
Electrical Design of Narrow Band Filters

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Introduction

- ❑ The design of a narrow band microwave filter starts with the assignment of the specifications
- ❑ According to these, a specific technology for the filter implementation must be selected. This choice is primarily driven by two opposing requirements:
 - ▶ Overall filter size (as small as possible)
 - ▶ Insertion loss in passband (as small as possible)
- ❑ In fact, for a given technology, the maximum realizable unloaded Q of the resonators is defined; the higher is the Q , the larger is the volume of the resonators and the smaller are the losses.
- ❑ Other aspects, application dependent, must be also considered in the selection of the implementing technology: maximum power handling; influence of environment parameters as temperature, humidity, vibrations...

From specs to design parameters

- ❑ The definition of electrical design parameters must take into account an adequate margin with respect to the given specifications
- ❑ The margin is requested to compensate possible variations of the filter response due to various causes (typically: working temperature range, mechanical tolerances, aging)
- ❑ The margin is typically taken on the passband and stopband limits, on the passband loss and on rejection at the specified frequencies
- ❑ A consequence of the margin introduction may be the increase of the filter order and of the filter size (with respect to the specs without margin)

Example: Design of a filter for Base Station of Mobile Communications

□ Specification:

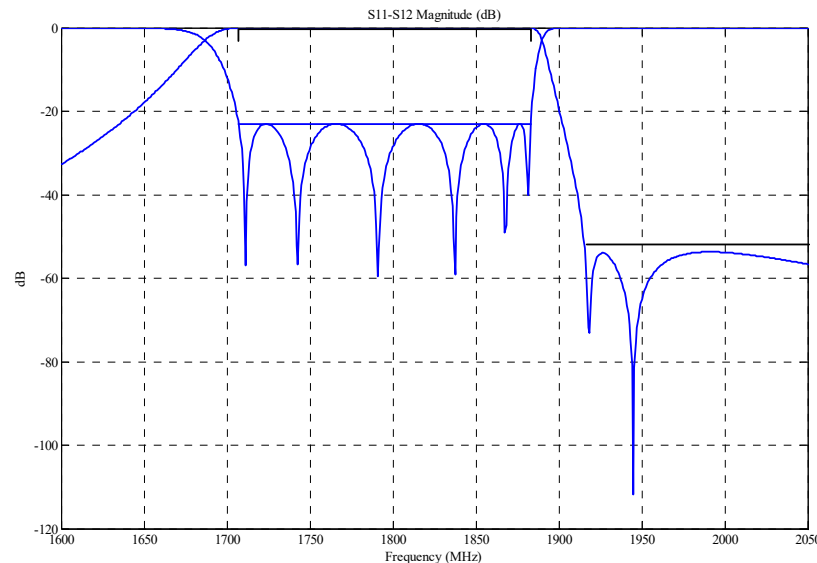
- ▶ Passband: 1710 - 1880 MHz (*DCS 1800 up link*)
- ▶ Return Loss: 23 dB
- ▶ Max attenuation in passband: 0.3 dB (A_{max})
- ▶ Stopband: 1920 - 2200 MHz (*UMTS Band I*)
- ▶ Minimum attenuation in stopband: 50 dB (A_{min})
- ▶ Temperature range: -20° to 50°

□ Technology:

- ▶ Coupled Coaxial cavities
- ▶ Max unloaded Q: 2000-4500 (depending on size)
- ▶ Material: aluminum or brass (silver plated)

Ideal design ($Q_0=\text{infinite}$)

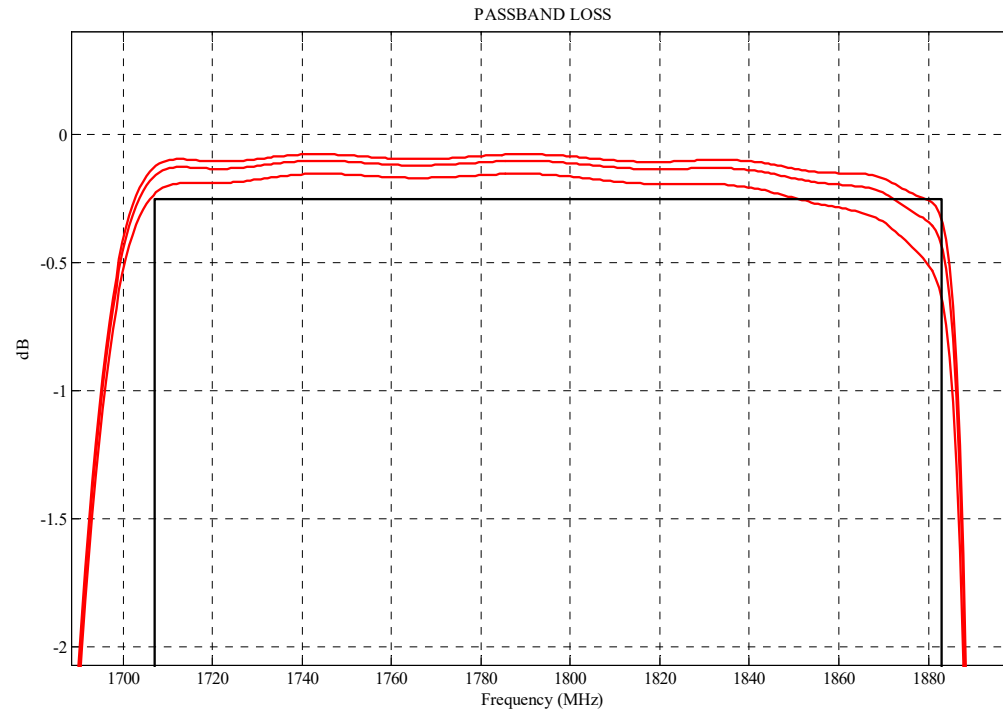
- Marginated specs:
Passband: 1707 - 1883 MHz , Stopband: 1917 - 2200 MHz
 $A_{\text{max}}=0.25$ dB, $A_{\text{min}}=52$ dB
- To satisfy the specifications in this case are sufficient 6 resonators with 2 transmission zeros (1918, 1945 MHz):



Passband losses are due only to reflection (0.012 dB)

But, what happens when the finite Q_0 is considered?

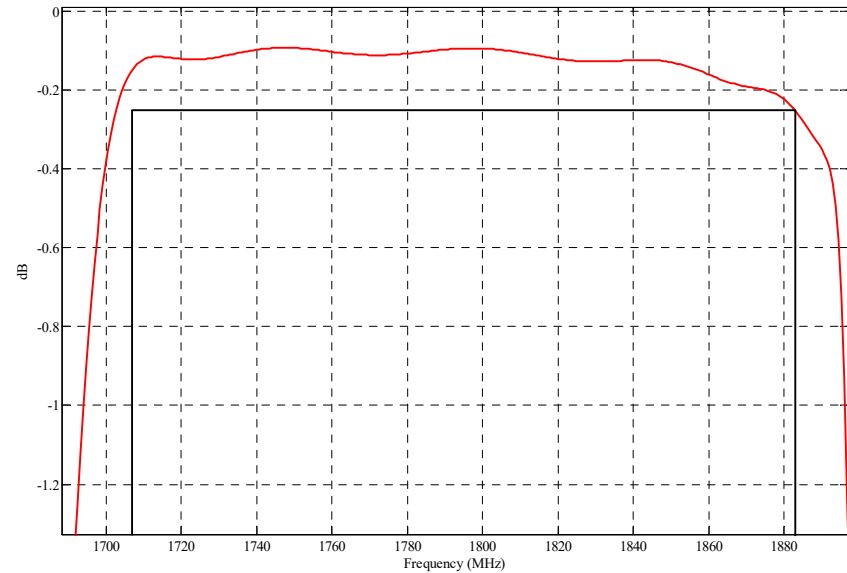
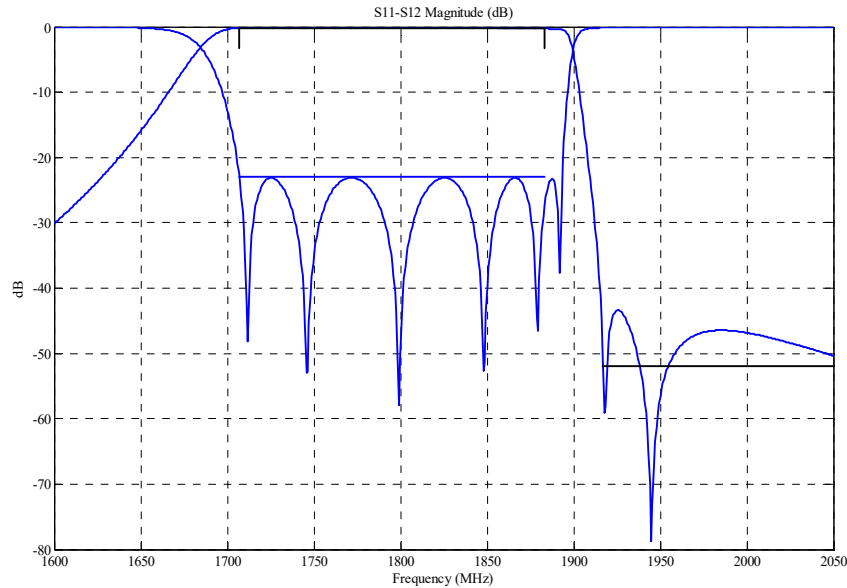
Passband Attenuation with $Q_0=2000, 3000, 4000$



Attenuation at upper passband limit does not satisfy the requirement even with $Q_0=4000$!

Let's increase the upper passband frequency (with $Q_0=3000$) and see what happens....

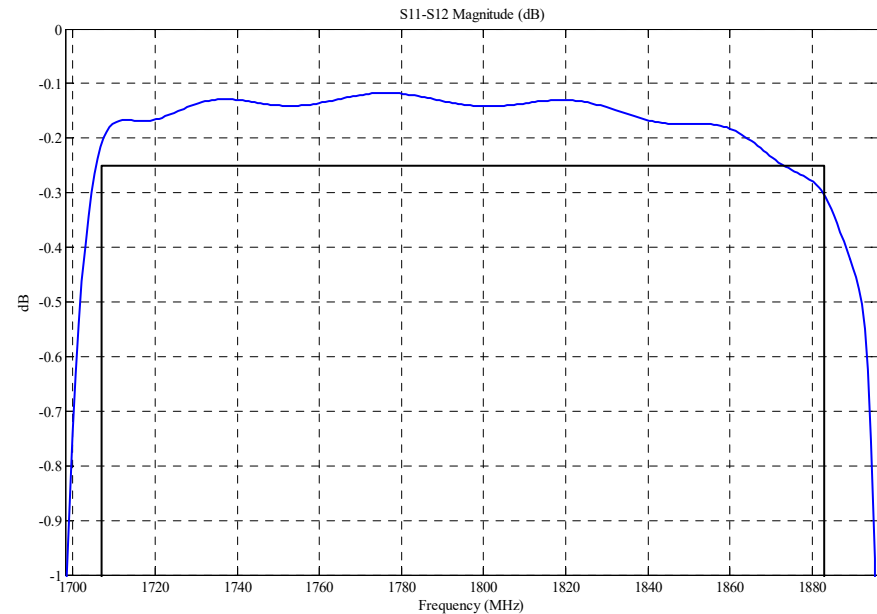
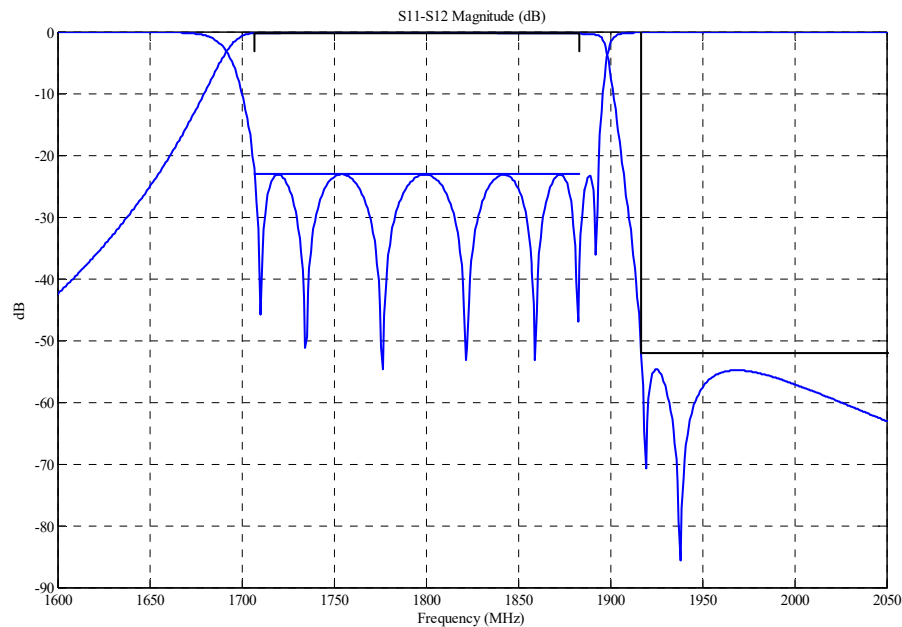
Upper passband frequency: 1883 → 1893.5 MHz



Now the attenuation requirement in passband is satisfied
but not the one in stopband!

We need to increase the order of the filter.
Let try with $n=7$...

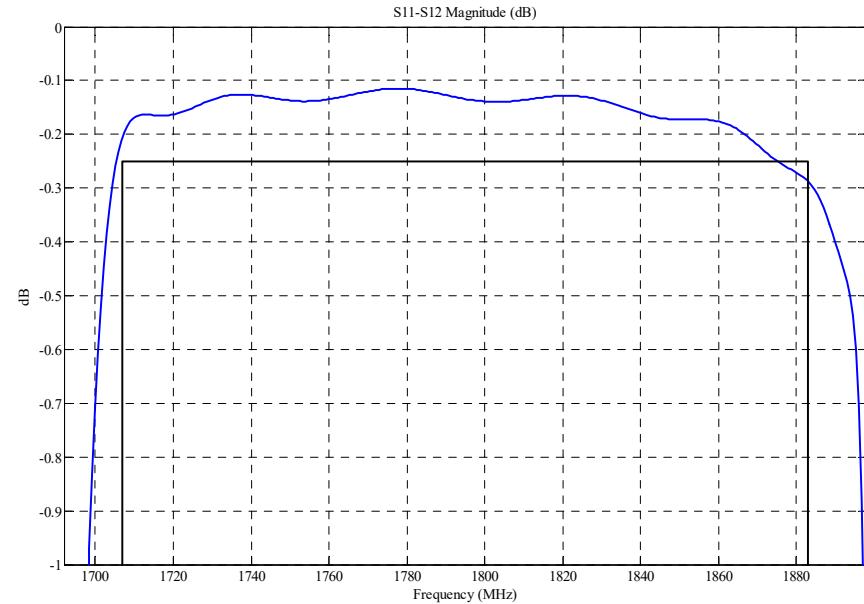
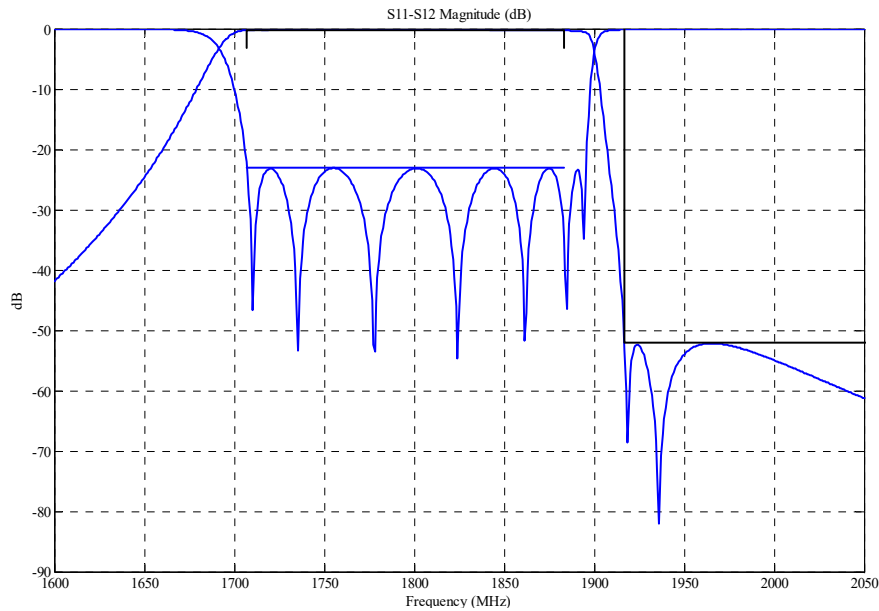
Response with $n=7$ (Zeros: 1919, 1938 MHz)



Stopband Attenuation OK, not the attenuation at the upper passband frequency!

As there is a little margin in the stopband attenuation (the value obtained is about 54.5 dB), let try to increase slightly the bandwidth....

Upper passband frequency: 1895.4 MHz (+1.9)



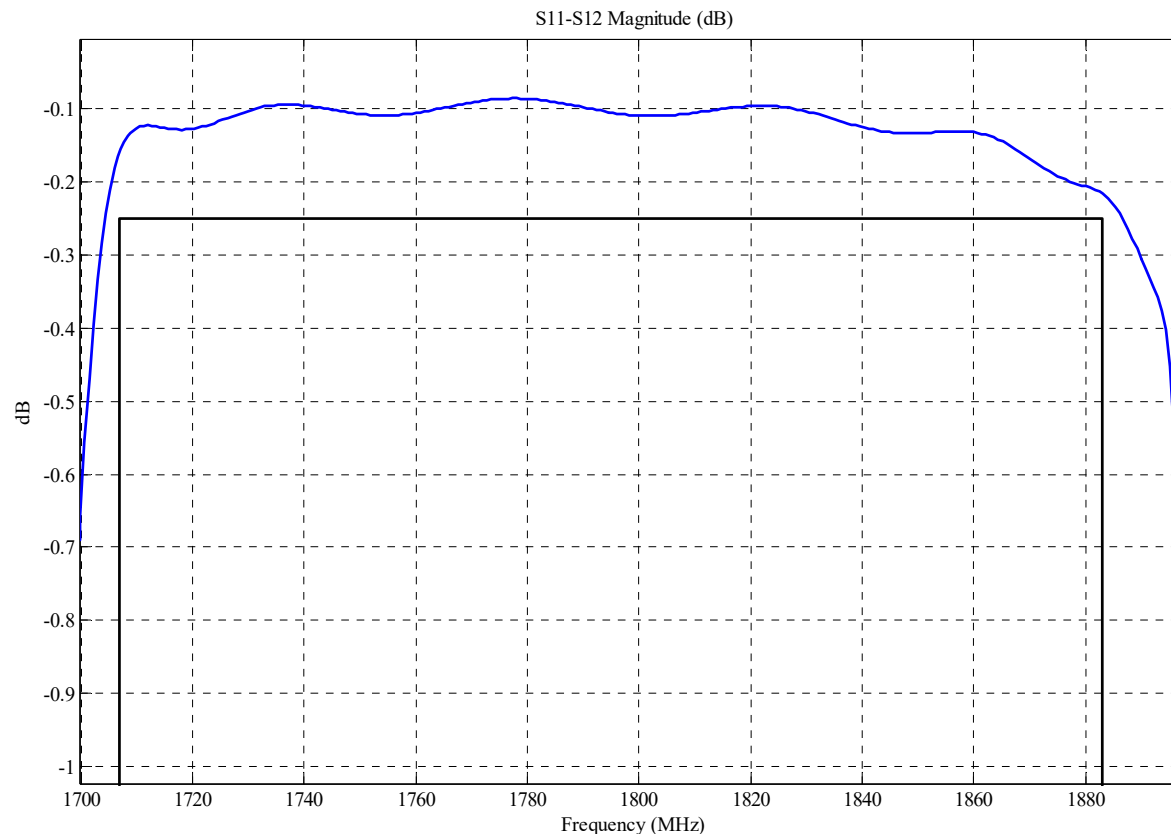
Nothing to do, passband attenuation is still larger than required...

Possible options:

- 1) Increase the filter order for the same band
- 2) Increase the number of zeros (3) reducing the band,
- 3) Increase the unloaded Q (to 4000).

Let try with the Q_0

Unloaded $Q=4000$



Finally we did it!

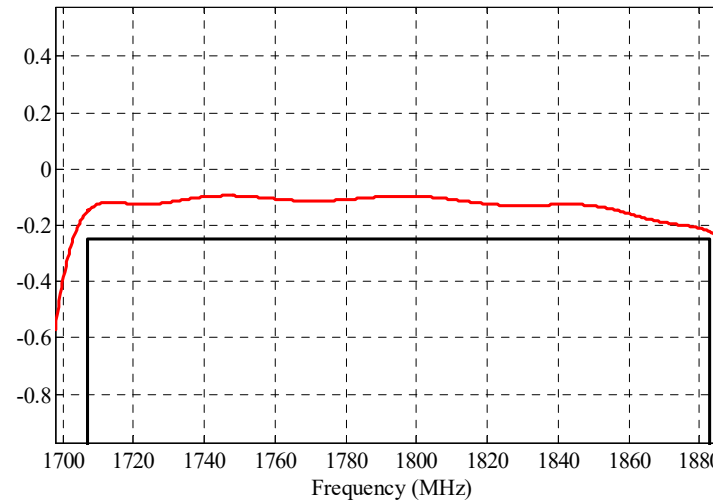
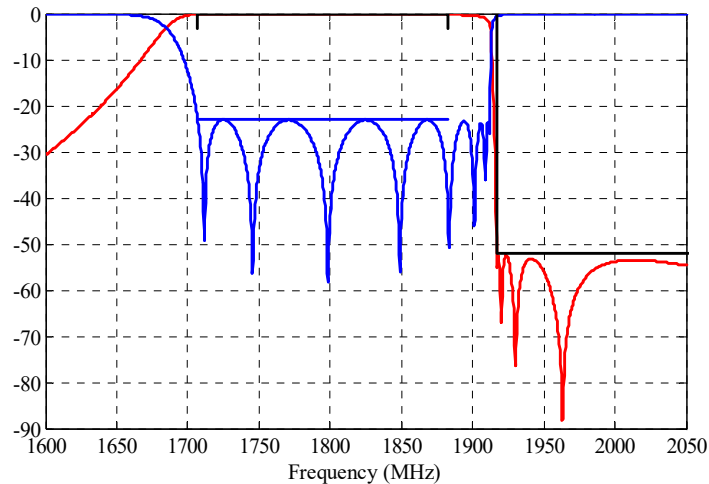
Note : This choice determines an increase of the overall filter size. Probably also the other options, if working, would produce an increase of size

Effect of increasing the bandwidth and filter order on the passband losses

- Passband losses:
 - ▶ decrease as the bandwidth increases
 - ▶ increase as the filter order increases
- The goal is to search for the best compromise between the increase of bandwidth and the increase of cavities number
- The larger is the number of cavities, the harder is to reduce losses by increasing of the bandwidth (if this latter is bounded by the stopband attenuation requirement). Additional transmission zeros may be requested

An other design result with $n=8$, $n_z=4$, $Q_0=3000$) (Bandwidth increased by 6.1 MHz)

Passband: 1707-1911.9, Zeros: 1917.1, 1920, 1930, 1963



Pro:

- The increase of bandwidth allows a smaller group delay in passband respect the previous design (13ns vs. 17ns)
- Better use of the overall volume (n even)

Con:

- More difficult to tune (4 zeros instead of 2)

Choice of the filter topology

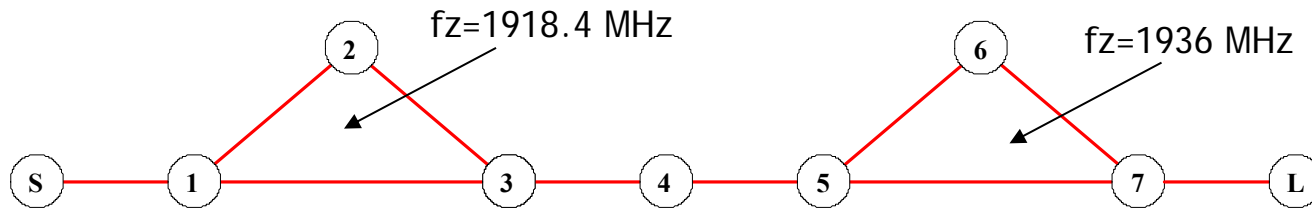
Cross coupled (cascaded-blocks) or extracted-pole?

- Both choices could be considered, however:
 - ▶ Implementation with cross-coupled topology is generally easier (NRN are not requested)
 - ▶ Extracted-pole allows an all in-line configuration (with 'appended resonators')
 - ▶ Filter tuning is more easy to do with simple cascaded-block structures (for instance, two triplets)

- In the following, both the choices will be illustrated

Synthesis with cascaded-block topology

- Selected structure:



Note: all the couplings have the same sign (zeros above passband)

Computed Coupling Parameters

Node	Resonance frequency (MHz)
1	1796.151520
2	1868.461765
3	1786.836246
4	1790.941048
5	1787.565007
6	1855.005106
7	1796.151520

Generalized coupling coeff.

0.07118
0.05904
0.04381
0.06042
0.06021
0.05140
0.04761
0.07928

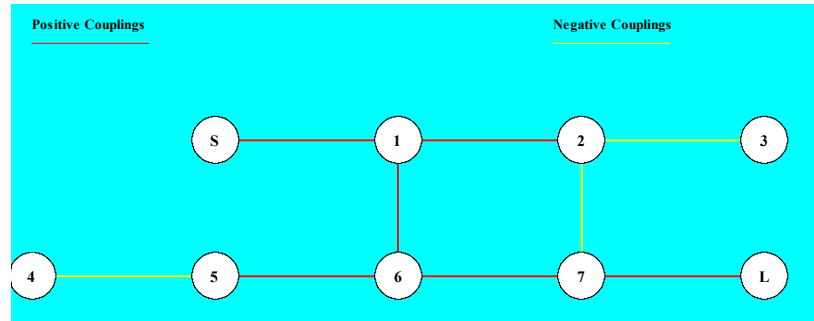
Node

1 - 2
1 - 3
2 - 3
3 - 4
4 - 5
5 - 6
5 - 7
6 - 7

Qext

8.59 (Source)
8.59 (Load)

Other cross-coupled topologies



Cul-de-sac

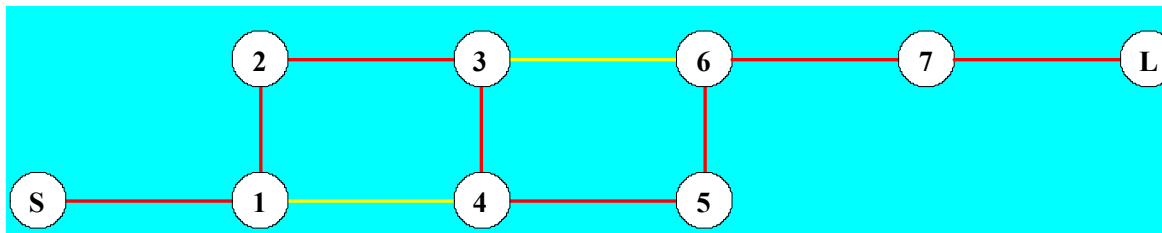
Node	Resonance frequency (MHz)
1	1796.151520
2	1795.686616
3	1852.976490
4	1882.707087
5	1762.549578
6	1795.686616
7	1796.151520

Generalized coupling coeff.

0.11640
0.06540
0.06540
-0.05474
-0.06540
-0.04098
0.07341
0.06540
0.11640

Node

S - 1
1 - 2
1 - 6
2 - 3
2 - 7
4 - 5
5 - 6
6 - 7
7 - L



Box Section 6

Node	Resonance frequency (MHz)
1	1796.151520
2	1894.529511
3	1861.755334
4	1784.979015
5	1753.734179
6	1795.686616
7	1796.151520

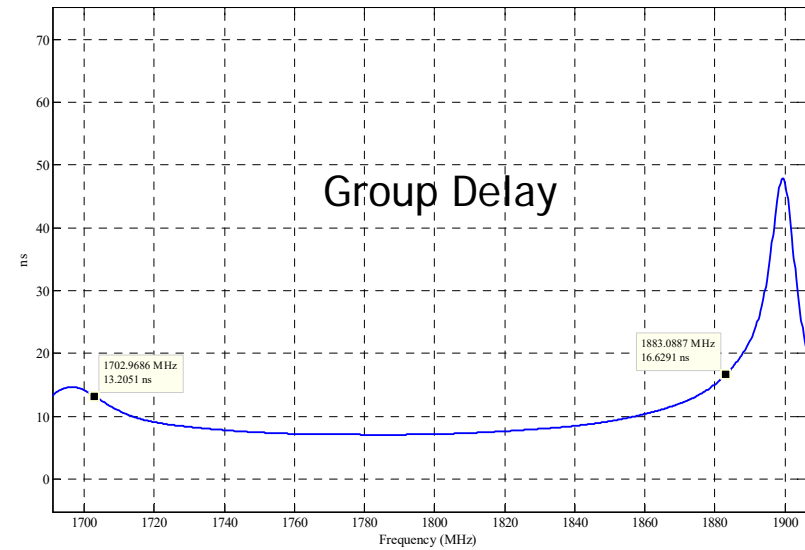
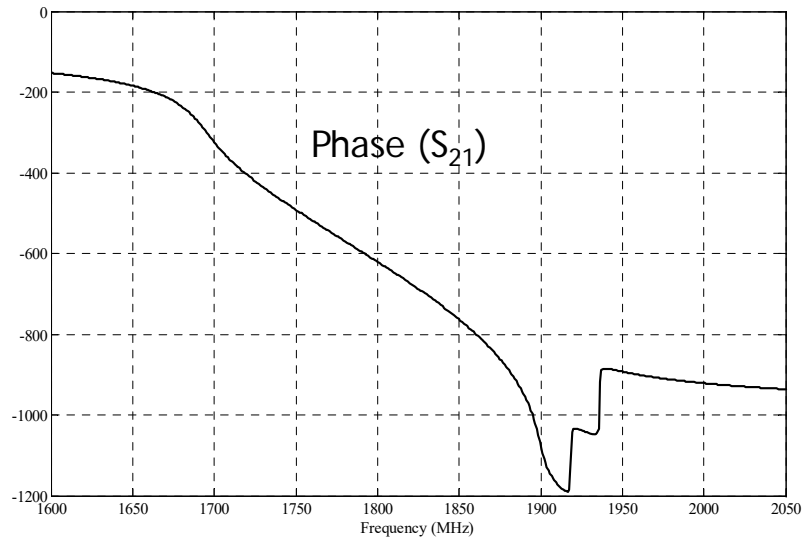
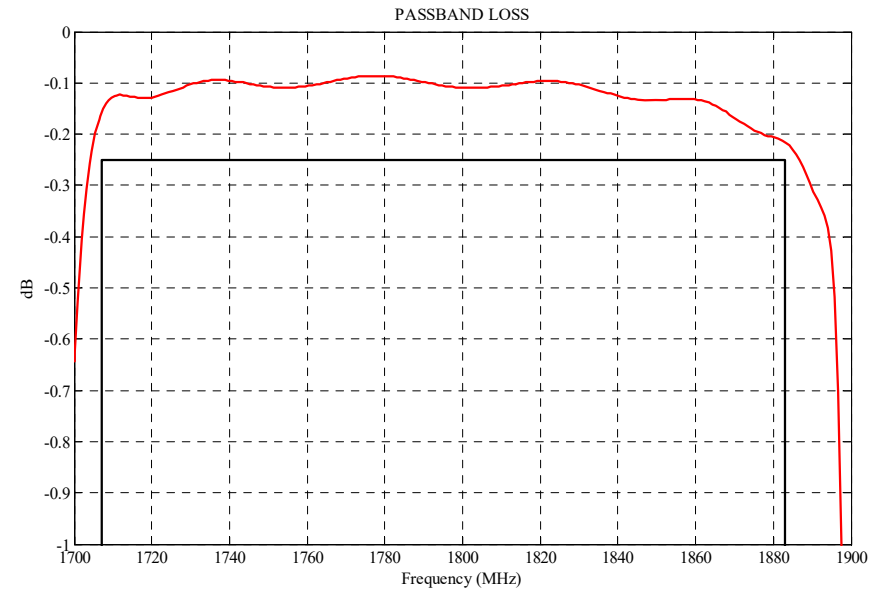
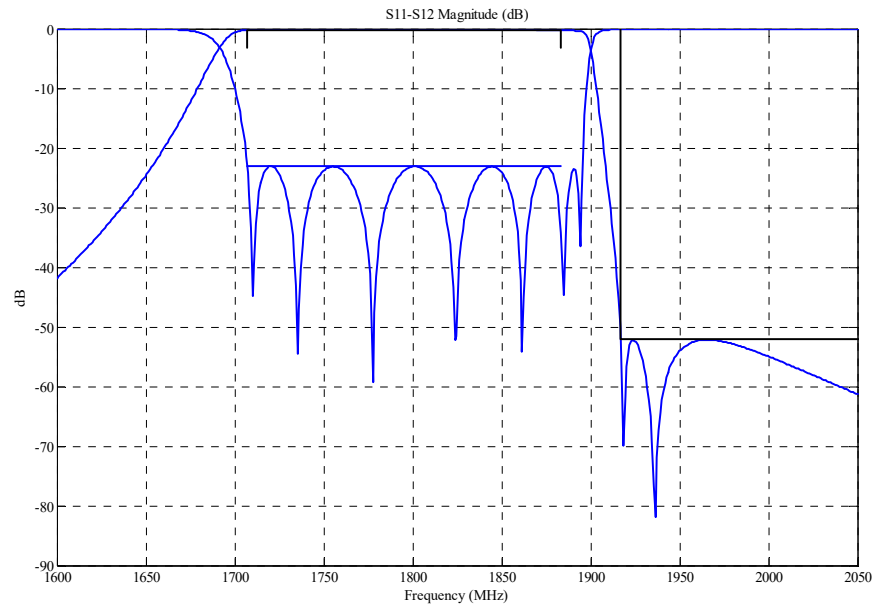
Generalized coupling coeff.

0.11640
0.02930
-0.08772
0.00816
0.03265
-0.04007
0.04837
0.05086
0.09248
0.11640

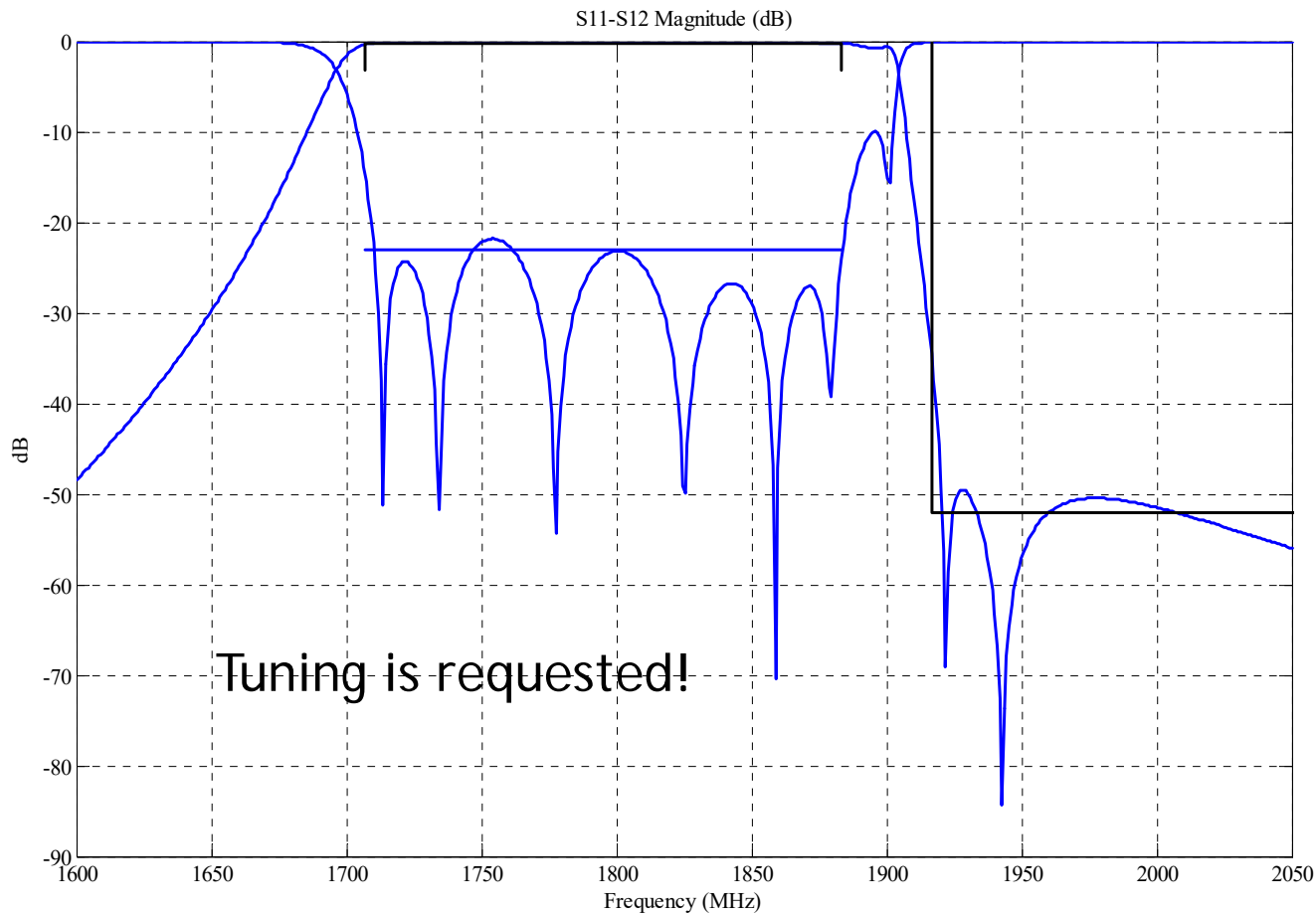
Node

S - 1
1 - 2
1 - 4
2 - 3
3 - 4
3 - 6
4 - 5
5 - 6
6 - 7
7 - L

Filter response (ideal elements)

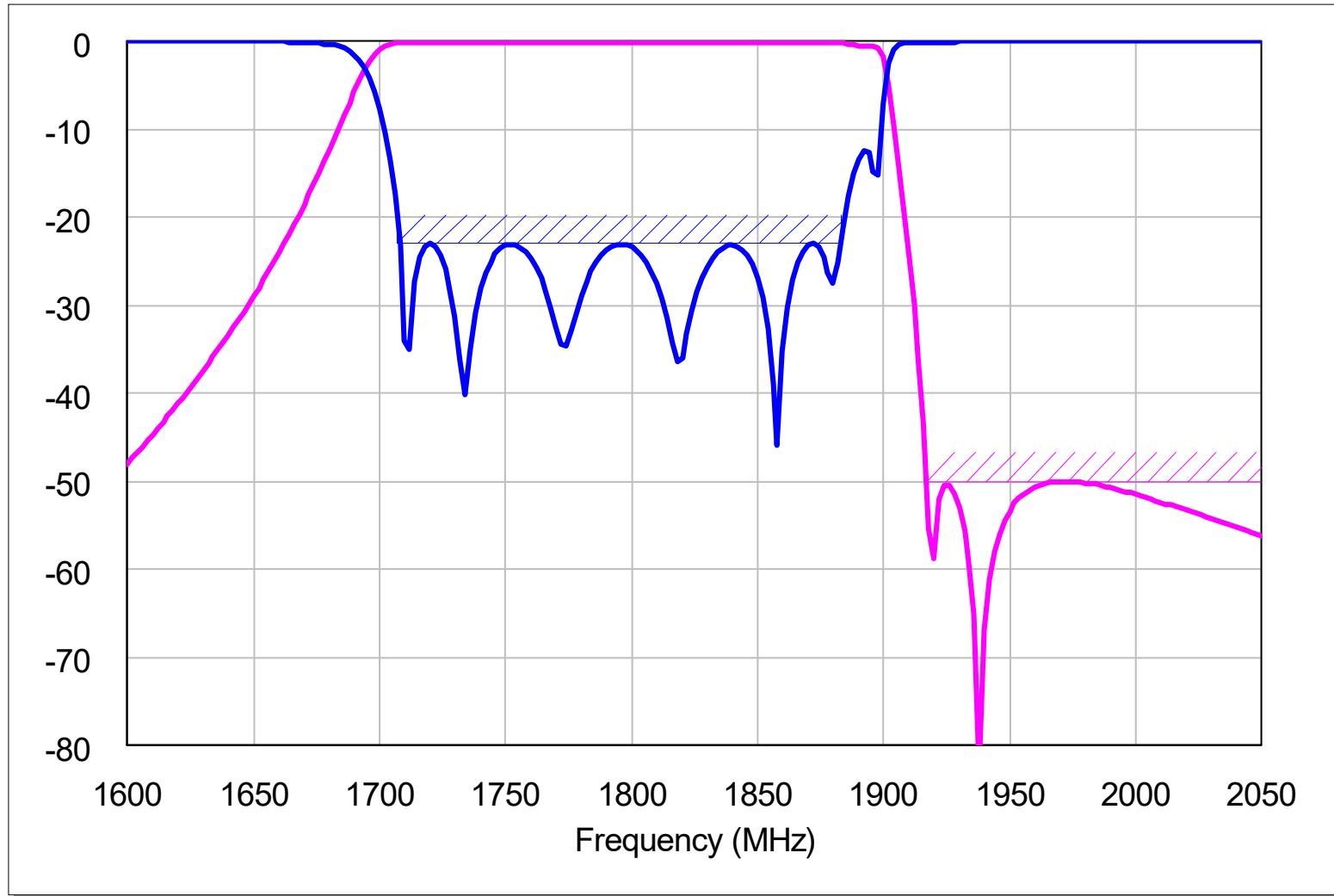


Response with couplings linearly variable with frequency



The normalized bandwidth is relatively large (10.5%)

Tuned response (circuit optimization)



Evaluation of temperature effects

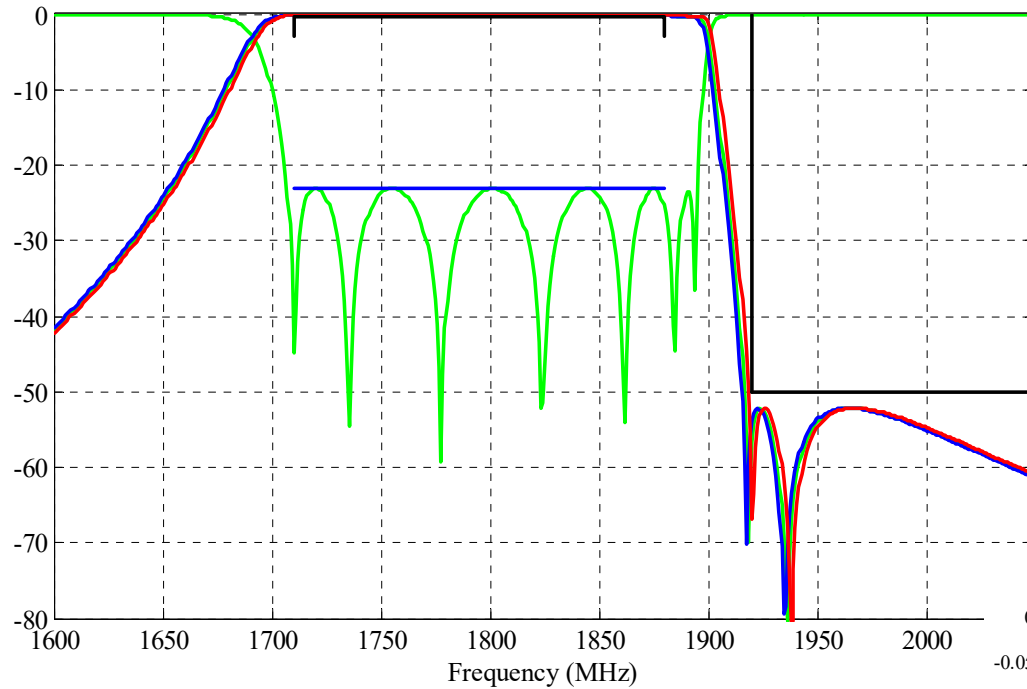
- Temperature variation affects the resonance frequency and the unload Q of the cavities
- A simple model for the coaxial cavities here used is represented by the following equations:

$$f_{ris,T} = f_{ris} (1 - K_f \cdot \Delta_T), \quad Q_{0,T} = \frac{Q_0}{\sqrt{1 + K_\rho \cdot \Delta_T}}$$

- For cavities made with one conductor, it has experimentally found that K_f has a value very close to the linear expansion coefficient of the cavity material ($23 \cdot 10^{-6} / ^\circ$ for aluminum).

- K_ρ is related to the resistivity increase with the temperature (for silver is around $6 \cdot 10^{-3}/^\circ$).
- Δ_T is evaluated with respect to room temperature (20°)
- The margin initially introduced on attenuation specifications (A_{min} , A_{max}) are needed for neutralizing these effects
- If a very strict control of temperature effects is needed, either special material (i.e. invar) or suitable combination of different conductors (for cavity, rod and screw) must be used

Filter response for $\Delta_T = -45^\circ, +25^\circ$



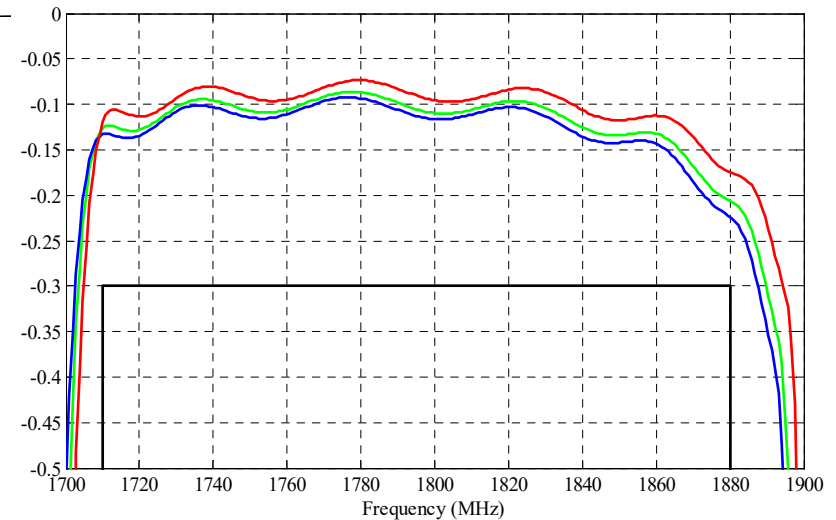
$$K_\rho = 6 \cdot 10^{-3} / ^\circ$$

$$K_f = 23 \cdot 10^{-6} / ^\circ$$

Green: Ref. Temp.

Red: -45°

Blue: $+25^\circ$

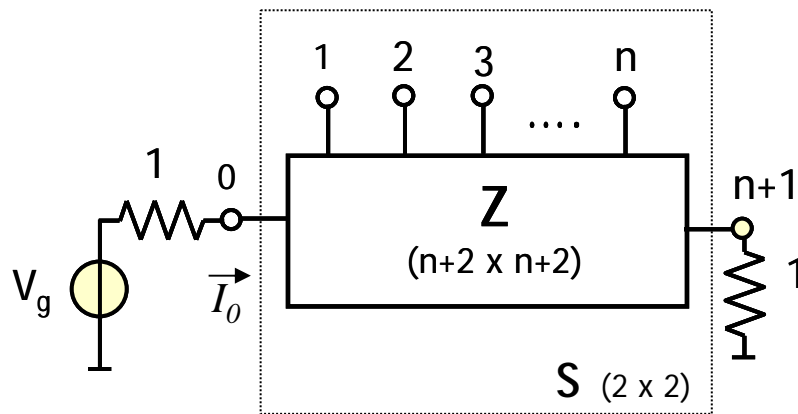


Voltages on resonators

- ❑ Power handling is a factor to be considered in base stations applications
- ❑ For a given input excitation, the voltages across the resonators vary along the filter
- ❑ Moreover, there is a multiplying effect (depending on the loaded Q of resonators) which increases the voltages with respect to the one at input and output of the filter
- ❑ The voltages may reach very high values, even with relatively small power level at the filter input
- ❑ The maximum admissible voltage across the cavities determines the filter power handling (above this voltage, breakdown phenomena may arise, quickly degrading the filter performances)

Evaluation of voltage on resonators (circuit model)

- Let consider the filter as a $n+2$ network (with the inner port placed in parallel to the shunt resonators):



$$V_k = I_0 \left(Z_{k,0} - \frac{Z_{k,n+1} Z_{n+1,0}}{1 + Z_{n+1,n+1}} \right)$$

$$V_0 = V_g \frac{1 + S_{1,1}}{2} = I_0 \left(Z_{0,0} - \frac{Z_{0,n+1} Z_{n+1,0}}{1 + Z_{n+1,n+1}} \right)$$

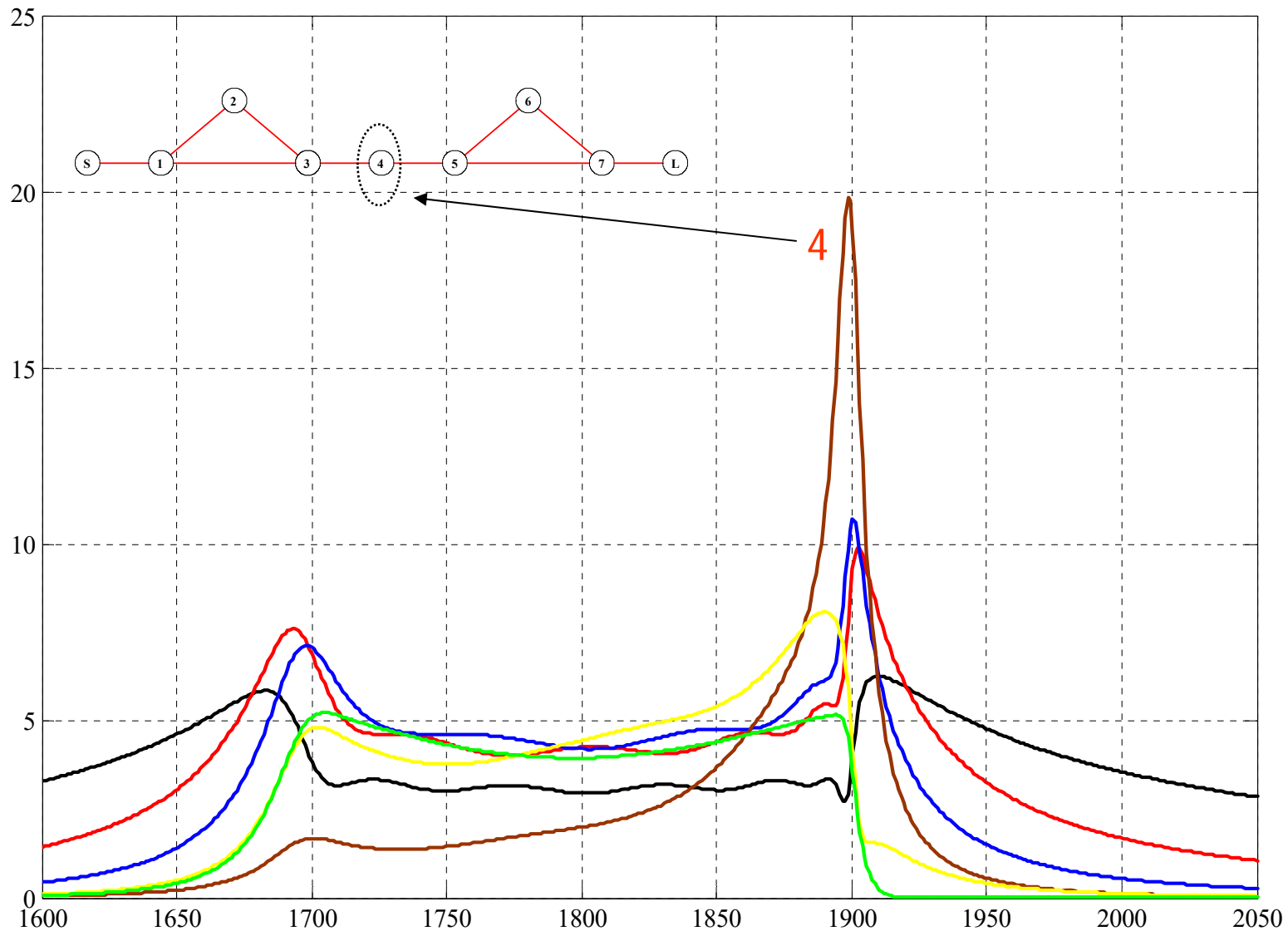
The maximum value of V_0 is obtained within the passband at frequencies where $S_{1,1}=0$: $V_{0,\max} = V_g/2$. Then, the following *excess voltage factor* F_k is introduced:

$$F_k = \frac{V_k}{V_{0,\max}} = (1 + S_{1,1}) \left(Z_{k,0} - \frac{Z_{k,n+1} Z_{n+1,0}}{1 + Z_{n+1,n+1}} \right) \left(Z_{0,0} - \frac{Z_{0,n+1} Z_{n+1,0}}{1 + Z_{n+1,n+1}} \right)^{-1}$$

Excess voltages on real cavities

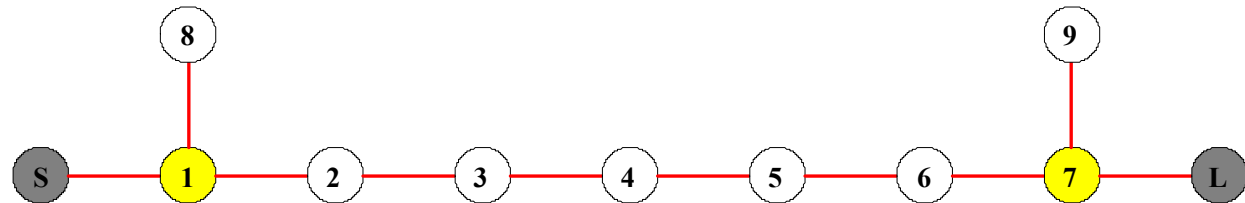
- The values of F_k depend on the equivalent parameter (C_{eq} or L_{eq}) of the resonators and on the source, load resistance. These parameters must be then specified when the lowpass to bandpass transformation is realized
- Voltages across real cavities are related to the voltages on the equivalent resonators through a **scaling coefficient**
- The scaling coefficient can be determined using EM simulations (once the reference section of the resonators has been defined)
- The values of F_k depends also on the specific topology selected (the sequence of blocks in the cascaded-block topology)

F_k vs. frequency for the synthesized (7+2) filter (coaxial cavities assumed with $Z_c=60$ Ohm)



Filter synthesized with extracted-pole topology

Topology:



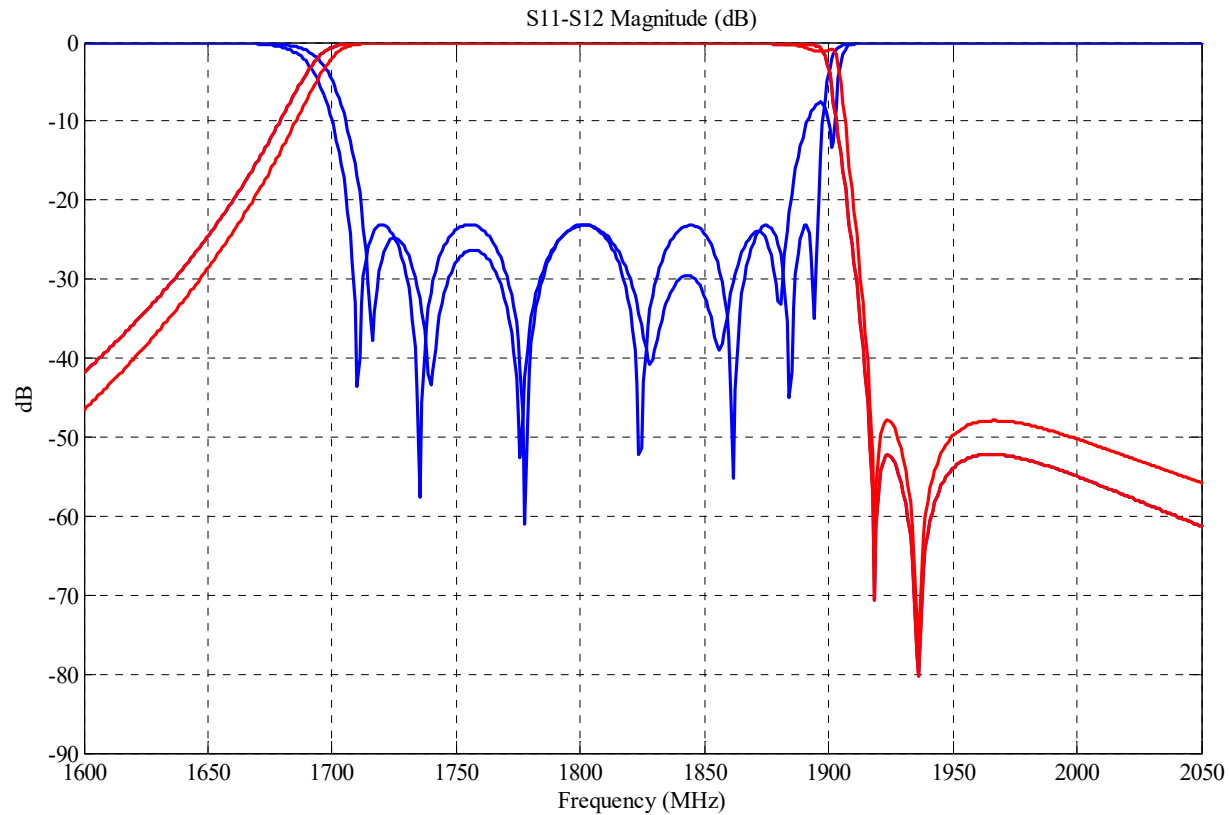
Node	Resonance frequency (MHz)	
1	0.000000	
2	1780.973826	
3	1786.836246	
4	1790.941048	
5	1787.565007	
6	1788.134537	Qext(source): 0.3080
7	0.000000	Qext(load): 0.6585
8	1918.400000	
9	1936.000000	

Generalized
Coupling Parameters:

Sign of Non Resonating Node:
-1 0 0 0 0 0 -1 0 0

Generalized Coupling Coeff.	Node
0.51657	1 - 2
0.07352	2 - 3
0.06042	3 - 4
0.06021	4 - 5
0.07007	5 - 6
0.34209	6 - 7
0.38332	1 - 8
0.36932	7 - 9

Filter response

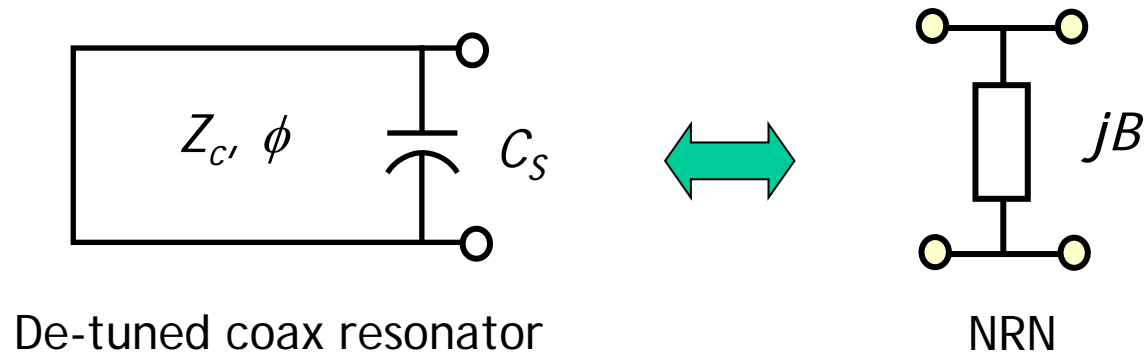


Solid line: ideal response

Dashed line: couplings and
NRN variable with freq.

Implementation of NRN in coax technology

- NRNs are practically implemented with de-tuned coax resonators:

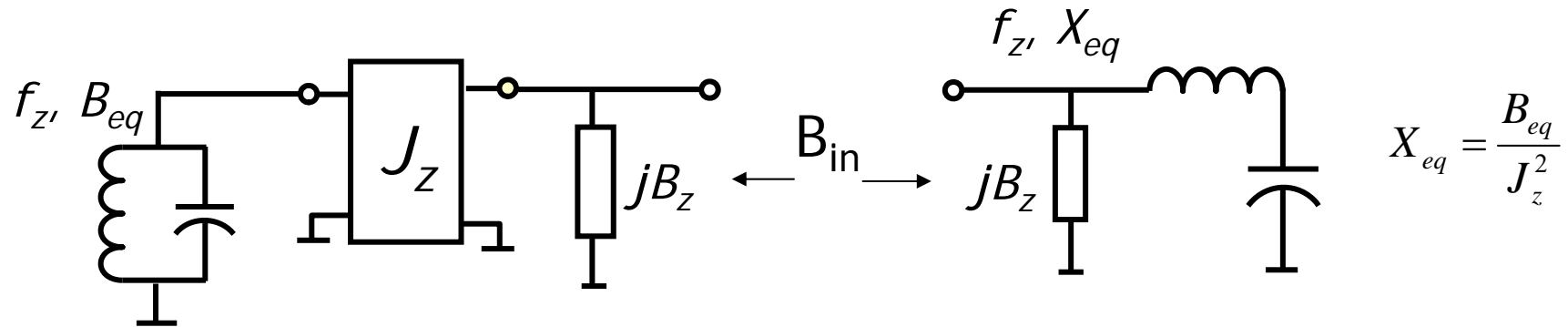


$$B = \omega_0 C_s - Y_c \tan(\phi_0)$$

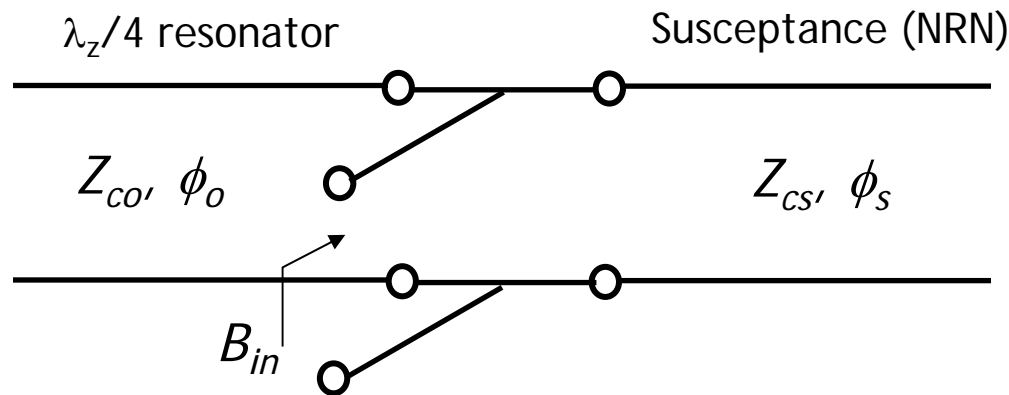
This is a gross approximation of a NRN. In fact B varies with frequency

Another topology for extracted-pole filters

- Transformation of extracted-pole block:



Using transmission lines:



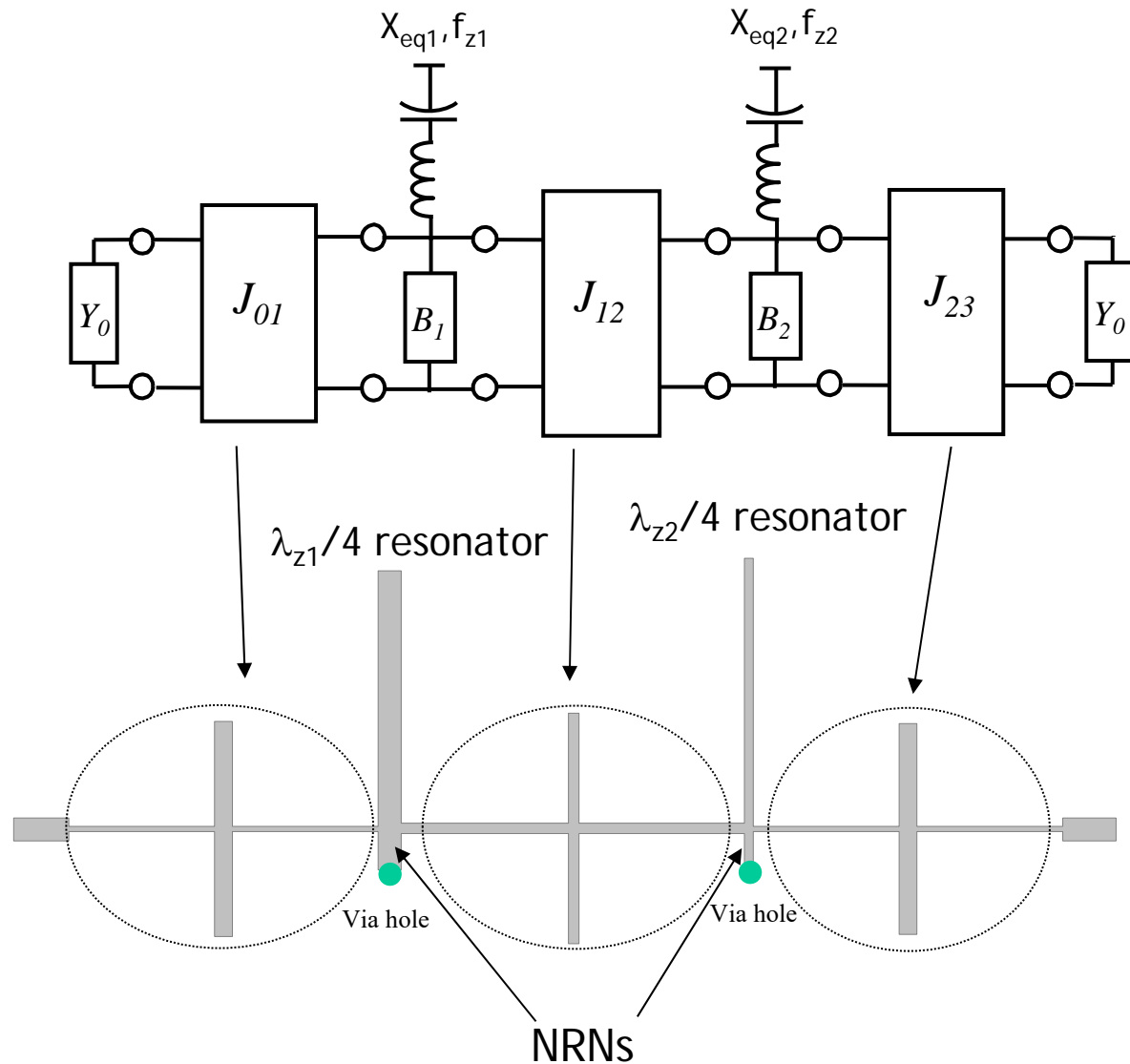
$$\phi_o = \frac{\pi}{4}, \quad Z_{co} = \frac{2}{\pi} X_{eq}$$

$$Z_{cs} = \frac{\tan(\phi_s)}{B_z}$$

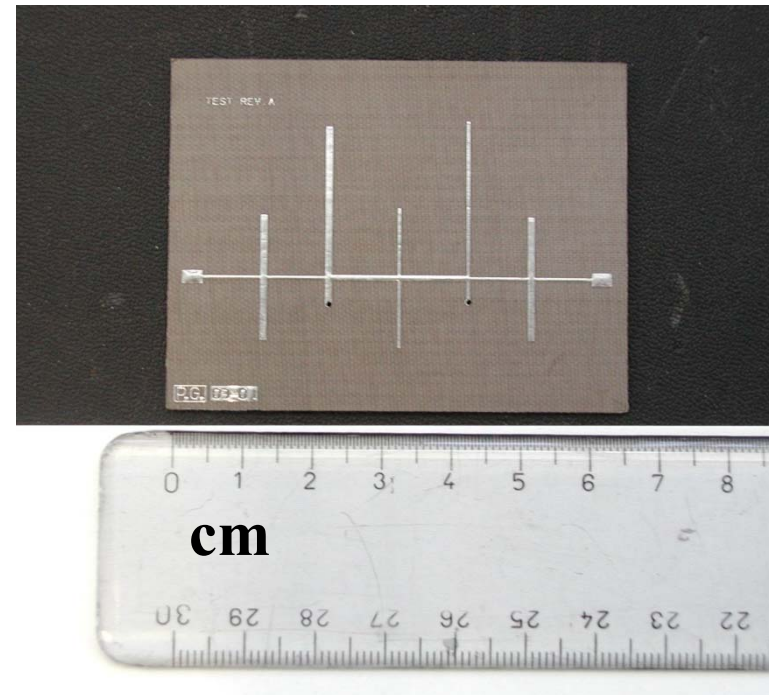
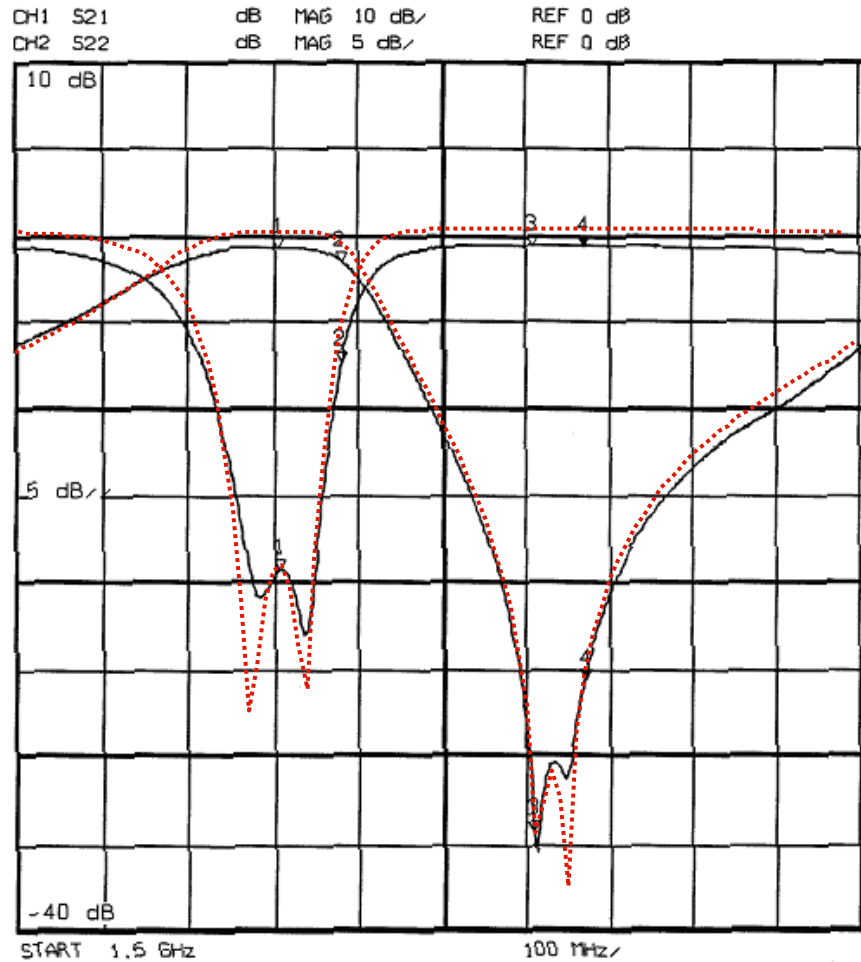
Line 2 is short circuited if $B_z < 0$

Fully canonical Microstrip Filter

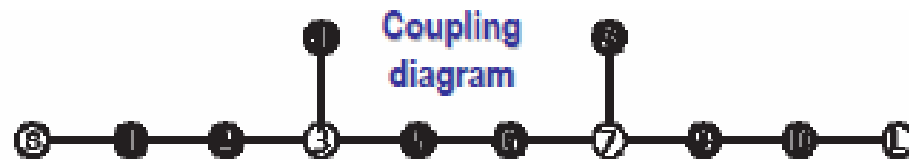
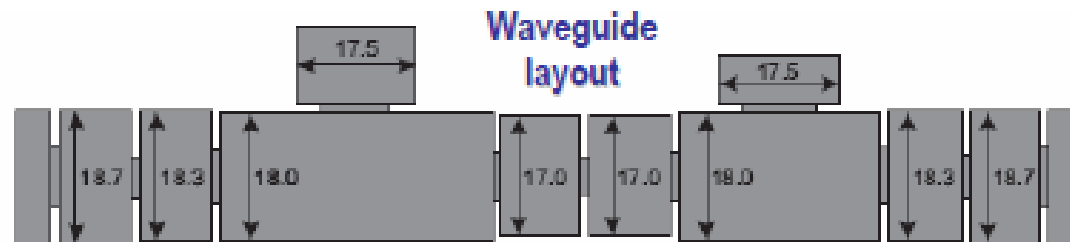
Filter Specifications	
Order	2
Pass Band (start)	1.805 GHz
Pass Band (end)	1.880 GHz
Stop Band (start)	2.110 GHz
Stop Band (end)	2.170 GHz
Return Loss	20 dB



Implemented filter & measured response



NRN Filter in rectangular waveguide



- Flexible transmission zeros
- Reduced frequency sensitivity
- Tapering: remove spikes

