Modelling of a Printed VHF Balun Using E-M Simulation Techniques

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Abstract—this paper describes a design approach for printed Baluns. One route to high efficiency amplifiers is to construct a balanced design where two transistors are driven 180° out of phase. Structures that can create these phase shifts are called Baluns, BALanced to UNbalanced. Commonly these are constructed from coaxial cables, often incorporating ferrites; an alternative approach using planar transmission lines has cost and assembly advantages but has suffered from being difficult to design, partly due to a lack of adequate linear models. Modern E-M simulations which can create an E-M simulation from a conventionally described microstrip circuit can be utilised to accurately analyse such structures.

Keywords – Balun, balanced amplifiers, push-pull, VHF power combiners, power dividers, suspended stripline, E-M CAD.

I. INTRODUCTION

At microwave frequencies (>1GHz) the use of printed microstrip couplers has been well established and numerous papers have been written about the design of such structures. Thus it came as something of a surprise when asked to update the design of a VHF power amplifier that little had been written about the subject of printed Baluns. Perhaps even more surprising was that the predominant approach to power combining at these frequencies used coaxial cables with inherent assembly, cost and thermal issues, Figure 2. There is some debate as to which approach is better at power handling; a useful reference on the subject [1] states that "planar structures are not able to handle as much average power as broadband transmission line transformers…". It is not entirely clear to the author why this should be the case, if good thermal design practice is followed and especially with
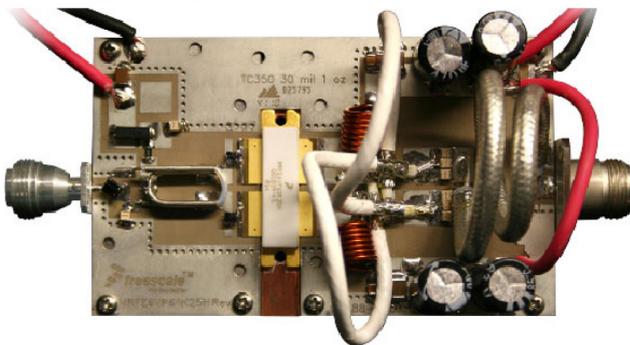

Figure 2, Freescale 1100W FM Broadcast Reference Design

the advent of higher thermal conductivity substrates. Indeed [1] itself demonstrates a planar combiner for a 1.1kW amplifier, Figure 1.

The advantages from a high power amplifier point of view of a balanced (or push-pull) design are significant:

- High efficiency
- 4x increase in device impedance compared to a single ended circuit.
- Excellent even harmonic rejection.
- Impedance transformation.
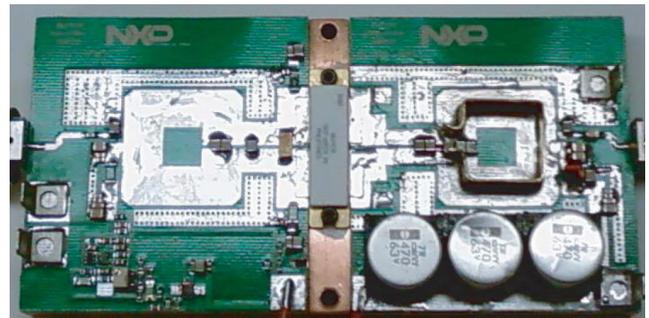- Increased reproducibility and lower assembly costs.

Figure 1, NXP DVB-T amplifier using planar combiners [1].

II. 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 modelling it is to use superposition principles, as has been described in detail [2]. It is beneficial to recall the basics of the ideal transformer behaviour in order to start to formulate a design method for planar couplers. In the ideal transformer two separate wires are coiled around a former 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. In the ideal transformer it is assumed that:

a) The magnetic flux is the same for both coils (there is no flux leakage.

b) Faradays electro-motive law applies: the voltage induced is proportional to the rate of change of current times the number of turns.
c) That there are no losses, i.e. the source and load power are the same.
d) That the permeability of the ideal transformer is independent of flux density, i.e. it is a linear device.

Further it can be shown, [3] 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:
i. There is flux leakage, which is represented by leakage inductance.
ii. The magnetising inductance is finite.
iii. There are copper and core (hysteresis and eddy currents) losses.
iv. The relative permeability of magnetic materials does change with DC and RF currents (and frequency and temperature).
v. There is parasitics 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 calculated the coupling coefficient, K, as described in {1}, and the relationship between secondary inductance L_S and the other key parameters are given in {3}, for a full derivation see [3]. The subscript "S" can be confusing for in the case of Z_S and R_S it refers to 'source' and L_S it refers to secondary. These relationships can be modelled using a linear simulator as shown in Figure 3.

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

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

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

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 narrow band match, as shown in Figure 4. The phase difference between the output ports remains a constant 180° in both the matched and unmatched case although the actual phase trajectory changes, Figure 5. The anti-phase performance is intuitive, the current flowing in the two output ports being in opposite directions. The impact of the input resonating capacitor can probably best be seen by looking at the changes to the impedances on a Smith Chart, Figure 6. The dashed lines shows 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.
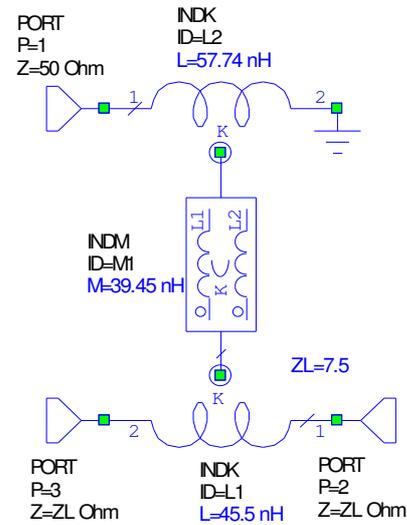

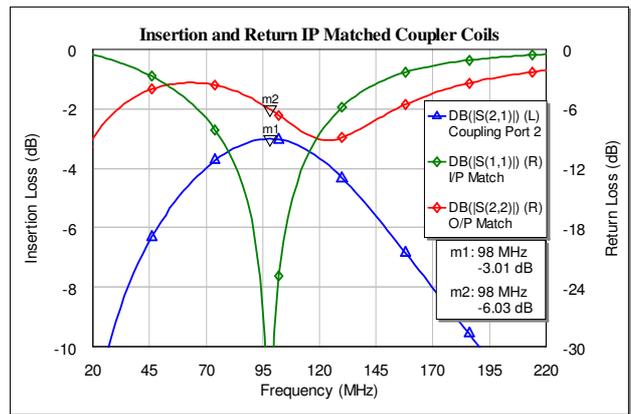Figure 3, Circuit Model of Mutually Coupled Inductor Balun

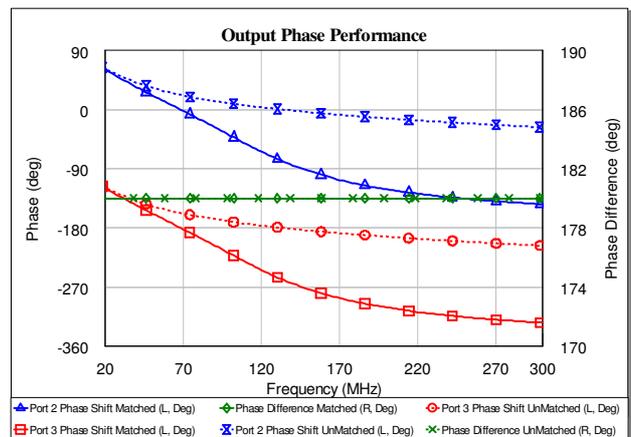Figure 4, Narrow band response of capacitively tuned input.

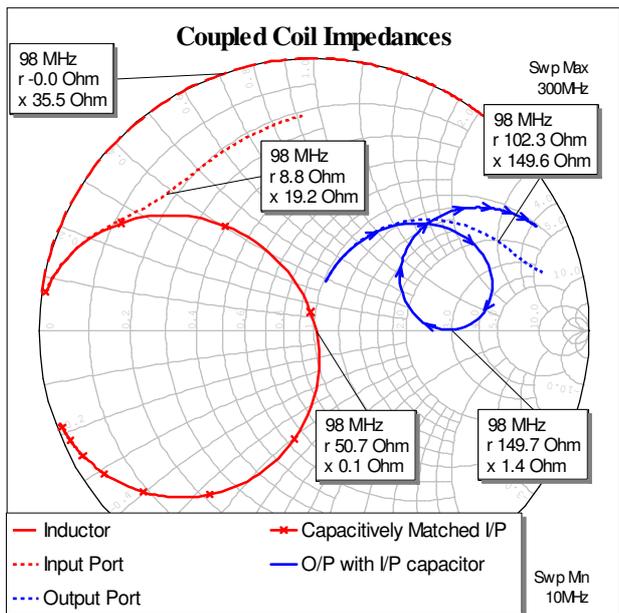Figure 5, 180° phase differential between Balun ports

Coupled Coil Impedances

98 MHz
r -0.0 Ohm
x 35.5 Ohm

Swp Max
300MHz

98 MHz
r 8.8 Ohm
x 19.2 Ohm

98 MHz
r 102.3 Ohm
x 149.6 Ohm

98 MHz
r 50.7 Ohm
x 0.1 Ohm

98 MHz
r 149.7 Ohm
x 1.4 Ohm

Swp Mn
10MHz

— Inductor   —×— Capacitively Matched I/P
···· Input Port   — O/P with I/P capacitor
···· Output Port

Figure 6, impedance of purely inductive coupled inductors (dashed) and effect of resonating input capacitor.

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.

The bandwidth can be improved by the addition of capacitance on the output ports; if this is tuned to a slightly different frequency to that at the input, then the effective bandwidth can be extended. However nothing in this life is free and the cost of this bandwidth is a reduction in input
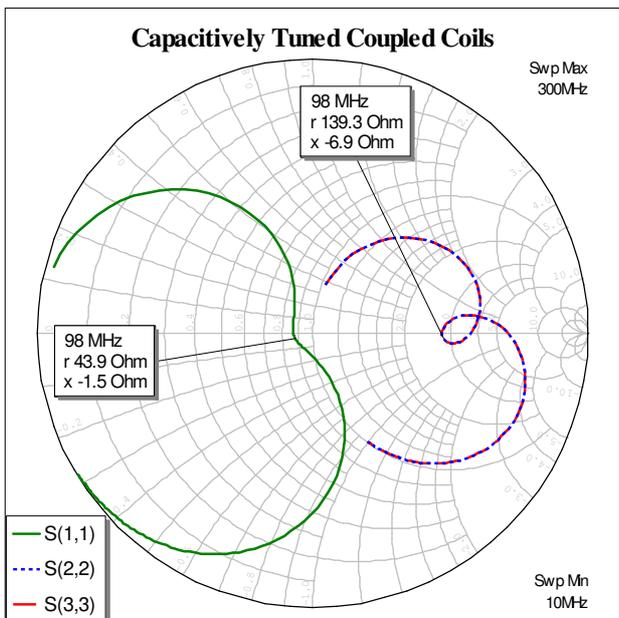
Capacitively Tuned Coupled Coils

Swp Max
300MHz

98 MHz
r 139.3 Ohm
x -6.9 Ohm

98 MHz
r 43.9 Ohm
x -1.5 Ohm

Swp Mn
10MHz

— S(1,1)
···· S(2,2)
— S(3,3)

Figure 7, Impedance trajectories after capacitive tuning at input and output.

match. This second capacitor not only 'tightens' the resonance at the output ports, but also introduces an inflection in the input impedance trajectory, which broadens the input match, Figure 7. An attractive part of this approach is that the parasitic capacitance of an amplifier's port impedance can be absorbed into this resonating capacitor.

The next stage in the design process is to attempt to realise these coupled inductances in suspended stripline. The problem then comes 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, one will get in practice at the outset (it can be calculated retrospectively). The authors of [1] found the mathematical approaches impractical and so settled for an iterative approach using 3D E-M analysis approach.

III.   SUSPENDED STRIPLINE IMPLEMENTATION

In its most basic implementation the printed balun consists of two tracks printed on opposite sides of a printed
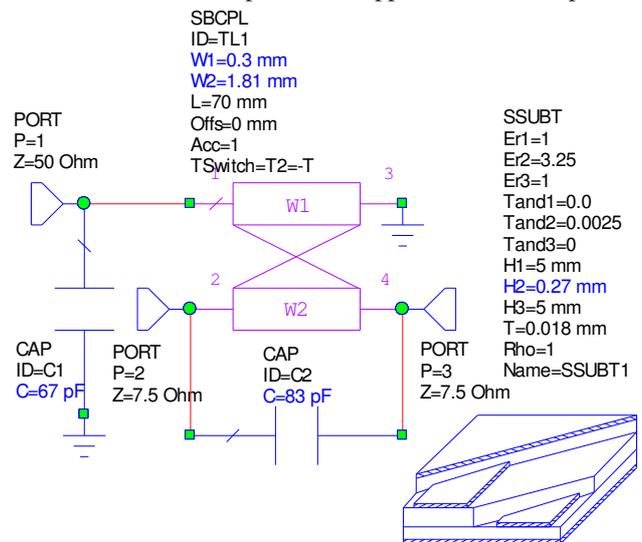
SBCPL
ID=TL1
W1=0.3 mm
W2=1.81 mm
L=70 mm
Offs=0 mm
Acc=1
TSwitch T2=-T

PORT
P=1
Z=50 Ohm

SSUBT
Er1=1
Er2=3.25
Er3=1
Tand1=0.0
Tand2=0.0025
Tand3=0
H1=5 mm
H2=0.27 mm
H3=5 mm
T=0.018 mm
Rho=1
Name=SSUBT1

CAP
ID=C1
C=67 pF

PORT
P=2
Z=7.5 Ohm

CAP
ID=C2
C=83 pF

PORT
P=3
Z=7.5 Ohm

Figure 8, Simple planar balun transformer.

circuit board. This is easily and quickly analysed [4] in MWO using the circuit as shown in Figure 8. This model provides a quick and useful starting point, as it can be quickly optimised and gives values for the line widths as well as giving an assessment of the impact of different substrate materials and thicknesses. However there are drawbacks. The model assumes that the tracks follow the same paths on opposite sides of the board (although they can be offset), and there are no models for bends so the layout is not practical. If we look at the balun implemented in [1] we can see that the underside track is a spiral whilst that on the top is a dimple 'C' shape, Figure 9. In this case the unbalanced signal enters (or leaves) the spiral inductor on the lower out end and the grounded end is connected through via holes to the 'virtual' ground at the mid-point of the upper
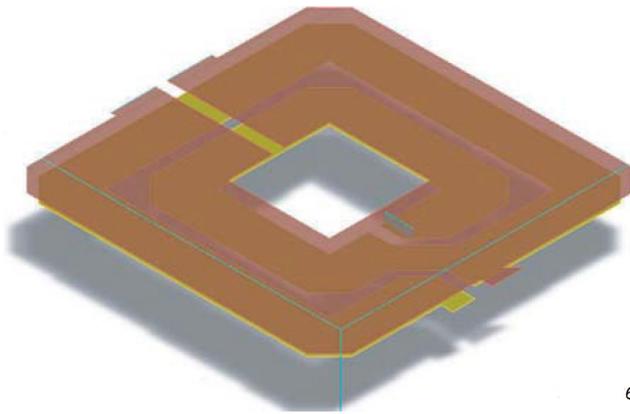
Figure 9, 3-D drawing of balun from [1].

trace. The space above and below the balun typically to >5x the substrate thickness, should be air of dielectrically filled, but a ground post could have been brought up into the centre. Alternatively the outer end could have been grounded and centre linked to the unbalanced port (e.g. via a jumper). From the designers perspective the problem we have is that for the layout of structures the microstrip/stripline models are the most convenient, whilst the accuracy comes from running an E-M simulation of a specifically defined structure.

Fortunately the Axiem simulator from AWR allows the best of both worlds, geometry can be defined using circuit models and the software uses these to create the shapes for a structure that can be analysed using the E-M engine.

IV. DESIGNING A PLANAR BALUN USING AXIEM

The first stage is to use the look at the impedance that you are trying to match to and separate the resistive and reactive components. It is not necessary to achieve the entire match with the balun, it may be advantageous to take it in stages, as the greater the impedance step the higher the Q and the narrower the bandwidth. For example with a balun centred at 98MHz, used to input match a very high power device such as the NXP BLF178, the resistive element is of the order of 2.5Ω and the reactive –j6.8Ω (equivalent to a 240pF capacitor), Table 1 gives the possible balun inductor values.

For a Z_S of 50Ω and a L_S of 11nH (to resonate with C_L 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 that we could take depending on how great an impedance transform we would wish to achieve with the balun. Note that we have only assumed a single frequency in the calculation and that the wider the bandwidth the larger will be the variation in K. As a general rule, for wider bandwidths we would attempt a lower impedance transformation, in order to minimise ripple. High Z ratios ~10 can be achieved is loss is not a critical factor, however particularly in power amplifiers it is unlikely that one would try to achieve more than a 5:1 transformation. In this example therefore we might select a value for n of 4, which still makes the job of matching to 2.5Ω from 12.5Ω challenging, (but not nearly as difficult as from 50Ω).

Table 1, provides starting values, which can be 'plugged into the equivalent circuit model, Figure 3, and then optimised for the required bandwidth. This then provides the template for optimising the simple balun, Figure 8, against. It may be possible to include variables such as substrate thickness and dielectric constant, although these may be determined by other factors, for example the line widths for matching, cost and availability. Higher thermal conductivity RF substrates (flexible/soft) are becoming available, but with limited choice in dielectric constant. Also the load capacitance can be included as a variable, but the next stage of matching will be needed to compensate for any remaining reactance.

From the simple model values, the more complicated shape desired can be created, again using the line widths and lengths of the simple circuit as a starting point. Typically the input line will have been optimised to a narrow width, whilst the output line will be much wider, as the inductance ratio will have been attempted to be achieved with the same length line. By making the two lengths different more practical line widths can be used.

The shape and size of the single turn is easiest to define 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, thus for higher currents wider lines will need to be longer to achieve the required inductance.

Conventional microstrip models are used to define the geometry. The shapes that are associated with the E-M analysis are defined and an E-M extraction block is added to the schematic, as shown in Figure 11. Those elements such as the ports, ground connection and the MSUB not part of the E-M analysis are in blue. The "Extract" block contains the E-M analysis parameters such as the cell size, simulation engine and which block defines the materials used. The "Stackup" parameter refers to a multiple substrate definition element which allows different material layers to be used, thus giving the user control of the normal substrate parameters, conductor materials and also the air gaps.

In the schematic both track patterns have been defined; one of the element parameters select which side of the board a particular section is on. It is essential to ensure that all of
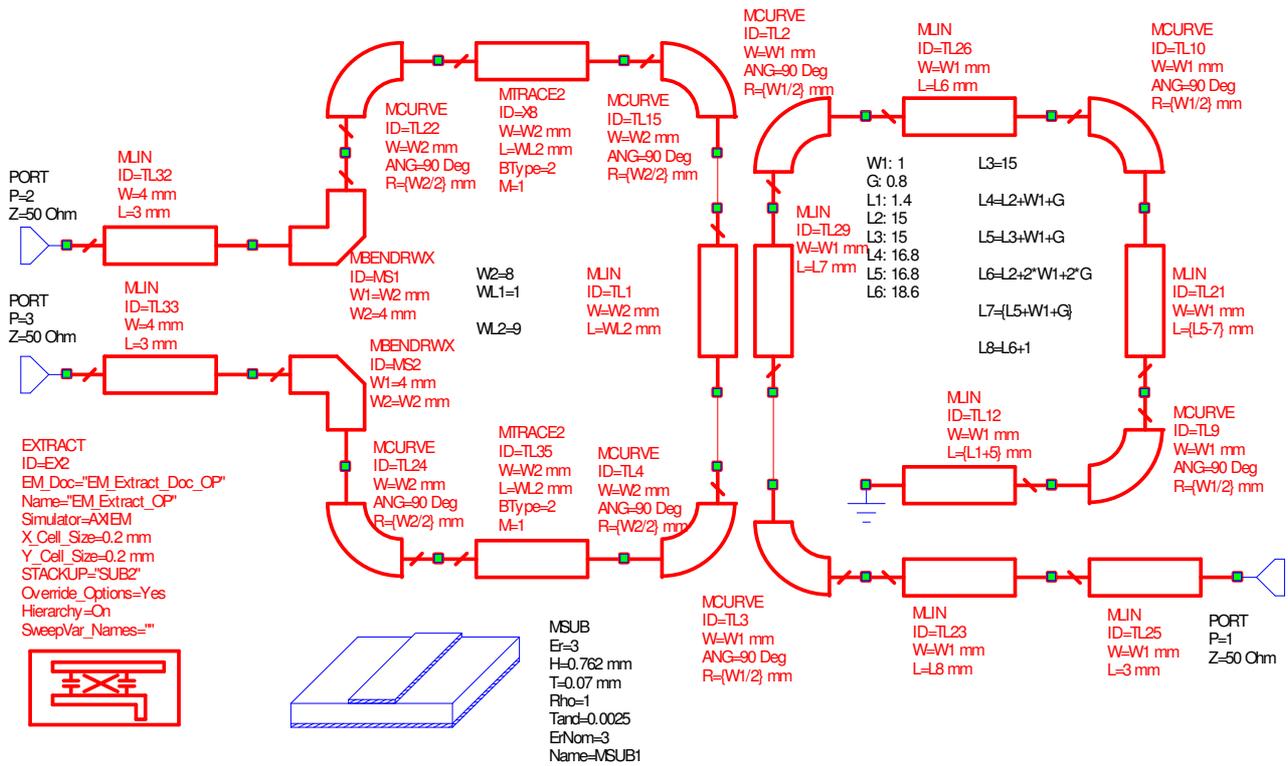
Figure 11, Balun circuit schematic, with elements associated with E-M analysis highlighted.

the sections are joined and for this reason it is useful to make sure that dimensions are multiples of the grid spacing. In this case a single turn of both primary and secondary are used, and the layout is shown in Figure 10. Note that this does not include the capacitors or grounding method, these must be added as in a normal layout.

The E-M section of the circuit is separated out for speed of analysis; when this part of the circuit does not change it does not get re-examined, thus saving time. Instead it is included as a block and the other circuit elements included.
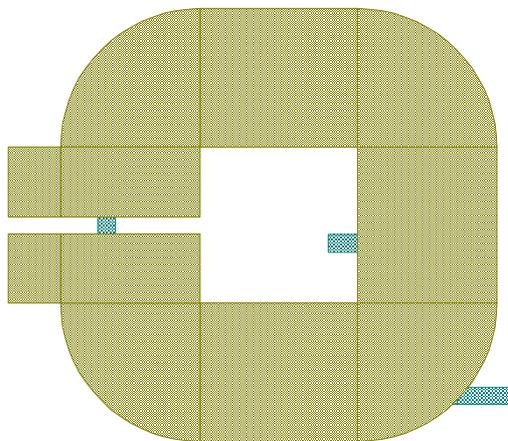
In this case we have just shown the resonating capacitors, Figure 12, but other matching elements could also have been included.

The results of optimising the capacitor values for a 3dB split are shown in Figure 13. 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 this simulation. Care must be taken to maintain the 180° phase difference with any correction, as this is the fundamental requirement of the balun. The anti-phase response required is met as shown in Figure 14.
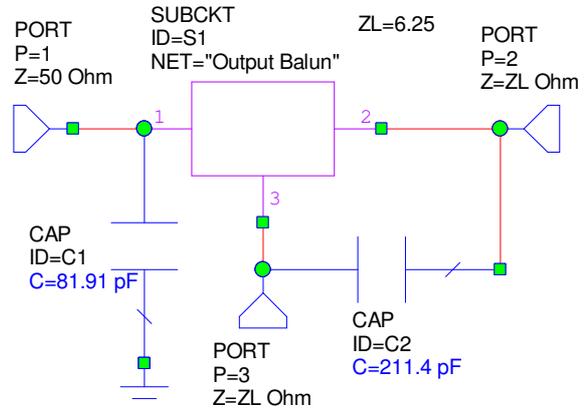

Figure 10, Layout of balun transformer.

Figure 12, Top level analysis schematic with E-M block contained in a sub circuit.

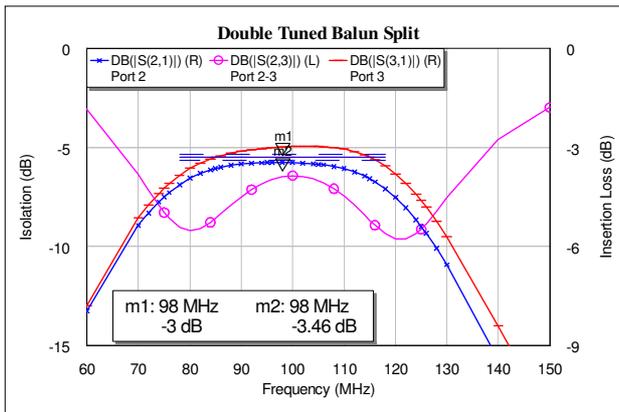Figure 13, Tuned Balun power split.

Figure 14, Tuned Balun Phase Response.

Simulation time for this structure is less than 1 minute on a fairly standard laptop and therefore tuning and optimisation of the design is practical. Automatic optimisation is feasible; however (as with most optimisation routines) a close eye needs to be kept on the design to ensure that the layout stays with realisable parameters. By setting the elements to "auto-snap" and limiting the optimisation to varying dimensions by the grid layout size should restrict this problem.

V. CONCLUSION

An approach has been demonstrated to both enhance the understanding and to enable the design of a useful balun construction that is suitable for volume, high power, VHF-UHF applications. The use of Axiem ™ E-M simulation engine from AWR has been found to be very useful in accurately creating suspended stripline layouts for which conventional linear circuit models do not exist.

VI. ACKNOWLEDGMENT

VII. BIBLIOGRAPHY

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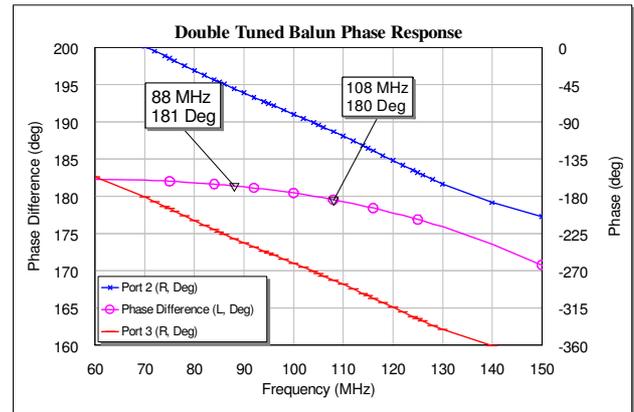[4] M. Bazdar, A. Djordjevic, R. Harrington and T. Sarkar, "Evaluation of quasi-static matrix paramters for multiconductor transmission lines using Galerkin's method," *IEEE Transactions Microwave Theory Techniques,* vol. 42, pp. 1223-1228, July 1994.