Designing Edge-coupled Microstrip Band-Pass Filters Using in Microwave Office[TM]

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The design of printed filters, which perform close to their modelled specifications, can be a difficult and time consuming process using traditional methods. This application note shows a straightforward and largely non-mathematical method of tackling this problem using Microwave Office (MWO) assuming little or no knowledge of filter design. A basic familiarity with the operation of MWO is assumed. Microstrip filters are the examples shown, but the same principles can be applied to other types of filter (stripline, suspended substrate etc) using any of the excellent models that are available in the MWO element catalogue.

Introduction

Band-pass filters require precise transmission characteristics to allow a desired band of signals to pass with minimum loss through a two-port network and reject unwanted signals at both higher and lower frequencies. They are generally characterised by such terms as bandwidth, centre frequency, insertion loss, selectivity or rejection, ripple and return loss. For instance a particular filter may have a centre frequency of 1000 MHz and bandwidth of 100 MHz or 950 –1050 MHz. The insertion loss may be required to be less than 1dB in this band. The rejection at 1100 MHz may be specified as 40 dB or greater. It is usually desirable to have a high return loss in band. Return loss and in-band ripple are directly and inversely related. For instance an in-band ripple specification of 0.1dB corresponds to a return loss of 16.4dB, and a ripple of 0.5dB corresponds to a return loss of 9.6dB. Chebyschev polynomials are often used to describe the mathematical relationship between these quantities but from a practical point of view we need only understand their general nature.

An important part of understanding any particular design problem is a consideration of the trade-offs that apply. Increasing the bandwidth means less loss in the passband but at reduced selectivity. Increasing the ripple will increase the selectivity but at the cost of reduced return loss.

An effective way to understand a microwave band pass filter is to consider it as a number of resonant elements, see [2] for more information on this, connected with impedance or admittance inverters, which are often referred to as the filter couplings. The resonant elements may be series or shunt and may be of any positive number N. This defines the order of the filter. Increasing N increases the selectivity but at the cost of size and insertion loss. For an N section two port filter, with no cross couplings, we need to define N resonant elements, N-1 internal couplings and two external couplings. Using symmetry we can reduce this by a factor of two (or nearly two if odd numbers are involved). This is quite an important concept, and is useful when optimising filter structures. We need to ensure that we have the right number of optimisable variables to achieve a solution but not too many which would

over complicate the problem and lead to non-unique solutions from the optimiser. At a later stage when using an em-simulator it is probably even more important to not have more variables than necessary.

Couplings are often realised using the electromagnetic interaction between adjacent transmission lines or, for external couplings, a tapped transmission line structure onto the first and last resonators.

Design Example: A Hairpin Microstrip Filter

This is a very popular type of filter and is essentially a broadside-coupled half wave section filter, with the resonators folded into a 'hairpin' configuration to conserve board space.
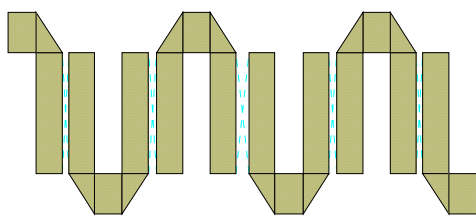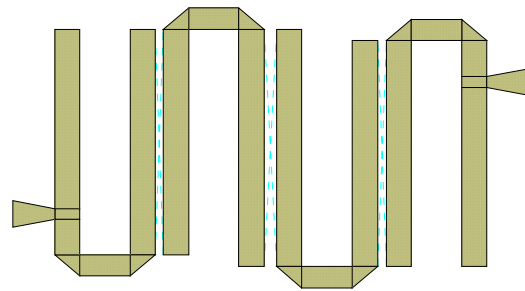

Fig1a

Fig 1b

Both Fig 1a and Fig 1b show filters of 4th order (N=4). Fig 1 a shows a coupled line input and fig 1b a tapped line input.

Let us give ourself some design parameters. Say we need a filter to pass frequencies from 2.4 to 2.6 GHz. We need to have 25dB rejection at 2.1 GHz and 30dB at 2.9 GHz. The in-band return loss should be 15dB or better with an insertion loss of less than 1.3dB.

We need to calculate the order N. A spreadsheet available from [3] is available for this. From it we can see that a 4 section filter can achieve this rejection even if we extend our bandwidth to 300MHz. Filters often show asymmetric responses and so the calculation of N should only be considered as a first estimate. Any previous knowledge can be useful and the optimiser can be used at the end to adjust the response to achieve a better fit for design specifications and allow an adequate margin. The more margin that we have, the higher our yield will be at the end. We next choose our substrate. To start with, we can choose a relatively inexpensive material such as Taclam[TM] TLC32-0310. This has a loss tangent of 0.003 and a dielectric constant of 3.2. If we run into insertion loss problems we may need to review this choice later.

Traditional design methods for this type of filter call for different line widths in the hairpin resonators. However, when we think of the filter in terms of its couplings and resonances we can consider the couplings to be determined by the line spacings and the resonant

frequencies to be determined by the resonator line lengths. The line width will have an effect on both of these but a solution should be possible for any arbitrary width. For now, let's keep it simple and make them all the same width at some convenient value of 60mils. (1mil = 0.001"). We need to calculate the dimensions of the hairpins to make them a half-wave long at 2.5 GHz. We do not have to be too exact at this stage. The built in TX line calculator in MWO gives us what we need
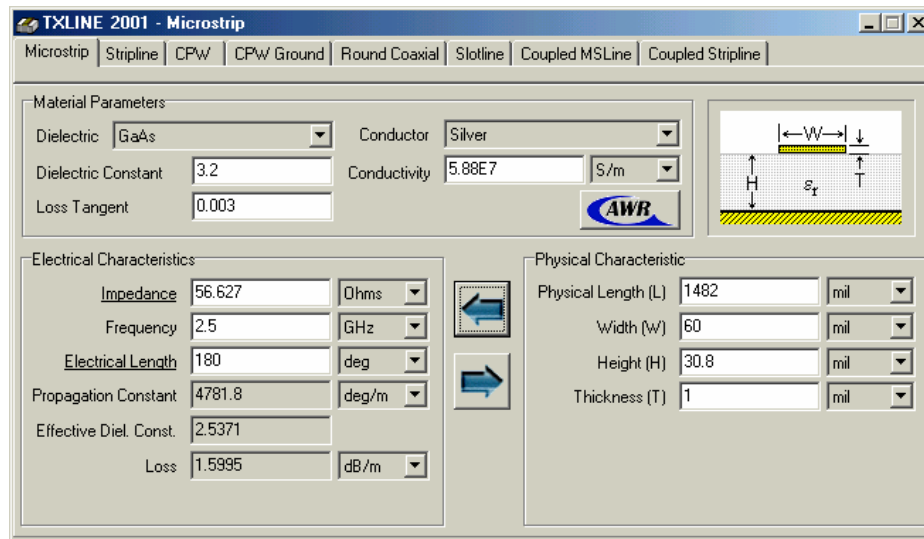

Fig 2. The MWO Transmission Line Calculator

Next, a schematic diagram of the filter is constructed, as in fig 3. Note that 'X' or em-based models are used wherever possible for better accuracy in the filter design. A tapped structure is chosen for the external couplings. Generally, coupled line inputs are fine for narrow band designs but the narrowing gap can be a problem as the bandwidth is increased and more coupling is required. We can choose a length for the horizontal sections of the hairpin to be 120mil. So, the other two arms will be approximately 690mils each. Taking a guess at the additional length in the right angle bends and the end effects we can say that these should be around 650mils long. A tapered line is used to give 75mil wide, corresponding to a 50ohm track width, input and output ports to the structure. We have defined this structure by its parameters, which will make it easy to change the substrate if we need to. Also, we can easily adapt the schematic to other frequencies and bandwidths if we need something similar in future.

We have a 4 section filter. We have to decide how many adjustable, either by the tuner or optimiser, parameters are necessary. We need to adjust 2 internal couplings, 1 external coupling and two resonant frequencies if we assume symmetry. The two resonant frequencies can be adjusted by lengths l2 and l3, which we initially set to 650 mils. The external coupling is adjusted by the position of the tap point "ltap". We have not calculated this but let's just guess a starting point of 300 mils. The two internal couplings are determined by the gaps between coupled lines, which we call s2 and s3. We can take a starting guess and make these 50mil. We would like to make the end hairpins symmetrical and so the remaining length l1 in these is defined by adding an equation

l1=l2-ltap-30

*It is important to either fix all other dimensions or define them by means of equations relating to the **5** variables l2, l3, s2, s3 and ltap.*

Let's take a look at what we have in the frequency range 2 to 3 GHz
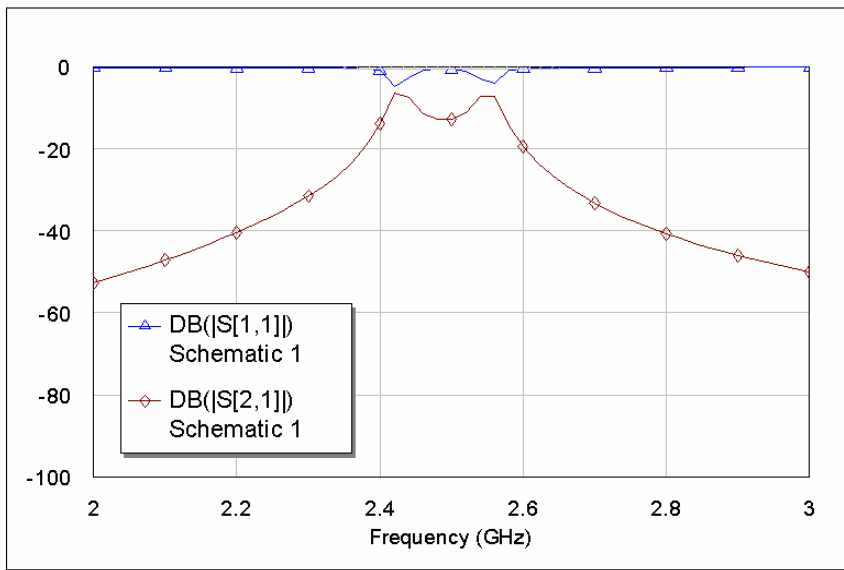

Fig 4. Filter Response with Initial Estimated or Guessed Values

We can add a single optimisation target to the graph of –21 dB for return loss in the pass-band of 2.35 to 2.65 GHz. It is better to leave the optimiser alone for now and get the feel of the circuit using the tuning tool. We can adjust the 5 parameters very easily and also adjust the constraints so that we have a sensible range of adjustment.

This is a 4 section filter so we need to tune for a reasonable return loss in the pass-band, *showing 4 distinct minima,* as shown in fig 5. It is pretty much a golden rule to always count your return loss minima. If these are fewer than the filter order N, then there is a possibility that one or more of the resonances are out of band. This will cause poor performance. At this point we need to check for physical realisability. If we have defined some parameters by equations, we have to check that they are not negative. Coupling gaps can be a problem if it looks like they are going to be too small for easy manufacture.
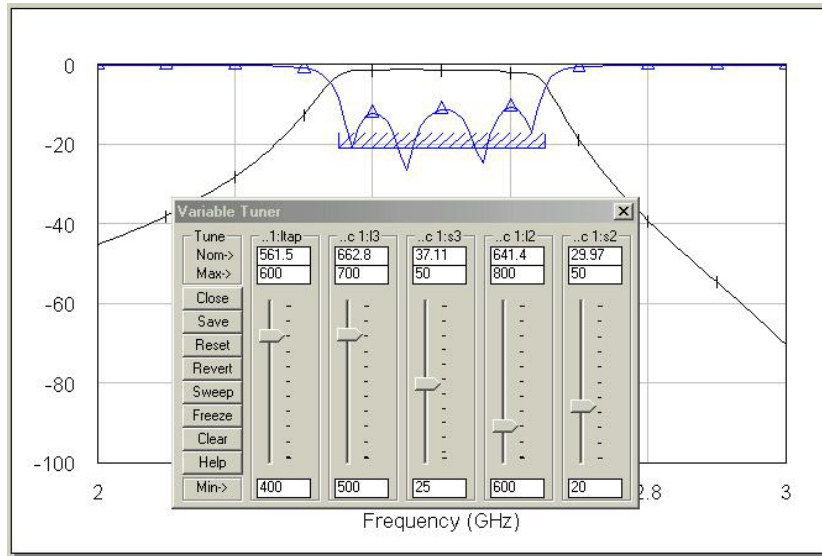
Fig 5   Filter Response After Tuning

As a final step we can run the optimiser. It is sometimes quicker to reduce the project frequency range during the optimisation to be just about twice the pass-band. The optimisation should only take a few minutes. Again we take care of our ripples. If one goes missing during optimisation then we need to backtrack until we find it. To prevent this happening, it is better to tune the filter initially to be slightly too narrow in bandwidth.
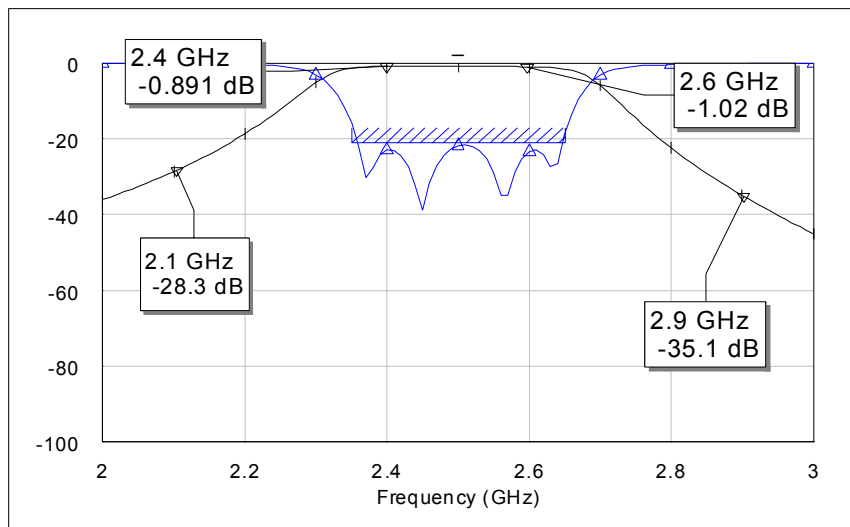

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Fig 6 Filter Response after optimisation and with Specification Values

Figure 6 shows the performance at the specified frequencies. The rejection levels at 2.1 GHz and 2.9 GHz are within specification. Filters can be particularly intolerant of parameter variation and as much margin as possible needs to be built into their design. In the above case it may be advisable to re-optimise to a slightly higher band centre to reduce the insertion loss at 2.6GHz and increase the rejection at 2.1GHz. Additional optimisation targets may also be added at this stage.

Using the layout facility in the EDA tool, a structure very similar to fig 1b will be seen which can usually be exported to other applications directly in a variety of formats including Gerber and DXF.

Two more examples of microstrip filters (figs 7-10) showing the possibilities of using cross couplings to achieve transmission zeros in the stopband are shown. The filter, in the first example, was initially suggested by Hong and Lancaster [4]. The same method of initially using the tuner and then the optimiser to achieve the desired response was followed.
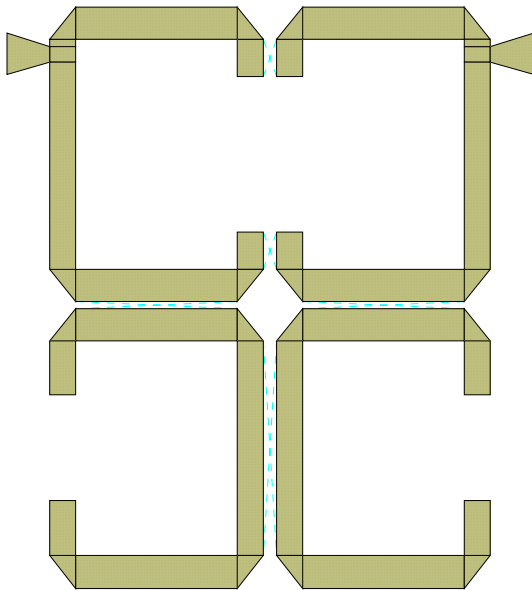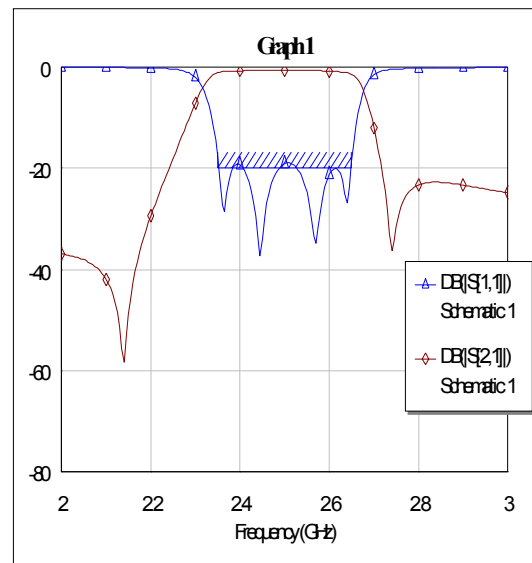

Fig 7

Fig 8

The filter circuit of fig. 7 shows how to implement a pair of transmission zeros. One is either side of the passband. Even though the symmetry is not perfect this is usually referred to as symmetrical response.

The filter in fig. 9 exhibits an asymmetrical response with a single transmission zero in the upper stopband.
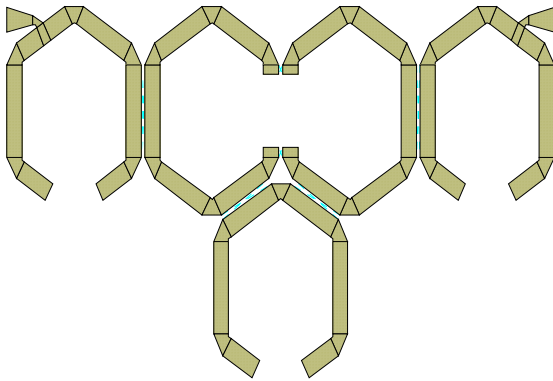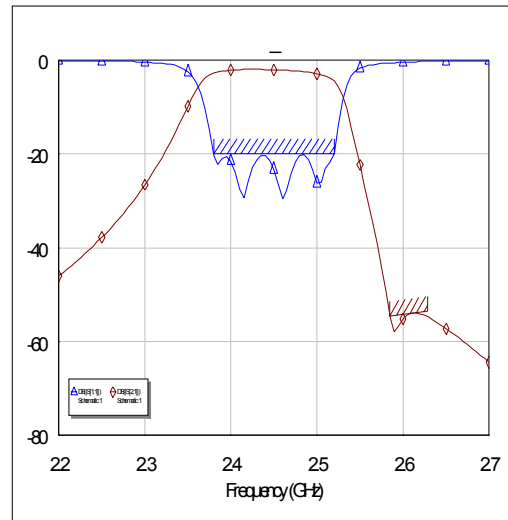
Fig 9

Fig 10

Using an Electromagnetic Simulator to Check and Correct for Design Inaccuracy.

With the availability of cost-effective electromagnetic simulators, it is now feasible to take the design process one step further and so avoid much of the costly iterative process of manufacturing prototype PCBs, testing, and reworking before being confident of the viability of a particular design.

The models in Microwave Office, especially X models or those are based on em-simulations, are very accurate but, of course, they cannot take into account the effects of the circuit enclosure. The unwanted effects of a metal cover over a microstrip circuit will be a familiar problem to most engineers. Often, schematic diagrams are somewhat idealised. For instance, the schematic as shown in figure 3 is fine for the purposes of understanding the circuit but it does not show the effects of non-adjacent coupling or coupling between the ends of the resonators in each hairpin. In fact if the circuit diagram had been constructed using multi-element coupled microstrip lines, rather than just pairs, the linear simulator results would have been more accurate.

Sometimes these secondary effects are small enough to be negligible, but for the design of circuits, particularly filters, to tight specifications, these effects need to be measured and compensated for in the design.

Modern em-simulators such as AWR's EMSight, sometimes referred to as 2.5 D because they are designed for the analysis of planar structures connected in the z dimension with vias, are very easy to use. Circuits can be directly imported or copied and pasted from the layout views in much the same way as any other Windows application. Multi-layer boards are described as a series of layers, which allows structures such as suspended substrate,

stripline, co-planar waveguide to be constructed. The height of a lid over microstrip is defined by the inclusion of an air dielectric layer.

Substrate Information

Enclosure			Dielectric Layers			Boundaries		

Dielectric Layer Parameters

Layer	Hatch	Via Hatch	Thickness mil	er	Loss Tangent	Bulk Cond. (S/M)	View Scale
1			250	1	0	0	1
2			30.8	3.2	0	0	1

		250	1	0	0	1

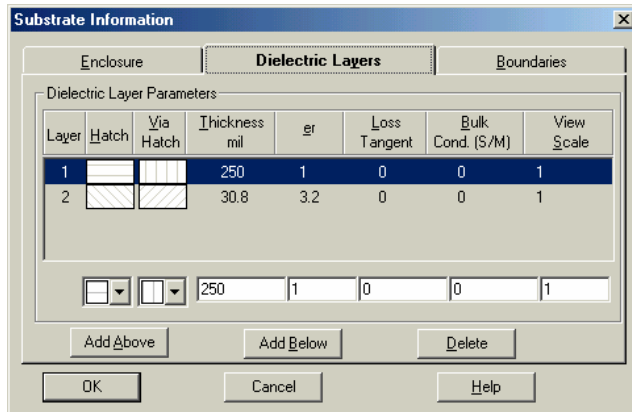Add Above   Add Below   Delete

OK   Cancel   Help

Fig 11 Setting up the substrate information.

In this example we are going to enclose the first circuit described previously in a box of size 1.5 x 1" with a metal cover situated at a height of 250mils over the microstrip circuit.
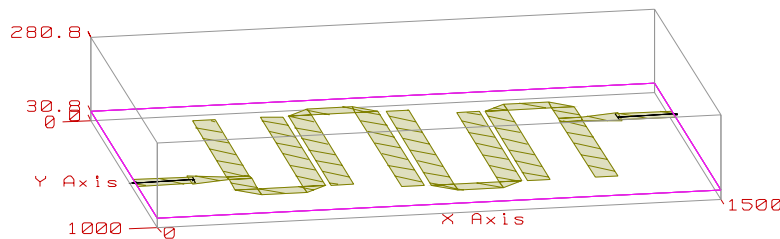
Fig 12   A 3-D view of the circuit is a useful check.

Some experience is necessary to set the grid and frequency points in an em-simulator to give the right compromise between speed and accuracy which will, of course, vary depending on PC processor speed and the available amount of RAM. There is usually a temptation to set the grid spacing to be too small and the number of frequency points to be too high. Before any em-simulation, it is advisable to check for details and an estimate of the solution time.

Fig 13. Know how much time your simulation requires.

In the design example shown previously we are going to check our design using 5 mil grid spacing. It is safer and more efficient to select all the copper in the layout and snap all dimensions to this grid spacing, but of course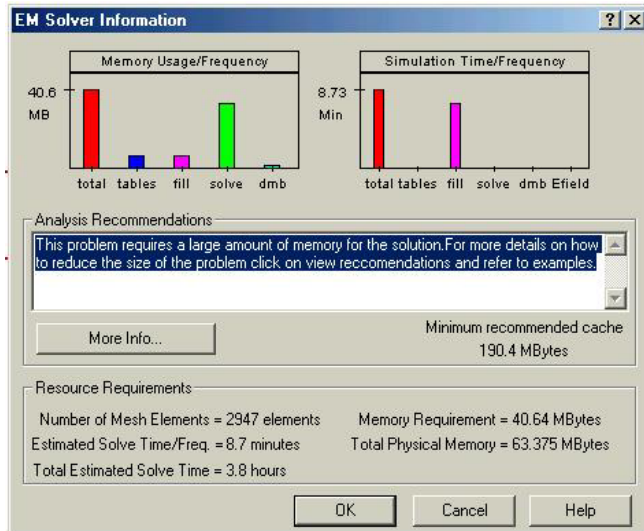 some unwanted rounding of values will occur. We do not expect to see a perfect response, the first time, from the em-simulator and because of the slow solution times we cannot expect to tune anything other than very simple structures directly. Instead, we use the em-simulator to calibrate our imperfect linear model using a curve fitting technique.

Parameter	Optimised Values from Linear Simulator	Rounded Values for Emstructure0	Curve Fit Values for Emstructure0	1st Correction Emstructure1	Curve Fit Values for Emstructure1	2nd Correction Emstructure2
ltap	505.2	505	489.0	520	504.1	520
l2	636.5	635	625.0	645	631.7	650
l3	668.2	670	652.5	690	676.6	685
s2	20.55	20	25.96	15	20.47	15
s3	30.00	30	39.24	20	26.79	22.5*

All values are in mils.

 * This value is obviously not on a 5mil grid and was effectively achieved with a stepped gap of 50% 20mils and 50% 25mils. (See fig16)

The curve fit for each em-structure was achieved initially by using the tuner to obtain the approximate shape of the mistuned response and finally by setting an optimisation target, of

zero, for the built-in function "smodel2" which gives a least squares error between two sets of s parameters.
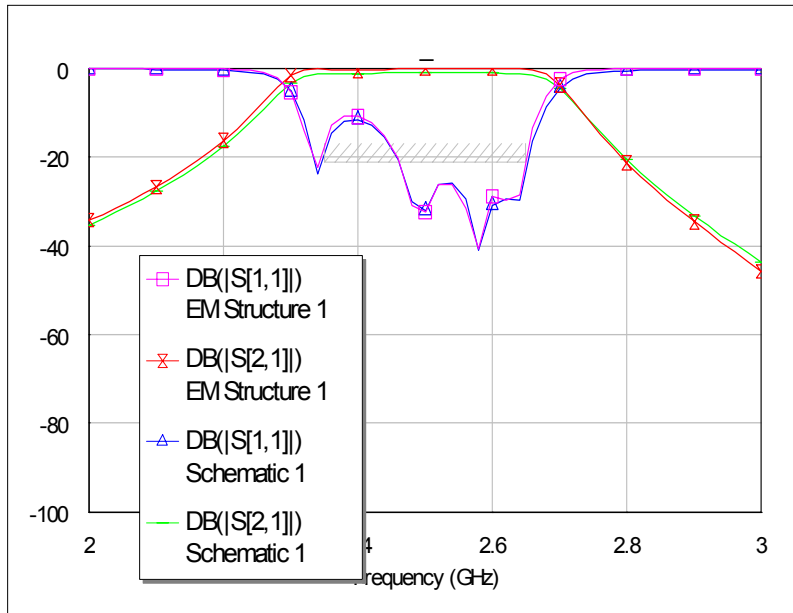
Fig 14 Curve Matching the linear simulator to the em-simulator results.

Each new layout can be conveniently generated from the schematic diagram by adjusting the 5 variables to new values. This is then copied into the em-simulator and snapped to the grid size.
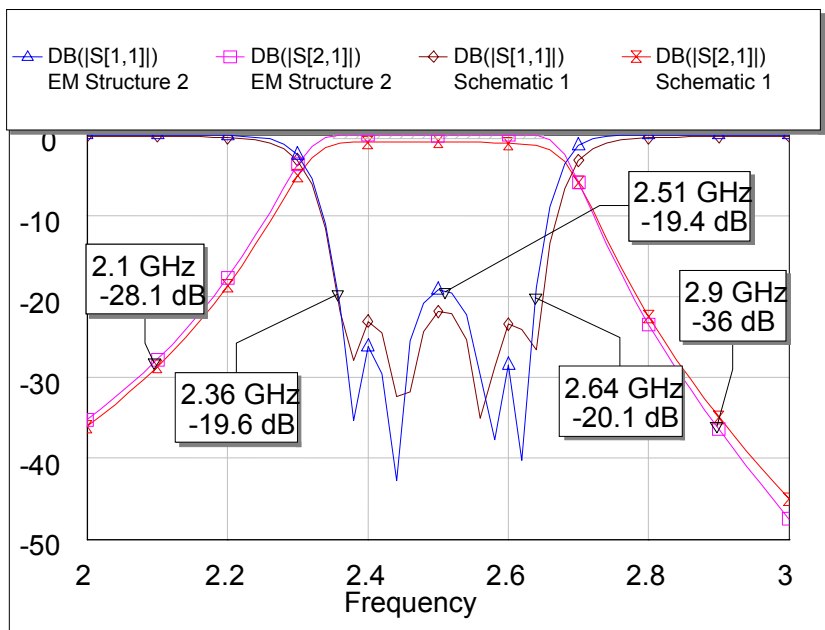
Fig 15. The Em-simulated results after 2 iterative corrections.

Fig 15 shows the corrected em-simulated results and the ideal response from the schematic with the five optimised variables. The remaining imperfections in the response of fig 15 are largely due to the limitations of the 5mil grid, except for optimistic loss shown in the

response. This was due to the em-simulator being set up for lossless conductors to minimise simulation times.
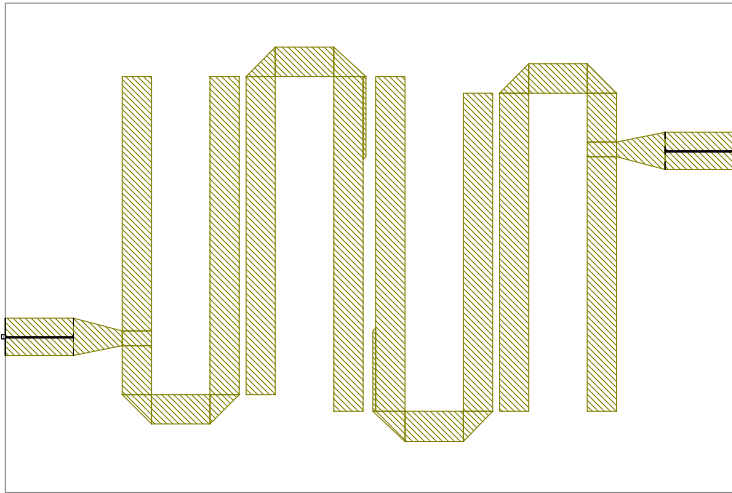
Fig 16. The final layout.

Note the two shims of metal, in Fig 16, drawn in between the second and third resonators to set the coupling to the correct level without reducing the grid size.


Conclusion

An effective method of designing and checking microstrip filters has been demonstrated with the minimum of mathematical involvement. The same process can be applied to other types of filter (stripline, suspended substrate etc) using any of the models that are available in the EDA tool element catalogue. The project files containing this worked example can be downloaded from [3].

Acknowledgement

I would like to thank Mr Lloyd Nakamura, AWR Inc, for suggesting the curve fitting technique using the built-in function "smodel2" in Microwave Office.

References

[1] GL Matthei, L Young, EMT Jones "Microwave Filters, Impedance Matching Networks, and Coupling Structures" Artech House 1980

[2]   P. Martin and J.B.Ness "Coupling Bandwidth and Reflected Group Delay Characterisation of Microwave Band-pass Filters", Applied Microwave and Wireless, May 1999

[3] http://www.rfshop.webcentral.com.au

[4] J.S. Hong & M.J. Lancaster "Couplings of Microstrip Square Open Loop Resonators for Cross Coupled Planar Microwave Filters" IEEE MTT Vol. 44 No12 Dec 1996

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Fig 3

MOPENX
ID=MO1
W=w1 mil

lt=t20
lh=t20

w1=60

ltap=505.2
l2=636.5
s2=20.55
s3=30.0
l3=668.2

l1=l230ltap

MLIN
ID=TL4
W=w1 mil
L=ltap mil

MTAPER
ID=MT3
W1=75 mil
W2=30 mil
L=100 mil

PORT
P=1
Z=50 Ohm

MTEEX
ID=MT1
W1=w1 mil
W2=w1 mil
W3=30 mil

MOPENX
ID=MO4
W=w1 mil

MBEND90X
ID=MS1
W=w1 mil
M=.5

MLIN
ID=TL9
W=w1 mil
L=lt mil

MBEND90X
ID=MS8
W=w1 mil
M=.5

MOPENX
ID=MO7
W=w1 mil

MOPENX
ID=MO6
W=w1 mil

MBEND90X
ID=MS5
W=w1 mil
M=.5

MLIN
ID=TL11
W=w1 mil
L=lt mil

MBEND90X
ID=MS4
W=w1 mil
M=.5

MLIN
ID=TL6
W=w1 mil
L=l1 mil

M2CLIN
ID=TL1
W1=w1 mil
W2=w1 mil
S=s3 mil
L=l2 mil
Acc=1

MOPENX
ID=MO8
W=w1 mil

M2CLIN
ID=TL3
W1=w1 mil
W2=w1 mil
S=s3 mil
L=l3 mil
Acc=1

M2CLIN
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W1=w1 mil
W2=w1 mil
S=s2 mil
L=l2 mil
Acc=1

MLIN
ID=TL7
W=w1 mil
L=l1 mil

MTEEX
ID=MT2
W1=w1 mil
W2=w1 mil
W3=30 mil

MLIN
ID=TL8
W=w1 mil
L=lt mil

MOPENX
ID=MO3
W=w1 mil

MOPENX
ID=MO8
W=w1 mil

MLIN
ID=TL10
W=w1 mil
L=lt mil

MOPENX
ID=MO5
W=w1 mil

MBEND90X
ID=MS3
W=w1 mil
M=.5

MBEND90X
ID=MS2
W=w1 mil
M=.5

MBEND90X
ID=MS7
W=w1 mil
M=.5

MBEND90X
ID=MS6
W=w1 mil
M=.5

MSUB
Er=3.2
H=30.8 mil
T=1 mil
Rho=1
Tand=0.003
ErNom=3.2
Name=tlc-1

MTAPER
ID=MT4
W1=75 mil
W2=30 mil
L=100 mil

PORT
P=2
Z=50 Ohm

MLIN
ID=TL5
W=w1 mil
L=ltap mil

MOPENX
ID=MO2
W=w1 mil