

## Helical Filters

For high Q value resonators at UHF frequencies, a cylindrical rod of one-quarter wavelength long is placed inside a cavity. This cavity can be cylindrical but in many cases, the cavity is rectangular for ease of construction. Such filters are called coaxial filters, as the cross-section of the filter is a coaxial transmission line.

At VHF frequencies, the size of the coaxial filter becomes too large for many applications. To reduce the length of the resonator, the centre quarter wave resonator is wound as a helix. Often the helix is wound on a former. One has to ensure that the material for the former is not lossy at the operating frequency. Polyethylene is suitable for low loss formers.

The resonators have shields between them to confine the fields for the resonators. The resonators are coupled by having the appropriate shield height, so that some of the field of a resonator couples to the next resonator. The coupling is controlled by adjusting the difference between the height of the shield and the top of the helix as shown by dimension "h" in figure 45. Alternately, the coupling can be achieved by having a coupling hole in the centre of a full-height shield or by having a slot at the bottom of the shield. The choice of coupling aperture depends on the construction of the cavity.

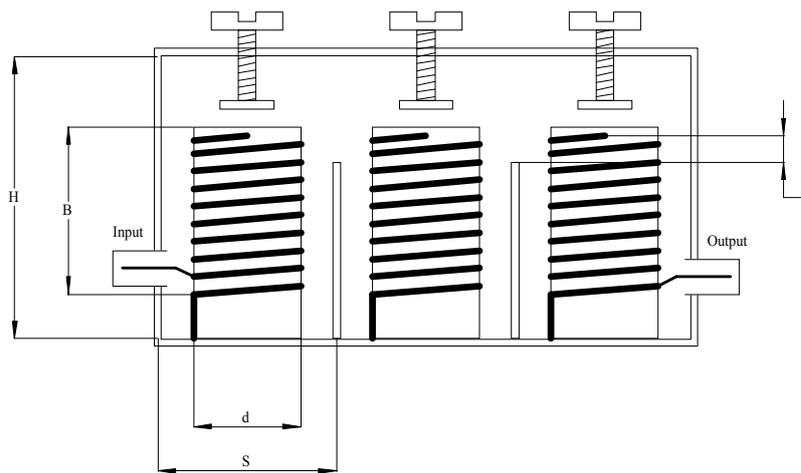


Figure 45. Drawing of a helical filter showing the sizes used in the formulae.

### Formula for Helical Filters (from Zverev)

S is the width of the rectangular resonator cavity. (units m)

H is the height of the resonator cavity (units m)

B is the height of the helix (units m)

d is the diameter of the helix (units m)

N is the number of turns of the helix

$Z_0$  is the characteristic impedance of the helical transmission line in the cavity.

$\delta$  is the skin depth. The wire diameter should be  $> 5$  times the skin depth.

$Q_0$  is the unloaded Q of the resonator cavity (f is in MHz).

If instead of a rectangular cavity, a cylindrical cavity is used, the diameter D of the cavity is  $D = 1.2 S$ .

For the cavity:

$$d=0.66S \quad \text{Eqn. 15}$$

$$B=S \quad \text{Eqn. 16}$$

$$H=1.6S \quad \text{Eqn. 17}$$

$$Q_0 = 2363 \times S\sqrt{f} \quad \text{Eqn. 18}$$

For the highest Q value, the wire diameter of the helix is chosen such that it is half the pitch of the helix. The gap between the wires is thus the same as the wire diameter.

For the helix:

$$N = \frac{40.6}{S \times f} \quad \text{Eqn. 19}$$

$$\delta = \frac{6.6 \times 10^{-5}}{\sqrt{f}} \quad \text{Eqn. 20}$$

$$Z_0 = \frac{2070}{S \times f} \quad \text{Eqn. 21}$$

$$\frac{R_b}{Z_0} = \frac{\pi}{4} \left( \frac{1}{Q_d} - \frac{1}{Q_0} \right) \quad \text{Eqn. 22}$$

The doubly loaded input  $Q_d$  is defined as  $\frac{1}{2} Q_1$  for the input tapping point calculation and  $\frac{1}{2} Q_n$  for the output tapping point calculation, and represents the total load seen by the resonator if there is a very small insertion loss as is normally the case. The electrical length  $\theta$  from the bottom of the helix to the point where the input or output connects to the helix is given by:

$$\text{Sin}(\theta) = \sqrt{\frac{R_b R_{tap}}{2Z_0^2}} \quad \text{Eqn. 23}$$

The gap between the shield and the top of the helix is given by:

$$K = 0.071 \left( \frac{h}{d} \right)^{1.91} \quad \text{Eqn. 24}$$

It should be noted that when the bandwidth increases the gap between the top of the shield and the top of the coil increases. For wideband filters no shield is required, thus reducing the cost of construction. In that instance, it is also possible to vary the coupling by varying the spacing between the resonators.

### **Design Example:**

Design a helical filter to have a 1 MHz bandwidth at 100 MHz. A diecast box of 112 mm width and 50 mm depth and 90 mm height is available. This can be split into two cavities 56 x 50 x 90 mm. The average side S is 53 mm. The H/S ratio is  $90/53 = 1.7$ , which is close enough to the  $h = 1.6S$  value from equation 17

The K and Q values required can be obtained from table 1. Since for two resonators and Butterworth filters the K and Q values are independent of the insertion loss, equations 4 and 5 can be used to calculate the required K and Q values.

Equation 4 is  $q_0 = q_n = 2\text{Sin} \frac{(2-1)\pi}{2n} = 2\text{Sin} \frac{\pi}{2n}$  and for a two-resonator filter  $n = 2$ , this gives  $q = 2\text{Sin}(45 \text{ deg}) = 1.4142$

Equation 5 is  $k_{ij} = \frac{1}{2\sqrt{\text{Sin} \frac{(2i-1)\pi}{2n} \text{Sin} \frac{(2i+1)\pi}{2n}}}$  and for a two-resonator filter this gives  $k = 1/(2\text{Sin}(45 \text{ deg})) = 0.7071$

Applying equation 18 gives the unloaded Q of the cavity as:  $Q_0 = 2363 * 0.053 * 10 = 1252$ . The normalised unloaded Q is  $1252 * 1/100 = 12.52$ . From the filter table 1, the resulting filter will thus have close to 1 dB insertion loss.

Since  $S = 53 \text{ mm}$ , applying equation 15 gives the coil diameter as:  $d = 35 \text{ mm}$ . From equation 17, the coil height is 53 mm and equation 19 indicates it requires 7.6 turns. The optimum wire diameter is thus  $53/(2*7.6) = 3.49 \text{ mm}$ . A one-eighth of an inch copper tube is 3.2 mm and is used for winding the coil. The minimum wire diameter of five times the skin depth (equation 20) is 0.033 mm. For practical windings, the minimum skin depth requirement is normally satisfied.

From the filter tables  $q_1 = q_n = 1.4142$  and  $k_{12} = 0.7071$ . Denormalising this gives  $Q_1 = Q_n = 1.4142 * 100/1 = 141.42$  and  $K_{12} = 0.7071 * 1/100 = 0.007071$ .  $Q_1$  and  $Q_n$  are the loaded Q values, which are due to the input and output load impedance being coupled into the resonators and thus appearing as a load to the resonators. The doubly loaded input Q is defined as  $\frac{1}{2} Q_1$  and represents the total load seen by the resonator if there is a very small insertion loss as is the case here. The doubly loaded output Q is defined as  $\frac{1}{2} Q_n$ .  $Q_d$  is thus 70.71 for both the input and output. Putting this in the expression for  $R_b/Z_0$  gives:  $R_b/Z_0 = 0.0105$ . Since we want the filter to be have 50  $\Omega$  input and output impedance,  $R_{\text{tap}} = 50 \Omega$  resulting in  $\theta = 1.48$  degrees. Since the coil wire length is  $7.6 * 35 * \pi = 842 \text{ mm}$  and corresponds to 90 electrical degrees, the tapping point should be  $1.48 * 842 / 90 = 13.9 \text{ mm}$  from the bottom of the helix. In practice, the tapping point needs to be fine-tuned, to match the actual conditions after construction and include losses of the former used for winding the helix on.

For this Butterworth filter, the tapping points for the input and output are the same. For Chebychev or Bessel filters, or for lossy higher order Butterworth filters, the input and output tapping points will be different, and the above calculation needs to be performed for  $Q_1$  and for  $Q_n$ .

The filter was constructed and performed exactly as expected. The measured bandwidth for the helical filter was 1 MHz and the insertion loss was 0.8 dB. A photograph of the filter is shown in figure 46 and the measured response is shown on figure 47.

The tuning procedure for adjusting the tapping point for the input and output, and for verifying and if needed adjusting the coupling, is described in Zverev on page 518 and Figure 9.23. This same procedure is used to determine the coupling factors for interdigital filters as shown in figures 51, 52, and equations 29 and 30 of these notes.

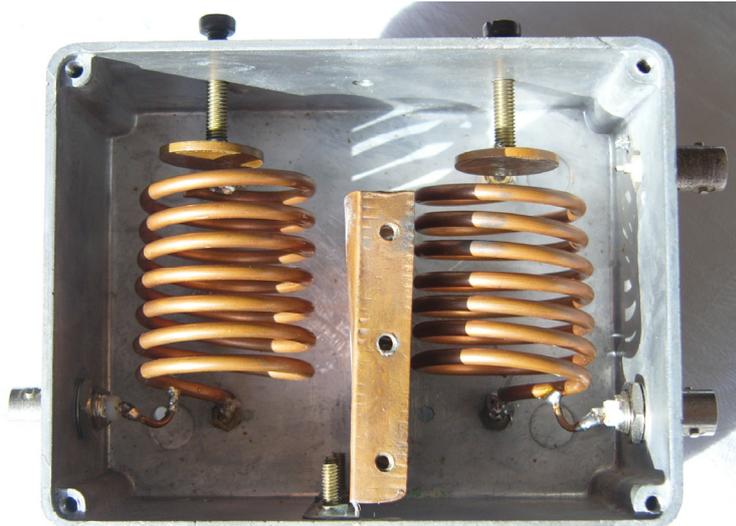


Figure 46. 101 MHz helical filter 1 MHz bandwidth

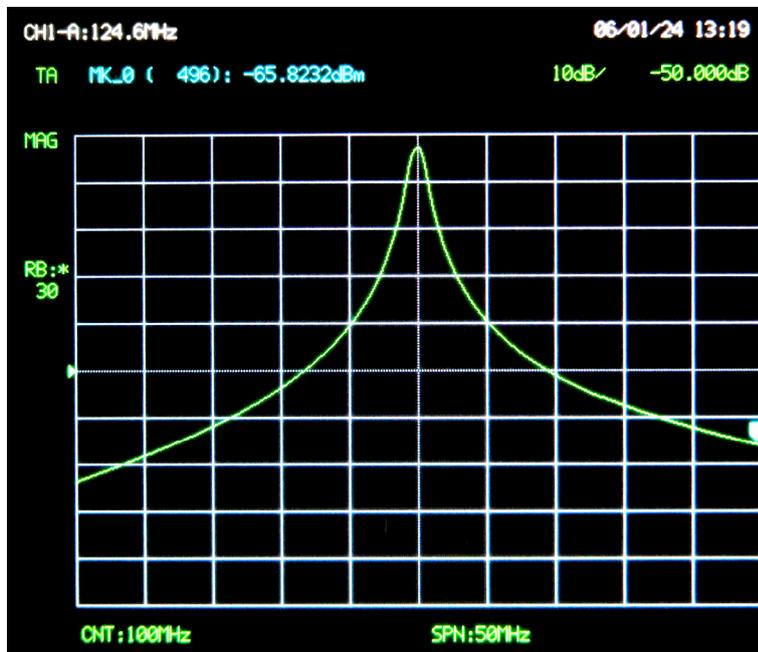


Figure 47. Frequency response of 101 MHz helical filter, vertical 10 dB/div.



Figure 48. Commercial Helical Filters. Temwell Left, Toko Right.

The left filter in figure 48 shows a 3-resonator 142 MHz helical filter with a bandwidth of 6 MHz from Temwell ([www.temwell.com.tw](http://www.temwell.com.tw)). Each resonator is 8mm wide and 12mm high. Temwell make filters from 45 MHz to 2.6 GHz. The right filter is an older, 3-resonator, 137 MHz helical filter with a bandwidth of 2 MHz made by Toko ([www.toko.co.jp](http://www.toko.co.jp)). The resonators are 12mm wide and 18mm high. Toko now only make helical filters for 350 MHz to 1.5 GHz. The measured dimensions of these filters are the outside dimensions of the cavities. The ratio of the internal dimensions thus satisfies equation 17.

## Interdigital Filters

With interdigital filters, the resonators are grounded at alternate ends. This reduces the coupling between the resonators and as a result, a compact filter can be constructed that has no walls between the individual resonators. A drawing of an interdigital filter is shown in figure 49a.

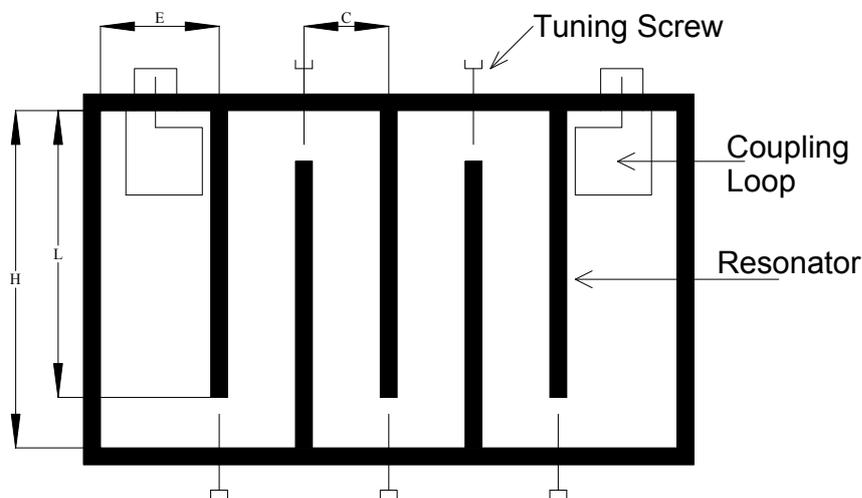


Figure 49a. Interdigital filter dimensions.

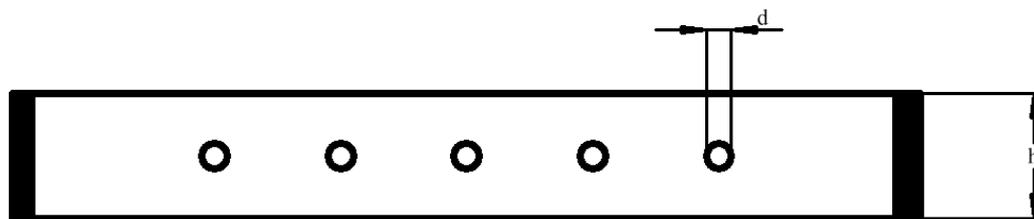


Figure 49b. Interdigital filter dimensions.

For a small bandwidth (<10%) M. Dishal in his paper, "A Simple Design Procedure for Small Percentage Bandwidth Round Rod Interdigital Filters" MTT, Vol 13, No5, Sep 1965, pp 696 – 698, provided the following equations for designing interdigital filters:

$$\frac{E}{h} = 0.6$$

Eqn. 25

$$\text{Log } K = \left[ -1.37 \frac{C}{h} + 0.91 \frac{d}{h} - 0.048 \right] \quad \text{Eqn. 26}$$

Where  $h$  (figure 49b) is the spacing between the walls and  $d$  is the rod diameter. Knowing  $K$  from the filter tables and  $d$  and  $h$  from the design specifications, allows the spacing  $C$  between the resonators to be calculated.

In addition, the impedance of the resonator is given by:

$$Z_0 = 138 \text{Log} \left[ \frac{4h}{\pi d} \right] \tanh \left( \frac{\pi E}{h} \right) \quad \text{Eqn. 27}$$

The singly loaded  $Q$  of the resonator is:

$$Q_1 = \frac{\pi R}{4Z_0} \frac{1}{\text{Sin}^2 \left( \frac{\pi \ell_1}{2L} \right)} \quad Q_n = \frac{\pi R}{4Z_0} \frac{1}{\text{Sin}^2 \left( \frac{\pi \ell_n}{2L} \right)} \quad \text{Eqn. 28}$$

Since  $Q_1$  and  $Q_n$  can be obtained by de-normalising  $q_1$  and  $q_n$  from filter tables,  $R$  is the desired termination impedance, (usually  $50 \Omega$ ) and  $Z_0$  is the characteristic impedance of the resonator as obtained from equation 27, that the tapping points  $\ell_1$  and  $\ell_n$  can be determined using equation 28. Interdigital filters can be constructed in two forms:

### **Round Rod Interdigital Filters**

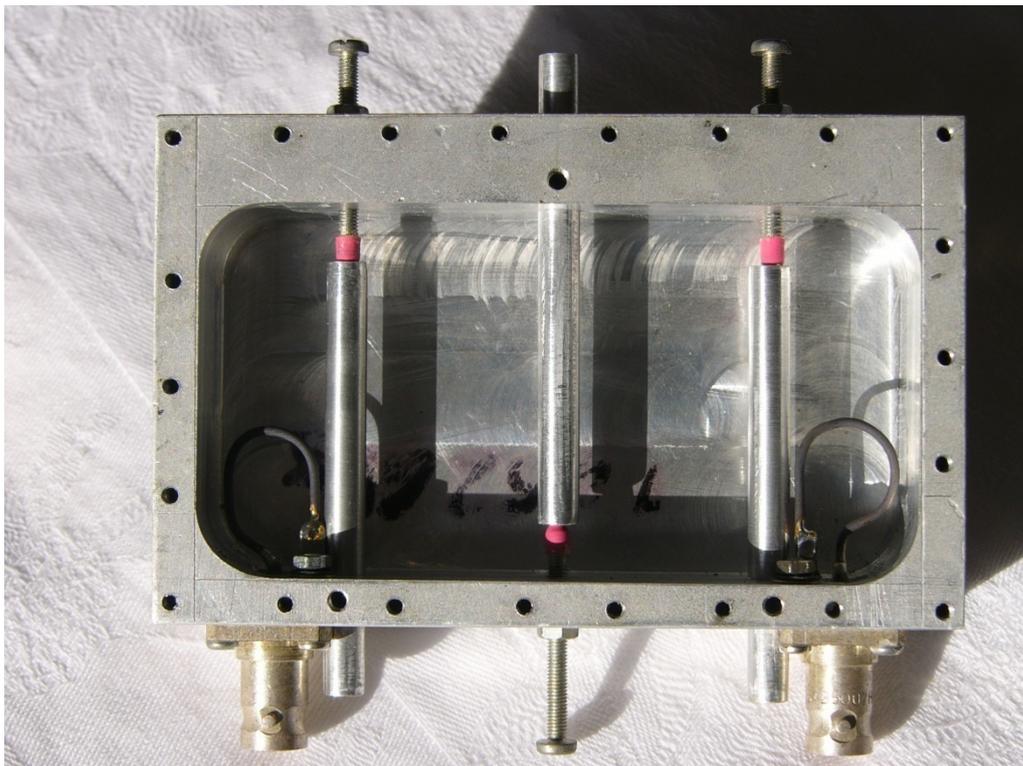


Figure 50. Interdigital filter for operation at 940 MHz, with top plate removed.

Similar to the Helical filter, the larger the filter cavity, the lower the insertion loss of the filter and the bigger the power handling capability of the filter. The Round Rod Interdigital filters are reasonably large and as a result have a low insertion loss and a

high power handling capability, permitting them to be used at the output of transmitters. These filters normally have bandwidths less than 10%. The filters become very large for frequencies below 300 MHz. The coupling into the filter can be achieved by coupling loops as shown in figure 50, or by direct tapping to the resonator rod. In most instances the resonator rods length can be adjusted. This provides a coarse tuning adjustment. In the figure 50, the Allan Key locking screws for the resonator rods can be seen at the base of each of the resonators allow the course adjustment to be performed. A small variable capacitor, in the form of an insulated screw that slides inside the hollow resonator rod, as shown in figure 50, result in filters that can easily be fine-tuned to the correct frequency.

The filter in figure 50 was designed using equations 25 to 28. Applying the course-tuning and fine-tuning and adjusting the coupling loops enabled a filter to be obtained, which satisfied the design specifications exactly.

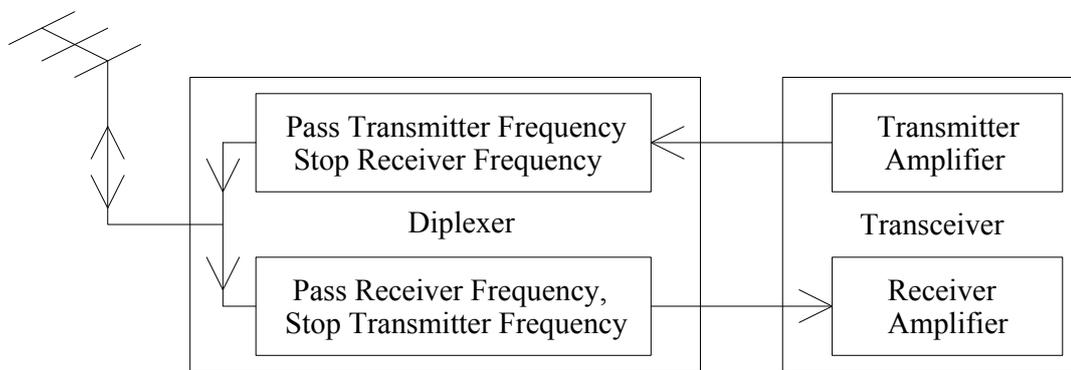


Figure 51. The use of a diplexer to provide full duplex operation.

Round rod interdigital filters are often used as a diplexer in mobile radio transceivers, as shown in figure 51. The filter connected to the receiver has a very high, typically greater than 60 dB, attenuation at the transmitter frequency. This ensures that the transmitter output, which is typically 25 Watt, does not damage the receiver input, which is typically designed to accurately demodulate signals of 1 nW. The use of the diplexer will thus permit full duplex operation, simultaneous transmission and reception, of the transceiver.

Mobile phones have similar diplexers, however to ensure a small size of the diplexer, the diplexer is made using a ceramic material with a typical dielectric constant of 30, allowing a  $\sqrt{30}$  reduction in each dimension of the filter. The resulting filter will thus have 0.6% of the volume of the air-filled filter and also have a fraction of the weight.

### **PCB Interdigital filters**

With circuit miniaturisation, and improved microstrip simulation tools, Interdigital filters are often designed using stripline or microstrip circuits for the resonating elements. The power handling capacity of these filters is small, so that they can only be used as filters in receivers or low power parts of communication equipment. The design procedure is illustrated with the following example.

**Example:**

Design a 5-resonator filter with a 70 MHz bandwidth centred at 1 GHz, using a RO4003 substrate for the PCB. The required  $k$  and  $q$  values can be obtained from table 1, the relevant part which is shown in table 4.

n	q0	I. L.	q 1	q n	k 12	k 23	k 34	k 45
	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

Table 4. Butterworth response, K And Q value filter table, From A. I. Zverev. "Handbook of Filter Synthesis", Wiley, pp 341.

Notice that the  $k$  and  $q$  values vary depending on the losses. One could guess the losses to be about 2 dB and use the corresponding  $k$  and  $q$  values. Note that for each value of insertion loss, there are two sets of  $k$  and  $q$  values. Since the tap coupling is normally used and the tapping points only cover a limited range, it is desirable to select the table values where the difference between the  $q$  values is small.

Alternately, the design is done assuming no losses, so that equations 4 and 5 can be used. After the design is completed, any variations in  $k$  and  $q$  values required to accommodate the losses can be achieved by optimisation. That process is implemented here. The lossless  $q$  values are thus  $q_1 = q_n = 0.6180$ , resulting in  $Q_1 = Q_n = 0.6180 \times 10^9 / 70 \times 10^6 = 8.828$ .

The bandwidth of the first and last resonator, without loading due to adjacent resonators is given by equation 9.4.1 in Zverev as:

$$\text{Resonator 1 } \Delta_{3dB} = \frac{\text{Filter } BW_{3dB}}{q_1} \quad \text{Resonator } n \Delta_{3dB} = \frac{\text{Filter } BW_{3dB}}{q_n} \quad \text{Eqn. 29}$$

In our design the filter  $BW_{3dB}$  is 70 MHz,  $\Delta_{3dB}$  is thus  $70/0.618$  MHz = 113.2 MHz. For this filter, the input and output tapping points are the same. The tapping point in the circuit of figure 52 is thus adjusted to result in a 113.2 MHz bandwidth.

The tapping point can be determined using Microwave Office and the two-resonator circuit of figure 52. To prevent the second resonator from influencing the tuning of the tapping point, a very large coupling gap is used and the resonance of the coupled resonator is placed far away from the centre frequency by disabling all the possible elements as shown in figure 52. To measure the voltage of the input resonator, a high impedance (1 M $\Omega$ ) port (port 2) is connected to the resonator. The signal at port 2 is then the voltage at the top of resonator and has the correct frequency response for determining the loading of the resonator, as shown in the graph of figure 52. After tuning the tapping point for the correct bandwidth, the correct  $q_1$  and  $q_n$  for the filter has been determined.

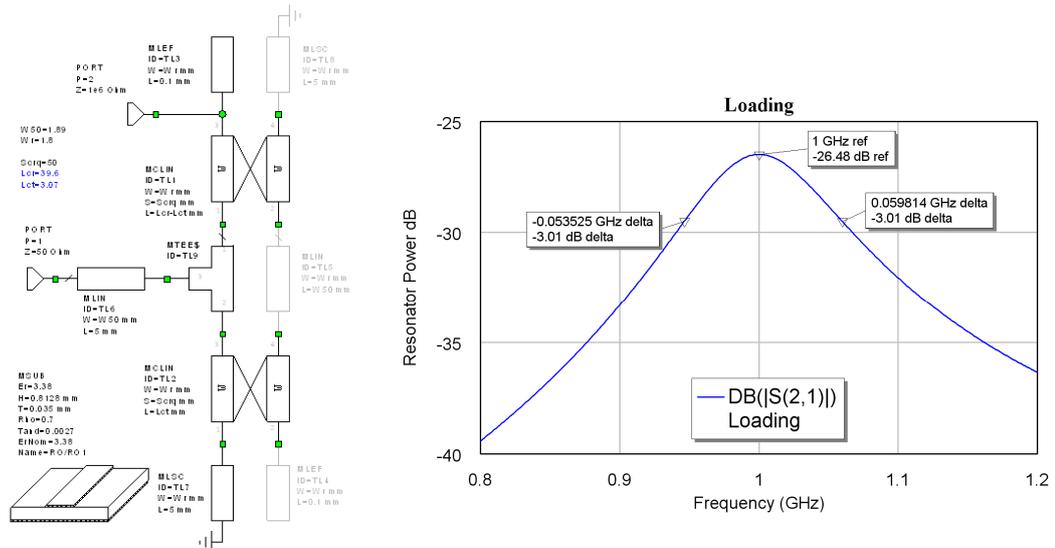


Figure 52. Left: Circuit for input tap and coupling determination, Right: Resonator loading

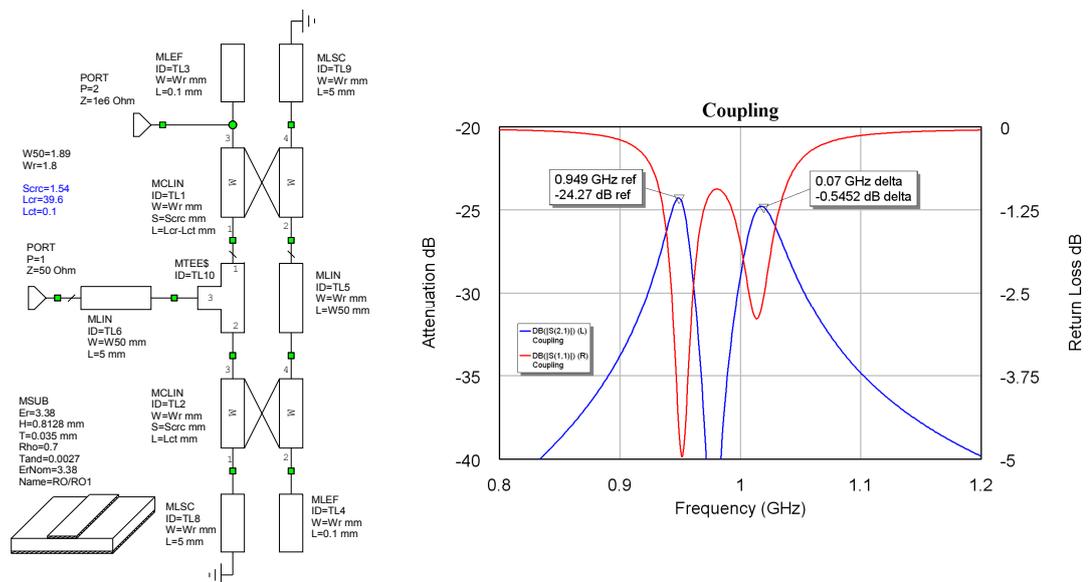


Figure 53. Interdigital filter coupling determination

The coupling gap is tuned to obtain the correct coupling between resonators. To reduce the effect of loading, the tapping length  $L_{ct}$  is made as small as possible, resulting in the test circuit shown in figure 53. When the second resonator is tuned and brought close to the first resonator, a double peak response will result as shown in figure 53. The frequency difference between the peaks is given by equation 9.4.3 in Zverev as:

$$\Delta_{fp} = k_{12} BW_{3dB} \tag{Eqn. 30}$$

For this filter  $k_{12}=1$ , so that  $\Delta_{fp}=70$  MHz for the peaks in frequency response of the test circuit, coupling between the first and second resonator as shown in figure 53. This corresponds to a coupling gap of 1.54 mm. The same test circuit is now used to determine the coupling gaps for the other resonators. For the coupling between the second and the third resonator  $\Delta_{fp} = 0.5559*70$  MHz = 38.9 MHz. The coupling gap is



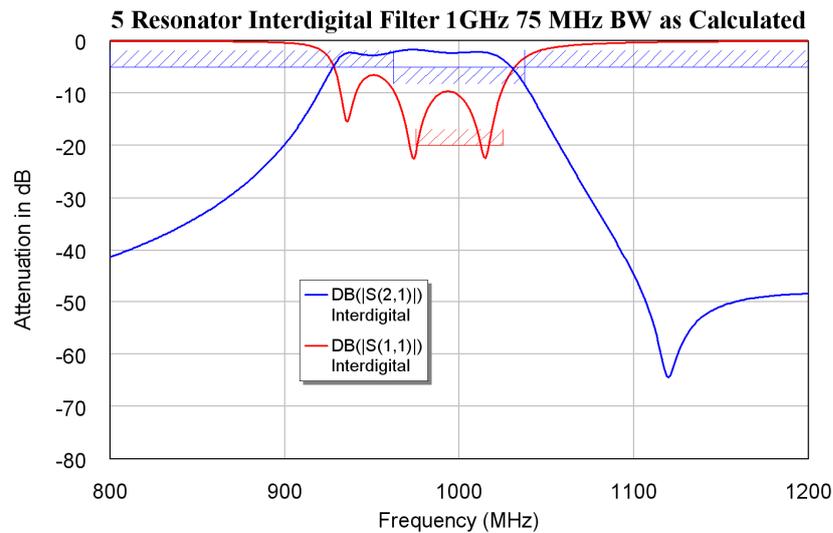


Figure 55. Frequency response of the filter of figure 54, as calculated.

Theoretically, all resonator lengths are the same. In practice it is however, better to allow the resonator lengths to vary slightly under optimisation in order to achieve the desired frequency response. The filter of figure 54 was optimised to give a 70 MHz bandwidth at a 1 GHz centre frequency, as shown in figure 56.

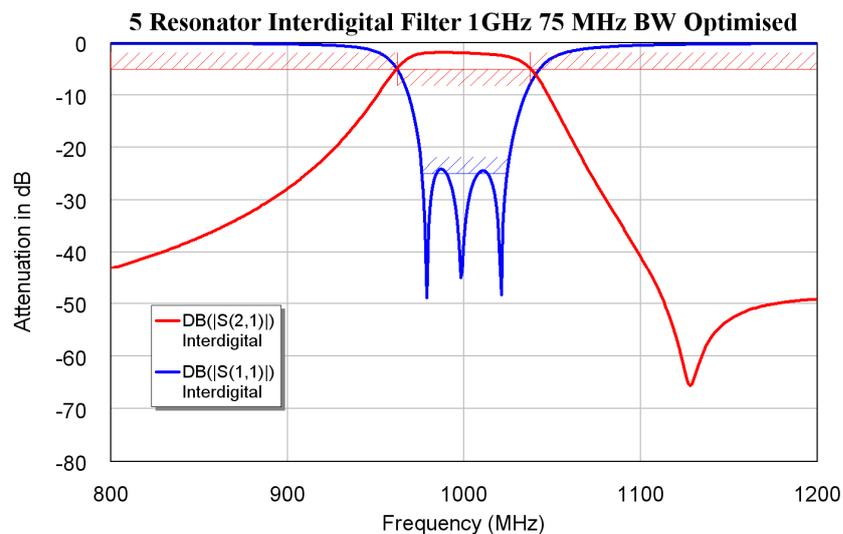


Figure 56. Frequency response of the filter of figure 54, after optimisation.

When the first prototype of the filter was produced, it was found to be 20 MHz (2%) low in centre frequency and have a passband amplitude response that is slightly sloping and a bandwidth that was too small. To compensate the centre frequency of the designed filter, was shifted to 1.02 GHz, the bandwidth was increased, and a slope was specified in the passband as shown in the red curve in figure 54. The measured performance of the second filter is shown in blue in figure 54 and it has a centre frequency of 984 MHz compared to the 1 GHz specification. The measured bandwidth is 92 MHz compared to the 70 MHz specification. This may be sufficiently close to satisfy the required specifications. If this performance is inadequate, then another filter needs to be designed.

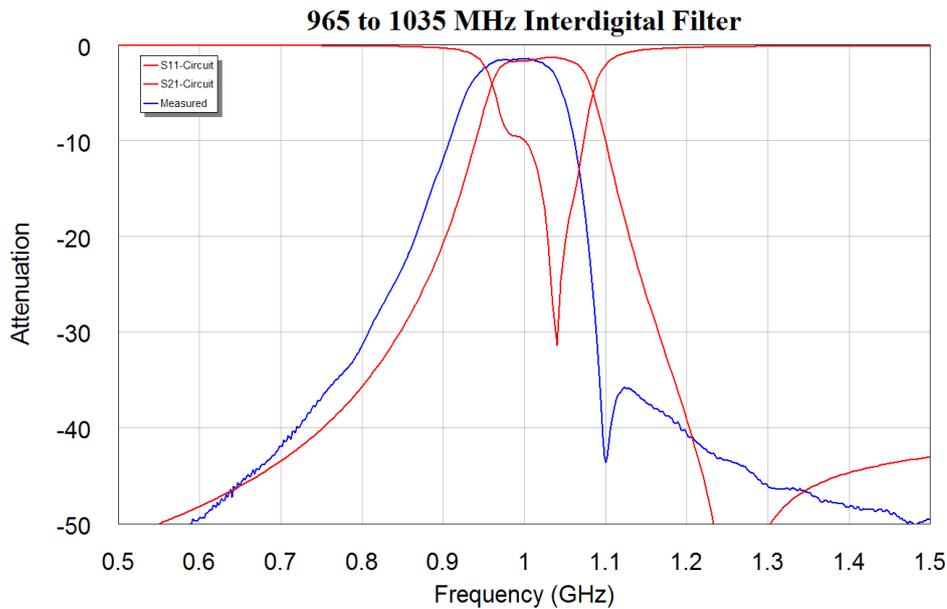


Figure 54. Frequency response of the 70 MHz bandwidth interdigital filter.

These values for tapping point and coupling gaps are then used for the design of the interdigital filter. The first prototype filter has a measured centre frequency, which is 2.2 % low. This difference could be due to a variety of factors, such as differences in PCB substrate, additional capacitance of the resonators to the ground-plane and a bigger end effect.

Figure 55 shows the resulting filter realisation. Interdigital filters are often used in commercial designs as can be seen in figure 56, which is part of a microwave layout produced by Codan in Adelaide, South Australia. Comparing figures 55 and 56, it can be seen that the procedure for earthing the grounded end of the resonator to the bottom layer ground-plane is slightly different. The positioning of the vias in figure 55 is the most likely cause of the centre frequency being 2% low.

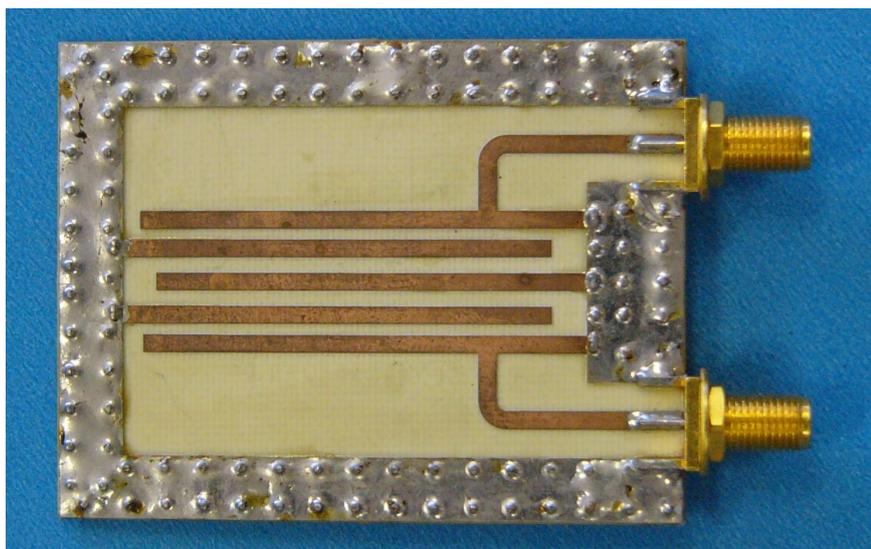


Figure 55. Microstrip interdigital filter 70 MHz bandwidth at 1 GHz

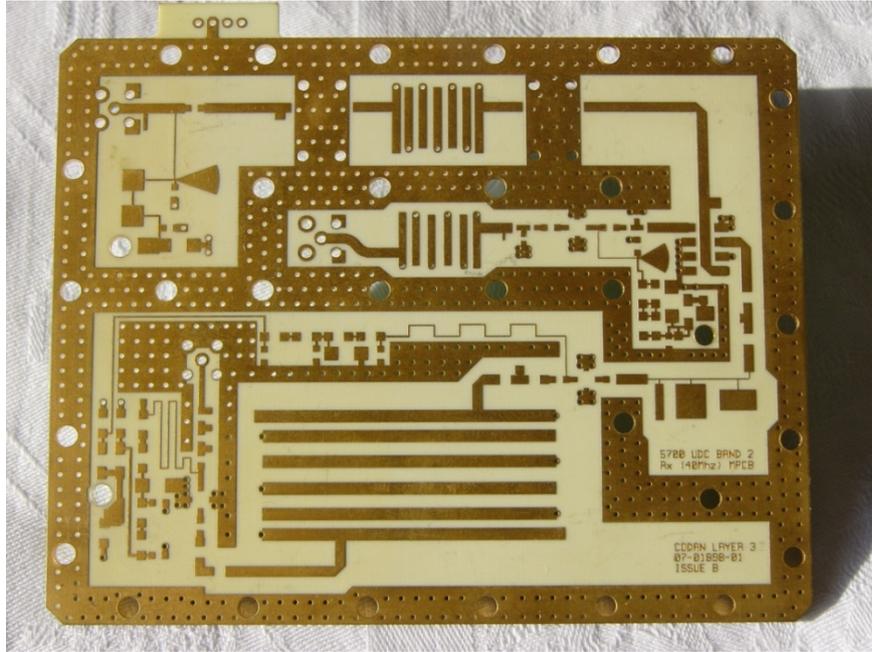


Figure 56. PCB showing three interdigital filters. (Courtesy Codan).

## Direct Coupled Resonator Filters

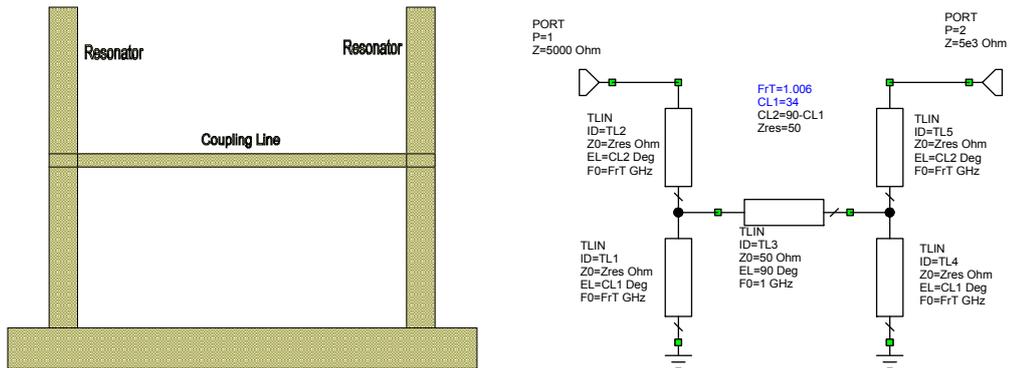


Figure 57. Transmission line coupled resonators.

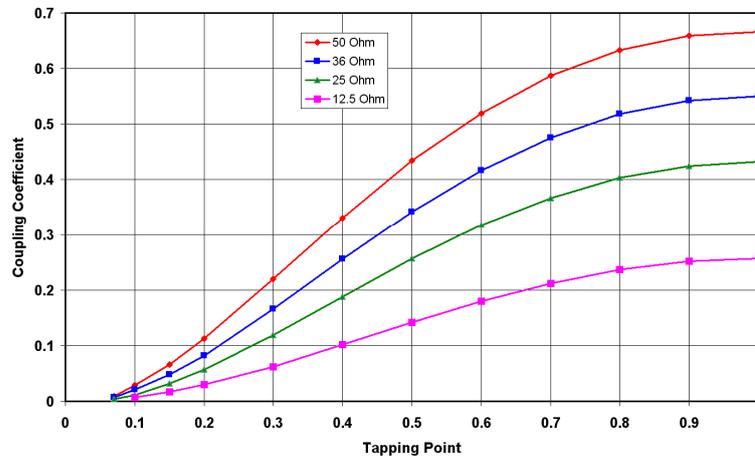


Figure 58. Coupling coefficient versus tapping point for different line impedances.

When the percentage bandwidth of the filter becomes greater than 20%, the coupling gaps for interdigital filters become too small to be produced using readily available technology. As a result, a different coupling technique is required. It is possible to couple two resonators using a quarter-wavelength long transmission line as shown in figure 57. The coupling can be varied by changing the tapping point. When this is done, graphs similar to figure 53 are produced by the circuit of figure 57 and from equation 30 the coupling coefficient can be determined as a function of tapping point as shown in figure 58, since the spacing between the peaks and the bandwidth are known..

## Filter Design Example

The coupling coefficients shown in Figure 58 can now be applied to the design of a wideband filter. As an example, the design of a 5-resonator Butterworth bandpass filter with a lower 3 dB cut-off frequency of 750 MHz and an upper cut off frequency of 1250 MHz is chosen the filter will thus have a 50% bandwidth.

Equations 4 and 5 give the normalized Q values and coupling coefficients as:

Normalised	Denormalised for 50% Bandwidth
$q_1 = q_n = 0.618$	$Q_1 = Q_n = 1.236$
$k_{12} = k_{45} = 1$	$K_{12} = K_{45} = 0.5$
$k_{23} = k_{34} = 0.5559$	$K_{23} = K_{34} = 0.278$

Figure 58 shows that different transmission line impedances should be used for different coupling coefficients in order to obtain realistic tapping points. For the filter, 50  $\Omega$  transmission lines are used for the input and output resonators, which require the highest coupling coefficients of 0.5. For the other resonators, 36  $\Omega$  transmission lines, which have lower losses, are used. Interpolating the tapping points from figure. 58, results in the following normalised tapping points, Txline results in the following resonator line lengths:

Resonator 1	50 $\Omega$	Tap = 0.487	Lr = 46.96 mm
Resonator 2	36 $\Omega$	Tap = 0.614	Lr = 44.99 mm
Resonator 3	36 $\Omega$	Tap = 0.426	Lr = 44.99 mm
Resonator 4	36 $\Omega$	Tap = 0.614	Lr = 44.99 mm
Resonator 5	50 $\Omega$	Tap = 0.487	Lr = 46.96 mm

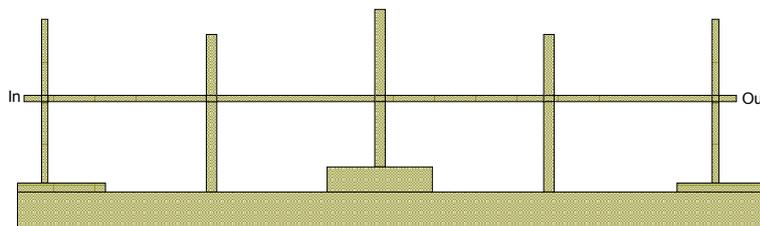


Figure 59. Filter layout for initial tap values.

Entering those values in the Microwave Office circuit schematic of the filter, results in the initial filter layout shown in Figure 59, with the corresponding frequency response shown in Figure 60. To highlight differences between the 750 MHz and 1250 MHz, corner frequency specifications and frequency-response obtained from the computer simulation, an optimisation mask is superimposed on Fig. 60. In addition, an optimisation mask for a -1 dB attenuation from 780 MHz to 1220 MHz is shown. It can

be seen that the design procedure results in a filter that has a bandwidth that closely matches the design specifications, but whose centre frequency is 3.5% low. This difference is due to the end effect of the open circuit resonators not being taken into account in the above calculations.

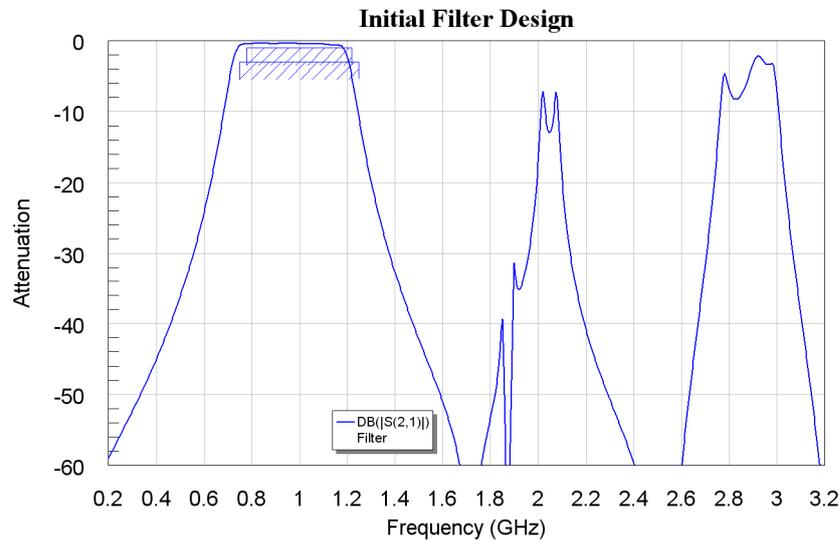


Figure 60. Frequency response for initial wideband filter.

### Fine Tuning the Filter

To shift the centre frequency, the length of all the resonators is changed slightly using the 'tuning simulation' capability of Microwave Office. In addition, it is desirable to fine-tune the filter to obtain the lowest insertion loss. This is best achieved by ensuring that the filter has a low return loss. Setting the optimiser constraints on  $S_{11}$  to be less than -20 dB and carrying out the first stage of the optimisation process of the filter to meet this return loss as well as the passband specifications results in the following tapping points:

Resonator 1	50 $\Omega$	Tap = 0.5147
Resonator 2	36 $\Omega$	Tap = 0.5969
Resonator 3	36 $\Omega$	Tap = 0.4304
Resonator 4	36 $\Omega$	Tap = 0.5969
Resonator 5	50 $\Omega$	Tap = 0.5147

Figure 61 shows the resulting frequency response of the filter after this first stage of optimisation. For a wideband filter implementation like this, the spurious response at the harmonics, particularly that at 2 GHz is unacceptable. To provide better attenuation at those frequencies, stubs need to be added to the filter. Those stubs will distort the frequency response of the filter. In order to minimise the effect of the stubs on the centre frequency of the filter, it is desirable to use two sets of stubs and vary the spacing between the two sets of stubs to provide a low return loss at the centre frequency of the filter. In addition, it is desirable to make the harmonic stubs slightly different lengths, to widen the bandwidth over which good harmonic attenuation is obtained.

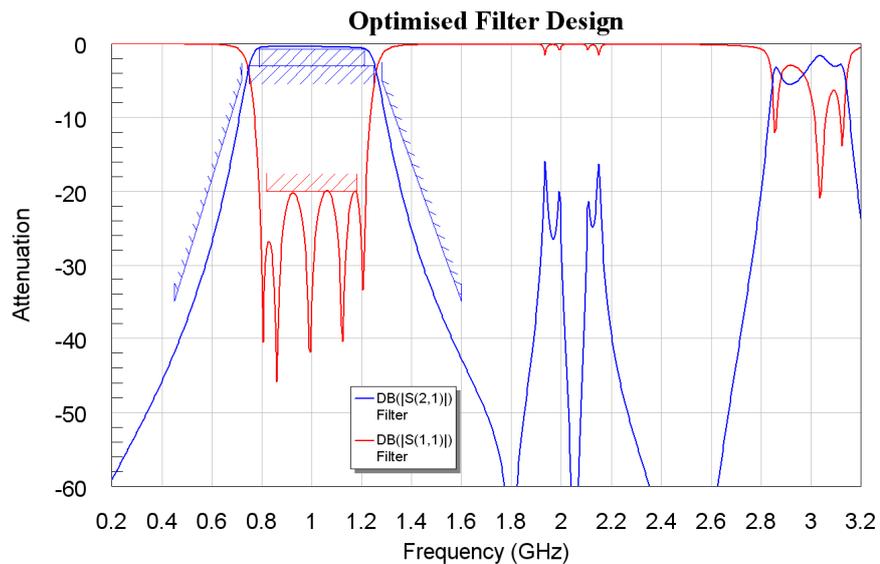


Figure 61. Frequency response of the filter after stage 1 optimisation.

The final step in the filter design is to change the layout from a simple design as shown in Figure 62, to a more compact and easier to manufacture design. To achieve this, the three tiered ground level shown in Figure 59 is made one level and the coupling transmission lines are folded and bent in order to take up as little space as possible, while connecting to the resonators at the correct tapping points. The harmonic stubs may be folded as shown in Figure 40 and 41. The final layout should also ensure that the input and output connectors are in the correct location for mounting the filter. Figure 62 is a photograph of the final filter.

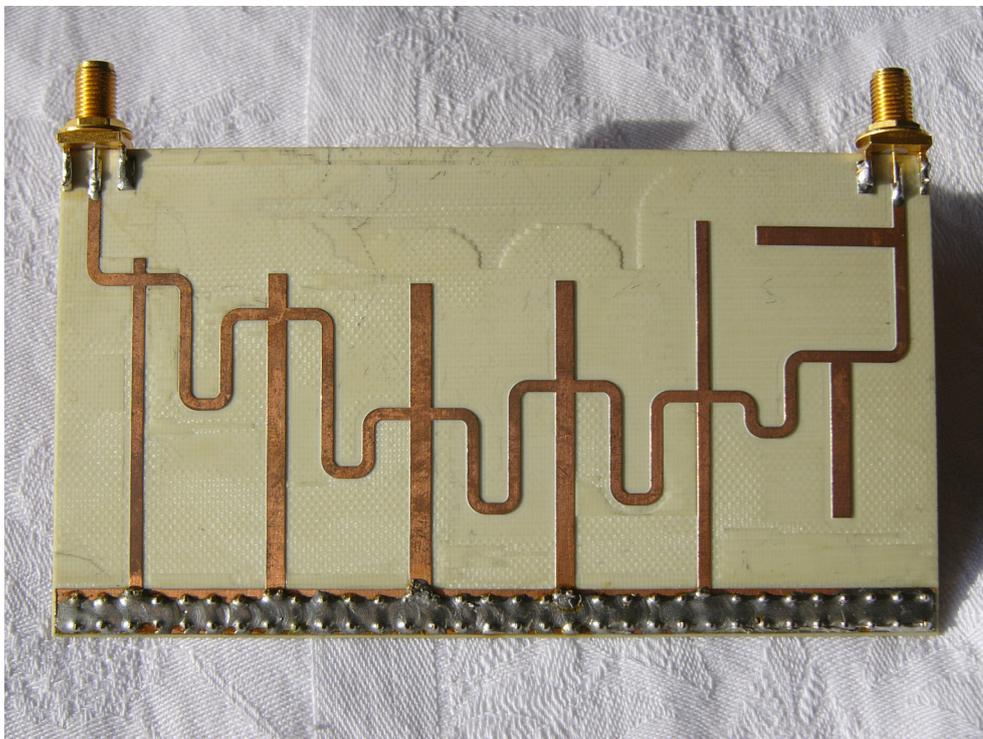


Figure 62. 750 MHz to 1250 MHz filter hardware.

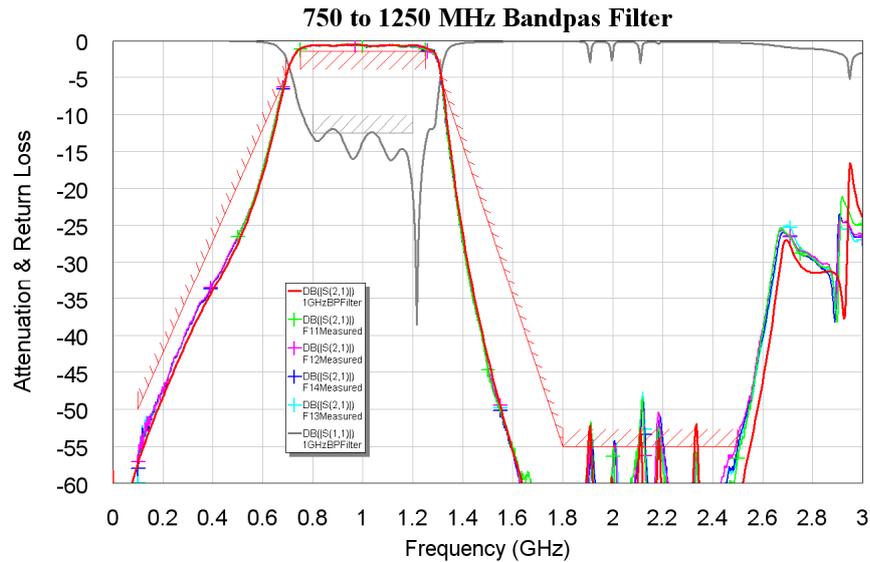


Figure 63. Frequency response of the filter of figure 61.

Fig. 63 shows the frequency response of the final filter after optimisation, it also shows the final optimisation limits used. To determine the reliability of the design technique, four filters were constructed. The measured frequency response of those filters is also shown in Figure 63. It can be seen that the design technique results in highly repeatable filters, whose measured performance agrees remarkably with the results from computer simulation. The agreement between the measured and simulated results for these filters is much better than those for the hairpin filter of Figure 40 or the interdigital filter of Figure 55. It should be noted that the right most resonator is significantly larger than the other resonators. This is required to compensate for the effects of the harmonic stubs on the passband response.

## Microstrip Filter Comparison

When designing a Microstrip filter the four common filter types described in these notes can be used. For a fair comparison, a 5-resonator filter with a centre frequency of 1 GHz and a 70 MHz bandwidth was designed as an Interdigital filter, a Compline Filter, a Hairpin filter and a Direct-coupled resonator filter. The filters were optimised using the same optimisation limits. Each filter has 10 mm tracks connecting to the tapping point to allow a connector to be soldered to that track. The same RO4003 substrate was used and all resonators are 3 mm wide. For each of the filter types, a simple test circuit like those shown in figures 52, 53 or 57 is used. The tapping and coupling test circuit for the compline filter is shown in figure 64 and that for the hairpin filter is shown in figure 65.

The input tapping for all these four circuits is adjusted to obtain frequency responses like figures 52 and 53, but with the bandwidths as per equations 29 and 30 for the relevant filter. Since the centre frequency, bandwidth and the number of resonators for these filters is the same as those of figures 52 and 53, exactly the same responses should be obtained. The coupling tuning, can be done most accurately when the end resonator loading is reduced as done in figures 64 and 65.

Figure 66 shows the initial frequency response of the four, 5-resonator filters, when the tapping and coupling gaps are set to those calculated using the test circuits.

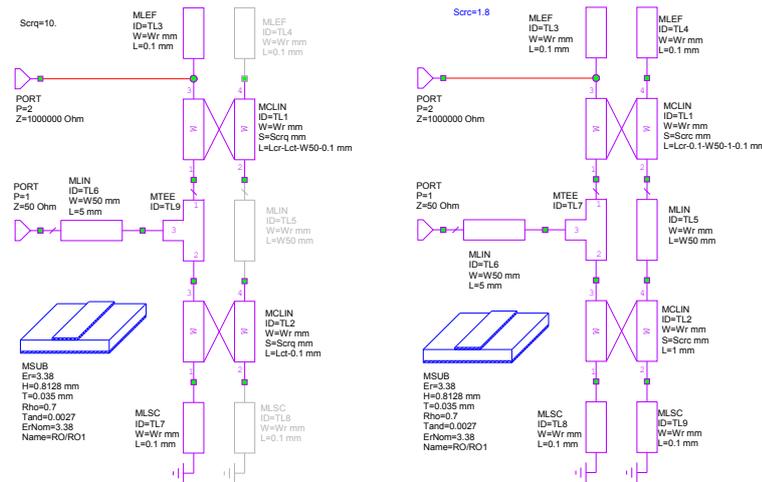


Figure 64. Resonator Loading and Coupling test circuit for Combine filters.

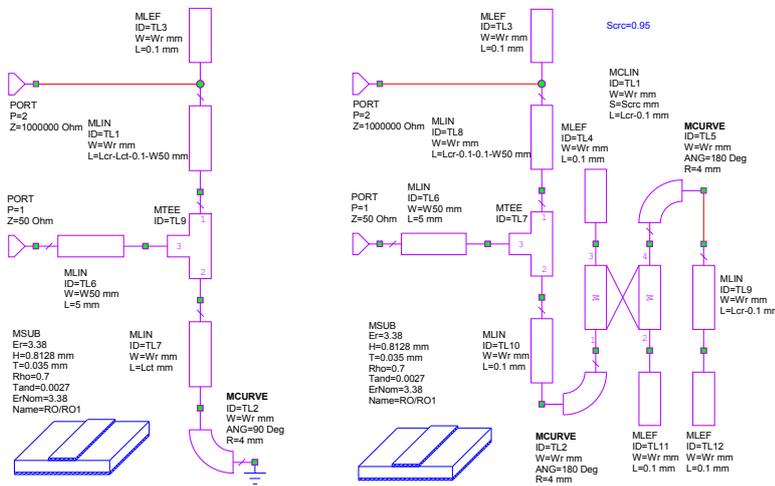


Figure 65. Resonator Loading and Coupling test circuit for Hairpin filters.

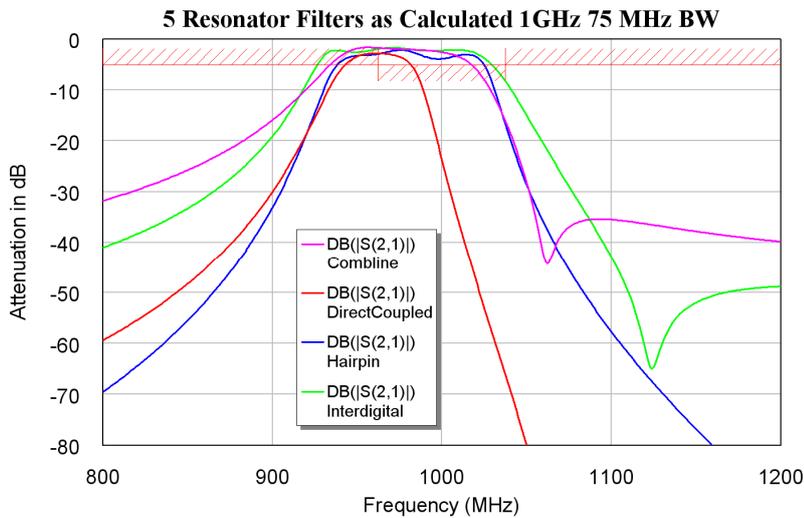


Figure 66. Filter performance with parameters as calculated.

The performance of all these filters is close enough that the specifications can be met using some simple optimisation. The tuning and optimisation is best done in three stages: Firstly adjust the resonator length using manual tuning, to give the correct centre frequency. The second step is to fine tune the coupling gaps and tapping points using optimisation, to get as close to the requirements as possible. The final step is to allow the individual resonator lengths to be optimised as well as further minor optimisation of the coupling gaps and resonators. Since there is a significant interaction between the parameters to be optimised, the local random optimisation is most likely to result in the best optimisation. If the optimisation gets stuck, changing the weight of the worst error may get the optimisation out of the local minima and thus result in lower error costs.

Figure 67 shows the final performance of the filters after optimisation. The Hairpin filter has a good symmetrical filter response and has good stopband attenuation. The Direct Coupled Resonator filter has a better higher frequency attenuation, but a poorer lower frequency attenuation. The interdigital filter has a poorer attenuation characteristic, compared with the other filters. In particular, the attenuation near 1 GHz is only 50 dB. The combine filter has the worst close-in stopband attenuation of these filters. Combine filters are used less often than interdigital filters.

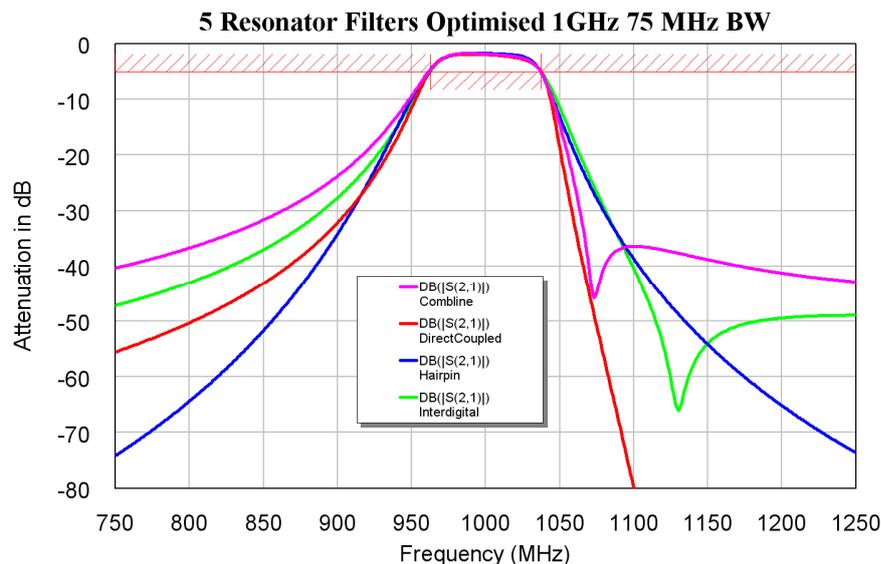


Figure 67. Frequency response of three 5-resonator filters.

Figure 68 shows the frequency response of the four filters at the harmonics of the 1 GHz passband. For the direct-coupled resonator filter the lengths of the coupling lines was 12.5% of a wavelength, which is half of the length of the coupling lines used for the filter in figure 62. This resulted in a lower harmonic attenuation. The direct coupled filter has the best close-in stopband attenuation, but the worst stopband attenuation above 2 times the centre frequency. The second harmonic response from the hairpin filter is significant, but can be removed using harmonic stubs. The interdigital filter does not have a significant harmonic response, but it does not have a very good stopband attenuation either. The stopband attenuation for interdigital filters does not significantly improve with filter order. The combine filter is the smallest, occupying 41x60 mm. The interdigital filter is slightly larger occupying 42x60 mm. The Direct Coupled Resonator filter, occupies 75 x 60 mm and the Hairpin filter is the largest, occupying 78 x 68 mm.

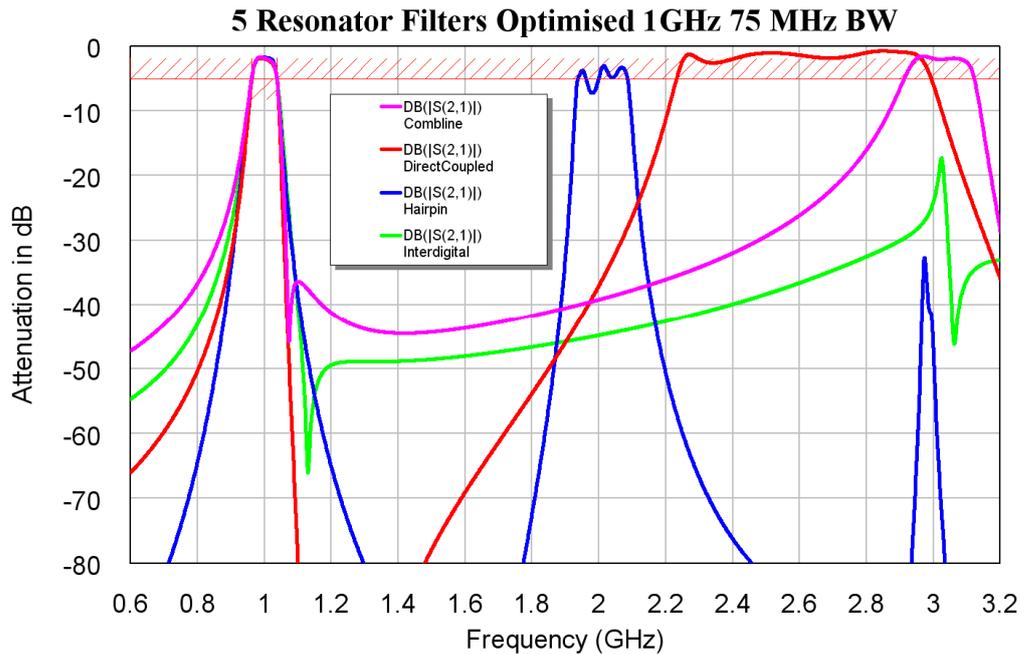
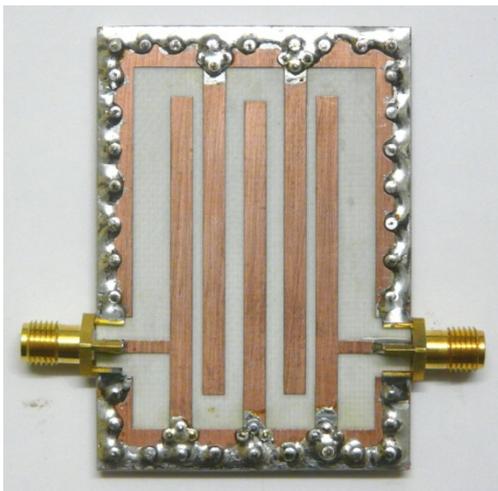
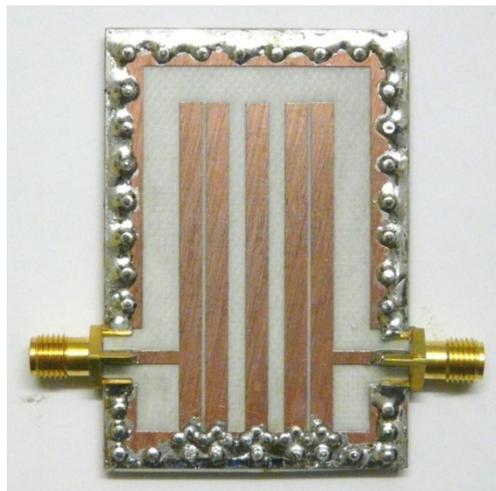


Figure 68. Harmonic response of the filters of figure 67.

Figure 69. Interdigital filter hardware.  
(42x60mm)Figure 70. Comblines filter hardware.  
(41x60mm)

Figures 69 to 72 show the hardware of the four filters, approximately to the same scale. The Comblines filter is slightly smaller than the interdigital filter, but in most instances, there is not enough space saving to warrant the worse stopband attenuation.

Figures 73 to 76 show the measured frequency responses of the filters. Comparing these with figures 67 and 68, shows the remarkable agreement between the calculated and measured results. The passband of the Interdigital, Comblines and Direct Coupled filters are a few percent lower than calculated. The difference matches the length of the pins used in connecting the grounded end of the resonators to the groundplane on the bottom of the PCB. This can be taken into consideration in the second iteration of the filter designs if needed.

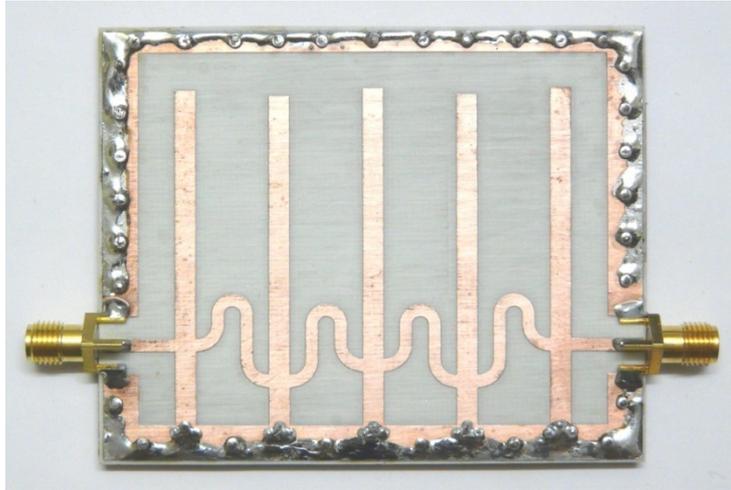


Figure 71. Direct Coupled filter hardware. (75x60mm)

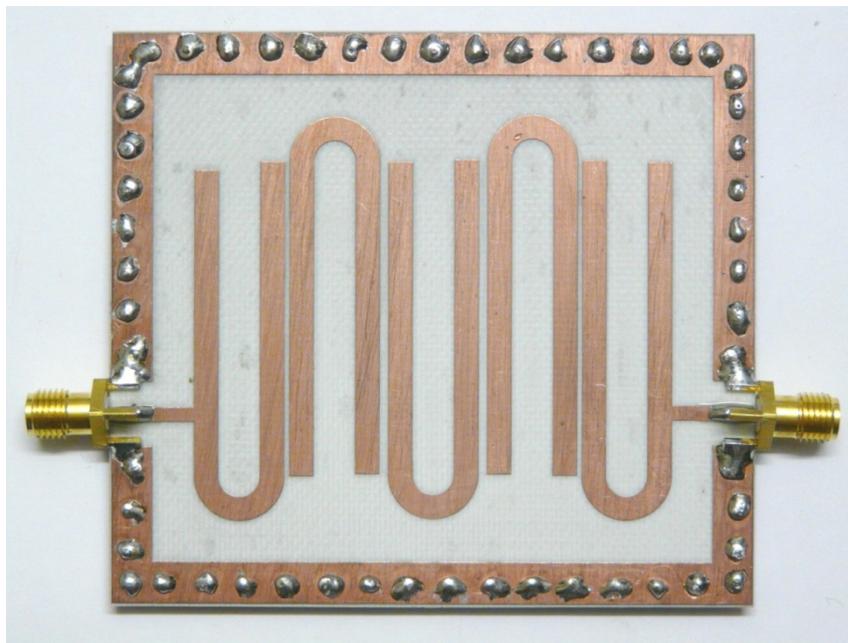


Figure 72. Hairpin filter hardware. (78x68mm)

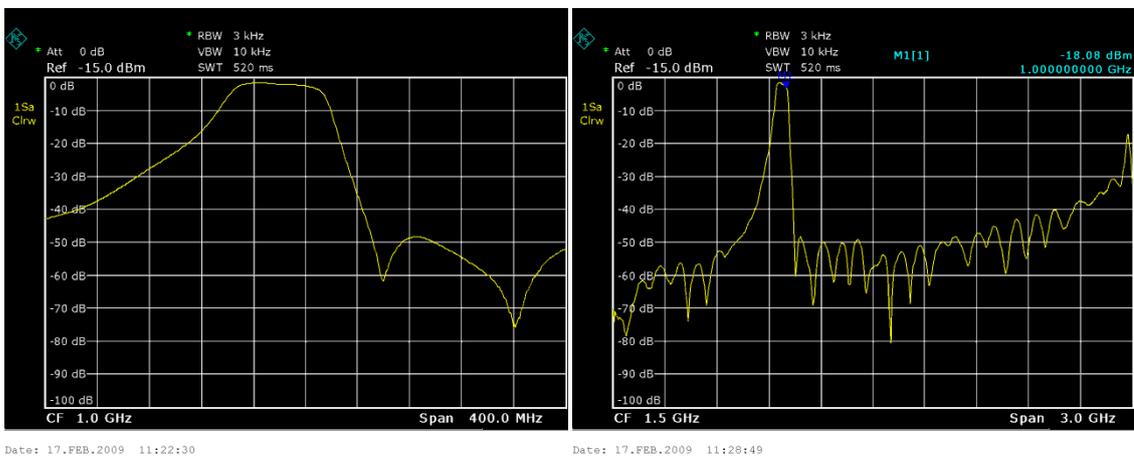


Figure 73. Measured frequency response of Interdigital filter.

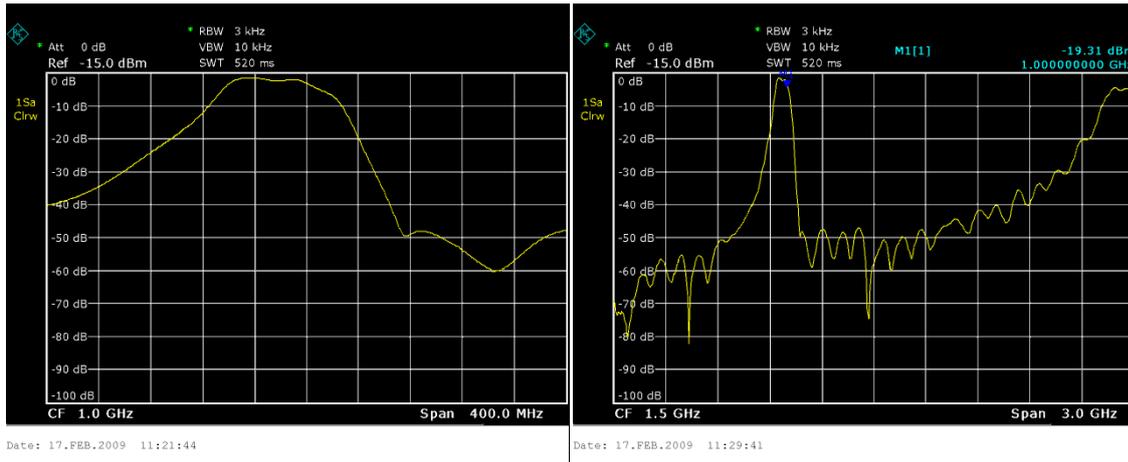


Figure 74. Measured frequency response of Combline filter.

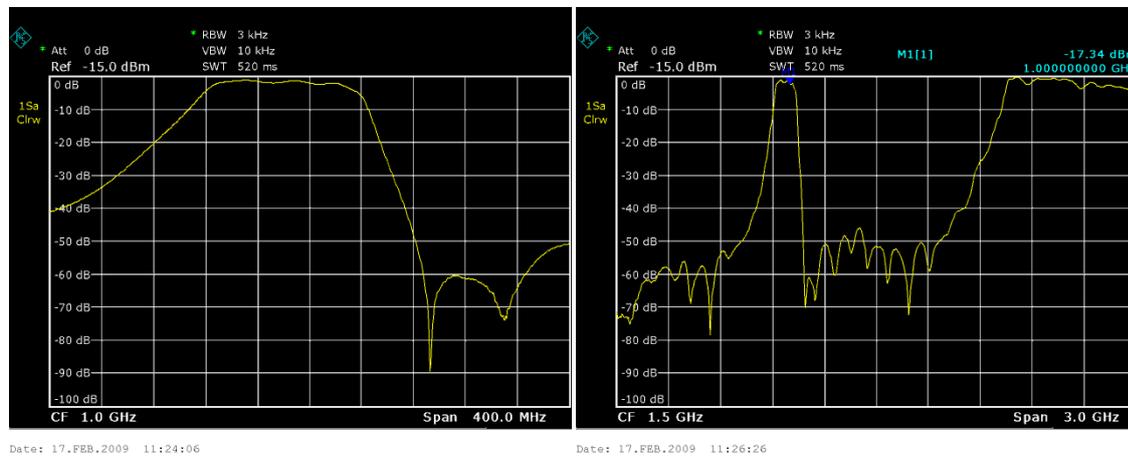


Figure 75. Measured frequency response of Direct Coupled filter.

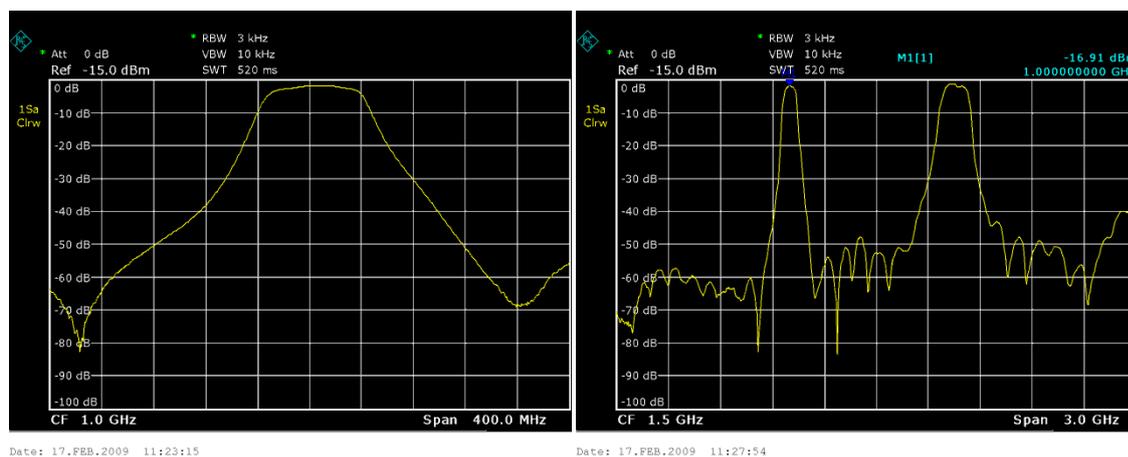


Figure 76. Measured frequency response of Hairpin filter.

The Hairpin filter has a measured centre frequency that is very close to the desired 1 GHz centre frequency. This is due to no ground connections being required, thus removing an uncertainty from the design. From the measured passband responses, Hairpin filters have a better close in stopband attenuation and a more accurate centre frequency are thus a good choice for one off designs, where the performance has to be correct first time.

## Coaxial Filters

Coaxial filters are similar in principle to helical filters, except that the resonator is a straight rod, rather than a wound helix as shown in figure 77. The length of the centre rod is a quarter wavelength. For the highest Q values, the ratio of the centre rod diameter to the inner diameter of the cavity is such that the characteristic impedance of the coaxial line forming the resonator is  $77 \Omega$ . The equation for the characteristic impedance of a coaxial transmission line is:

$$Z_c = \frac{1}{2\pi} \sqrt{\frac{\mu_0}{\epsilon_0}} \ln\left(\frac{B}{A}\right) = \frac{60}{\sqrt{\epsilon_r}} \ln\left(\frac{B}{A}\right) \quad \text{Eqn. 30}$$

So that the maximum Q is obtained when  $B/A = 3.6$ .

Typical Q values are in the range of 2000 to 10000. Because cavity filters have such a high Q, they are used in many applications requiring a low insertion loss. Air filled coaxial filters have a high power handling capability and are typically used for filters and combiners for radio and TV transmitters and in mobile radio and mobile phone base stations.

Figure 78 shows a typical coaxial filter used in a mobile radio talk-through repeater operating at 80 MHz. The total height of the filter is 1.2 m and the cavities are 200 mm in diameter. The cavities used in this filter have a Q of 5000. The round cylinders at the top of the cavities are isolators, which ensure that the input impedance to the filter is perfectly matched over the entire bandwidth of the transmitter.

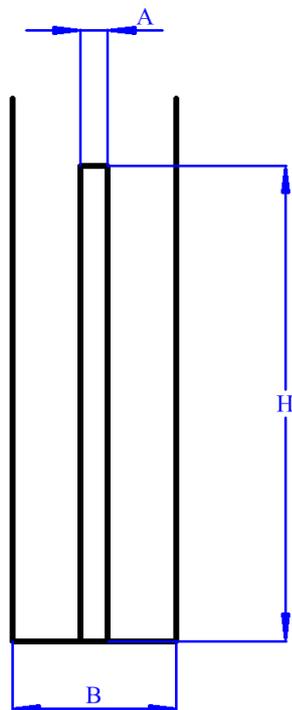


Figure 77. Coaxial cavity dimensions.



Figure 78. Coaxial cavity diplexer for 80 MHz.

Some coaxial cavities are much smaller. Most Mobile Phones have diplexers with ceramic filled coaxial resonators. Figure 79 shows a surface mount, ceramic 3-resonator coaxial filter for mobile phone applications. Using a high dielectric constant material for

the cavity gives a significant reduction in filter size. The centre frequency of this filter is 881 MHz. The filter is 14 mm long.

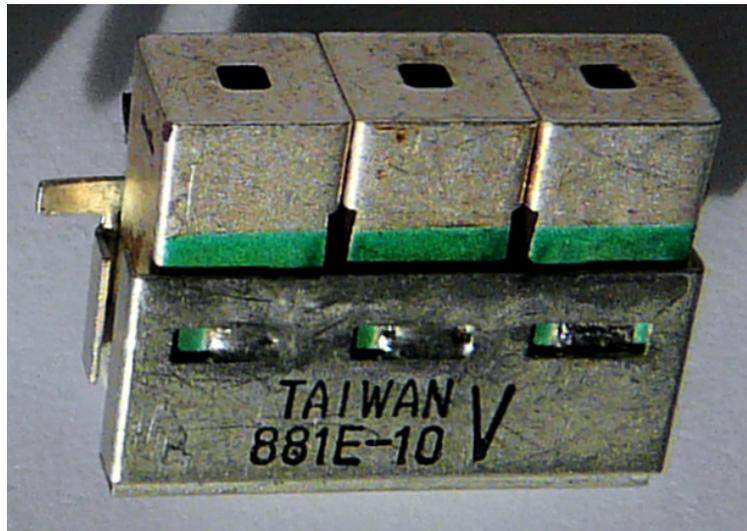


Figure 79. Coaxial cavity filter

For Mobile Radios, used by trucks or taxis to talk to their base, a coaxial diplexer is often used to provide isolation between the transmitter and the receiver in their Mobile Radio. Figure 80 shows a typical coaxial diplexer. The receive frequency is 450.625 MHz and the transmit frequency is 460.125 MHz. These radios typically have a transmission power level of 25 Watt (43.98 dBm). The receiver typically has signals of -60 dBm applied to it, so that the difference between the transmitted and received signal is more than 100 dB. Connecting the transmitter directly to the receiver, will result in severe damage to the input circuitry.



Figure 80. Cavity filter diplexer for mobile radio applications.

As shown in Figure 81, at the receiver frequency of 450.625 MHz, the diplexer of figure 80 causes a less than 1 dB loss between the antenna and the receiver ports and more than 60 dB isolation between the antenna and the transmitter ports, thus protecting the input from any spurious signals generated by the transmitter at the receiver frequency.

At the transmitter frequency of 460.125 MHz, the diplexer provides less than 1 dB loss between the transmitter port and the antenna port, so that virtually all the 25W output signal is transmitted. At this frequency, there is more than 60 dB isolation between the antenna and receiver ports, thus protecting the input from the high power levels generated by the transmitter.

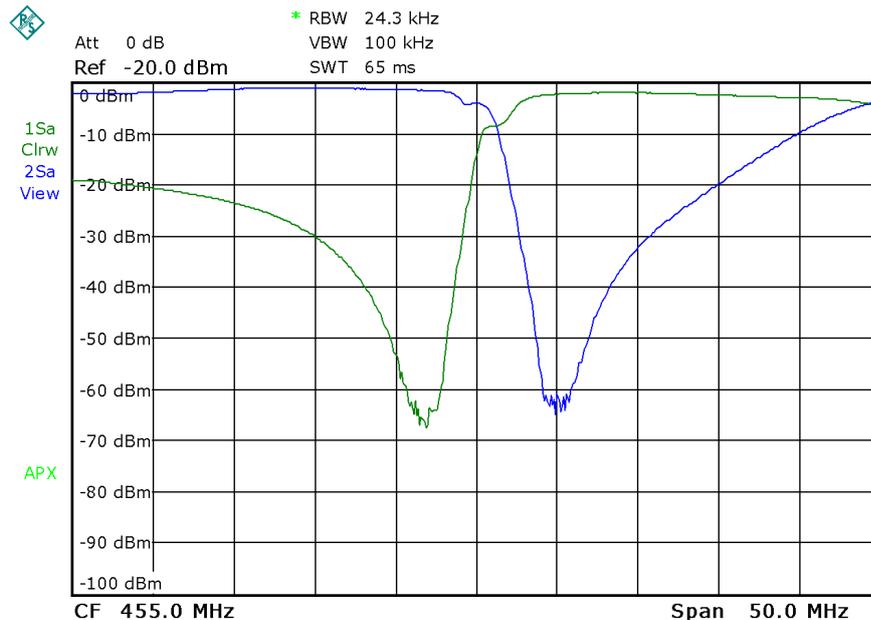


Figure 71. Frequency response of the diplexer of figure 70.

For low loss, high power microwave filters; it is often convenient to mill the whole filter out of a solid block of aluminium, as shown in figure 72. By including coupling between non-adjacent cavities, zeros of transmission can be included. In the filter of figure 72, non-adjacent cavity coupling has been applied to the first and last resonator.

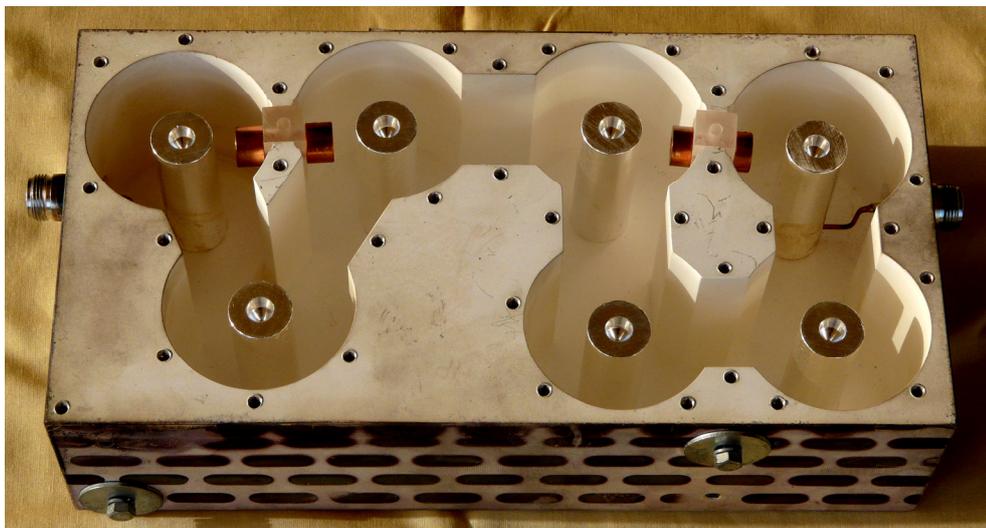


Figure 72. 1.8 GHz Coaxial Cavity Filter.

These filters are tuned by capacitive top loading on each resonator. Each of the resonators is thus a fraction shorter than a quarter wavelength, with the required top-loading capacitance making up this shortfall.

## Ceramic Filters

For many consumer applications IF filters are required at only a few specific frequencies. For example, all FM radios have an IF of 10.7 MHz. All AM radios use a 455 kHz IF. As a result, commercial filter manufacturers like Murata have a wide variety of low cost filters available for those frequencies. To achieve small size, low cost and freedom from microphony, ceramic materials are used for the resonators. Typical 10.7 MHz ceramic IF filters are shown in figure 73. Many of the latest versions are surface mount instead of the through-hole filters shown in figure 73. The right hand filter contains two filters like the left hand one, inside the one casing and is used to provide more filtering of unwanted signals. Most of the FM IF filters are designed to provide a flat group delay over the passband. This minimised the distortion caused by the IF filter to the demodulated audio signals.



Figure 73. 10.7MHz ceramic IF filters.

In addition there are many filters available for specific IF applications, such as TV sound IF, 38.9 MHz for Digital Audio Broadcasting (DAB), 44 MHz for digital TV receivers etc.

Ceramic resonator filters are also used for RF filtering. The Murata Gigafil range for instance contains many different RF filters covering the 800 to 950 MHz mobile phone band and the 1.4 to 2.4 GHz mobile phone and WLAN bands.

Since these ceramic filters are mass-produced, their cost is very low and it may be worthwhile to modify one's design to be able to incorporate these filters.

## SAW Filters

The resonators considered so far have electromagnetic or acoustic resonances with the resonator being a quarter wavelength. Voltages across an input transducer on the surface of a piezo-electric substrate causes an acoustic wave to travel along the surface of the substrate. These travelling waves then induce voltages on a second transducer, which is used as an output transducer. By correctly designing the shape of these transducers, filtering can be achieved, with independent control over both the amplitude and group delay. SAW filters are thus normally used for TV IF filters, IF filters for Radars and other applications where sharp filtering is required and where the group delay must be absolutely flat.

Since the input transducer generates a travelling wave in the direction of the output transducer as well as one in the opposite direction, which must be absorbed, SAW filters have a high insertion loss, typically 16 to 20 dB.

The design of SAW filters is outside the scope of these lectures.

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