EE 172 Final Project

Evanescent Mode Filter

Svetoslav (Sam) Mandev

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Introduction

There are various types of microwave filters being used in the RF industry, each with its own advantages and shortcomings. Some examples include coaxial, stripline, microstrip, cavity, and waveguide filters. Among them, waveguide filters are known to have very low loss and high Q factors, while their application is mostly limited by their relatively large size compared to standard transmission lines. One way to implement waveguide filters with smaller dimensions is to use them in evanescent mode, or below cutoff frequency. For example, if we wish to implement a 2 GHz circular waveguide filter, we could either use a 4-inch circular waveguide (f_c=1.73GHz) in a propagating mode, or we could use a 3-inch circular guide (f_c=2.31GHz) in evanescent mode, for a 25% reduction in diameter. Waveguides operating in evanescent mode are well-suited to operate as filters, due to their high attenuation in the bandstop range.

Theory

1. Lumped element model of evanescent mode filter

By modeling the waveguide with lumped elements, the filter can be designed using standard design techniques. A waveguide in evanescent mode can be represented by equivalent T or π-sections of lumped inductances [Fig. 1 (b) and (c)] according to [1],[2]. For moderate bandwidths (<20%), Craven and Mok [2] show that the quantity $\frac{\gamma * l}{2}$ is virtually independent of frequency, and the evanescent waveguide can be approximated as in Fig. 1 (d) using inductively coupled LC resonators.

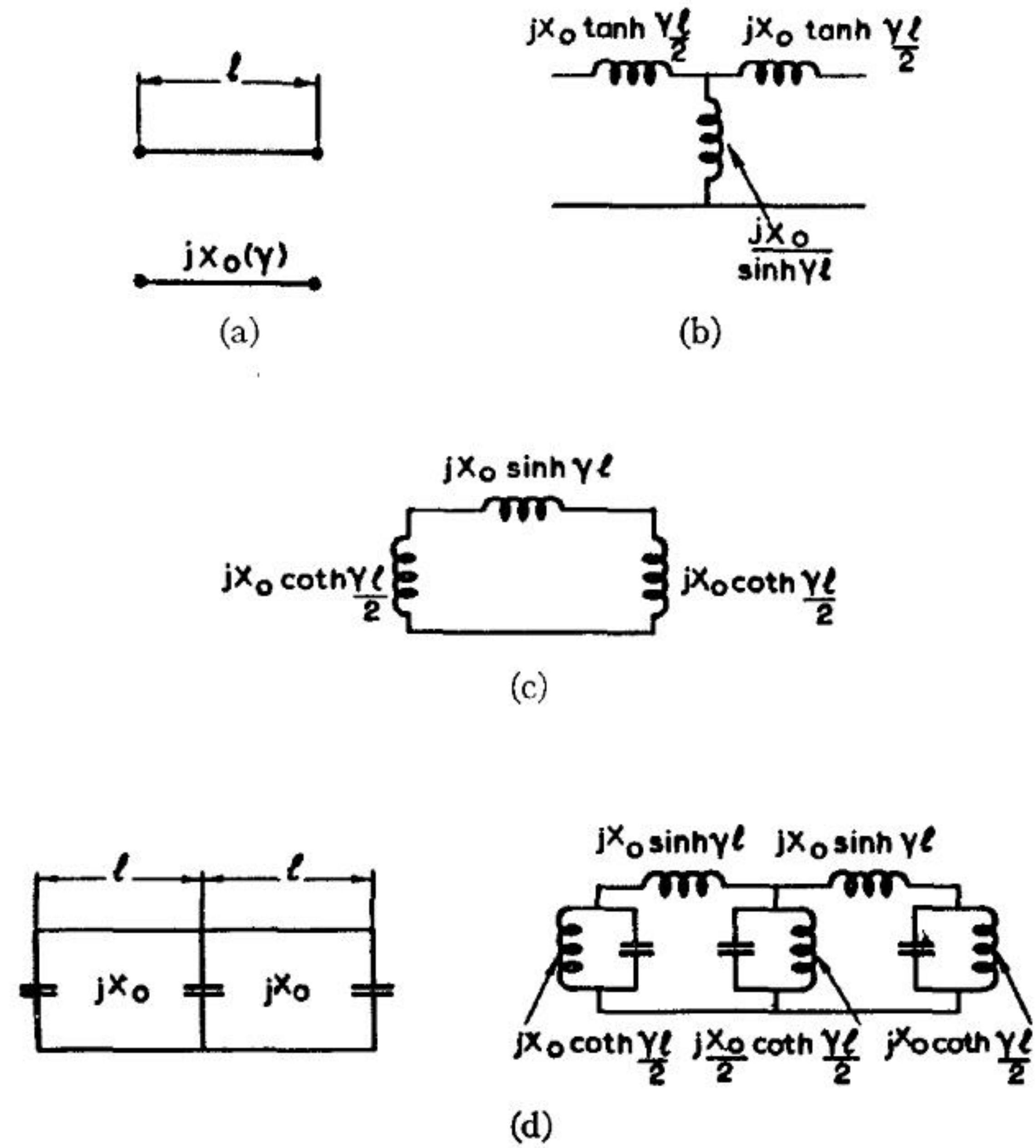


Fig. 1. (a), (b), and (c) Equivalent circuits of evanescent mode waveguide. (d) Lumped equivalent circuit of evanescent mode waveguide filter.

Here, $\gamma = \frac{2\pi f_0}{c}\sqrt{\left(\frac{f_c}{f_0}\right)^2 - 1}$ is the propagation constant, or in this case – the attenuation of the waveguide below cutoff.

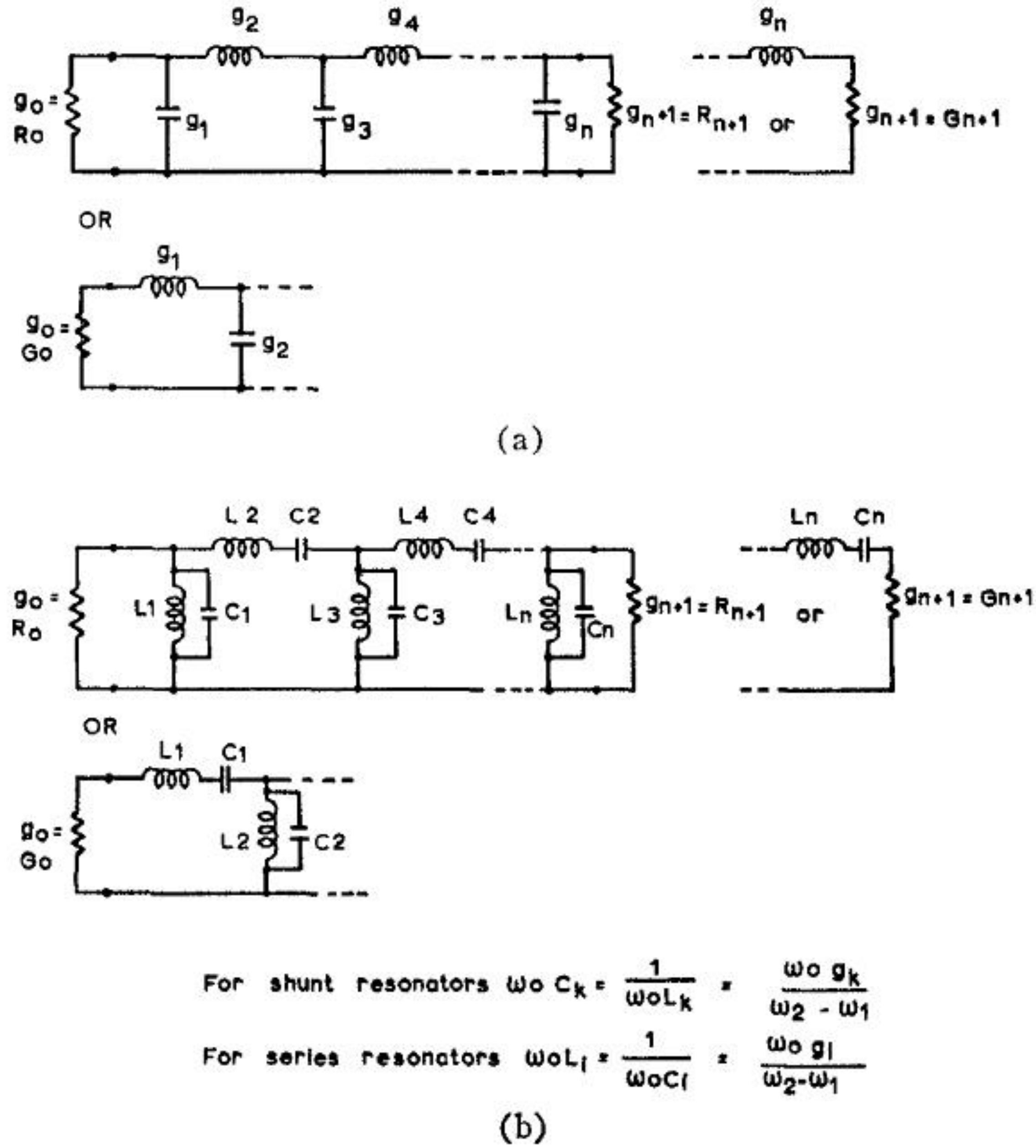


Fig. 2. (a) Low-pass prototype. (b) Bandpass ladder network derived from low-pass prototype.

Furthermore, the bandpass ladder network [Fig. 2 (b)] can be derived from the low-pass prototypes [Fig. 2 (a)], using basic filter transformations as shown in Table 8.6 in Pozar's text [4].

The equivalent lumped reactance and susceptance values can be obtained as follows:

- a shunt capacitor, C_k, in the low-pass prototype is transformed to a shunt LC circuit with element values,

$L'_k = \dfrac{\Delta}{w_0 C_k}$, $C'_k = \dfrac{C_k}{\Delta w_0}$, where $\Delta = \dfrac{w_2 - w_1}{w_0}$ is the fractional BW of the bandpass , and

$$C_k = g_k$$

- series inductors, L_k, of the low-pass prototype are converted to parallel LC circuits having element values,

$$L'_k = \frac{\Delta L_k}{w_0}, \qquad C'_k = \frac{1}{w_0 \Delta L_k}, \text{ where } L_k = g_k$$

G-values for Butterworth (Maximally flat) and Chebyshev (Equal-ripple) low-pass filters, are tabulated in various texts including Pozar's and Collin's [4],[5]. This forms the basis for the lumped-element model of the Craven-type evanescent mode filters. Using the filter transformations, the lumped model can be accurately simulated, and filter response obtained, using proper reactance and susceptance values. I have presented a Microwave Office simulation and filter response for my particular implementation, further in the paper.

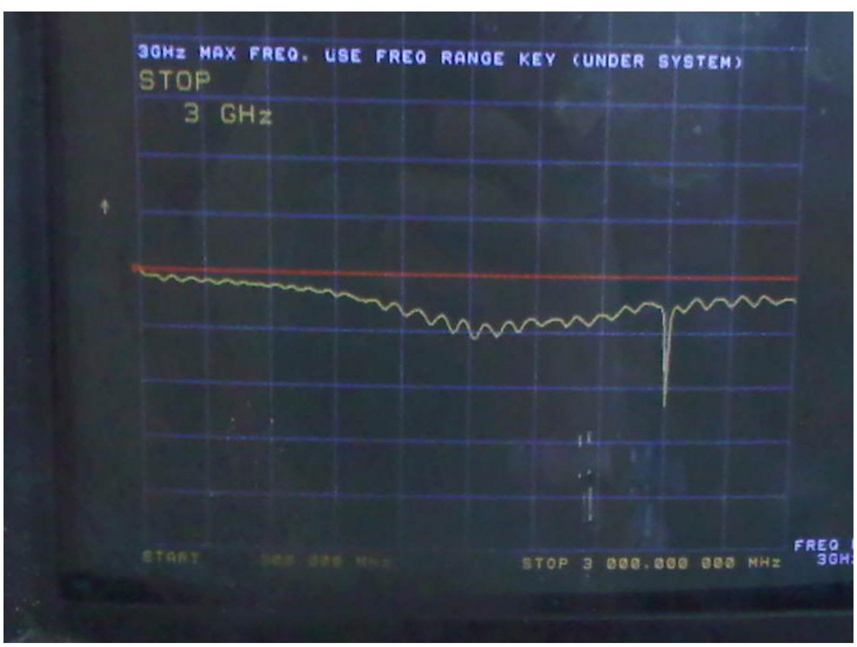
2. Designing and building the waveguide filter

The cited papers by Craven and Mok [2], and Howard and Lin [3], explain in detail the relationship between the lumped element values and the dimensions of the waveguide filter. They present different methods for implementing resonating obstacles into the waveguide, various coupling types, Q-factor calculation, and other interesting considerations. In this project however, I have used the "quick and dirty" method developed by Dr. Raymond Kwok, a professor of Physics and Electrical Engineering at San Jose State University.

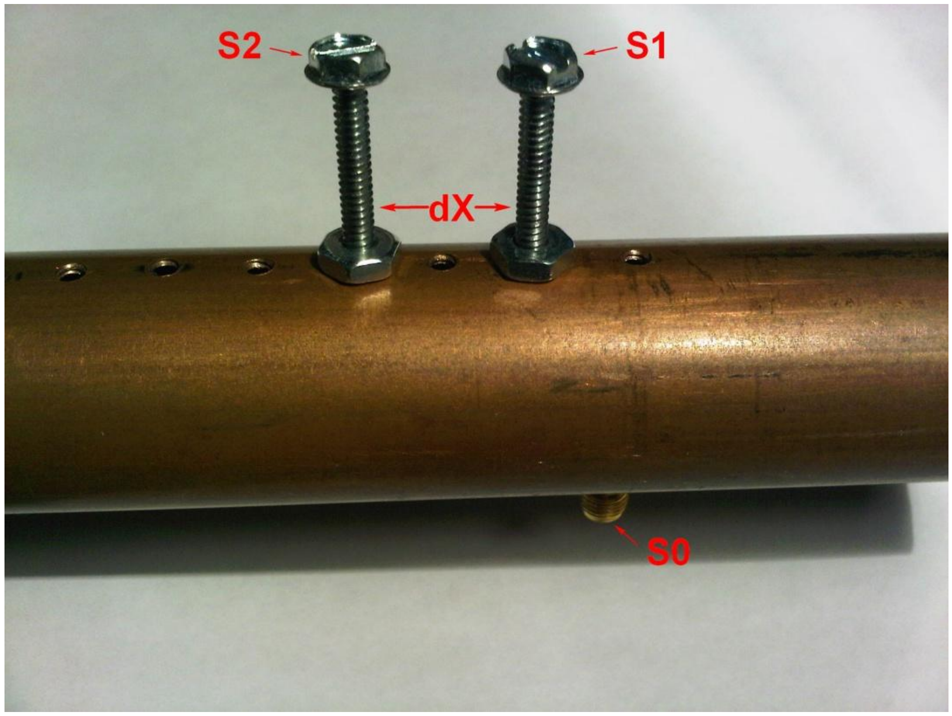
In this method, two identical waveguide sections are used. Length will depend on filter requirements. The first section of waveguide is used to obtain the coupling curve of the waveguide, which is an exponential relationship between bandwidth and waveguide length.

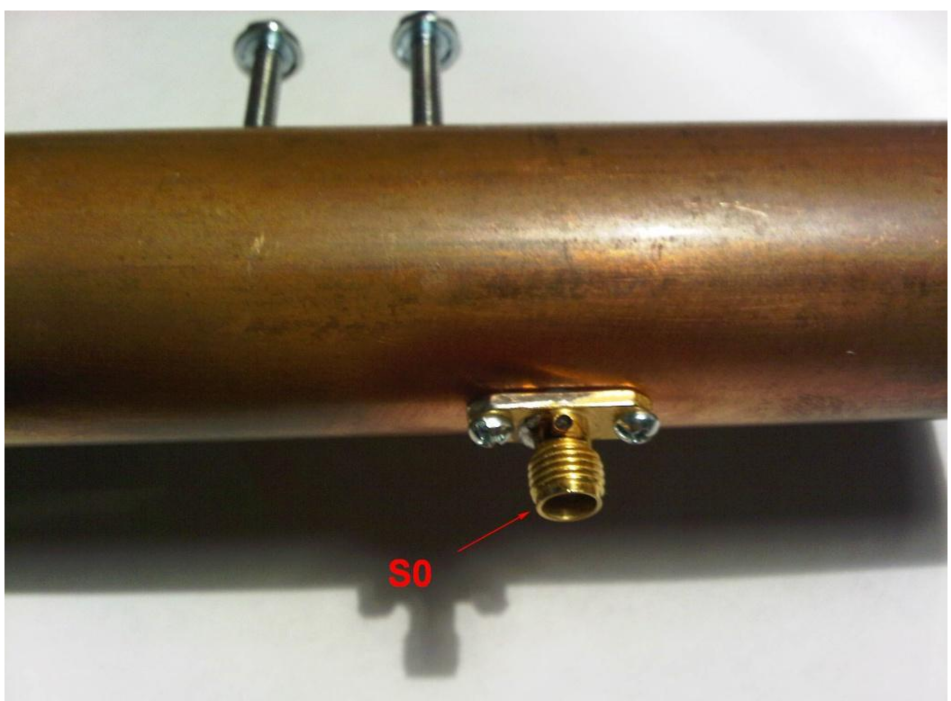
The basic design steps of the quick and dirty method are as follows:

1. Choose a waveguide such that $f_0 < f_c < 2 f_0$. Howard and Craven recommend, as a rule of thumb, $f_0 = 0.85 f_c$, for highest possible Q-factor.
2. After connecting the first coaxial connector S_0 somewhere on the waveguide, place the 1st capacitive screw S_1 near the connector. The initial distance is somewhat arbitrary. As a rule of thumb, for a WR28 waveguide, place the screw about 0.3" away from connector. Tune the screw until the return loss S_{11}, on the network analyzer, is centered at the desired center frequency. The return loss should be between 5 to 10 dB down.

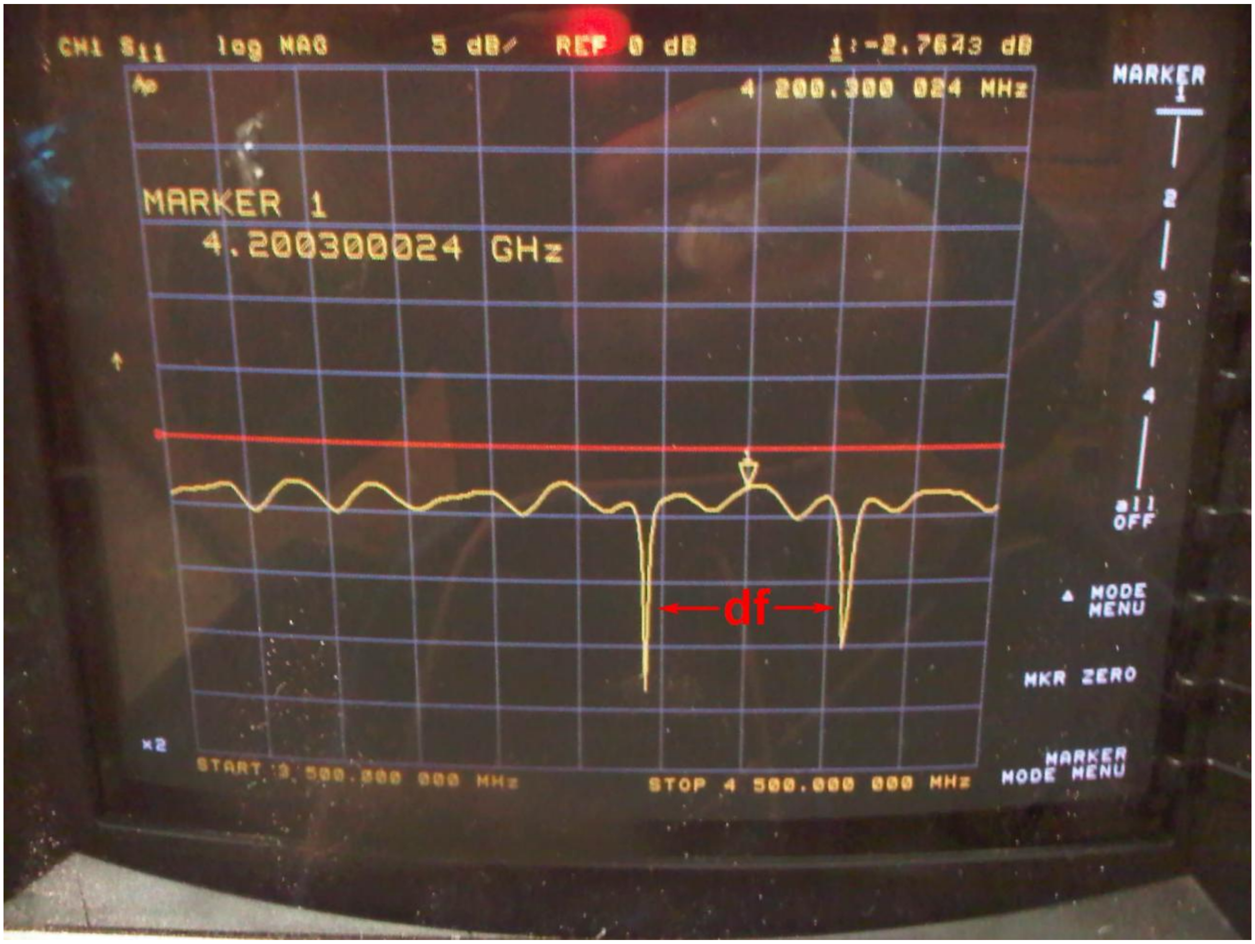


3. Connect screw S_2 next to S_1 (arbitrary distance) and tune until two peaks are visible on the screen.





Keep tuning until both peaks are at about equal distance from f_0, placed symmetrically around the f_0. Here, $f_0 = \frac{1}{2}(f_1 + f_2)$. At this point measure and record ΔX and Δf.



4. Move S_2 and repeat measurement, until enough data is obtained in order to plot coupling curve. The farther S_2 is from S_1, the smaller the peaks (dibs) will be, since the signal is being attenuated fast.
5. Plot coupling curve "ln(Δf) vs. ΔX" on semi-log paper. Plot should be a straight or near straight line. Extract slope and y-intercept of plot.

6. Once the required filter specs are determined (BW, number of sections, ripple dB), the new ΔX distances between the tuning screws can be calculated from the coupling curve, as follows:
    a. Using the published g-values for the specific dB ripple and BW, calculate the new Δf's from the equation $\Delta f_{i,i+1} = \dfrac{BW}{\sqrt{g_i g_{i+1}}}$
    b. From the calculated Δf's, extract the distances between the resonators from the coupling curve.
7. Design the distances between the connectors and the first screw. The process is somewhat arbitrary, and experimentation may be necessary based on desired coupling, but here are some suggestions:
    a. Guideline: 0.15" is sufficient for WR28 waveguide (Dr. Kwok's example)
    b. Craven and Mok [2] suggest the following approximation: $\coth(\gamma l_0) = 1$ for up to four significant figures.
    c. Dishal Method – design the connector-to-screw length based on desired external Q-factor. The method entails making several test models with different coupling positions 'h', tuning S1 screw to the same frequency, and then plotting Q(external). vs 'h' [8],[9]. Then design based on plotted curve. Here, $Qext = \dfrac{f_0}{\Delta f}$

Any of the above suggestions are good starting points, but further measurements may be needed for the connector couplings. Also, the length of the connector pin is an important consideration.

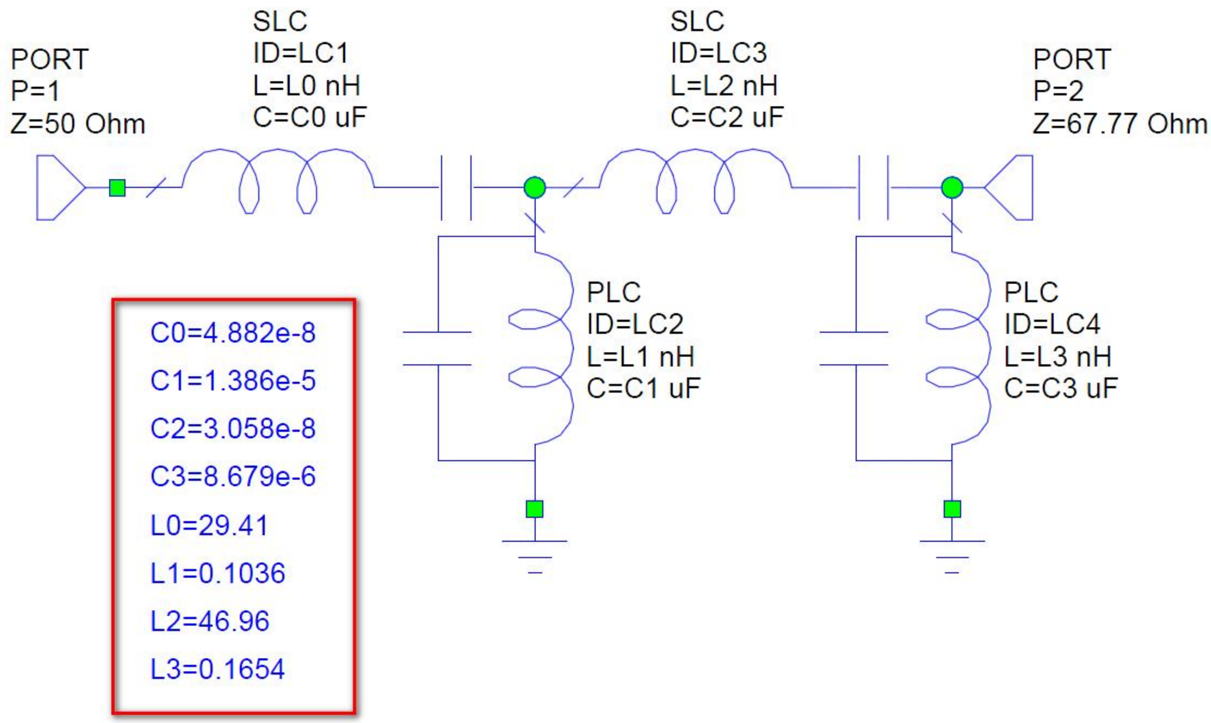
8. Build and tune filter. Fine tune using interstage screws, roughly midway between the resonating screws. Tune the coupling pins if needed.

Design Example

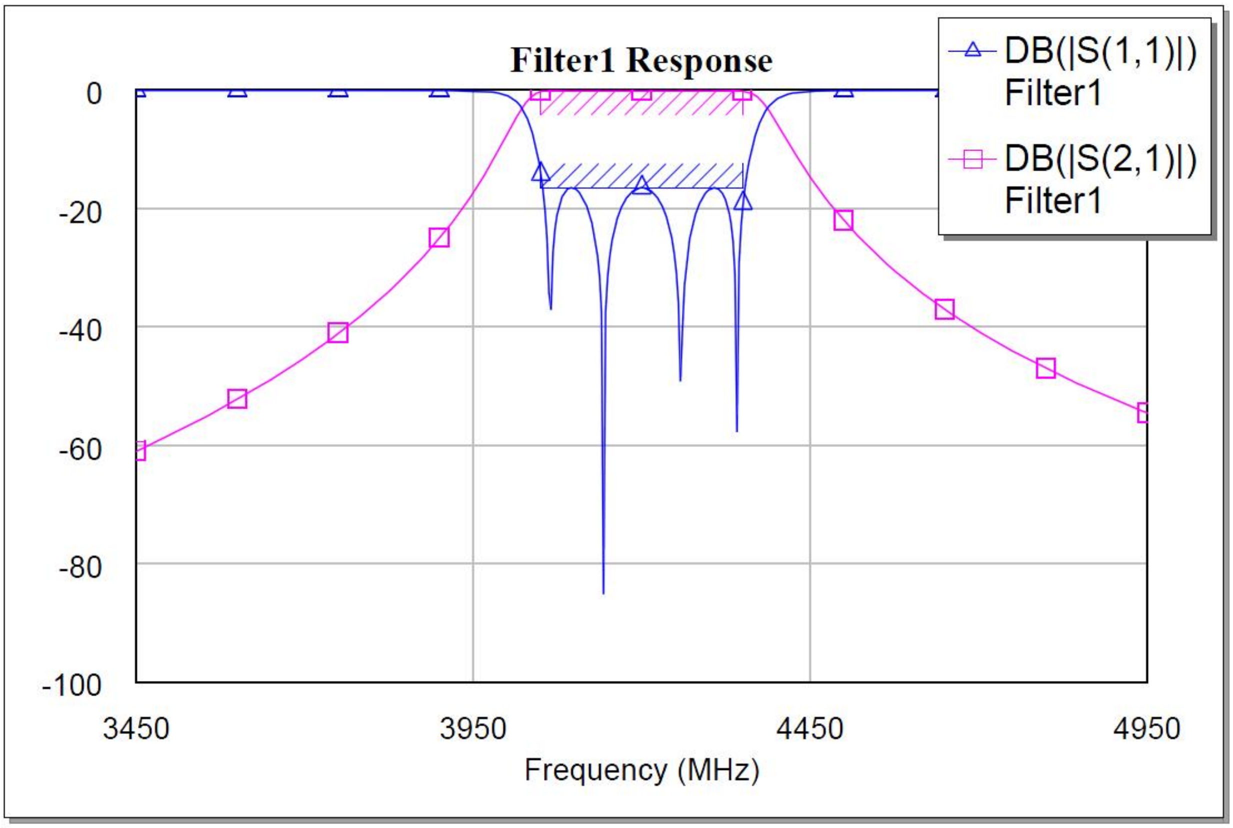
My particular design is for a 4-section, 0.1dB Chebyshev Equal-ripple filter, operating at center frequency 4.2GHz with 300MHz bandwidth.

1. MWO Simulation

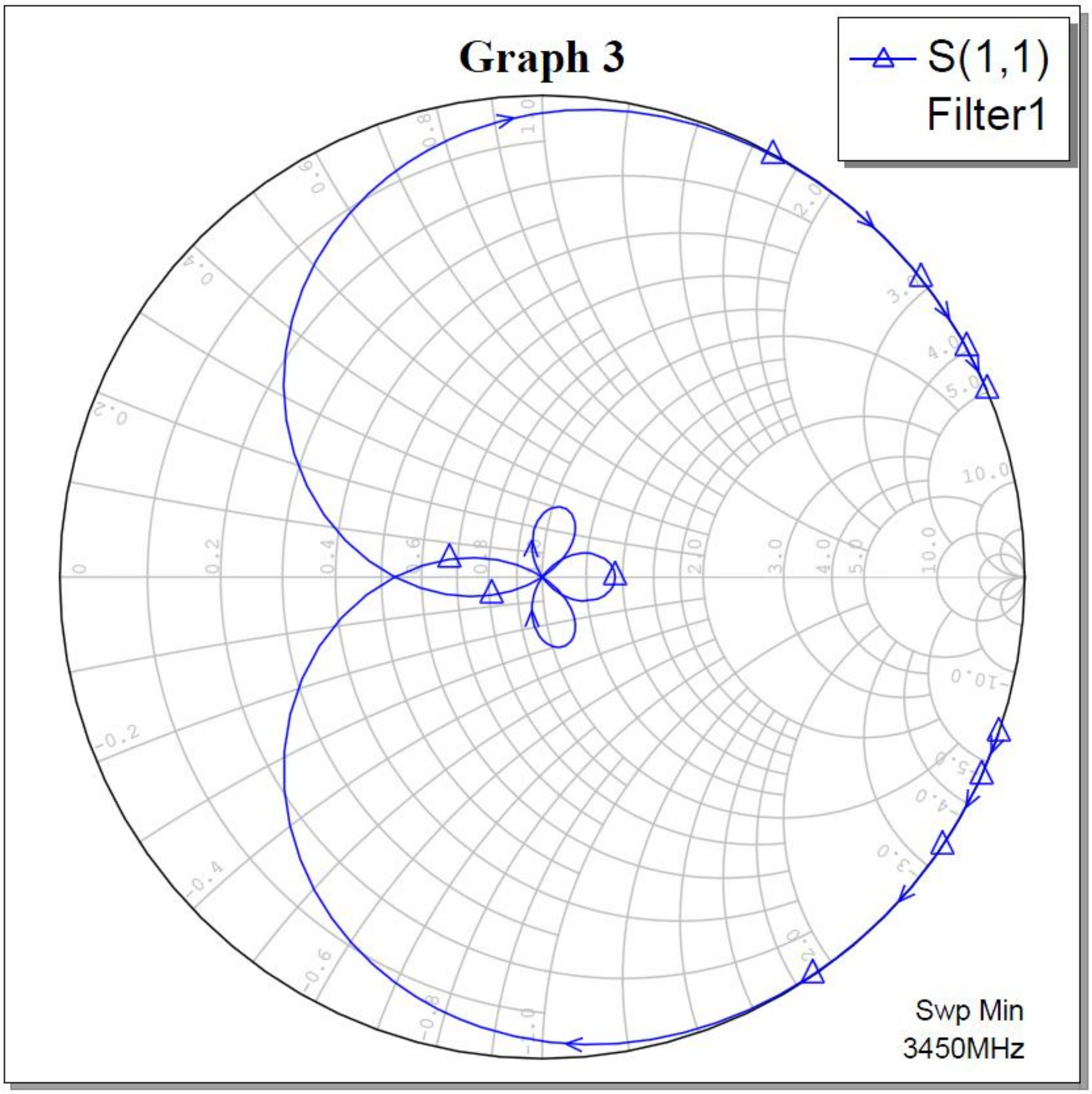
The model has four LC resonators, equivalent to the resonating screws on the waveguide.



The L and C values in the circuit were calculated using the filter transformation equations mentioned earlier, with the published g-values for a 4-pole, 0.1 dB Chebyshev ripple [5].



BW = 300 MHz, RL ≈ 16dB

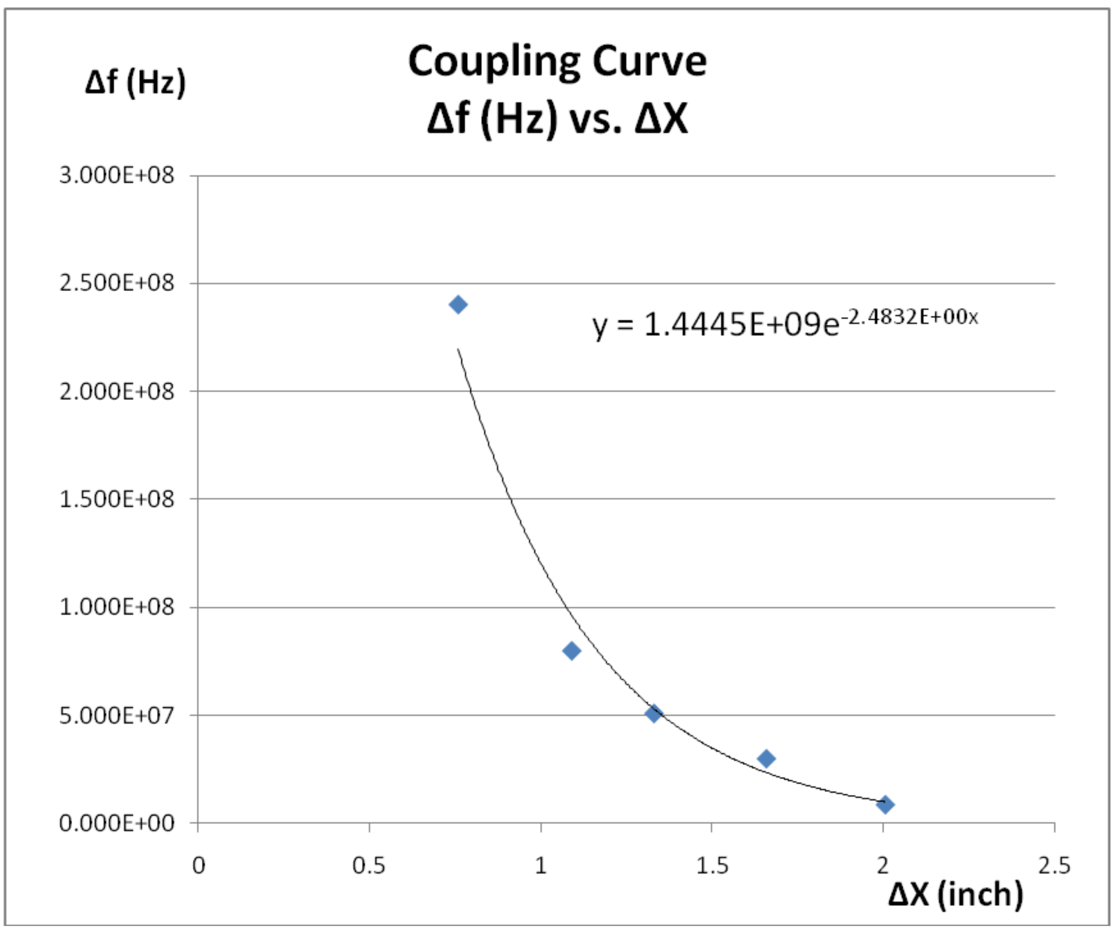


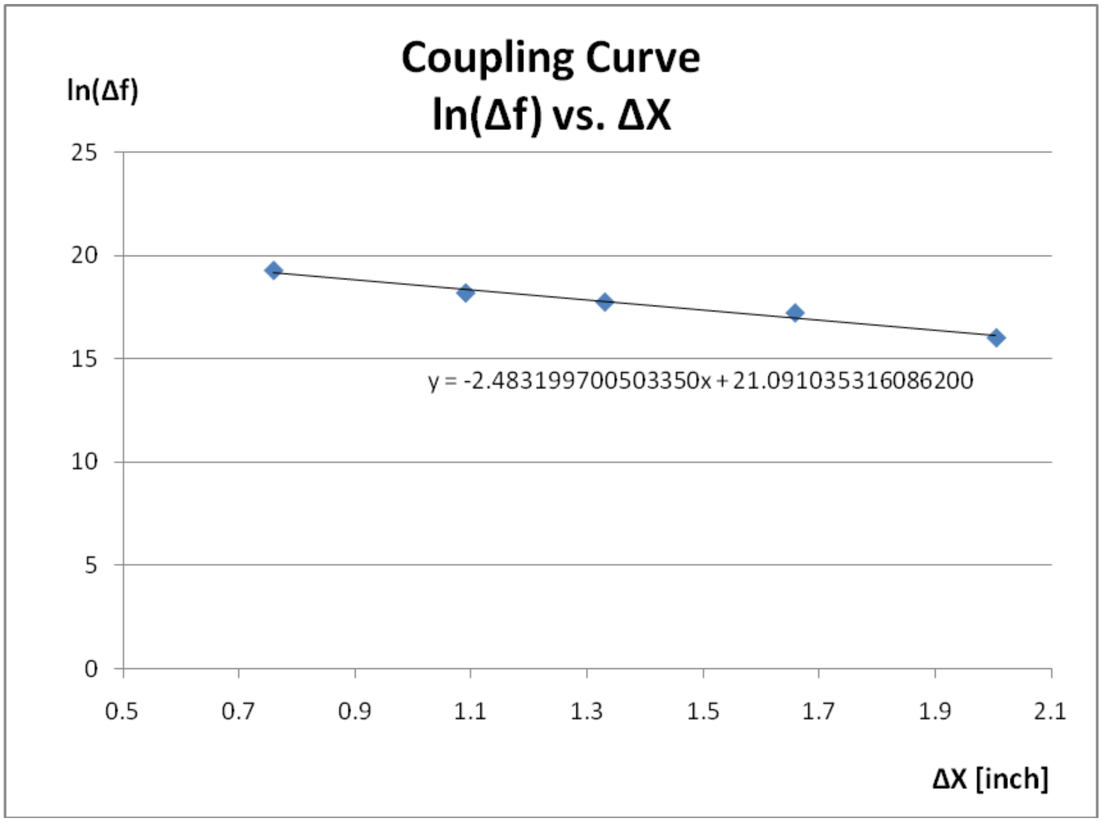
Graph 3
S(1,1)
Filter1
Swp Min
3450MHz

2. Waveguide measurements, coupling curve

- 1-inch diameter copper pipe, $f_{c_{nm}} = \frac{1.841c}{2\pi a} = 6.922GHz$

Measurements			
ΔX (inch)	Δf (GHz)	Δf (Hz)	ln(Δf)
0.759	0.24	240000000	19.296149
1.09	0.08	80000000	18.197537
1.33	0.05106	51060000	17.748512
1.658	0.03014	30140000	17.221364
2.005	0.0089	8900000	16.001562



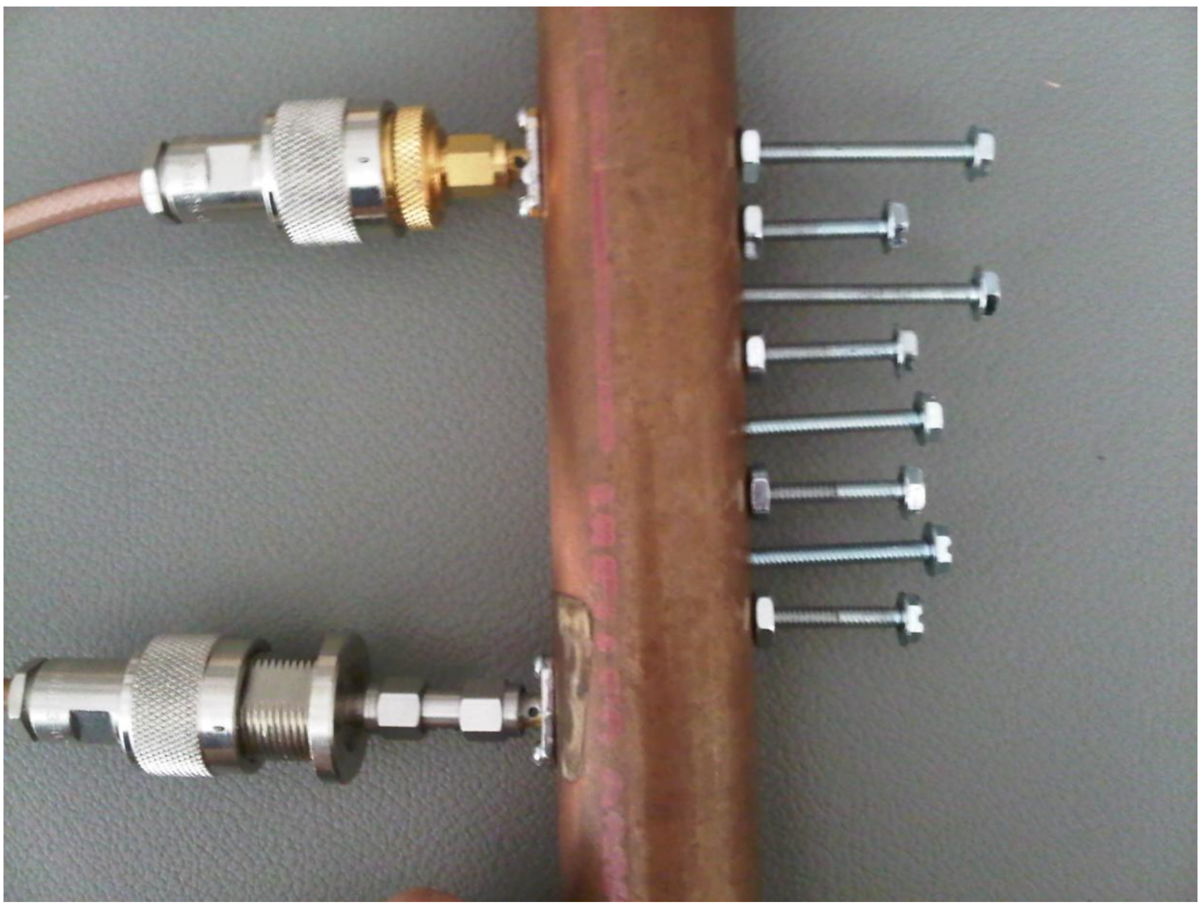


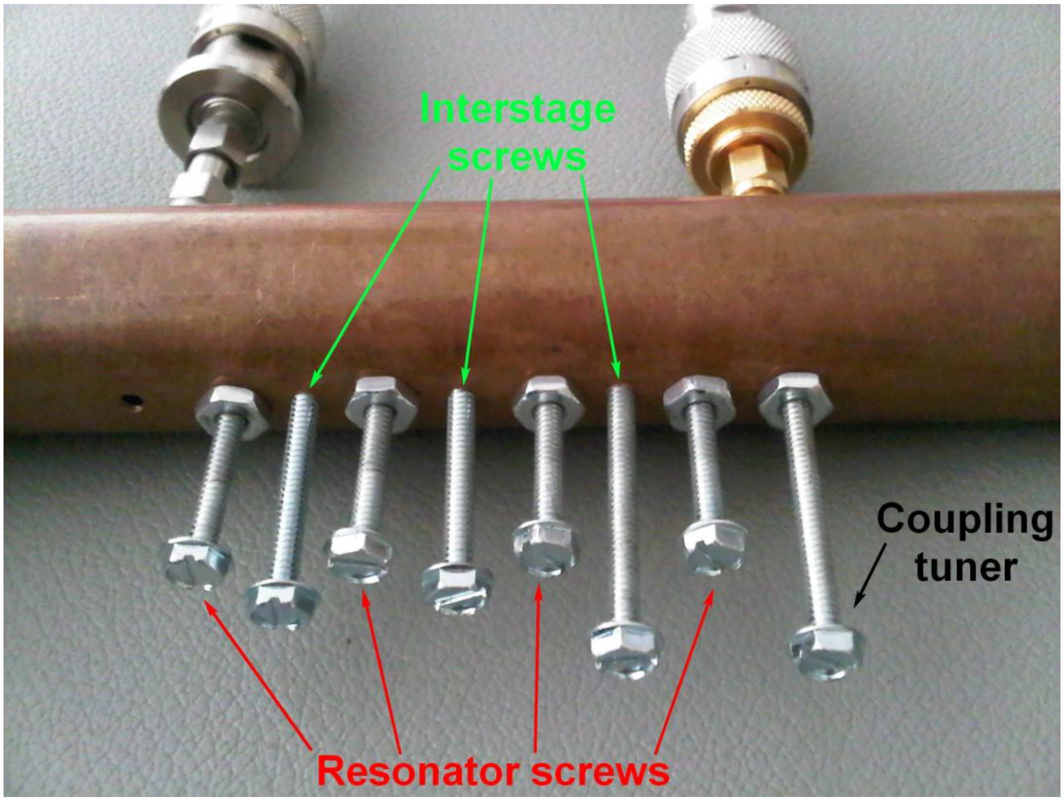
Attenuation		
α (meas)	α (calc)	
2.4832	2.9257	Np/inch
21.5690752	25.41263	dB/inch
97.7637795	115.185	Np/m

3. Calculate Δf and extract ΔX from coupling curve

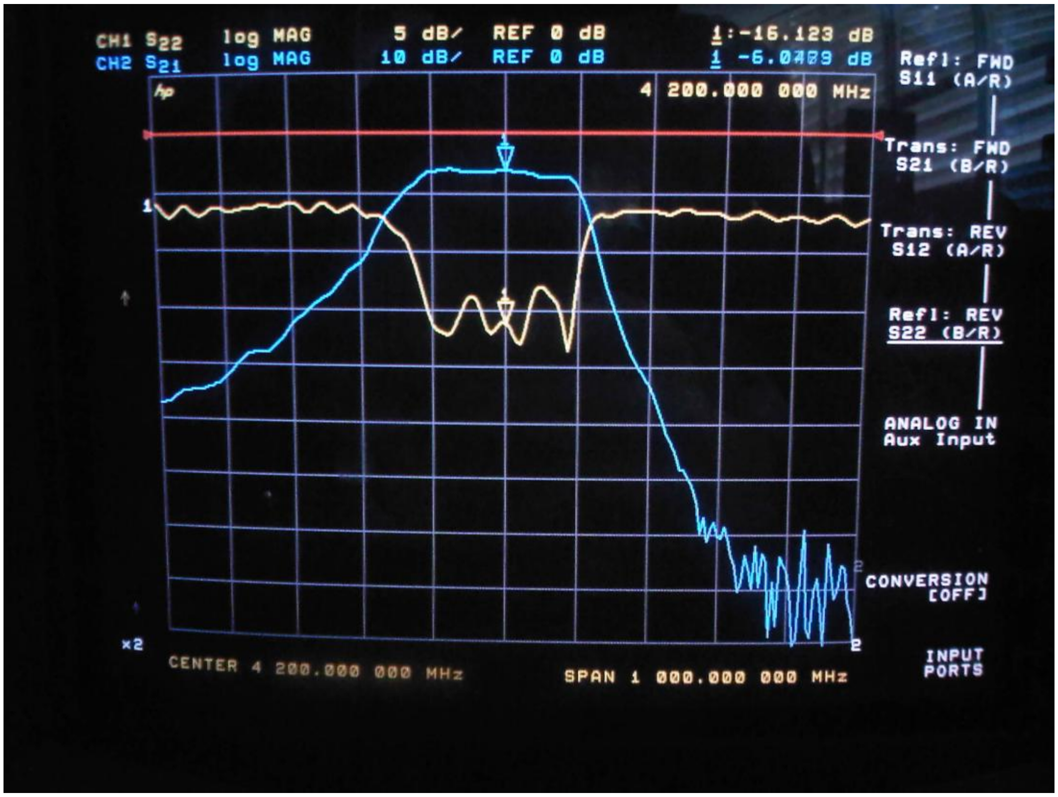
Chebyshev 4-section, 0.1dB ripple, BW = 300MHz				
	$g_{i,i+1}$	$\Delta f_{i,i+1}$ (MHz)	$\ln(\Delta f_{i,i+1})$	$\Delta X_{i,i+1}$ (inch)
g1	1.1088	249.29	19.334	0.70752
g2	1.3061	197.29	19.100	0.80172
g3	1.7703	249.30	19.334	0.70750
g4	0.8180			
g5	1.3554			

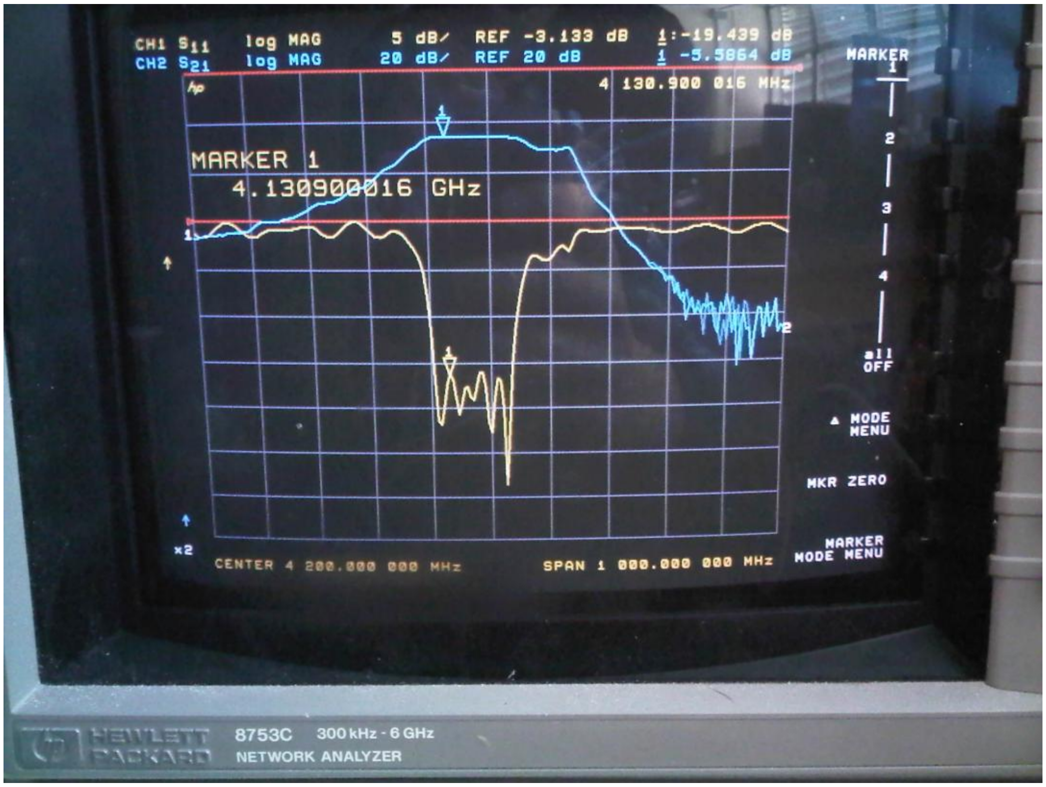
4. Building the filter
   - Connector pin length ≈ 5/6 of diameter (SMA pin extended with a 3mm wire for better coupling)
   - Did not change the distance between S_0 and S_1 (≈0.5”)
   - Tuning screws / nuts : #6-32 (tapped)
   - Connectors: SMA panel mounts + SMA to N-type adapters
   - SMA connectors bolted with #2-56 screws (tapped and bolted)

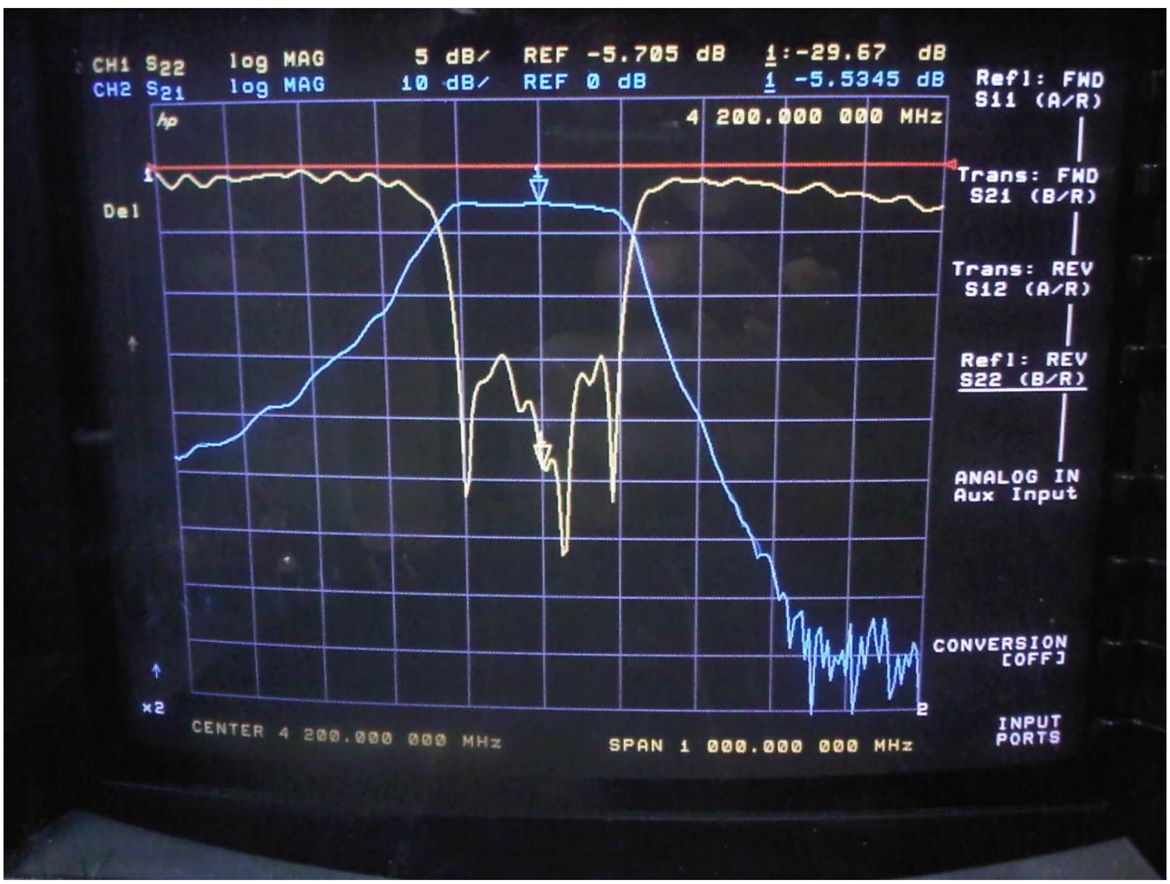



Interstage
screws
Coupling
tuner
Resonator screws

5. Tuning the filter – for 0.1 dB ripple, need to achieve ≈16dB Return Loss




CH1 S11 log MAG 5 dB/ REF -3.133 dB 1:-19.439 dB
CH2 S21 log MAG 20 dB/ REF 20 dB 1 -5.5864 dB
4 130.900 016 MHz
MARKER 1
4.130900016 GHz
CENTER 4 200.000 000 MHz
SPAN 1 000.000 000 MHz
MARKER
all OFF
MODE MENU
MKR ZERO
MARKER MODE MENU
HEWLETT PACKARD
8753C 300 kHz - 6 GHz
NETWORK ANALYZER


CH1 S22 log MAG 5 dB/ REF -5.705 dB 1:-29.67 dB
CH2 S21 log MAG 10 dB/ REF 0 dB 1 -5.5345 dB
4 200.000 000 MHz
Del
CENTER 4 200.000 000 MHz
SPAN 1 000.000 000 MHz
Refl: FWD S11 (A/R)
Trans: FWD S21 (B/R)
Trans: REV S12 (A/R)
Refl: REV S22 (B/R)
ANALOG IN Aux Input
CONVERSION [OFF]
INPUT PORTS

6. Conclusion, observations, and suggestions

- ~15dB of Return Loss (S_{22}) achieved, BW = 200MHz
- Only one of the couplers could be tuned for a small difference. That means that the connectors may have been over or under-coupled. For better result, need to design coupling positions as described in design steps - Craven/Howard approximation [2],[3] or Dishal's method [8],[9].
- When tuning the first screw by itself, it produced a low and wide peak, which also indicates that poor coupling, as mentioned above.
- For easy calculation of external and unloaded Q factors, refer to [7] by Dr. Ray Kwok

References

[1] G. Craven, "Waveguide Below Cutoff: A New Type of Microwave Integrated Circuit," Microwave Journal, Vol. 13, August 1970, p. 51.

[2] G. Craven and C. Mok, "The Design of Evanescent Waveguide Bandpass Filter for a Prescribed Insertion Loss Characteristics," IEEE Trans. MTT, Vol. MTT-19, March 1971.

[3] Howard, John and Wenny C. Lin. "Evanescent mode filter: design and implementation." Microwave Journa/32.10 (1989): 121+. Academic OneFile. Web. 18 May 2011.

[4] D. M. Pozar, *Microwave Engineering*, 3rd ed., John Wiley & Sons, New York, 1998.

[5] R. E. Collin, *Foundation for Microwave Engineering*, 2nd ed., McGraw-Hill, New York, 1992.

[6] Kwok, Raymond S. "Notes on Evanescent Bandpass Filter". 1991

[7] Kwok, Raymond S. and Liang, Ji-Fuh "Characterization of High-Q Resonators for Microwave-Filter Applications". 1999

[8] M. Dishal, "Alignment and Adjustment of synchronously tuned multiple-resonant -circuit filters" Proc I.R.E. vol 39 November 1951

[9] M. Dishal, "A Simple Design Procedure for Small Percentage Bandwidth Round Rod Interdigital Filters" MTT Vol13 Sept 1965 p 696