Modeling a Printed VHF Balun Leveraging EM Simulation Techniques

INTRODUCTION

The use of printed microstrip couplers in power amplifiers operating at microwave frequencies above 1GHz has been well established and numerous papers have been written about the design of such structures. It is surprising, however, that little has been written about the subject of printed baluns (BALanced to UNbalanced)—electrical devices that convert between a balanced signal and an unbalanced signal. Even more surprising is that the dominant approach to power combining at VHF and UHF is via coaxial cables (Figure 1), which comes with inherent assembly, cost, and thermal issues.

A balanced amplifier is particularly desirable in the design of high-power, high-efficiency amplifiers and is based upon driving two transistors 180° out of phase. This has thermal and 2nd harmonic rejection benefits. The alternative to coaxial baluns, which can overcome the cost and assembly issues using printed planar transmission lines, can be more difficult to design, which may be one reason why it has not been more widely adopted. A key issue has been the lack of adequate linear transmission line models.

A 1.1kW amplifier that uses a planar combiner is shown in Figure 2 (courtesy of NXP). From a high-power amplifier designer point of view, the advantages of a balanced (or push-pull) design are many and significant. They include:

- High efficiency
- Four-fold increase in device impedance (compared to a single-ended circuit) [1]
- Excellent even-harmonic rejection
- Impedance transformation
- Increased reproducibility
- Lower assembly costs

To overcome the difficult-to-design disadvantage referenced above, modern electromagnetic (EM) simulators can be leveraged to accurately analyze such structures. This detailed white paper describes a design approach for printed baluns that achieves design success using AWR's AXIEM® planar EM simulation software within its Microwave Office® design environment.

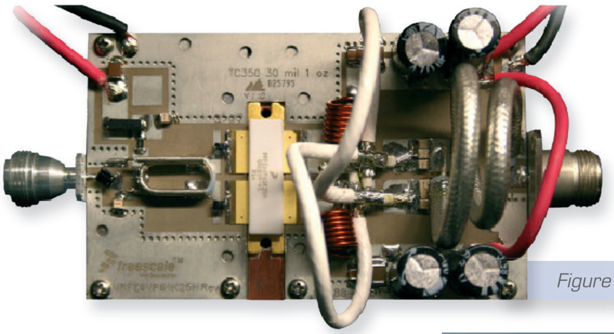


Figure 1: 1100W FM broadcast reference design using coaxial cables, courtesy of Freescale.

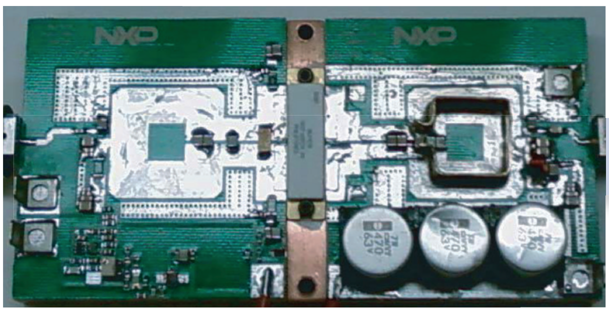


Figure 2: NXP's DVB-T amplifier using planar combiners [2].

PLANAR COUPLER THEORY

A common error is to forget that the planar balun is not dependent on fractions of a wavelength, but rather relies on inductive coupling. The theory of the coupling is complex and one approach to modeling it is to use superposition principles, as has been described in detail by D. Jaisson in "Planar Impedance Transformer" (D. Jaisson, 1999 [3]).

It is beneficial to recall the basics of the ideal transformer (Figure 3) behavior in order to start to formulate a design method for planar couplers. In the ideal transformer, two separate wires are coiled around a transformer, whereby the flow of current through the one wire (the primary) causes electric and magnetic fields, which interact with the other wire (the secondary), to cause a current to flow in this wire.

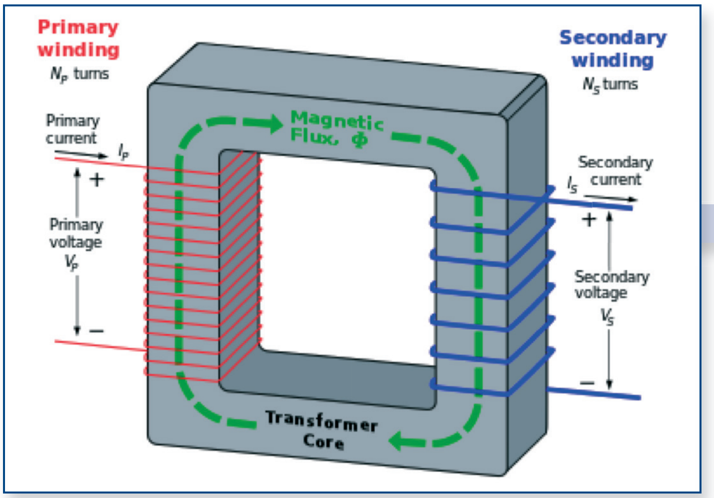


Figure 3: An ideal transformer representation courtesy of Wikipedia.

In the ideal transformer it is assumed that:

- The magnetic flux is the same for both coils (there is no flux leakage)
- Faraday's electro-motive law applies: the voltage induced is proportional to the rate of change of current times the number of turns
- There are no losses, i.e., the source and load power are the same
- The permeability of the ideal transformer is independent of flux density, i.e., it is a linear device

Further, it can be shown (Abrie, 2009 [4]) that the ratio of the source and load impedance is proportional to the square 'of the turns' ratio. This is important as it highlights the second function of the balun—that of an impedance transformer.

In practice however:

- There is flux leakage, which is represented by leakage inductance
- The magnetizing inductance is finite
- There are copper and core (hysteresis and eddy currents) losses
- The relative permeability of magnetic materials does change with DC and RF currents (frequency and temperature)
- There is parasitic capacitance between coil windings

The coupling between the two coils is described in terms of the mutual inductance, M, and this in turn can be used to calculate the coupling coefficient, K, as described in {1}. The relationship between secondary inductance, L_S, and the other key parameters is given in {3} (for a full derivation see [Abrie, 2009 (4)]). The subscript "S" can be confusing, as in the case of Z_S and R_S it refers to source and in L_S it refers to secondary.

$$K = \frac{M}{\sqrt{L_P \times L_S}} \qquad \{1\}$$

$$\frac{Z_S}{Z_L} = \left[\frac{n_P}{n_S}\right]^2 = n^2 \qquad \{2\}$$

$$L_S = \frac{R_L}{R_S} \times \frac{L_P}{K^2} \qquad \{3\}$$

These relationships can likewise be modeled with a linear simulator such as AWR's Microwave Office, as shown in Figure 4.

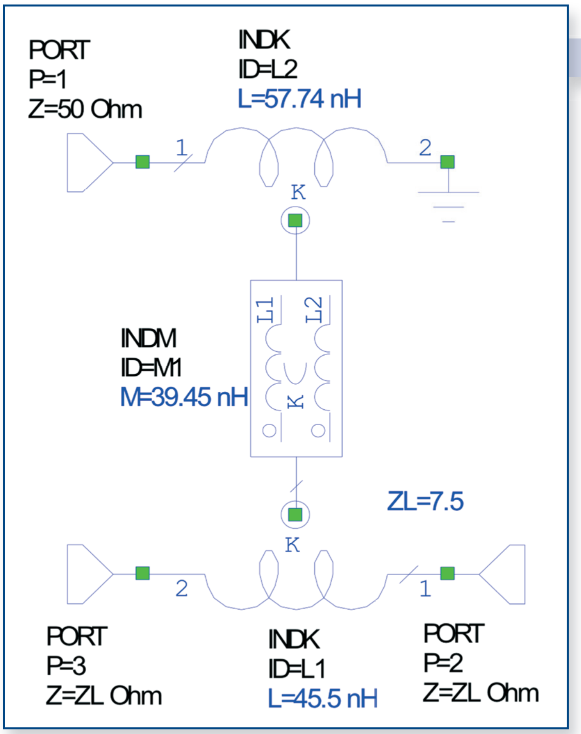
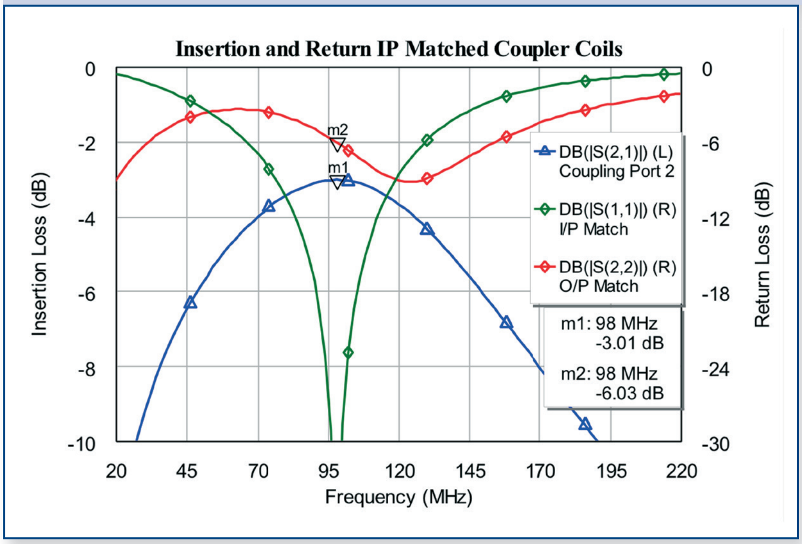


Figure 4: Equivalent circuit model of a mutually-coupled inductor balun.

Figure 5: Narrowband response of capacitively-tuned input.



In order to match the inductance of the input of the transformer, a shunt capacitor is added. This resonates with the inductance of the primary to give a good (if narrowband) match, as shown Figure 5.

The phase difference between the output ports remains a constant 180° in both the matched and unmatched case, even though the actual phase trajectory changes (refer to Figure 6).

The anti-phase performance is intuitive, as the current flowing in the two output ports is in opposing directions. The impact of the input-resonating capacitor can best be seen by looking at the changes to the impedances on a Smith Chart.

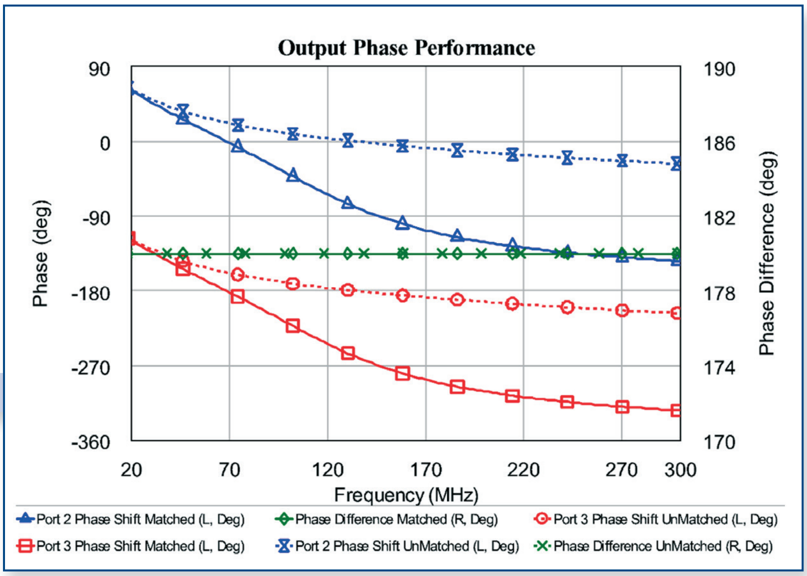


Figure 6: 180° phase difference between balun ports.

Referring to Figure 7, the dashed lines show the impedances of the purely inductive-coupled coils and, for reference, the solid red line along the perimeter of the chart is an inductor of the same value as the primary. The mutual inductance, secondary, and lower-load impedances reduce the effective reactance of the primary coil. The effect of the resonating capacitor matches the primary at the fundamental frequency.

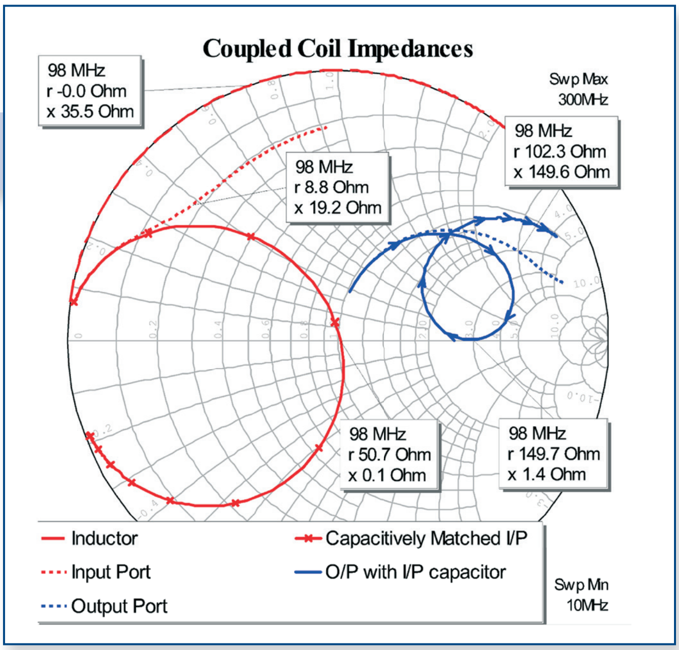


Figure 7: Impedance of purely inductive coupled inductors (dashed) and effect of resonating input capacitor.

The bandwidth can be improved with the addition of capacitance on the output ports. If this is tuned to a slightly different frequency than that at the input, then the effective bandwidth can be extended. The trade-off, however, is a reduction in input match. This second capacitor not only "tightens" the resonance at the output ports, it also introduces an inflection in the input impedance trajectory, which broadens the input match, as depicted in Figure 8. An attractive part of this approach is that the parasitic capacitance of an amplifier's port impedance can be absorbed into this resonating capacitor.

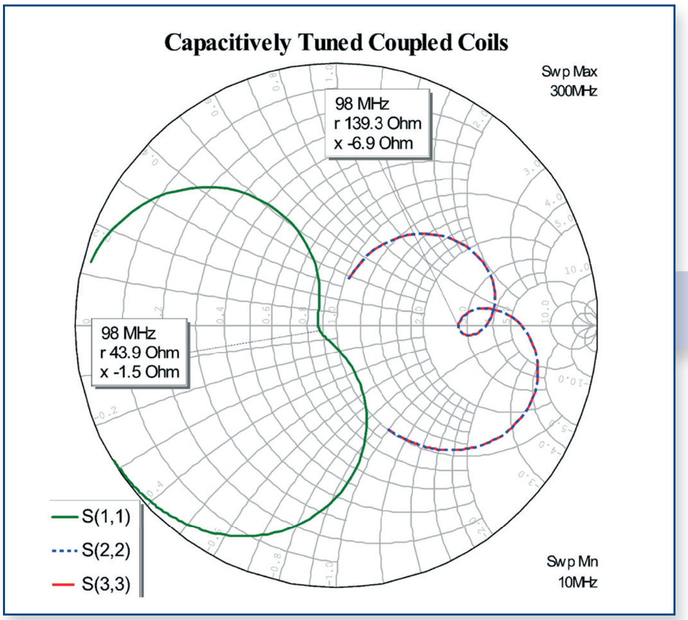


Figure 8: Impedance trajectories following capacitive tuning at input and output.

The next stage in the design process is to attempt to realize these coupled inductances in suspended stripline. The challenge then becomes translating these values into "real" circuit values, i.e., the width and length of the tracks and the impact of the substrate thickness and dielectric. It is not clear what coupling factor, K, will be achieved from the outset, but it can be calculated retrospectively. As a point of reference, the authors (NXP, 2010 [2]) found mathematical approaches impractical and converged upon an iterative approach using 3D EM analysis.

SUSPENDED STRIPLINE IMPLEMENTATION

In its most basic implementation, the printed balun consists of two tracks printed on opposite sides of a printed circuit board (PCB). This is easily and quickly analyzed (Bazdar, Djordjevic, Harrington, & Sarkar, 1994 [5]) in Microwave Office software using the circuit shown in Figure 9.

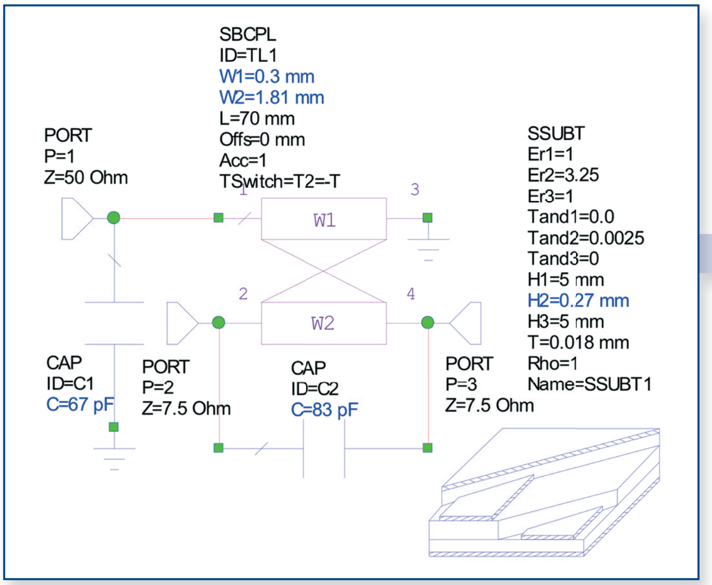


Figure 9: Simple planar balun transformer.

This model provides a quick and useful starting point. It can be then optimized to provide values for the line widths, as well as to assess the impact of different substrate materials and thicknesses. There are, however, limitations of which to be aware. The model assumes that the tracks follow the same paths on opposite sides of the board (although they can be offset) and as there is no accounting for bends, the layout is impractical.

Switching instead to the balun implemented in (NXP, 2010 [2]), it can be seen that the underside track is a spiral, whilst the track on the top is a simple "C" shape, as depicted in Figure 10.

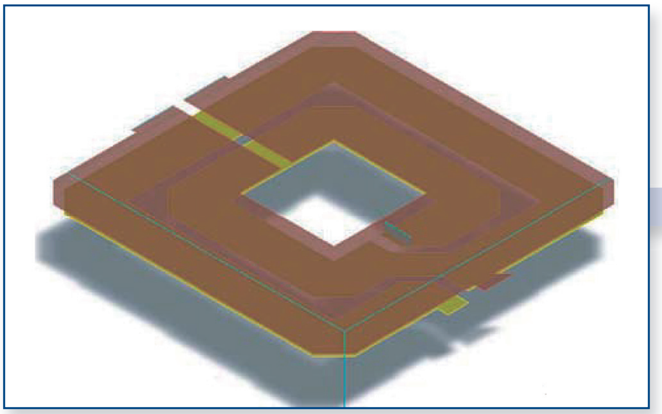


Figure 10: 3D drawing of balun from [2].

In this scenario, the unbalanced signal enters (or leaves) the spiral inductor on the lower outer end and the grounded end is connected through via holes to the 'virtual' ground at the mid-point of the upper trace. The space above and below the balun (typically 5x the substrate thickness) should be air or dielectrically filled. A ground post could also have been brought up into the center. Alternatively, the outer end could have been grounded and the center linked to the unbalanced port (e.g., via a jumper). From the designer's perspective, the conundrum is that for the layout of the structure microstrip/stripline, models are most convenient, whilst the accuracy results from running a full EM simulation of a specifically defined structure.

Fortunately, AWR's AXIEM enables the best of both worlds: geometry can be defined using circuit models and the software uses these to create the shapes for a specific instance of the structure that can be analyzed using the EM simulation engine.

DESIGNING A PLANAR BALUN USING AXIEM

The first stage is to look at the impedance that is to be matched and separate the resistive and reactive components. It is not necessary to achieve the entire match with the balun—it may be more advantageous to do this in stages, as the greater the impedance step, the higher the Q, and the narrower the bandwidth.

For example, with a balun centered at 98MHz and used to input match a very high-power device such as the NXP BLF178, the resistive element is in the order of 2.5Ω and the reactive element –j6.8Ω (equivalent to a 240pF capacitor). Table 1 illustrates the possible balun inductor values.

FOR A Z_S OF 50Ω AND A L_S OF 11NH (TO RESONATE WITH CL OF 240PF) FOR A K=0.8				
Z_L	N	L_P	M	C_P
50	1	7.0	7.0	375.0
25	2	14.1	9.9	187.5
12.5	4	28.1	14.1	93.8
10	5	35.2	15.7	75.0
8.3	6.0	42.4	17.3	62.3
4	12.5	87.9	24.9	30.0
2	25	175.8	35.2	15.0

Table 1: Possible transformer parameters for different impedance ratios.

It is clear that there are a number of options, depending on how great an impedance transform is desired. Note that only a single frequency has been assumed in the calculation and that the wider the bandwidth, the larger will be the variation in K. As a general rule of thumb for wider bandwidths, a lower impedance transformation minimizes ripple. High Z ratios of ~10 can be achieved if loss is not a critical factor. However, for power amplifiers it is unlikely (and unwise) to strive to achieve more than a 5:1 transformation. In this example, therefore, a value for n of 4 might be selected, which still makes the job of matching to 2.5Ω from 12.5Ω challenging, but not nearly as difficult as from 50Ω.

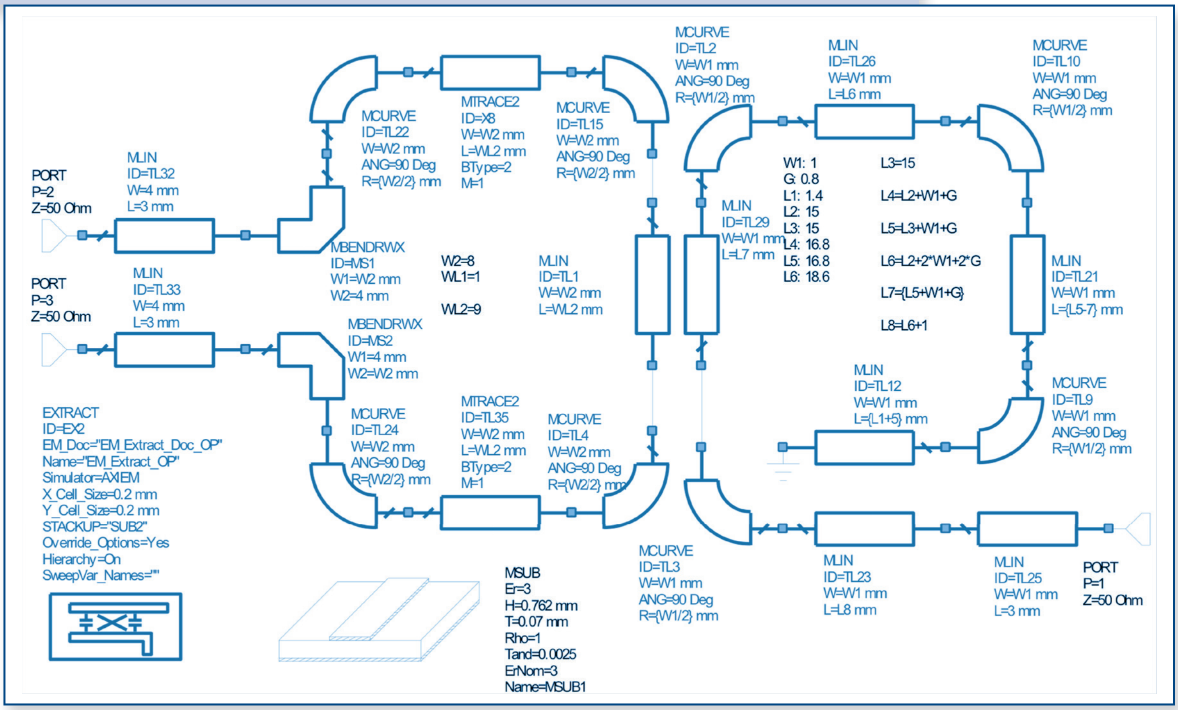
Again, Table 1 provides starting values. These can be plugged into the equivalent circuit model shown in Figure 4, and optimized for the required bandwidth. This then provides the template for optimizing the simple balun of Figure 9. While it is possible to include variables such as substrate thickness and dielectric constant, these may also be impacted by other factors such as line widths for matching, cost, and availability. And while higher thermal conductivity RF substrates (flexible/soft) are becoming available, they are restricted to a limited choice in dielectric constant.

Emerging from the simple model values, a more complicated shape can be created. Again, using the line widths and lengths of the simple circuit as a starting point, the input line can be optimized for a narrow width, while the output line can be wider in order to achieve the inductance ratio with the same length line. By making the two lengths different, more practical line widths can be used.

It is easiest to define the shape and size of the single turn first. That of the spiral is more complicated and the line lengths are best described in relation to one side of a turn taken as a reference. In this way only one parameter need be changed to alter all the others. If the impedance ratio chosen is not too large, a single turn may be sufficient. It is, of course, necessary to consider the current handling of the tracks when deciding on the line width to use. For higher currents, wider lines will need to be longer to achieve the required inductance.

In Figure 11 the conventional microstrip models are used to define the geometry. The shapes that are associated with the EM analysis are defined as an EM extraction block within the AWR environment and added to the schematic.

Figure 11: Balun circuit schematic, with elements associated with EM analysis highlighted in the EXTRACT block (lower left corner of schematic).



The ports, ground connection, and the multiple substrate (MSUB) elements that are not part of the EM analysis are shown in blue in Figure 11. The EXTRACT block contains the EM analysis parameters such as the cell size, simulation engine, and material definitions. The STACKUP parameter refers to an MSUB definition element, which allows different material layers to be used, thus giving the user control of the normal substrate parameters, conductor materials, and air gaps.

Referring again to the Figure 11 schematic, both track patterns have been defined and one of the element parameters selects on which side of the board a particular section is placed. It is essential to ensure that all of sections are joined. In this case a single turn of both primary and secondary are used, and the resulting layout is shown in Figure 12. (Note: capacitors and/or grounding method are excluded.)

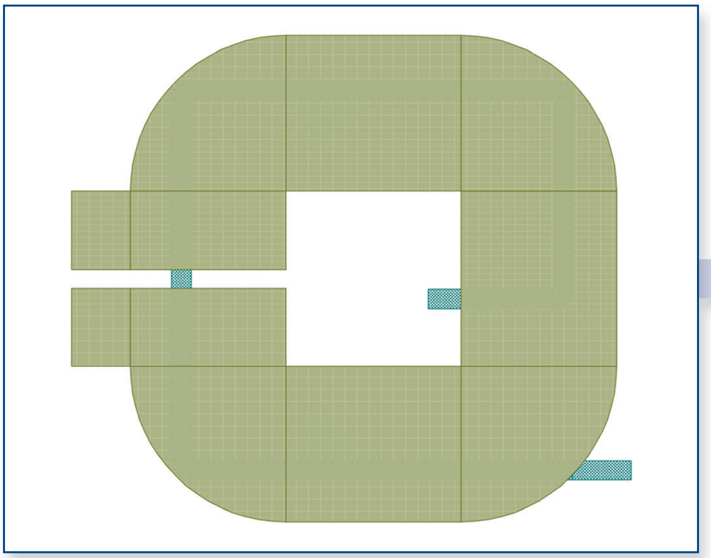


Figure 12: Layout of balun transformer.

The EM section of the circuit is separated out for speed of analysis; when this part of the circuit does not change it is not re-examined (re-analyzed), thus saving time. Instead, it is included as a block and the other circuit elements are included. In this case the resonating capacitors have been added, as shown in Figure 13. (Note: other matching elements could also have been included.)

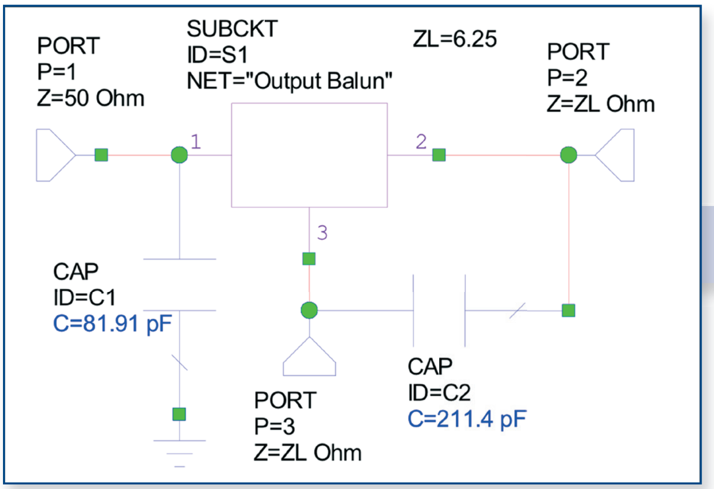


Figure 13: Top-level analysis schematic with EM block contained in a subcircuit.

The results of optimizing the capacitor values for a 3dB split are shown in Figure 14.

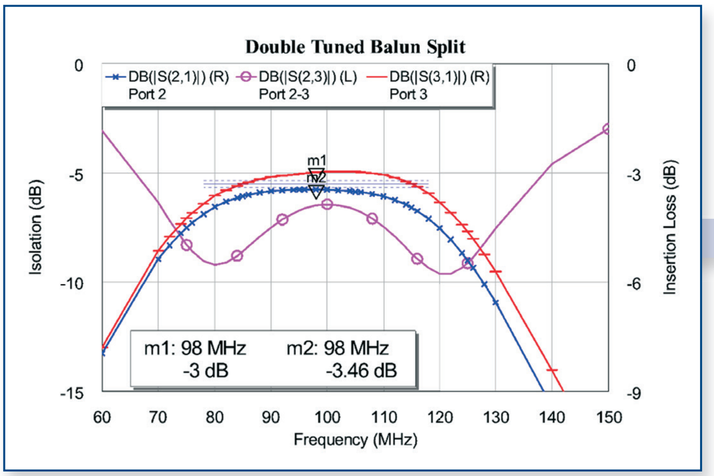


Figure 14: Tuned balun power split.

As can be seen with this construction, there is an imbalance of 0.5dB between the ports. There are a number of techniques that can be employed to correct for this and their effects can be quickly appreciated using simulation. Care must be taken to maintain the 180° phase difference with any correction, as this is the fundamental requirement of the balun. With this in mind, the anti-phase response required can be met and is shown in Figure 15.

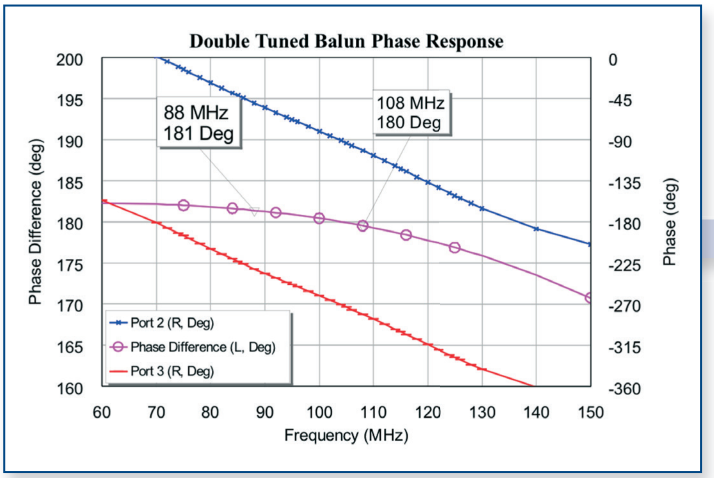


Figure 15: Tuned balun phase response.

Simulation time for this structure is less than one minute on a typical laptop configuration and, therefore, tuning and optimization of the design is practical. Automatic optimization is also feasible but as with most optimization routines, a careful eye must be kept on the design to ensure the final proposed optimized layout maintains realizable parameters. Setting the elements to “auto-snap” and limiting the optimization to varying dimensions by the grid layout size helps to mitigate this potential problem.

Understanding to what level of detail the modeling should be taken is another common concern with matching simulation and measured results. For example, often the box housing is either ignored or only the lid height is considered. In the case of the output balun, the cavity—a rectangular space under the board—has been considered. In practice, also, steps may be included at one end where there is a transition from suspended stripline to microstrip, and care must be given to where the spiral tracks under the board should be grounded, as it may be more desirable to bring up a "post" from the box floor. These can all be modeled quite simply with AXIEM by dividing the air space of the cavity into several layers. The steps can then be added to these layers; hence quite intricate cavity shapes can be modeled and simulated, as shown in Figure 16.

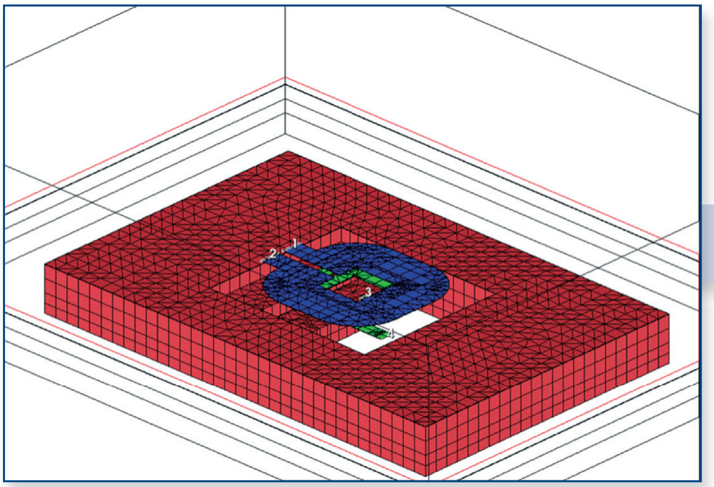


Figure 16: AXIEM 3D mesh layout view of the enhanced model with cavity, including steps and center pillar.

CONCLUSION

An approach has been demonstrated that enhances the understanding of a useful balun construction and enables a balun design that is suitable for volume, high-power, VHF/UHF applications. AWR's AXIEM EM simulation engine has been demonstrated to be very useful in accurately creating suspended stripline layouts, for which conventional linear circuit models do not exist. This approach enables both the EM and layout to be largely automated from standard microstrip circuit elements.

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