Break and Interpolate Technique: A Strategy for Fast EM Simulation of Planar Filters

by Shawn Carpenter
Sonnet Software, Incorporated, Liverpool, NY USA

In this article, we examine a strategy for using high frequency EM field solvers to accomplish high accuracy simulations with dramatically reduced analysis times. This paper introduces what we call the "Break and Interpolate" (BI) technique, where a passive high frequency resonant circuit is subdivided into several small, fast EM analyses over a small number of frequencies. EM analysis data from the subcircuits is recombined and interpolated in a linear circuit simulator to provide a detailed analysis result. Analysis time savings with the BI Technique can range from a factor of 10x to well over 100x over full structure EM analysis.

In the typical high frequency RF design cycle, linear and non-linear circuit simulators are used to create a first-pass design. These simulators usually have library elements for microstrip, stripline, CPW or other topologies, and provide fast simulation of prototype designs. Sophisticated commercial circuit theory software packages may have schematic entry, design optimization (based on library models), companion layout design, and the ability to export physical designs in a variety of formats. In general, the linear circuit analysis programs are fast, enabling a frequency sweep analysis in seconds to minutes.

However, there are limits to the accuracy of the linear circuit analysis tools. They are not able to model arbitrary circuit cross-coupling beyond individual circuit elements — especially important when circuit compaction is required — and assume each element in the circuit to be uncoupled from every other element. The models for various circuit parts, such as microstrip bends, tees, and junctions can have limited frequency and physical material property ranges of validity, and may themselves carry a significant degree of model error. They cannot generally account for package effects, such as sidewalls or package resonance effect on circuit behavior.

To overcome these limitations, high frequency EM simulators, based on computational electromagnetic formulations of Maxwell's equations have gained great popularity. These solvers take as inputs a geometric description of the circuit along with material properties for metal and dielectrics, and develop results in the form of S-, Y-, Z-parameters or possibly an extracted model in SPICE format. They can be used to extract an extremely accurate model for an individual circuit element, or for general circuit parts for which no library model exists. An EM simulator can analyze virtually any circuit configuration you can imagine; if you can lay it out, an EM simulator can usually simulate it. For microstrip, stripline, coplanar waveguide or other multilayer circuit designs, planar EM simulators (sometimes called 2.5D or 3D Planar simulators) are generally most efficient for accurate simulation of planar circuits, as opposed to using full 3D EM simulation. Planar circuits include single and multi-layer structures and vias.

Today's high frequency EM simulators can provide results that are very accurate — often scattering parameters with error that may be less that 1%. Some, like the Sonnet EM Suite, are even capable of reliable error bounds below 0.1% or even 0.01% — a very valuable tool for reducing costly design cycle time forced by redesign. Look at most designs above 1 GHz that met specifications on the first pass, and you will usually find an EM solver was used in some phase of its design.

However, the general downside of using an EM simulator is computation time and sometimes computer resources. It is not uncommon for an EM simulator to require on the order of minutes, or even hours for a highly converged result. For the class of circuits that involve high Q responses, a large number of frequency points is often desired for finding in-band match characteristics, ripple, transition band response and stopband performance. Circuit response can change very quickly, and usually a large number (50 to 100) of points are needed to adequately map out the response of the circuit. This is especially true for bandpass filter designs.

The Break and Interpolate (BI) Technique

In this article, we'll show a simple strategy for significantly reducing simulation time for filters that involve resonant coupled line structures. This strategy, which we call the "Break and Interpolate" (BI) technique involves breaking resonant structures up into two or more smaller non-resonant structures for EM simulation at a small number of frequencies. The separate EM simulation results are cascaded in order using a simple linear circuit simulator, which will interpolate the EM results to provide a sweep with many more frequency samples. As we'll see, a very accurate, detailed filter response will result from only 5 or 6 EM simulation frequencies of the subdivided circuit components. The basic flow is described in the flowchart in **Figure 1**:

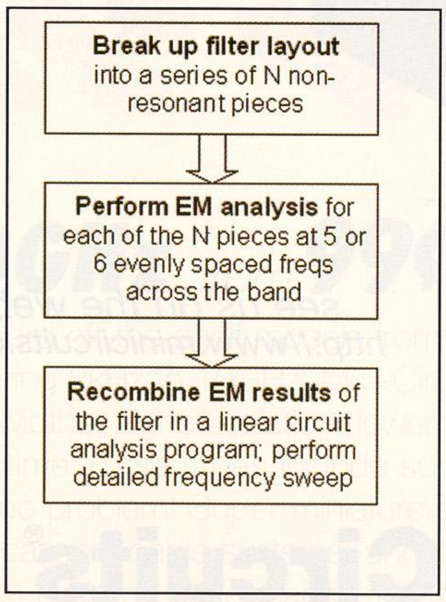

Figure 1 – Break and Interpolate (BI) Technique Process Flow for EM Analysis

An added benefit for the BI technique is that the subcomponent circuits are usually much smaller than the overall circuit. In a typical Method of Moments planar EM field solver, the memory requirements increase with the square of the number of unknowns, and the processing time increases with the cube of the number of unknowns. For example, if you have two component sub-circuits, each with half the number of unknowns involved in the full circuit, your memory requirements will be about _ that of the full circuit analysis (if you run the jobs one at a time) and the analysis time will be about _ that of the full circuit (1/8 time for each subcircuit). If you subdivide the circuit into more than 2 pieces, the savings in memory and simulation time are even more dramatic. Many complex filters can be subdivided into jobs small enough for analysis in the free Sonnet Lite simulation suite.

EM Field Solver Calibration Noise Floor

Good results using this technique depend on very accurate EM simulation, and require a very clean de-embedding algorithm in the EM simulator. Much like your vector network analyzer, your EM field solver also exhibits discontinuities associated with the ports. These discontinuities need to be removed, or undesired parasitics and discontinuities will introduce error into your simulation response. When you are cascading successive EM analysis results from several subcircuits, the phase of the S-parameter data at each of the ports must be very accurate, or the interpolation process will add unacceptable error to the result when you build the full filter back up — especially important when parts of the combined circuit are resonant structures, because their phase characteristics can change very quickly at resonance.

De-embedding error sources in planar EM solvers can come from three major areas:

1. General numerical processing error (happens most often in low frequency analysis)
2. Failure to adequately remove the port discontinuity itself
3. Failure to properly remove the extra phase length of added transmission lines to the de-embedding reference plane, or the cross-coupling between a pair of transmission lines leading to a reference plane.

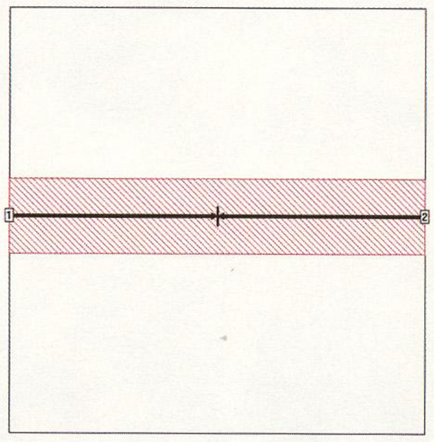

Figure 2 – Example circuit for the zero-length thru line test for evaluating the de-embedding noise for an EM solver

A high frequency planar EM field solver must have a very low de-embedding noise floor for this technique to work reliably. Similar to evaluating the quality of a calibration on a vector network analyzer, a "zero length thru test" is a good means for evaluating an EM field solver's calibration noise floor, and it works well for general classes of transmission lines and coupled transmission lines. These tests are easy to set up and quick to run in nearly any high frequency EM solver. An example of such a calibration check circuit is shown in **Figure 2** for a check of the calibration noise floor in the Sonnet® 3D planar EM simulator, ***em***®.

To check your EM solver's cal noise floor, create a series of through line examples for a series of impedances. We suggest a set of three standards, with characteristic impedances of 25, 50 and 100 ohms. Be sure your solver is providing renormalized 50 ohm S-parameters, and not generalized S-parameters. Set ports at opposite ends with 50 ohm sources to create a through line for each example, and instruct the solver to de-embed to a plane in the middle of the line, touching from each end. The result of the simulation of this structure, after calibration, should be a perfect through ($|S21| = 0$ dB, $Ang\{S21\} = 0$), regardless of the transmission line characteristic impedance. A relative indicator of the noise floor for the solver's de-embedding algorithms is provided by observing the input reflection that results, $|S11|$. We use this indicator as

Continued on page 22

Continued from page 18

the "calibration noise floor," and this should be at least –60dB for the BI technique to work reliably.

In this paper, we consider designs on two substrate types. Two designs employ microstrip lines on alumina substrates with a dielectric constant of 9.9. The third involves stripline structures on a substrate with a dielectric constant 2.94. To check our calibration quality, we investigate three standards for each substrate type: through lines with characteristic impedances of 25, 50 and 100 ohms for each material type.

The calibration results for each of the 6 standards are shown in **Figure 3**, and the |S11| data is plotted across the band of 1 to 15 GHz. As the plot shows, the de-embedding noise floor is below –100 dB in all cases for the Sonnet em solver.

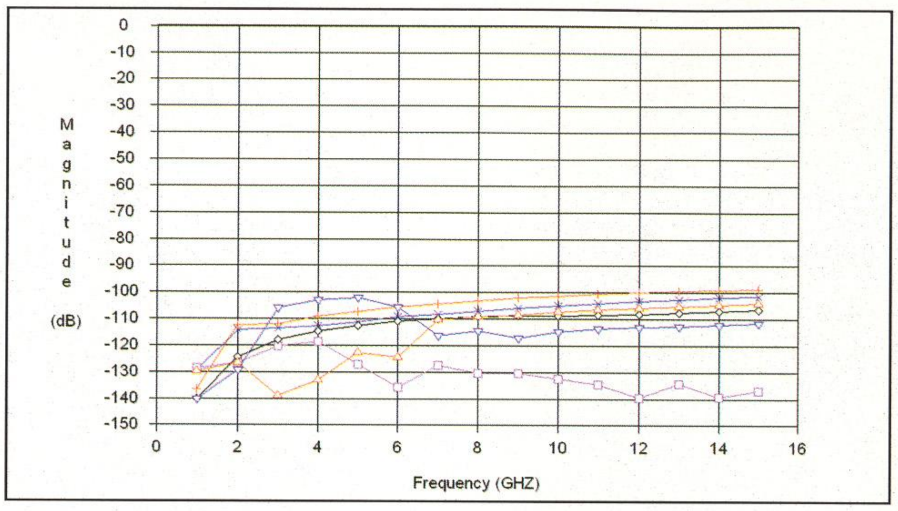

*Figure 3 – Zero-length through line test noise floor for Sonnet **em** for 25,50,100 ohm lines on substrates with dielectric constants of 9.9 and 2.94.*

The evaluation of coupled line structures is also at the heart of the BI technique, and often you will need to be able to de-embed coupled line structures that involve moving the reference plane forward on the coupled lines (as an example, look ahead at **Figure 10**). Some EM field solvers perform acceptably on the zero length through line test, but cannot adequately remove cross-coupling between parallel coupled lines that lead up to a reference plane. To check the quality of the de-embedding noise floor for your EM solver, we suggest you try a zero-length coupled line standard, as portrayed in **Figure 4**.

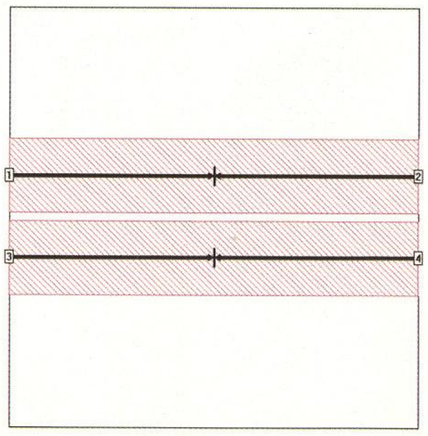

Figure 4 – Zero-length coupled line test standard example

Ideally, the results for this standard should show perfect through transmission and no input reflection if the solver is properly de-embedding the effect of parallel line coupling from the problem. In addition, all of the cross-coupling S-parameter magnitude terms should be zero. Any value other than zero provides a relative indicator of the noise floor for calibration of coupled lines.

Consider a zero-length coupled line test example using coupled microstrip lines on alumina. The lines have widths corresponding to 25 ohm single line characteristic impedance, and have 1/9 linewidth separation. Results for input reflection coefficient and all cross-coupling terms are presented in **Figure 5** as computed by the Sonnet em solver. Input and cross-coupling terms are all under —90dB, indicating good calibration for coupled line structures. Again, for completeness we suggest you try at least three different relative linewidths and spacings to check the range of your solver's accuracy, and perhaps an example where each of the coupled lines has a different width.

One should be especially careful to perform these tests on an EM solver prior to using BI subdivision. Not all

Continued on page 24

Continued from page 22

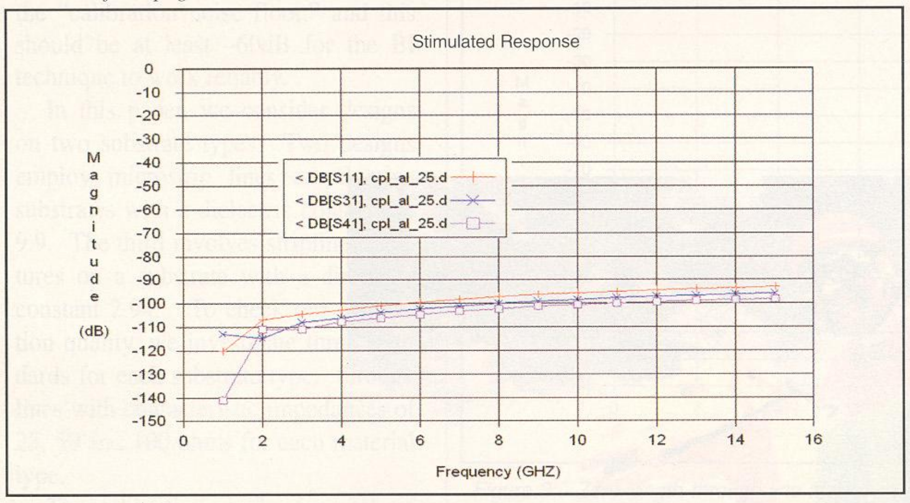

Figure 5 – Sonnet ***em*** *calibration noise floor indicators for zero-length through coupled line tests on alumina*

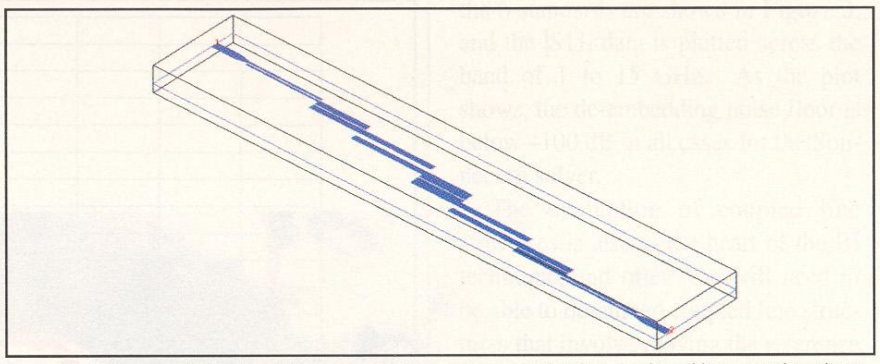

Figure 6 – Edge-coupled microstrip bandpass filter design on alumina, enclosed in a metal package

EM solvers can successfully use this technique. Now that we've demonstrated the evaluation of calibration noise floor for the Sonnet EM solver, let's move on to consider some examples of the BI technique.

Examples of the BI Technique

Now we consider three filter examples to illustrate the (BI) technique:

1. An edge-coupled microstrip filter
2. An interdigital microstrip filter, and
3. A stripline hairpin filter

All three filters involve coupled resonators; the first and third involve edge-coupled pairs, while the second uses multiple coupled lines.

Edge-Coupled Microstrip Filter

The layout for our first example is shown in **Figure 6**. This design (courtesy of ITT Industries Advanced Engineering and Sciences Division of State College, PA) is an edge-coupled bandpass filter design, with 7 coupled sections. The filter is fabricated on 15 mil alumina (ε_r = 9.9) in a chamber with a 230 mil cover height. The filter is designed for a passband of 7.5 GHz to 10.5 GHz. An EM simulation of the full filter using Sonnet **em** is compared in **Figure 7** (shown on *Page 26*) with filter measurements.

Next, we apply the BI technique to the simulation of the filter. There are a couple of different ways that one can break this filter down into separate non-resonant pieces. The first approach, which we'll call Method 1 involves sectioning the filter into 4 pieces, and simulating each piece separately in the EM solver, as illustrated in **Figure 8** (shown on *Page 26*). The circuit is divided at the dashed lines, and each of these subcircuits is simulated in the solver at only 5 frequencies each, spaced at 1.6 GHz intervals across the band of 5 to 13 GHz. The subcircuits are illustrated in **Figure 9(a)-(d)** (shown on *Page 26*), and half the filter is subdivided since the filter is symmetric about a vertical line through the middle of the fourth coupled line section.

Once the subcircuits of **Figure 9(a)-(d)** (shown on *Page 26*) are simulated in the EM solver, the results can be combined in nearly any linear circuit simulator. The Sonnet EM Suite has an integrated netlist analysis tool that can be used for this purpose. The coarse EM data (1.6 GHz frequency increments) is combined in the linear simulator, and the linear simulator interpolates the response between each of the EM simulation data file points yielding an exceptionally fine and detailed frequency sweep.

Continued on page 26

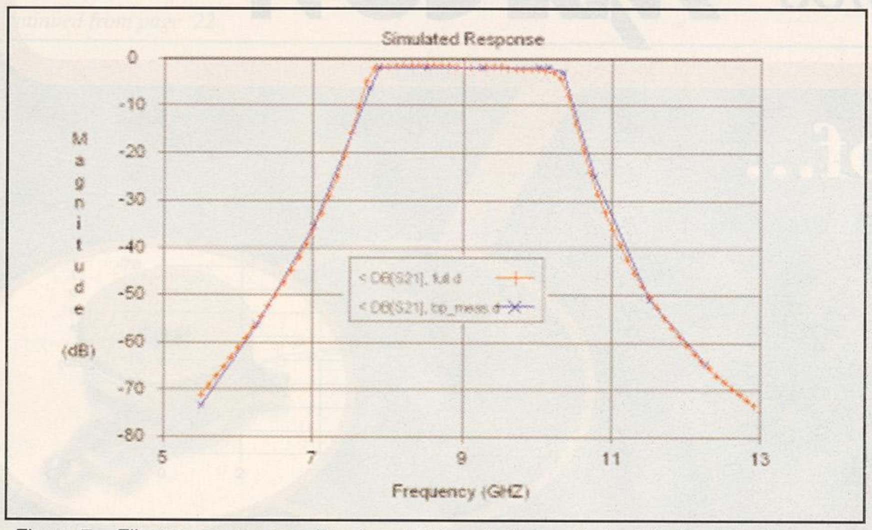

Figure 7 – Filter measurements (blue) and EM analysis results (red) for the full edge-coupled microstrip bandpass filter

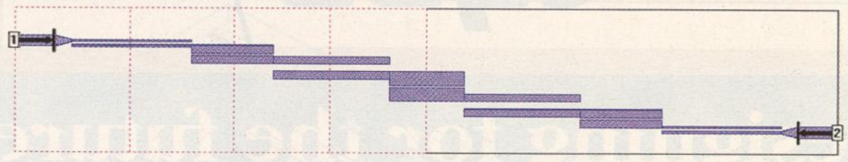

Figure 8 - Full filter layout, with dashed lines showing the filter subdivision

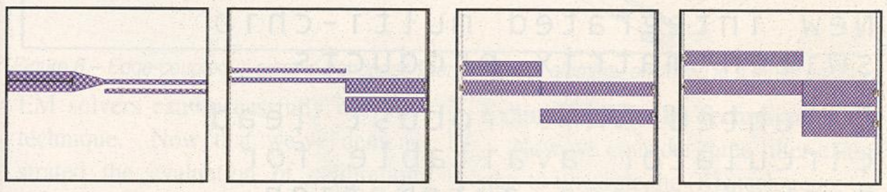

Figure 9(a)-(d): Filter subdivision into 4 unique subcircuits

The second way that we can subdivide the filter (Method 2) involves simulating coupled line sections and the transition discontinuities between these sections separately. Modeled this way, we can subdivide and model the component parts of the filter in **Figure 8** as shown in **Figure 10(a)-(f)**.

Probably the greatest deficiency in modeling this type of filter in a linear circuit analysis tool is the inability of that tool to properly account for non-adjacent resonator coupling; in **Figures 10(c),(e)** and **(g)**, the end effects and the coupling between the non-adjacent open ends is modeled. While this coupling may in many cases be quite small, it is possible for it to modify the overall filter response. **Figure 10(a)** shows the model for the input section, and the reference planes are indicated by the dotted lines at the ends of the dark arrows. The end discontinuity models in **Figures 10(c),(e)** and **(g)** appear to have zero length; in fact the result of these analyses contain an exact model for the complete discontinuity — open ends, cross coupling and the step in the through transmission line.

The coupled line sections that are shown in **Figures 10(b), (d), (f)** and **(h)** are exactly half the actual length of the actual coupled line sections. In reality, any length of this coupled section may be analyzed, as long as the model cascaded with a suitable number of models of itself in the linear circuit simulator add up to the right total coupled line length. We usually use a half-length to make the EM analysis of this section analyze even faster, while keeping the number of sections cascaded to make up the coupled section to a minimum (2).

Figure 11 shows the results of the filter measurements (red), full filter EM analysis (blue), the 4-section filter division method (magenta) and the 8-section filter subdivision method (black).

The different methods of analyzing this filter appear to agree very well with the filter measurements, well into the filter stopband. The real punchline, however, is the savings in time. Using

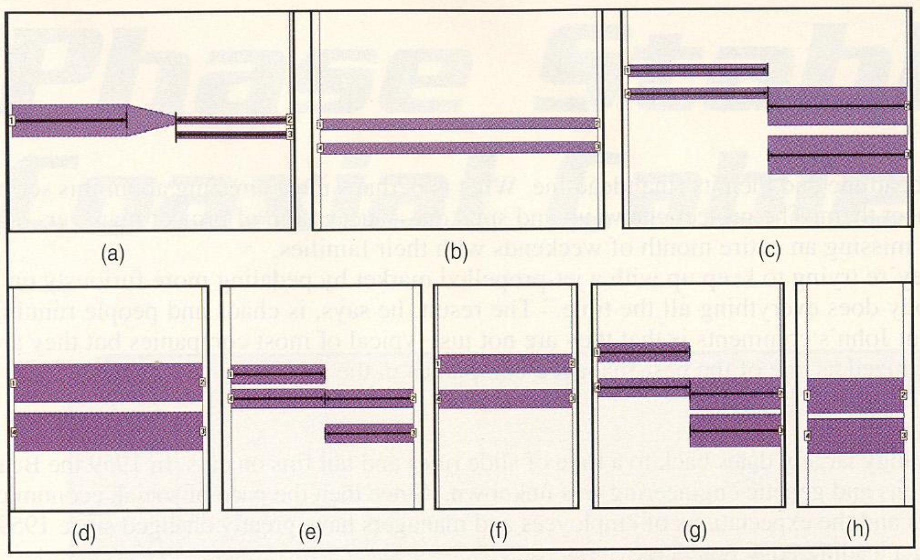

Figure 10(a)-(f) - Method 2 of edge-coupled microstrip filter subdivision

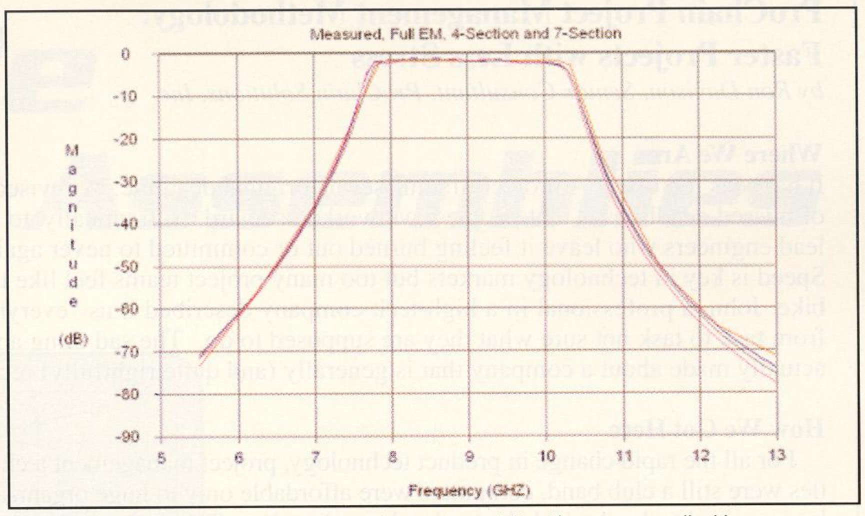

Figure 11 – BI method analysis results (magenta and black) compare well with measurement (red) and full EM analysis (blue)

the Method 1 BI approach, breaking the filter into 4 sections yields excellent results with a 50x factor time reduction, while the Method 2 approach of breaking the filter into 8 sections yields results with a 73x factor time reduction over analyzing the full filter at 80 frequency points. The faster analysis time by Method 2 is offset somewhat with the additional setup time for creating 8 separate analyses instead of 4. The more interpolated points we request from the linear circuit simulator for the overall filter response, the wider this performance improvement grows.

The computation resources for the EM analyses required by the BI technique are small enough for this example that even the free Sonnet Lite toolkit could be used. Sonnet Lite is a free version of the Sonnet EM Suite, and can be downloaded from http://www.sonnetusa.com.

Interdigital Filter

A second class of filter that we'll consider with the BI method is the interdigital bandpass filter. This filter employs multiple parallel-coupled lines as resonators, with vias to ground on alternating ends of each adjacent microstrip line. The filter considered here is shown in **Figure 12** (Page 56).

The filter is fabricated on 28 mil alumina ($\varepsilon_r = 9.9$) in a chamber with a 180 mil cover height. The filter is designed for a passband of 1.2 GHz to 1.3 GHz. Gold conductors are assumed, with appropriate models for ohmic and skin effect losses.

For this filter, the BI method is employed using the division scheme shown by the dashed lines in **Figure 13** (shown on *Page 56*).

The separate EM analysis files for the two parts are shown in **Figures 14(a) and (b)**. The selection of a cut location was made to cut the filter in half or close to half. The exact placement of a cut plane is not critical. The major concern is to cut it far enough from the feed taps (input and output connections) so that the cross-coupling

Continued on page 56

Continued from page 27

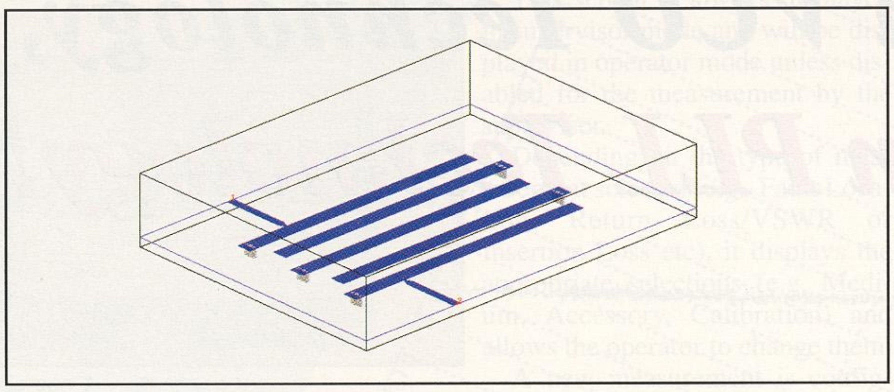

Figure 12 – Interdigital filter implemented with microstrip in a metal housing

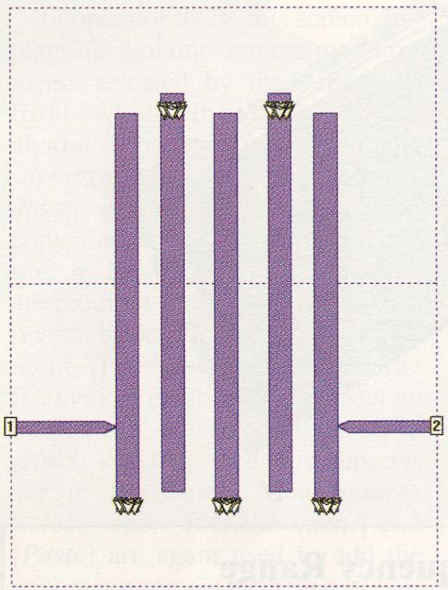

Figure 13 – Interdigital microstrip bandpass filter and subdivisions for BI method EM analysis

from the tap discontinuity to the nearby resonator lines is not removed from the problem. In general, we suggest locating such cut planes at a distance of at least 3 to 5 times the substrate height to be sure to include coupling effects of discontinuities to other nearby circuit entities. Also, for filters incorporating multiple coupled lines, we suggest maintaining the simulation model sidewall distance equal to that of your enclosed design. Sidewall coupling is often significant for this class of filters.

While the EM analysis of each of these parts will probably not be much faster than the analysis of the full filter, we still realize significant analysis time savings because we only use 6 frequency points from the EM analysis of each part. With the circuit recombination in the linear circuit simulator, 100 points are interpolated and computed for the full filter response.

The response for each piece of the filter is simulated from 1.0 to 1.5 GHz in 0.1 GHz steps in the EM solver. As mentioned before, we can direct the Sonnet EM simulator to parse a netlist with references to geometry input (GEO) files. The Sonnet netlist for the recombination of this simple example is shown in **Figure 15**. The netlist node numbers appear in port order for each GEO element. For instance, the node connections for top.geo (which corresponds to **Figure 14(a)** are shown in the GEO file port order. Port 1 of the geometry in top.geo is connected to netlist node 3, port 2 of the geometry to netlist node 4, and so forth. When the

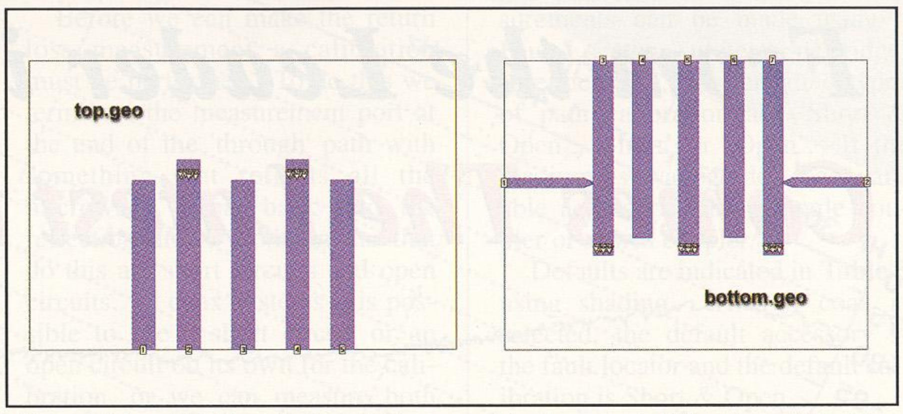

Figure 14(a) and (b), BI method subdivision geometries for separate EM analyses of the interdigital filter parts

```
GEO     3 4 5 6 7     top.geo
GEO   1 2 3 4 5 6 7  bottom.geo

DEF2P 1 2 interdig
```

Figure 15 – Sonnet netlist section for circuit combination

EM simulator encounters a GEO element in the netlist, it determines whether an EM analysis is yet required for the geometry file, executes that analysis, and cascades the data as ordered within the netlist.

The results of a full EM analysis and the BI Method analysis with 6 EM analysis frequencies are presented in **Figure 16**.

The EM analysis for the full filter required 8200 sec for 100 frequencies, while a total of 940 sec was required

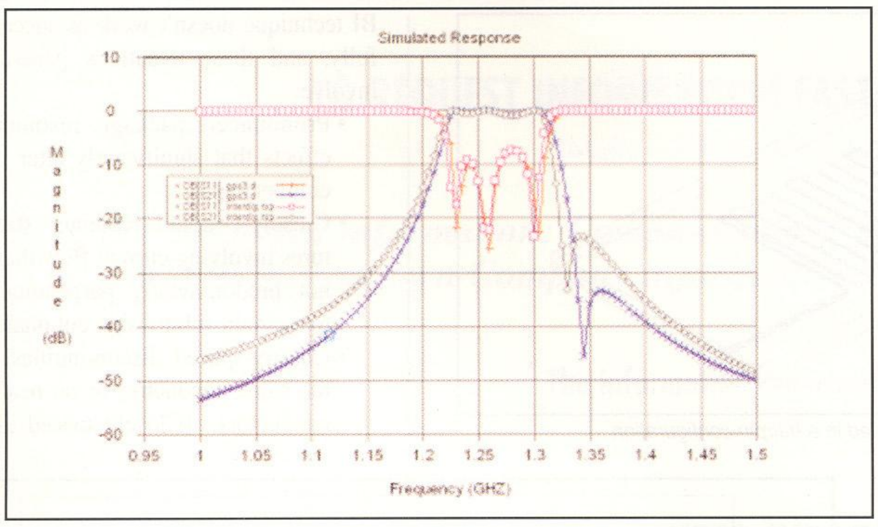

Figure 16 – EM analysis results for the full filter and the filter using the BI Technique

using the BI Method, yielding a total analysis time reduction of 8.7x. These timings were developed using a 300 MHz Pentium II laptop PC. The BI technique provides a good agreement to full filter analysis for the passband response, but the skirts of the filter do experience minor differences. The BI technique analysis (magenta/black curves) show a passband that is shifted about 5 MHz lower than the full filter EM analysis (red/blue curves). This is probably due to the removal of cross-talk between the two ends of the filter with the subdivided model.

An investigation of the current density of the filter provides useful insight into why this technique works for cutting the coupled interdigital sections across the middle. The current density on the filter at 1.25 GHz is presented in **Figure 17**.

The Sonnet emvu current density visualization interface can break this current up into X-directed and Y-directed components, and these are shown in **Figure 18** with same scale as in **Figure 17**.

As we see in the figures, nearly all of the current on the resonator lines is Y-directed, perpendicular to the direction in which we cut the circuit into two parts. Therefore, the significant current propagation on the middle of the resonators is vertical—parallel to the resonators, and crowded to the edges of the conductors. The field lines that comprise the coupling for these long resonator sections are then parallel to our cut planes, and we experience minimal loss in significant coupling when we subdivide the circuit across the coupled lines. Note, however, that there are both X and Y current density components near discontinuities like the taps and the vias. These entities will provide some degree of cross-coupling to other parts of the circuit, and it is best to do your subdividing at a distance of at least 3 to 4 substrate thicknesses back from the nearest discontinuity — sometimes even more for materials with lower dielectric constants (1-4).

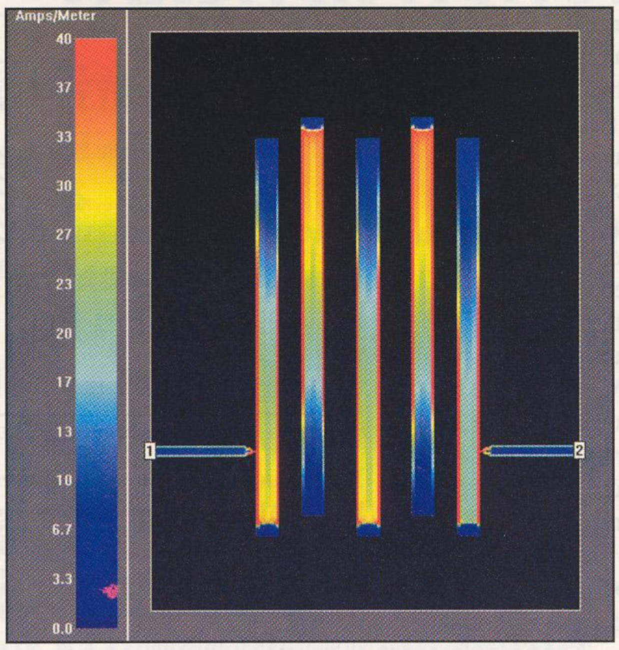

Figure 17 – Computed current density on the interdigital filter at 1.25 GHz

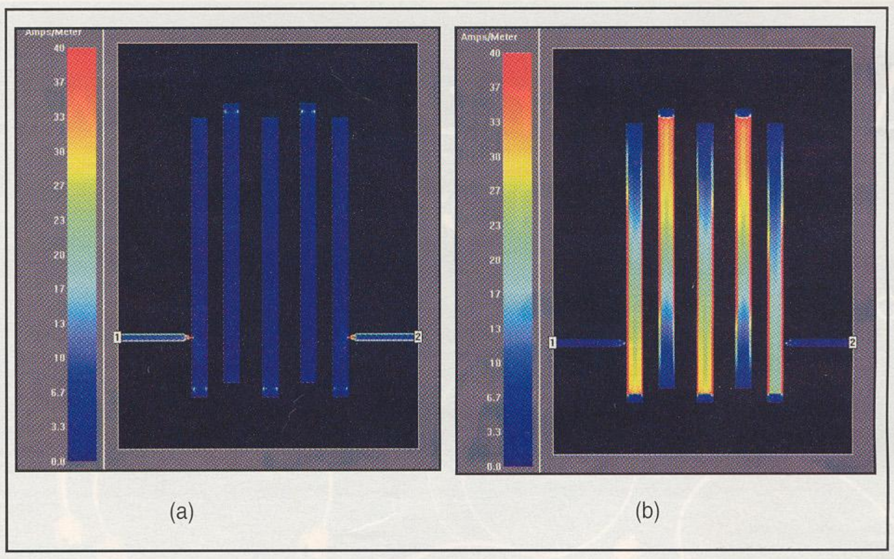

Figure 18 – Current density on the interdigital filter at 1.25 GHz, (a)X-directed current density components, and (b) Y-directed current density components

Hairpin Filter

A third and final example presented here is that of a hairpin filter implemented in stripline. This filter contains 7 resonant sections, and is fabricated between two 20 mil (ε_r = 2.94) substrates, with ground plane wrapped around the outside of the package. The filter is designed for a passband of 4 GHz to 4.125 GHz. Lossless conductors are assumed for this example, but ohmic and skin effect losses can easily be included. The filter is shown in **Figure 19**.

This hairpin filter has the unique property in that it exhibits an odd symmetry about a plane through the middle of the structure, at a right angle to the resonator lines. Figure 20 shows the filter layout, and a cut plane through which a half-filter is created for analysis using the BI technique. Since the filter has this symmetry, we will analyze the top half of the filter as a 15-port circuit, and cascade it with itself in reverse order. The EM analysis file for the BI technique is shown in **Figure 21**.

The EM netlist for this example is a bit more complicated because of the additional ports, and is shown in **Figure 22**. Since we are using exactly half of the filter, we only need to simulate that one half, and cascade it properly with itself.

The results for EM analysis of the full filter in 0.002 GHz steps, and the BI technique analysis are displayed in Figure 23. Here, we see very good agreement between full filter analysis (red, blue) and the BI method analysis (magenta, black) for both input reflection and transmission.

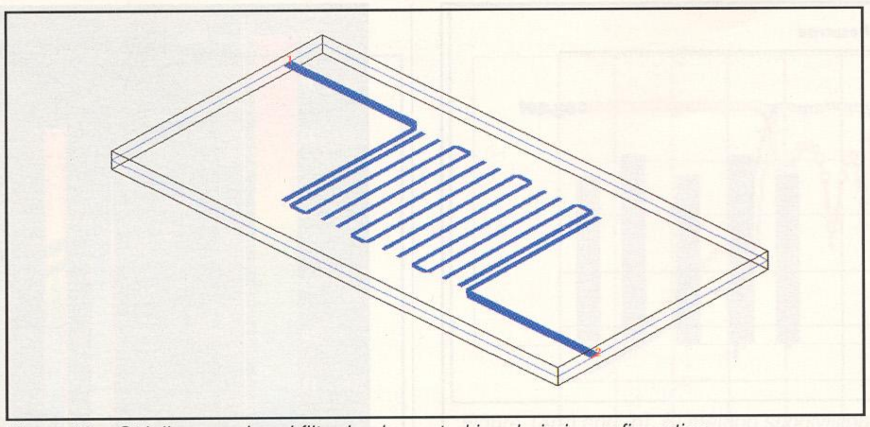

Figure 19 – Stripline passband filter implemented in a hairpin configuration

Caveats

There are some situations where the BI technique doesn't work as successfully, and these situations generally involve:

- Pronounced package resonance effects that significantly alter circuit behavior.
- Closely-coupled resonant structures involving current flow that is not predominantly perpendicular to a circuit subdivision cut plane.
- Closely-spaced discontinuities on the same conductor, or on nearby conductors. (Closely-spaced usu-

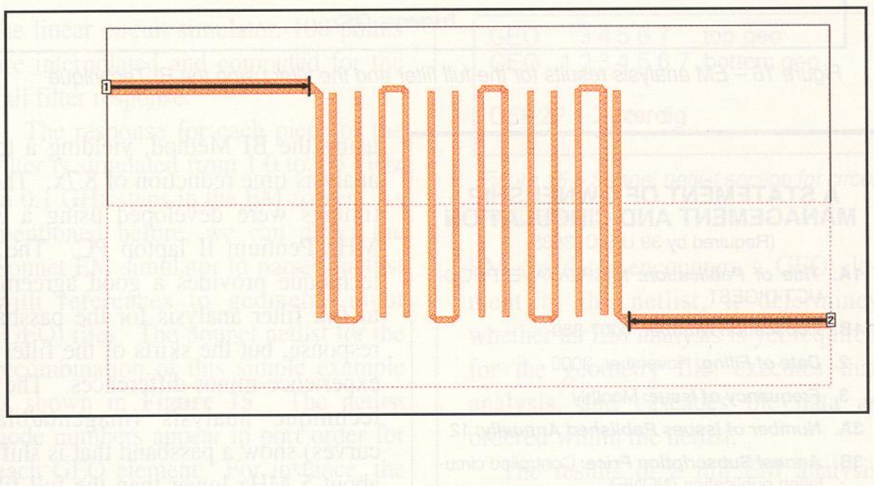

Figure 20 – Hairpin filter geometry showing reference planes and cut plane for creating a BI technique EM analysis file

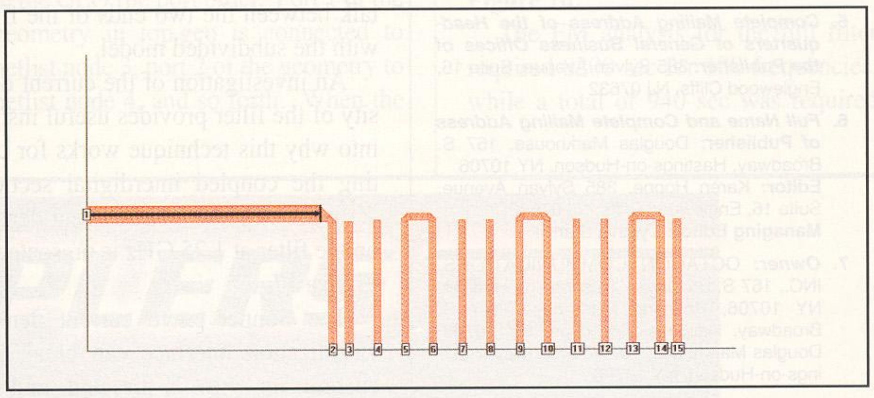

Figure 21 – EM analysis file to be used for the BI technique analysis

Continued on page 61

Continued from page 58

```
GEO  1   2 3 4 5 6 7 8 9 10 11 12 13 14 15 hairpin_half.geo
GEO  16 15 14 13 12 11 10 9 8 7   6 5 4 3 2 hairpin_half.geo

DEF2P 1 16 hairpin
```

Figure 22 – Sonnet netlist for BI method EM analysis of the hairpin filter

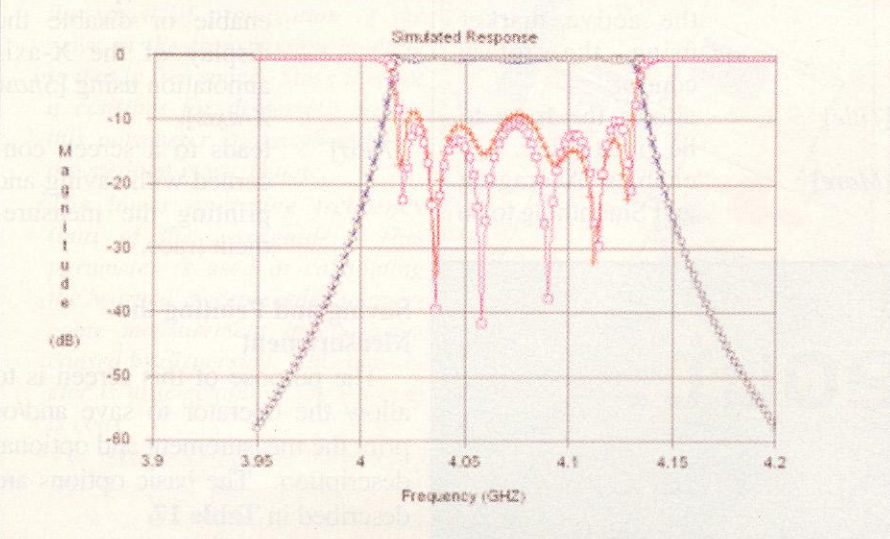

Figure 23 – Comparison between full filter EM analysis (red/blue curves) and BI method EM analysis (magenta/black) results

ally means 2–4 substrate thicknesses) Discontinuities may include impedance steps, vias or notches. BI subdivision should not be done between closely-spaced discontinuities as the higher-order modes that exist on each can cause discontinuity interaction.

The Break and Iterpolate (BI) technique has proven very useful to designers using high frequency EM software, and should be generally applicable to a variety of EM solvers and linear circuit analysis tools for dramatically faster electromagnetic simulation of resonant circuits.

Sonnet Software, Inc. develops, sells and supports high frequency 3D planar EM analysis software, known as the Sonnet EM Suite. Sonnet's planar solver, **em** has been used by customers worldwide since 1989 and is recognized as an industry standard.

Sonnet also provides a full range of high frequency EM analysis solutions, and also offers products to North American customers for full 3D EM analysis.

Technical information may be found on Sonnet's web site: http://www.sonnetusa.com. A free version of the Sonnet EM Suite, called Sonnet Lite is available as a download on the web site.

Example files are available via email. Contact the author by email at: scarp@sonnetusa.com

SONNET SOFTWARE

CIRCLE READER SERVICE NO. 60