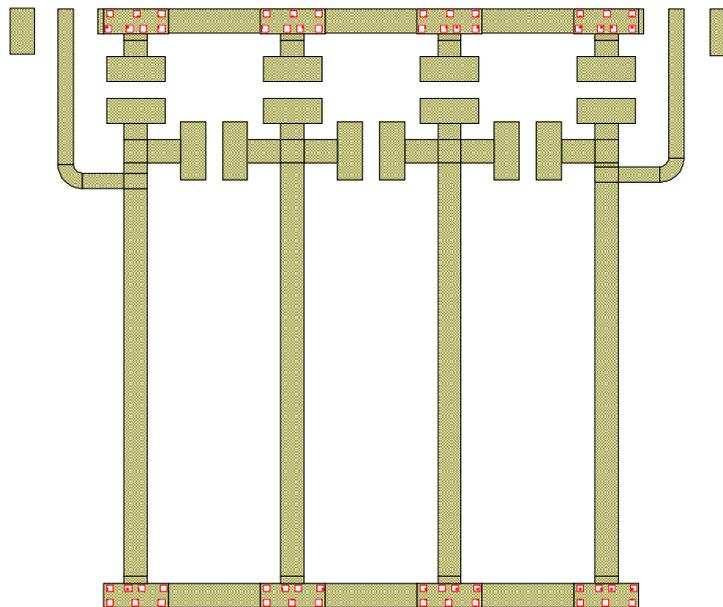


Figure 32. Coupled Resonator Filter using Microstrip Inductors.

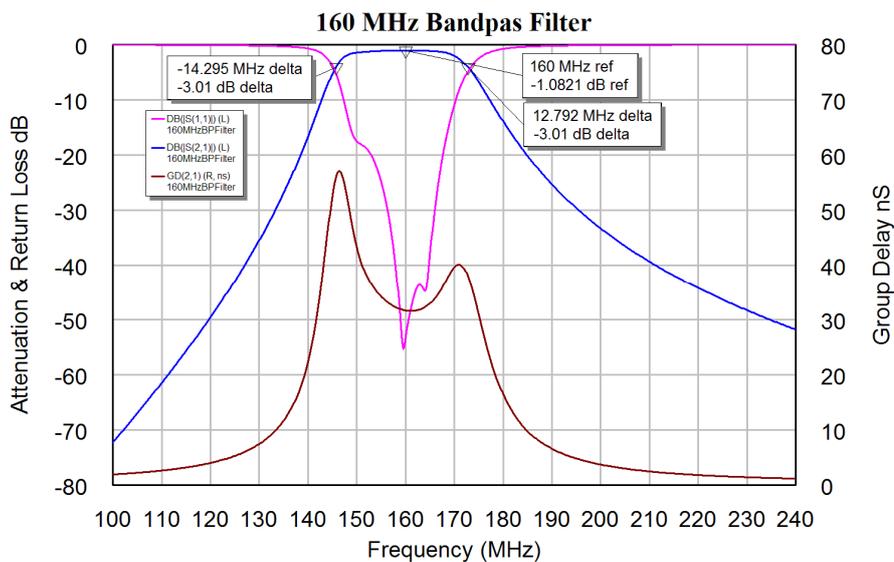


Figure 33. Frequency Response of the Filter of Figure 24.

Cauer-Chebyshev Bandpass filters

Cauer-Chebyshev filters have zeros in the stopband resulting in an equal ripples in the stopband as well as having ripples in the passband. It is more difficult to include zeros in the stopband of coupled resonator RF filters, as this requires coupling between adjacent resonators. It is however possible to include zeros of transmission by including open circuited stubs of one quarter wavelength at the appropriate frequency. Such stubs are normally used to suppress the unwanted passbands occurring at the second and third harmonic of the passband frequency. In other RF filters, transmission zeros are placed at frequencies that must be rejected, such as the transmission frequency in the receive input of a diplexer. In these cases an equal stopband ripple design as specified by the Cauer-Chebyshev filters is not implemented.

Parallel Coupled-Line Filter

This filter type is also known as a Stripline Filter. For this filter the design equations 11 to 14 can be used. However one can also use the generalised coupled resonator filter equations shown in equations 29 and 30 and applied to the interdigital filter design example on page 6.33.

The filter parameters $a(i)$ and the resulting even and odd mode Characteristic impedances are given as:

$$a(0) = \sqrt{\frac{\pi BW}{2q_1 F_c}} \quad \text{Eqn. 11}$$

$$a(i) = \frac{\pi BW}{2F_c} k_{i,i+1} \quad \text{Eqn. 12}$$

$$Z_{oo} = Z_{in} (1 - a(i) + a(i)^2) \quad \text{Eqn. 13}$$

$$Z_{oe} = Z_{in} (1 + a(i) + a(i)^2) \quad \text{Eqn. 14}$$

In these expressions BW is the Bandwidth of the filter, F_c is the centre frequency, q_1 , q_n are the normalised input and output loaded Q values and $k_{i,i+1}$ is the coupling coefficient between the i^{th} and the $(i+1)^{\text{th}}$ resonator as obtained from filter tables.

Example

Design a 4 resonator filter with a bandwidth of 70 MHz and a centre frequency of 1 GHz. The attenuation should be as low as possible in the passband. The input and output impedances should be 50 Ω . The filter is made using 0.82 mm thick RO4003 substrate with 35 micron copper.

A Butterworth filter type is selected for the initial design. Equations 4 and 5 can thus be used for obtaining the K and Q values, or they can be taken from table 1 to give:

Normalised	De-normalised
$q_1 = q_n = 0.7654$	$Q_1 = Q_n = 0.7654 * 1000 / 70 = 10.93$
$k_{12} = k_{34} = 0.8409$	$K_{12} = K_{34} = 0.8409 * 70 / 1000 = 0.05886$
$k_{23} = 0.5412$	$K_{23} = 0.5412 * 70 / 1000 = 0.03788$

Substituting this into equations 13 and 14 and substituting the resulting $a(0)$ and $a(i)$ into equations 11 and 12 gives the values for Z_{oe} and Z_{oo} shown below. Those values are then entered into the TxLine program (or Linecalc for ADS) as shown in figure 34.

Line	a	Z _{0e}	Z _{0o}	Width	Space	Length
0	0.3902	76.13	38.23	1.26 mm	0.12 mm	47.3 mm
1	0.09246	55.05	45.80	1.71 mm	1.08 mm	46.2 mm
2	0.05951	53.15	47.20	1.74 mm	1.57 mm	46.0 mm
3	0.09246	55.05	45.80	1.73 mm	1.2 mm	46.2 mm
4	0.3902	76.13	38.23	1.26 mm	0.12 mm	47.3 mm

Table 3 Parallel Coupled-Line Filter Parameters.

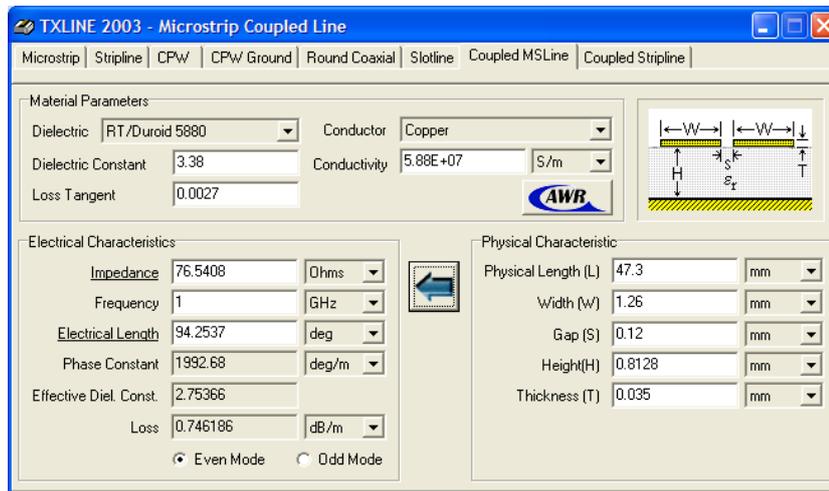


Figure 34. Calculating Line Widths and Coupling Gaps.

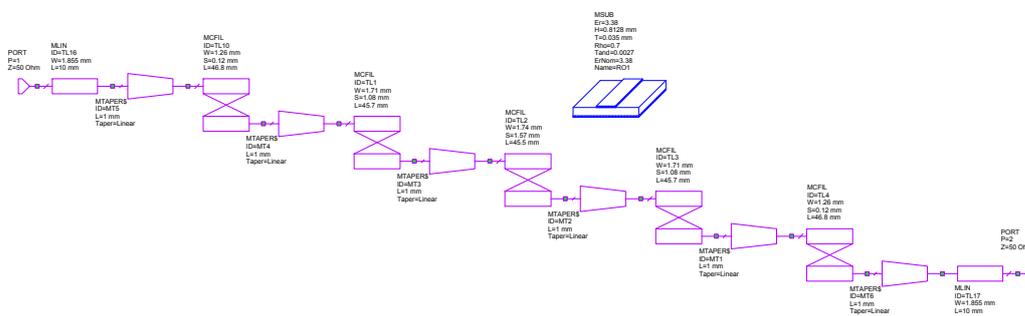


Figure 35. Microwave Office Schematic for the 1 GHz filter

Those line geometries are then entered in a circuit schematic to give the circuit shown in figure 35. End effects have been included by using the Microstrip filter element “MCFIL”. A 1 mm long taper has been added between each coupled section to ensure that the coupled lines do not produce a short circuit at the transition in width. As a result all resonators are shortened by 0.5mm to keep the total resonator length the same. The computer simulation software will produce a layout from the schematic as shown in figure 36. This layout can be plotted and produced as a PCB.

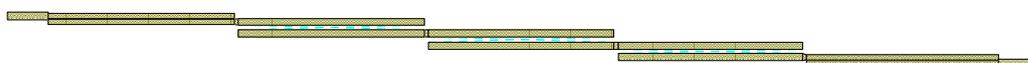


Figure 36. PCB layout for this filter produced by Microwave Office.

Figure 37 shows a close-up of the centre of the first resonator, and shows TL10, MT4 and TL1 of figure 35. It can be seen that without the tapered transition a short circuit would be produced. The filter from figure 35 as designed using the above procedure has a centre frequency of 981 MHz, and is 1.9% low. The bandwidth is 62 MHz and is 12 % low. The resonator lengths and coupling gaps are optimised, using the optimisation routines in MWO, to achieve the correct centre frequency and bandwidth. Since the design procedure gives a result very close to the specifications, the optimisation procedure is quick. The resulting frequency response is shown in figure 38. The insertion loss of the filter is 1.9 dB. The return loss can be improved if needed by setting optimisation goals on S_{11} and using the optimiser to automatically fine tune the elements.

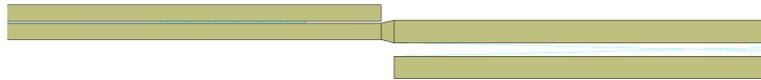


Figure 37. Close-up of part of figure 36, showing TL10, MT4 and TL1.

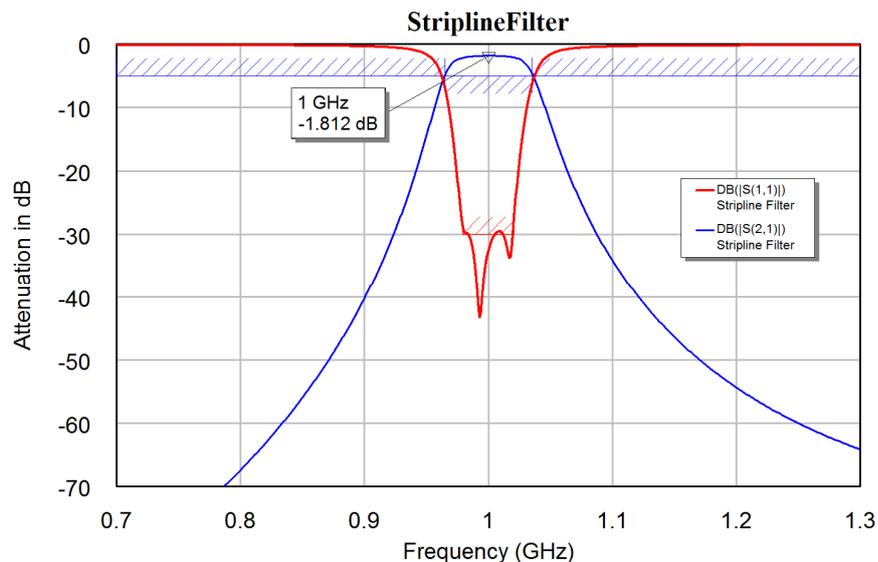


Figure 38. Frequency response of the filter from figure 35 after optimisation.

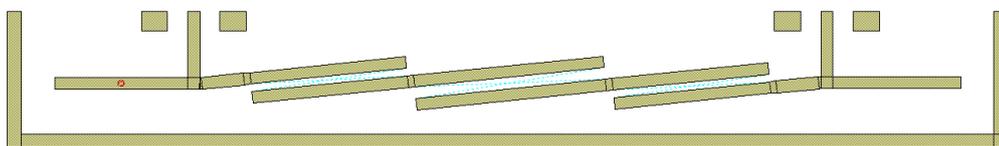


Figure 39. PCB layout for coupled line filter with tap coupling into and out of the filter.

For the filter in figure 35, 36 and 37, the input coupling gap is 0.14 mm. This gap is close to the limit of reliable manufacturing capability at JCU. To avoid this limitation, tap coupling is used to connect to the input and output resonators, as shown for the 100 MHz bandwidth at a 1.7 GHz centre frequency filter shown in figure 39. For wider bandwidth filters the input and output coupling gaps are very small and tap coupling is best. For very narrow bandwidth filters, the connection point for tap coupling becomes

too close to the effective ground point of the first resonator, so that line coupling as shown in figure 35 is more practical.

The actual filter, when produced, will have a centre frequency and bandwidth that will be slightly different from the simulated results. These differences are due to the computer simulation not being completely accurate, variations in PCB substrate materials, etching or milling errors in producing the PCB and so on. In many instances, the filter performance will be good enough. When the filter specifications are critical, a second iteration of the filter can be produced, where the simulation is biased to correct for the discrepancy between the measured and simulated results.

Hairpin Filter

The filters of figure 36 and 39 are long and skinny and are difficult to mount securely. To obtain a better form factor for the filter, the resonators can be folded as shown in figure 40. Coupled Resonator filters where the resonators are made up from quarter wavelength transmission lines will have additional passbands when the wavelength of the resonators are three quarter of a wavelength long, five quarter of a wavelength long and so on. These harmonic resonances are normally undesirable. To avoid these harmonic stubs can be added.

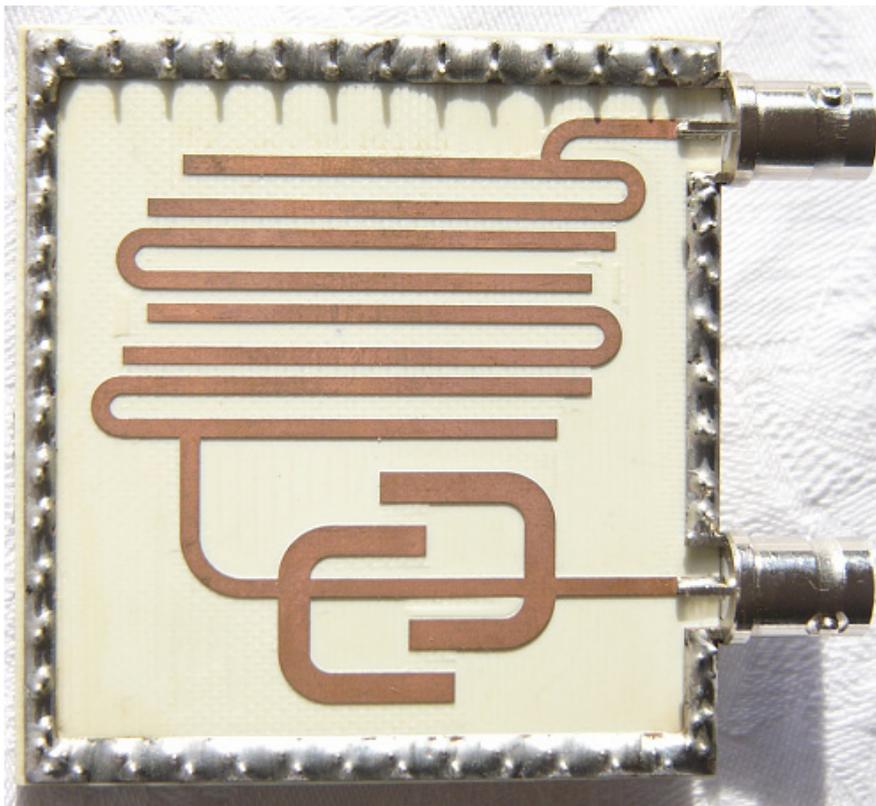


Figure 40. Hairpin filter.

The hairpin filter in figure 40 has two second and third harmonic suppression stubs incorporated as part of the design. The design procedure consists of firstly designing and optimising a basic hairpin filter as outlined in the previous example. Second and Third harmonic suppression stubs are then designed as a separate circuit, as shown in figure 41. The second harmonic stubs are one quarter wavelength long at the second harmonic and the third harmonic stubs are one quarter wavelength long at the third

harmonic. The spacing between the stubs is then adjusted such that the return loss of the circuit is large at the centre frequency of the filter, so that the harmonic stub circuit has the least effect on the filter performance. The stubs are folded to minimise the space used by the whole filter.

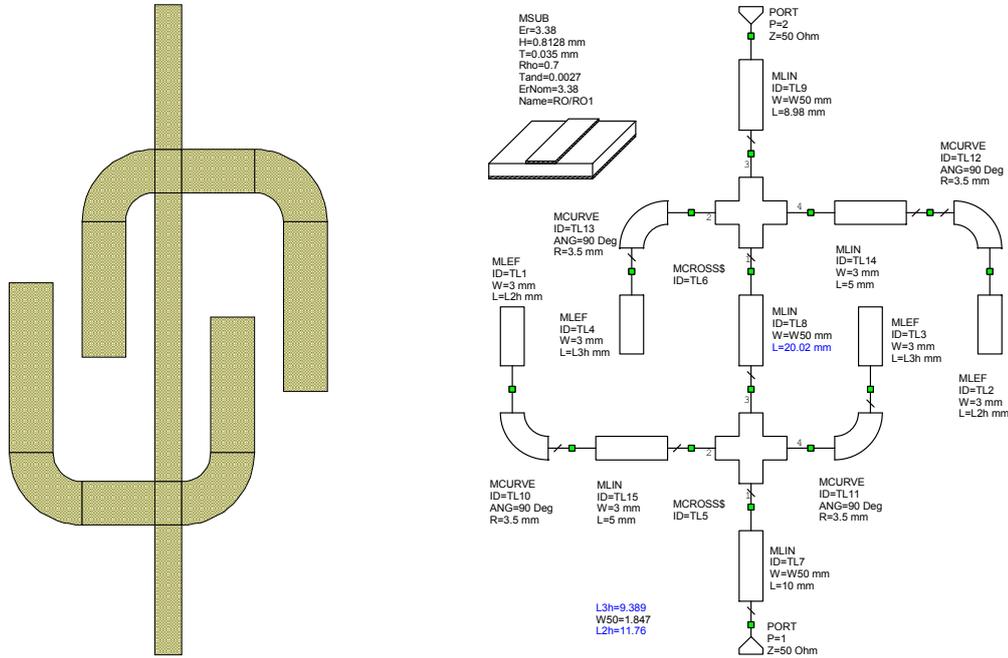


Figure 41. Folded Harmonic Stubs.

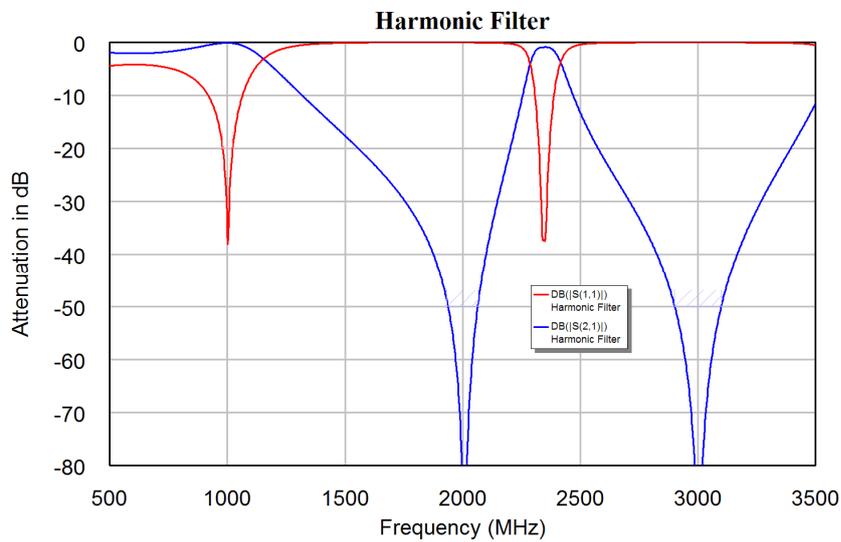


Figure 42. Frequency Response of Harmonic Stubs.

Figure 42 shows the frequency response of the harmonic stub circuit. It can be seen that there is some effect on the passband of the filter, so that further optimisation must be carried out in order to ensure that the combination of the hairpin filter and the harmonic stubs have the required passband response. Figure 43 shows the passband frequency response of the actual filter of figure 40 and compares it with the frequency response of the simulation. The filter was realised using the “MCFIL” elements used in figure 35

and using the coupled line element “MCLIN” with “MLEFX” or “MOPEN” elements compensating for the end effect of the open circuited ends of the coupled line elements. In addition an Electro Magnetic simulation is carried out on the coupled line sections. It can be seen that all the simulations give a good agreement with the measured results. The EM simulation in blue gives the poorest accuracy and requires the most computations.

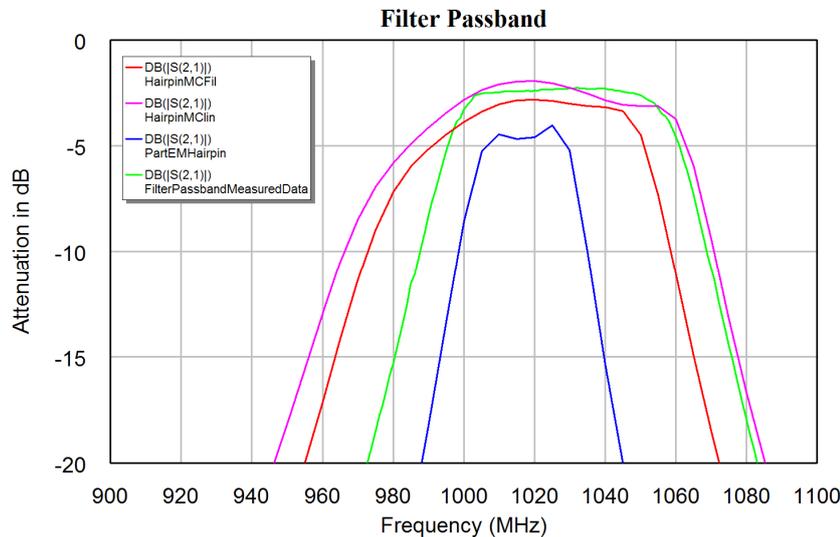


Figure 43. Frequency response of Hairpin filter.

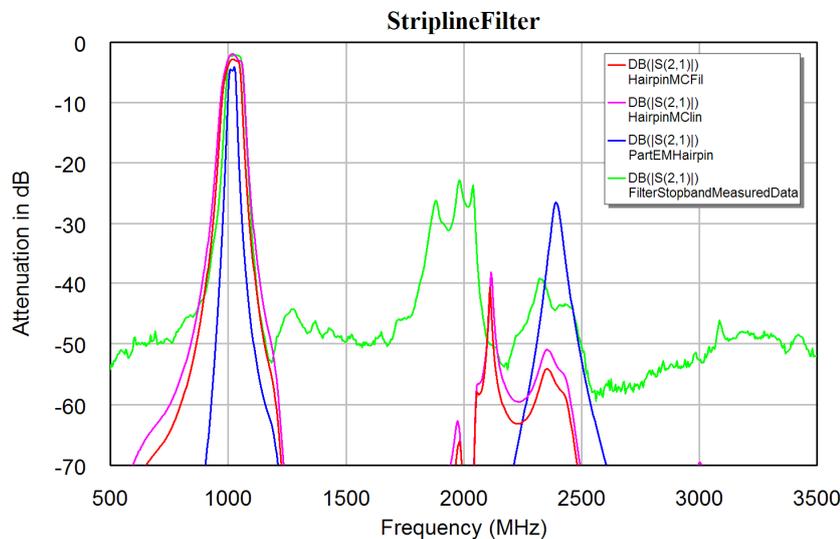


Figure 44. Frequency response of Hairpin filter of figure 40.

Figure 44 shows the stopband performance of the filter for both the measured performance and the computer simulation. The most significant difference is the reduction in attenuation in the stopband. This is due to Electro-Magnetic radiation between parts of the circuit and the input and output leads. The stopband attenuation can be optimised by mounting the filter in a conductive (aluminium) enclosure and applying microwave absorbing material in the appropriate places, to be determined experimentally. In many cases, applying the absorber to the back groundplane of the filter improves the stopband performance sufficiently.