# A Design Technique for Microstrip Filters.


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***Abstract*— Most communication systems contain an RF front end, which performs analogue signal processing with RF filters. Microstrip filters are a low cost means of doing this. This paper describes a general design technique for microstrip or stripline filters. Four different microstrip filters are designed. The simulated and measured responses agree closely over a wide range of frequencies. This technique allows new filter topologies to be investigated.**


## I INTRODUCTION

Most communication systems require an RF front end, where RF filters and low noise amplifiers perform analogue signal processing. Microstrip RF filters are commonly used in receivers and transmitters operating in the 800 MHz to 30 GHz frequency range. The two most common types used are the parallel coupled line filter and the interdigital filter. A hairpin filter is a variation of the parallel coupled line filter, where the resonators are bent into a hairpin shape in order to achieve a more convenient aspect ratio. The design of these filters is well known [1-4] and generally involves the use of empirical relations. Microwave RF filters are designed using either low pass filter equations [1] with suitable transformations or using coupled resonators design procedures [2].

In this paper a novel technique is presented for determining the PCB layout required for the end resonator loading and coupling factors for any stripline or microstrip realisation. The design technique presented here is based on the adjustment procedure for helical filters, described in Zverev [2]. This book contains tables for normalised coupling (k) and normalised loaded Q (q) values and gives the following equations for these for Butterworth filters:

$$q_0 = q_n = 2Sin\frac{(2-1)\pi}{2n} = 2Sin\frac{\pi}{2n} \qquad (1)$$

$$k_{ij} = \frac{1}{2\sqrt{Sin\frac{(2i-1)\pi}{2n}\,Sin\frac{(2i+1)\pi}{2n}}} \qquad i=1,3,5\ldots. \qquad (2)$$

## II DESIGN PROCEDURE

Four filters with a centre frequency of 1 GHz and a 75 MHz bandwidth are designed using this technique. Figure 1 shows the resulting interdigital filter.

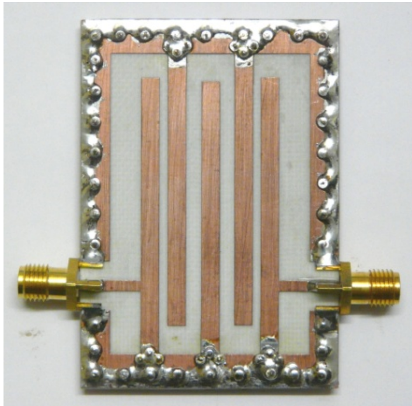

Figure 1. 1GHz interdigital filter.

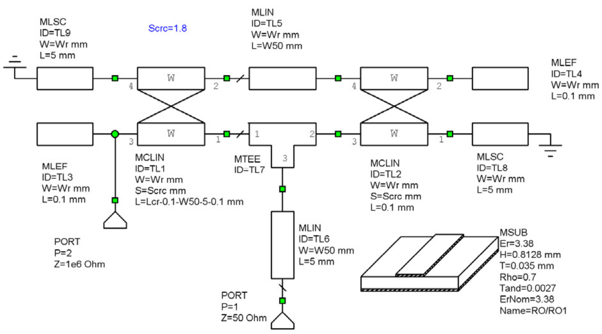


Figure 2. Test circuit for determining the required coupling gaps and resonator loading.

Figure 2 shows the Microwave Office [5] realisation of a coupled line structure, using two resonators. The structure corresponds to the first two resonators of the interdigital filter if figure 1.

The author has simulated many microstrip-line filters using Microwave Office [5] and ADS [6] with both circuit simulation and electromagnetic simulation. The author has found that all give accurate results, however the circuit simulation from Microwave Office gave the best agreement with the measurements on the actual filters produced and as a result that circuit simulation is used throughout this paper. Circuit simulation also has the advantage of being much faster.

The resonator of figure 2 is made up of different coupled line sections, the length Lct of one of these is made variable to enable the input tapping point, and thus the loaded Q of the first resonator, to be varied. An equation is used for the other coupled line length, to permit independent control of the input tapping point as well as the centre frequency by varying the total length of the resonator. To determine the resonator loaded Q and set the correct tapping point, the coupling distance between the coupled resonator sections (Scrc) is made large and the coupled resonator is split into two unconnected parts by disabling TL5 and TL9, to ensure that this coupled resonator does not effect the end resonator, as shown in figure 3. An end-effect, ground connection and T section is used to allow the model to be realised accurately. The resonator line width is a compromise between filter size, radiation losses and resistive losses. A resonator line width of 3 mm is chosen for these designs.

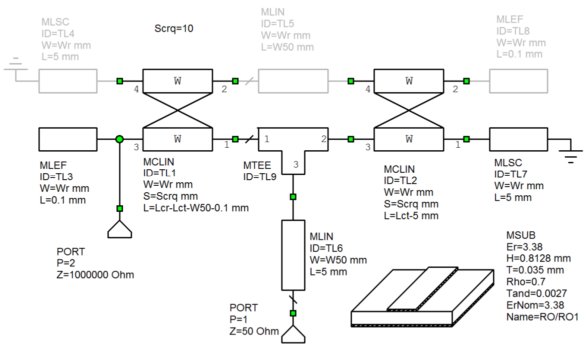


Figure 3. Test circuit for determining the resonator loading.

For helical filters, the adjustment of the loaded Q values for the end resonators involves the measurement of the 3 dB bandwidth of the field in the end resonator. During simulation of a microstrip filter, this loaded resonator bandwidth can be obtained by measured the voltage at the top of the resonator, (Port 2 of figure 1). Equation (3) shows the relationship between the 3dB bandwidth of this voltage and the loaded normalised q of the end resonators of the filter as:

$$\text{Resonator1 } \Delta_{3dB} = \frac{Filter\ BW_{3dB}}{q_1} \qquad (3)$$

For a 5 resonator filter, equations (1) and (2) or filter tables give $q_0 = q_n = 0.6180$, $k_{12} = k_{45} = 1.0$ and $k_{23} = k_{34} = 0.5559$. From equation (3), the 3 dB resonator voltage bandwidth should thus be 121 MHz. The input tapping point and the line length are then tuned to achieve the correct bandwidth and centre frequency. When Lct = 9.7 mm and Lcr = 45.3 mm, the frequency response of figure 3 is obtained.

From Zverev [2], when observing the fields inside the end resonator of a helical filter, a double humped response as shown in the blue curve in figure 4 results, with the distance between the peaks being related to the coupling coefficients as follows:

$$\Delta_{fp} = k_{12} BW_{3dB} \qquad (4)$$

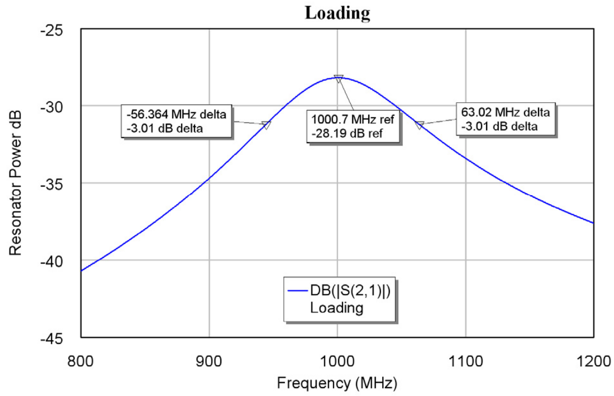


Figure 4. Frequency sweep of loaded end resonator of figure 2.

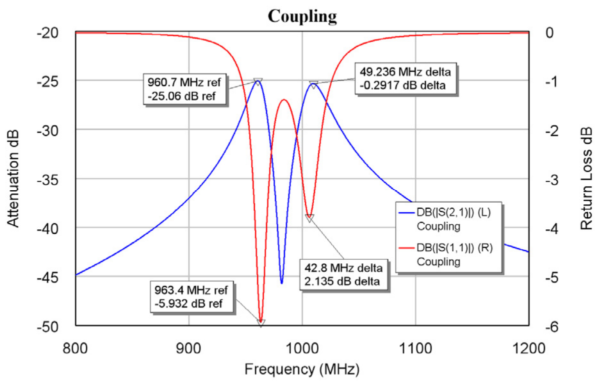


Figure 5. Frequency sweep of coupling test circuit.

This equation can also be used to determine the coupling gaps required. For the required filter $k_{12} = k_{45} = 1.0$, so that a distance between the peaks of 75 MHz is required. For the coupling between the inner resonators, $k_{23} = k_{34} = 0.5559$, corresponding to a 41.7 MHz distance between the peaks as shown in figure 5 is required.

In figure 2, elements TL5 and TL9 are enabled to determine the required coupling gaps (Scsc) by observing the voltage at port 2. Scsc is tuned to obtain the frequency response shown in figure 5. In order to make the peaks of the response as sharp as possible and thus allow an accurate determination of the peak values, the tapping point is made as small as possible. In addition, there will be minima in the $S_{11}$ plot shown in red in figure 5. The $S_{11}$ plot is sharper and provides a more precise but slightly different frequency spacing. As shown in figure 5, a 1.8 mm coupling gap results in a frequency spacing of 42.8 MHz when $S_{11}$ is used and 49 MHz when the resonator voltage is used. The result from $S_{11}$ is close enough to the required 41.7 MHz. The coupling for a 100 MHz frequency difference requires a

1.15mm coupling gap. Minor errors in the coupling gaps are not critical, as these values are used for the starting values for the filter optimization process, which then results in the final filter parameters.

The same design process can be applied for other filter types. The test circuits must be adapted for the different layouts. The combline filter layout is similar to that of the interdigital filter, however all the grounded connections are on the same side. The input tapping is the same as the interdigital filter, but the coupling gaps are 0.2 mm for the outer resonators and 1.8 mm for the inner resonators. The coupling gaps for the outer resonators are thus a lot smaller and that may limit the practicality of the filter. For the hairpin filter, the required tapping point is 3.7 mm from the start of the hairpin bend and coupling gaps of 0.45mm are required for the outer resonators and gaps of 0.95 mm are required for the inner resonators.

The test circuit for determining the tapping points for direct coupled filters is outlined in [7]. Using the procedure above, tapping points of 9.7 mm are required for the outer resonators and tapping points of 4 mm are required for the other resonators. For the direct coupled filter, the coupling lines are chosen to be 12.5% of a wavelength. This length has been found to give a reasonable stopband performance, whilst maintaining reasonable coupling tapping points.

## III Filter Comparison

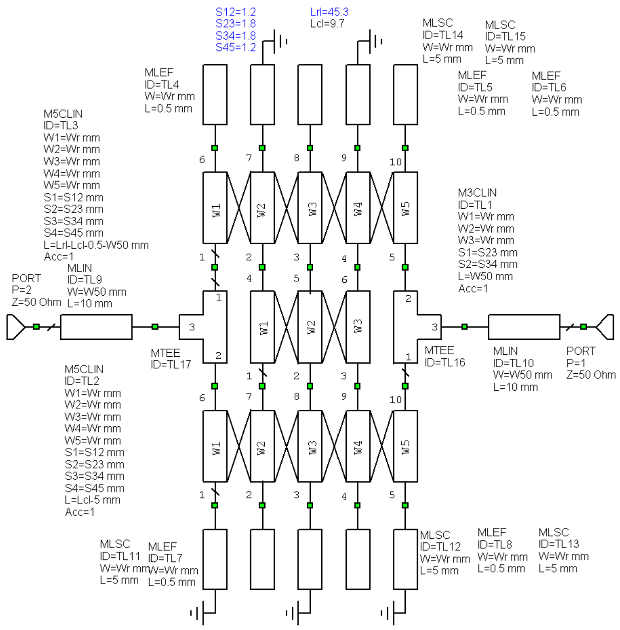


Figure 6. Circuit schematic for interdigital filter.

Once the coupling factors and tapping points have been determined, they are entered into the schematic circuit for each of the filter types. Figure 6 shows the schematic for the 5 resonator interdigital filter. Figure 7 shows the frequency response for the 4 different filter types, with the tapping points and coupling gaps indicated above. It can be seen that the initial performance of the filters is close to specification. To complete the design procedure, the filters are optimised to provide the fine tuning required to fully meet the design specification. In addition some manufacturing constraints can be included. For instance for the filters designed here, the minimum coupling gap size was set at 0.5 mm, which is larger than the coupling gap of 0.2 mm calculated for the combline filter.

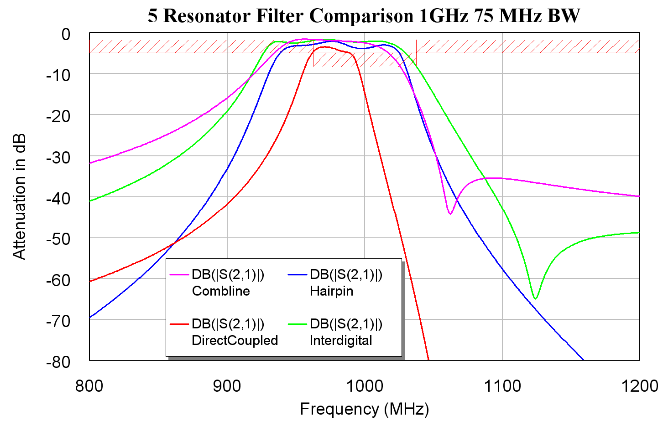


Figure 7. Frequency response of filters from design calculations.

The optimization goals should be kept as simple as possible to maximize the speed of the optimisation. The corner frequencies of the filter are specified by setting three optimization goals as shown in figures 7 and 8.

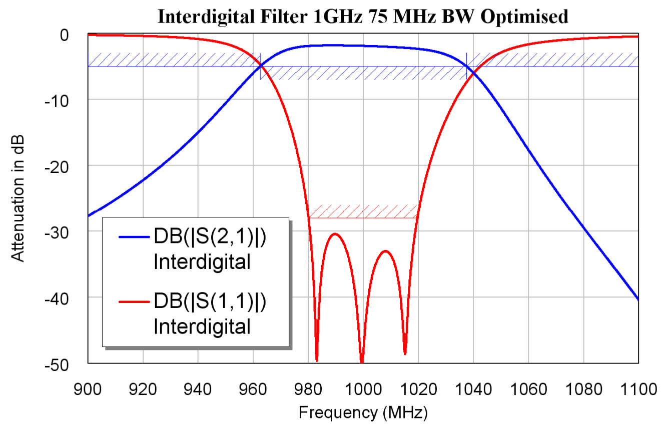


Figure 8. Interdigital filter after optimisation.

The insertion loss of the filter is close to 1 dB, so that the filter is to have less than 4 dB attenuation from 962.5 MHz to 1037.5 MHz, and more attenuation elsewhere. In addition an optimization goal with $S_{11}$ to be less than -25 dB from 980 MHz to 1020 MHz is added, to ensure that the filter has the lowest possible attenuation in the passband. For the interdigital filter, the frequency response after optimization is shown in figure 8.

The same optimization process is applied to the other 3 filters. Figure 9 shows the passband response of the 4 filters after optimization. Figure 10 shows the frequency response of these filters over a wide frequency range. The hairpin filter has a high stopband attenuation for frequencies less than the second harmonic, but has a harmonic response at that frequency. The direct coupled filter has a high stopband attenuation but has little attenuation at the third harmonic frequency. The combline and interdigital filters have smaller harmonic responses but have less stopband attenuation. The filter type to be used will thus depend on the stopband specifications. The direct coupled filter allow larger bandwidths to be realized [7] than the other filters.

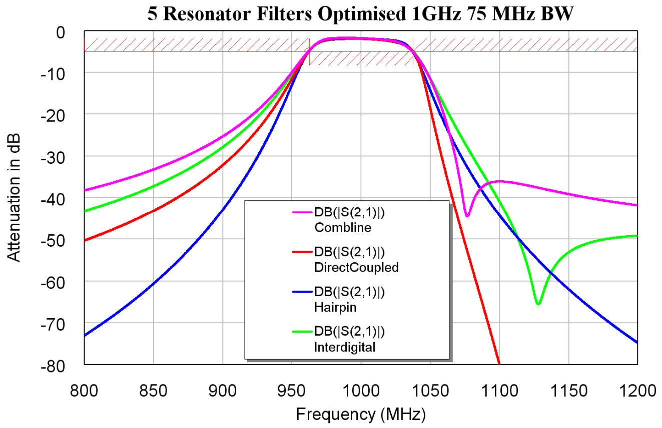


Figure 9. Comparison of filters, passband after optimisation.

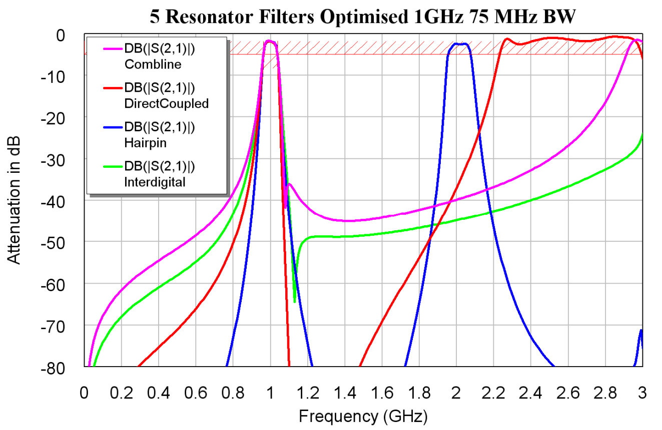


Figure 10. Comparison of filters, stopband after optimisation.

## IV Filter Measurements

The 4 filters with the simulated performance shown in Figures 9 and 10 were built. Figure 1 shows the photograph of the interdigital. Figure 11 shows the combline filter, figure12 shows the hairpin filter and figure 13 shows the direct coupled filter. The interdigital filter is 42 x 60 mm in size, the combline filter is 41 x 60 mm, the direct coupled filter is 75 x 60 mm and the hairpin filter is 78 x 70 mm. The photographs are reproduced to approximately the same scale

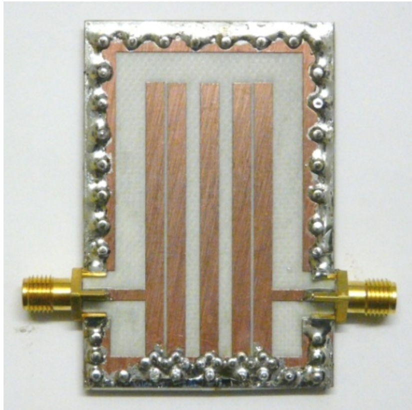

Figure 11. 1GHz combline filter.

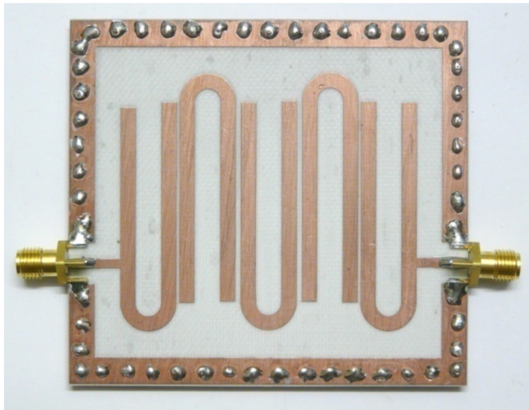

Figure 12. 1 GHz hairpin filter.

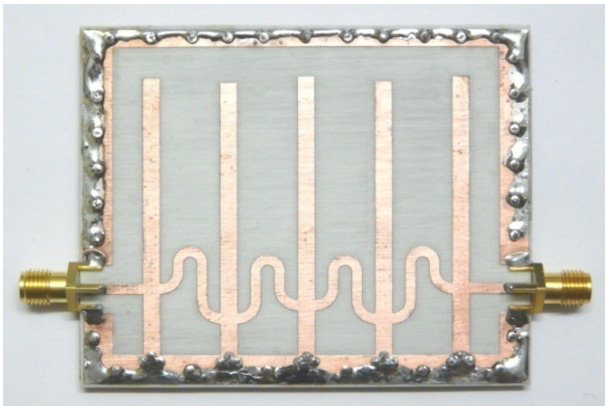

Figure 13. 1 GHz direct coupled filter.

Figures 14 to 15 show the measured frequency response of the filters. There is a remarkable agreement between the calculated and measured performance. The measured passband centre frequency of the combline, interdigital and direct coupled filters is 20 MHz or 2% lower than the design value. The resonators are thus 0.8 mm, or the substrate thickness, too long. This additional length is due to the via connecting the grounded end of the resonator to the bottom ground-plane. A second realization of those filters can take this effect into consideration to produce the correct centre frequency.

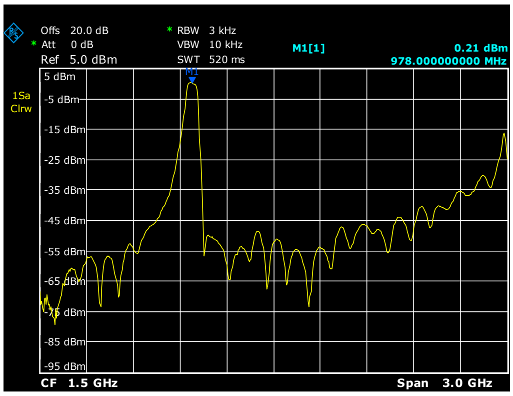


Figure 14. Interdigital filter frequency response.

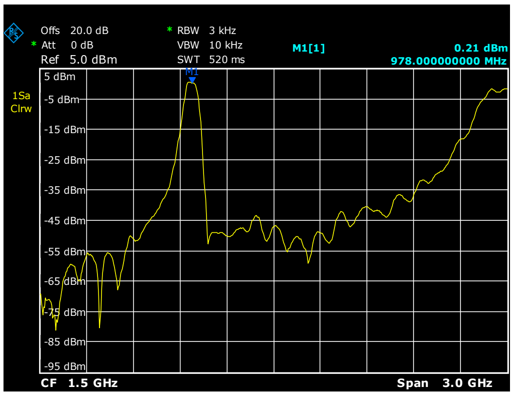


Figure 15. Combline filter frequency response.

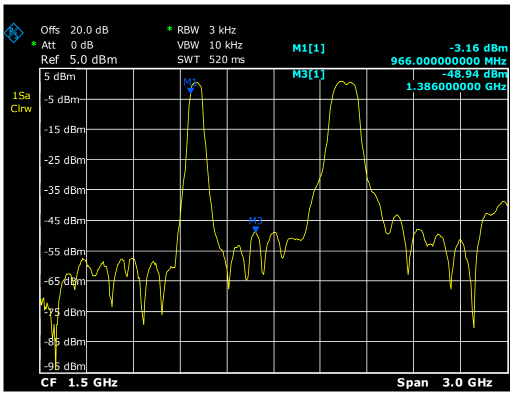


Figure 16. Hairpin filter frequency response.

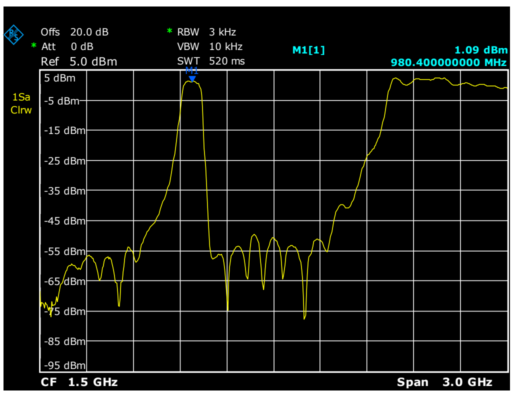


Figure 17. Direct coupled filter frequency response.

## CONCLUSION

This paper describes a design procedure that can be used to design any coupled resonator filter, whose layout can be simulated. Four different filters are designed, each with a similar passband response, but a very different out of band response. This allows the appropriate RF filters to be selected, such that unwanted RF signals are filtered out effectively. This design technique can be used to design new filters topologies, which can provide cost effective analogue signal processing by the receiver front end. One example would be a filter with some hairpin resonators, to achieve a high stopband attenuation, coupled to interdigital resonators to remove the harmonic responses from the hairpin filter.

The measured filter performances closely match those obtained by computer simulation.


## ACKNOWLEDGEMENT

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