CPW to CPS balun design (300 MHz - 1 GHz)

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Study, design and simulation of a coplanar CPW to CPS balun, for frequencies from 300 MHz to 1 GHz, including impedance transformer. Construction and tests.

1 Introduction

The main objective of DS4-T4 design task is to design a tile of antennas for SKA project. In this moment, those researchers in CAY and CIII involved are exploring all potential antenna elements in a band of frequencies from 300 MHz to 1 GHz (or wider if possible), looking for a solution that fulfils to be wide-band, low-noise and low-cost.

Not only the antenna is being studied, but a joint optimization of the antenna, interconnections and the low noise amplifier is required. In this configuration, there are two main possibilities for the LNA:

- *Single-ended LNA*. In this case, it is necessary to use a transition from a balanced structure to an unbalanced one, in order to connect the antenna and the amplifier.
- *Differential LNA*. Balun is not necessary for the structure, and could be possible to reach a much better noise level. Anyway, the balun would be needed once the amplifier and preliminary designs are fabricated, in order to measure and characterize them in laboratory.

Therefore, it's necessary to design a balun component in parallel with antenna element and LNA study, and try to reach a good performance (low noise and good input reflection coefficient).

2 Transmission Lines

Any two-conductor lossless transmission line placed in a homogeneous dielectric medium supports a pure TEM mode of propagation. For example, coaxial line, twin wire line and shielded stripline.

If the transmission line is enclosed in an inhomogeneous dielectric medium, the mode of propagation is pure TEM only in the limit of zero frequency, but if the separation between the conductors is very small compared to the wavelength (quasi-static limit: ?$_g$/20 or smaller), the mode of propagation is quasi-TEM (close to TEM). For example, microstrip line, CPW and slotline.

The characteristic impedance and complex propagation constant of this transmission lines can be described in terms of basic parameters of the line (R, L, C, G):

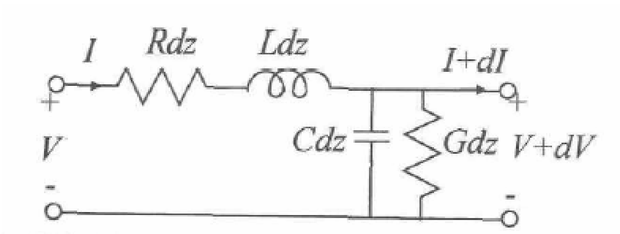


Fig. 1 – Equivalent circuit of TEM or quasi-TEM transmission line of length dz

$$Z_0 = \sqrt{\frac{R + jwL}{G + jwC}}$$ For high conductivity and low loss dielectric: $$Z_0 \approx \sqrt{\frac{L}{C}}$$

$$g = a + jb = \frac{1}{2}\left(\frac{R}{Z_0} + GZ_0\right) + jw\sqrt{LC}$$ Used in $$\begin{aligned} V^{\pm}(z) &= V_0^{\pm} e^{\pm g \cdot z} \\ I^{\pm}(z) &= I_0^{\pm} e^{\pm g \cdot z} \end{aligned}$$

$$\begin{cases} \text{Phase}: b = \dfrac{w}{v_p} = w\sqrt{LC} \ [\text{rad/unitlength}] \Rightarrow v_p = \dfrac{1}{\sqrt{LC}} \\ \text{Attenuation}: a = \dfrac{1}{2}\left(\dfrac{R}{Z_0} + GZ_0\right) [\text{Np/unitlength}] = 4.343\left(\dfrac{R}{Z_0} + GZ_0\right) [\text{dB/unitlength}] \end{cases}$$

So we find a relationship between characteristic impedance and line capacitance; therefore, the characteristic impedance can be determined finding the capacitance of the structure.

The phase velocity can be expressed in terms of the ratio of the actual capacitance of the transmission line to the capacitance it would have in a medium with a dielectric constant of 1.

$$\boxed{Z_0 = \frac{1}{v_p C}}$$

Therefore the attenuation constant is due to the conductor loss $\alpha_C = \dfrac{R}{2Z_0}$ and the dielectric loss $\alpha_d = \dfrac{GZ_0}{2}$.

The Q-factor of a half-wavelength transmission line resonator is:

$$Q = \frac{\beta}{2\alpha} = \frac{\beta}{2(\alpha_c + \alpha_d)} \Rightarrow \frac{1}{Q} = \frac{1}{Q_c} + \frac{1}{Q_d}$$

In dispersive lines, the term β must be replaced by ω / v_g.
This is not valid for non-TEM modes.

2.1 TEM Modes

The transmission line is placed in a homogeneous dielectric medium.

The velocity of propagation v_p is independent of the type of line and frequency, and only depends on the relative permittivity of the dielectric medium.

$$v_p = \frac{c}{\sqrt{\varepsilon_r}}$$

Therefore, the attenuation and phase constant, and the characteristic impedance are:

$$\beta = \frac{\omega}{v_p} = \frac{\omega\sqrt{\varepsilon_r}}{c} = k_0\sqrt{\varepsilon_r}$$ where k_0 is the free-space propagation constant

$$Z_0 = \frac{1}{v_p C} = \frac{\sqrt{\varepsilon_r}}{cC} = \frac{\sqrt{\varepsilon_r}}{c\varepsilon_r C_0} = \frac{1}{c\sqrt{\varepsilon_r}C_0}$$

where C_0 denotes the capacitance between the conductors if they are placed in a medium of a unity dielectric constant

$$\alpha_d = \frac{\beta}{2}\tan\delta = \frac{k_0\sqrt{\varepsilon_r}}{2}\tan\delta \ [\text{Np/unit length}] = 27.3\sqrt{\varepsilon_r}\,\frac{\tan\delta}{\lambda_0} \ [\text{dB/unit length}]$$

Conductor loss factor α_c depends on the type of line, the conductivity of conductors, the frequency and geometrical parameters of the line.

2.2 Quasi-TEM Modes

Effective dielectric constant is defined as $e_{re} = \frac{c^2}{v_p^2}$

In the quasi-static limit it can be assumed to be $e_{re} = \frac{C}{C_0}$

So then:

$$b = \frac{w}{v_p} = \frac{w\sqrt{e_{re}}}{c} = k_0\sqrt{e_{re}}$$

$$Z_0 = \frac{1}{v_p C} = \frac{\sqrt{e_{re}}}{cC} = \frac{\sqrt{e_{re}}}{ce_r C_0} = \frac{1}{c\sqrt{e_{re}}C_0} = \frac{Z_{0a}}{\sqrt{e_{re}}}$$

where Z_{0a} denotes the characteristic impedance of the line if it's placed in a medium of unity dielectric constant

$$a_d = 27.3\frac{e_r}{\sqrt{e_{re}}}\frac{(e_{re}-1)}{(e_r-1)}\frac{\tan d}{l_0} \quad [\text{dB/unit length}]$$

We can define the effective filling fraction as: $q = \frac{(e_{re}-1)}{(e_r-1)}$

Conductor losses of TEM and quasi-TEM mode lines can be determined by the incremental inductance rule, that is valid only if the thickness of conductors of the line and the radius of curvature o its surface are at least 5 or 6 times the skin depth. Then:

$$Q_C = \frac{Z'_{0a}}{(Z'_{0a} - Z_{0a})}$$

where Z_{0a}' is determined reducing the thickness of all conductors by $d_s/2$.

$$a_c = \sqrt{e_{re}}k_0\frac{(Z'_{0a}-Z_{0a})}{2Z'_{0a}} = \frac{p\sqrt{e_{re}}f}{c}\frac{(Z'_{0a}-Z_{0a})}{Z'_{0a}}$$

2.3 Coplanar lines

The term *coplanar lines* is used for those transmission lines where all the conductors are in the same plane; namely, on the top surface of the dielectric substrate.

These transmission lines include slotline, coplanar waveguide (CPW) and coplanar strips (CPS).

In our case, CPW and CPS are used, and the balun structure to be designed is a transformer between these two kinds of transmission lines.

Some of the advantages of these two lines arise from the fact that the mounting of lumped components is much easier, and drilling of holes through the substrate is not needed to reach the ground plane. For this reason, it is ideal for use with surface components.

The performance of coplanar lines is comparable to and sometimes even better than microstrip line in terms of guide wavelength, dispersion, and losses.

Active elements such as amplifiers or antennas can easily be connected to coplanar lines when they are also coplanar in nature. Consequently, coplanar lines are used commonly in MMICs.

Some disadvantages of coplanar lines are parasitic modes, lower power-handling capability, and field non-confinement.

2.4 CPW line

The coplanar waveguide consists of two slots of width *w* printed on a dielectric substrate, with a spacing between them denoted by *s* (Fig. 2).

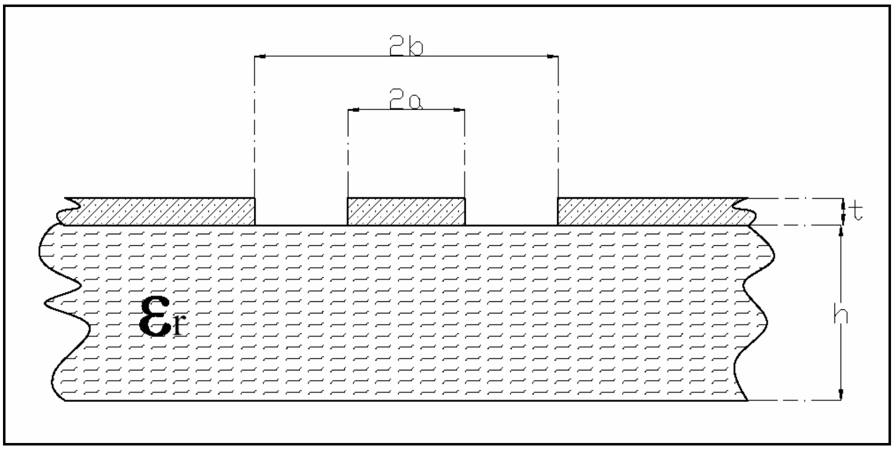


Fig. 2 – Coplanar waveguide (CPW) geometry

The notation for CPW geometry is:

$$a = \frac{s}{2}$$

$$b = \frac{s}{2} + w$$

For the dimensions of the line in this design, we can consider the quasi-static approximation. At higher frequencies, the mode of propagation in the CPW becomes non-TEM because a longitudinal component of the magnetic filed exists.

A quasi-static analysis of these transmission lines can be made based on the conformal mapping method. This method's basic approach is to assume that all the dielectric interfaces in the structure, including slots, can be replaced by magnetic walls. Then, half-planes above and below the metallization plane of the structure can be analyzed separately for line capacitance. The total line capacitance is then the algebraic sum of the two capacitances.

This assumption is strictly valid for those structures for which the electric field lies along the dielectric interfaces. The conformal mapping of coplanar lines gives rise to analytical expressions for the effective dielectric constant e_{re} and characteristic impedance Z_0.

The effective dielectric constant e_{re} is the ratio between the total capacitance per unit length *C*, and the capacitance of the corresponding line with all the dielectrics replaced by air C^a.

$$e_{re} = \frac{C}{C^a}$$

The values of phase velocity v_{ph} and the characteristic impedance Z_0 can be written as:

$$v_{ph} = \frac{c}{\sqrt{e_{re}}} \qquad\qquad Z_0 = \frac{1}{C \cdot v_{ph}} = \frac{1}{c\sqrt{e_{re}}C^a}$$

With the assumption that the dielectric substrate is thick enough to be considered infinite, and metallization thickness is considered negligible, the contribution of the lower half-plane to the line capacitance can be determined as the sum of the free space capacitance (obtained by replacing dielectric by air medium) and the capacitance of the dielectric layer alone assumed to have the permittivity (e_r – 1).

In this case, next expressions can be derived:

$$\boxed{Z_0 = \frac{30p}{\sqrt{e_{re}}} \frac{K'(k_1)}{K(k_1)} \quad [\Omega]} \qquad\qquad \boxed{e_{re} = \frac{e_r + 1}{2}}$$

where

$$k_1 = \frac{a}{b} = \frac{s}{s + 2w} \qquad\qquad k_1' = \sqrt{1 - k_1^2}$$

$$\frac{K(k)}{K'(k)} = \begin{cases} \dfrac{p}{\ln\left[2\frac{1+\sqrt{k'}}{1-\sqrt{k'}}\right]} & \text{if} \quad 0 \le k \le 0.707 \\[3em] \dfrac{\ln\left[2\frac{1+\sqrt{k}}{1-\sqrt{k}}\right]}{p} & \text{if} \quad 0.707 \le k \le 1 \end{cases}$$

The ratio *K/K’* varies from 0 to 8 as *k* varies from 0 to 1.

For practical applications the substrate thickness of a CPW has to be finite. To take into account the effects of the finite thickness of the dielectric substrate, a conformal transformation has been applied to the geometry. This transformation is

$$t = \sinh\left(\frac{p \cdot z}{2h}\right)$$

and the resulting configuration is a coplanar line with infinitely thick substrate.

In this case, the free-space capacitance for the lower substrate half-plane is exactly the same as that for the upper half-plane.

We obtain the following expressions:

$$\boxed{Z_0 = \frac{30p}{\sqrt{e_{re}}} \frac{K'(k_1)}{K(k_1)} \quad [\Omega]} \qquad \boxed{e_{re} = 1 + \frac{e_r - 1}{2} \frac{K(k_2)}{K'(k_2)} \frac{K'(k_1)}{K(k_1)}}$$

where k_2 is a factor that introduces dependence with substrate thickness:

$$k_2 = \frac{t_1}{t_2} = \frac{\sinh\left(\frac{p \cdot a}{2h}\right)}{\sinh\left(\frac{p \cdot b}{2h}\right)}$$

If also the lateral extent of the ground planes of a CPW is considered finite and denoted by c_0, the free-space capacitance for the upper-half plane is determined by transforming the first quadrant of the CPW into the upper half of the transformed plane by using the mapping function $t = z^2$. Via this transformation, ground planes with infinite extension result.

And the capacitance of the dielectric layer is computed by transforming the dielectric region into the upper half of a transformed given by using the mapping function

$$t = \cosh^2\left(\frac{p \cdot z}{2h}\right).$$

After the analysis, these are the expressions for the characteristic impedance and the dielectric constant:

$$\boxed{Z_0 = \frac{30p}{\sqrt{e_{re}}} \frac{K'(k_3)}{K(k_3)} \quad [\Omega]} \qquad \boxed{e_{re} = 1 + \frac{e_r - 1}{2} \frac{K(k_4)}{K'(k_4)} \frac{K'(k_3)}{K(k_3)}}$$

The factor k_3 introduces dependence with the ground planes width, so does k_4, that also introduces dependence with substrate thickness:

$$k_3 = \frac{a}{b}\sqrt{\frac{1 - b^2 / c_0^2}{1 - a^2 / c_0^2}}$$

$$k_4 = \frac{\sinh\left(p \cdot a / 2h\right)}{\sinh\left(p \cdot b / 2h\right)}\sqrt{\frac{1 - \dfrac{\sinh^2\left(p \cdot b / 2h\right)}{\sinh^2\left(p \cdot c_0 / 2h\right)}}{1 - \dfrac{\sinh^2\left(p \cdot a / 2h\right)}{\sinh^2\left(p \cdot c_0 / 2h\right)}}}$$

The values obtained so far are valid for infinitesimally thin metallic strip conductor plane. But in practice, the metallization has a finite thickness *t* that affects the characteristics; an increase in metallization thickness is accompanied by a corresponding decrease in e_{re} and Z_0. The decrease in e_{re} is larger for the case of a substrate with higher dielectric constant and lower aspect ratio *t/w*. On the contrary, the decrease in Z_0 is smaller for the higher dielectric constant substrate. e_{re} and Z_0 increase can be explained from the observation that a higher metallization thickness gives rise to additional concentration of the electric field between the metalized portions of the slots, thus increasing the value of C^a, specially for narrow slots.

The dielectric constant can be calculated by adding a term due to metal thickness to the expression obtained earlier:

$$e_{re}^{t} = e_{re} - \frac{0.7(e_{re} - 1) \cdot t / w}{\left[K(k) / K'(k)\right] + 0.7t / w}$$

The effect of metallization thickness on the characteristics of coplanar lines can be taken into account empirically by defining effective values of strip width and slot width:

$$\begin{cases} s_e = s + \Delta \\ w_e = w - \Delta \end{cases}$$

where

$$\Delta = (1.25t / p)\left[1 + \ln(4ps / t)\right]$$

Effective values of geometric factors can be defined as:

$$\begin{cases} a_e = \dfrac{s_e}{2} \\ b_e = \dfrac{s_e}{2} + w_e \end{cases}$$

From them, a new effective factor k_e can be derived, that takes into account metallization thickness:

$$k_e = \frac{a_e}{b_e} = \frac{s_e}{s_e + 2w_e}$$

And the characteristic impedance is found as

$$Z_0 = \frac{30p}{\sqrt{e_{re}^t}} \frac{K'(k_e)}{K(k_e)}$$

2.5 CPS line

Coplanar strips configuration is complementary to the CPW, as shown in Fig. 3.

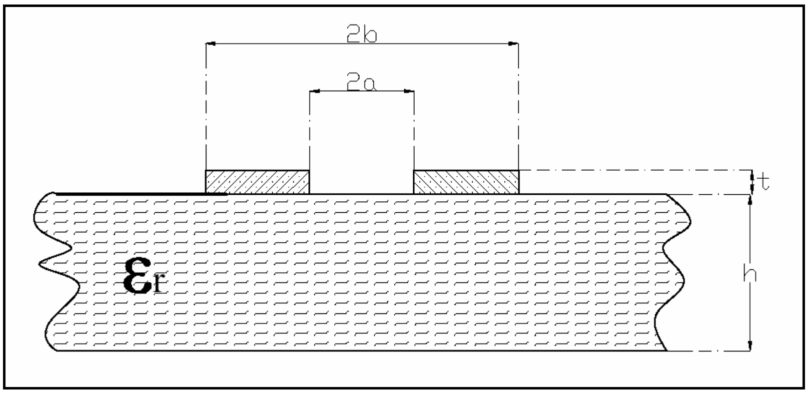


Fig. 3 – Coplanar Strips (CPS) geometry

It consists of two strips of equal width w on a dielectric substrate, with spacing between them which is denoted by s.

The notation for CPS geometry is:

$$a = \frac{s}{2}$$

$$b = \frac{s}{2} + w$$

The electric and magnetic field configurations are shown in Fig. 4.

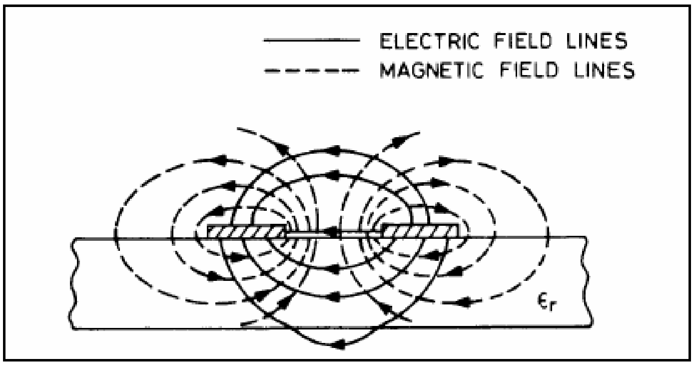


Fig. 4 –Coplanar strips (CPS) electric and magnetic field distributions

It is easier to realize high impedances with CPS than CPW, and further, it is a balanced transmission line, so it’s useful in balanced circuits (such as dipole antennas).

CPS can also be analyzed using the conformal mapping method, and as it’s complementary to CPW in the metallization plane, the transformation used for CPW can also be applied to the CPS. Just the electric and magnetic walls of the parallel plate geometry derived earlier will get interchanged.

If substrate thickness is considered infinite, the total line capacitance is sum of capacitance C_1 contributed by the electric field in the air and C_2 by the electric field in the dielectric region. Therefore, an effective dielectric constant can be derived:

$$e_{re} = \frac{e_r + 1}{2}$$

And the characteristic impedance is

$$Z_0 = \frac{120p}{\sqrt{e_{re}}} \frac{K(k_1)}{K'(k_1)}$$

where

$$k_1 = \frac{a}{b} = \frac{s}{s + 2w}$$

When substrate thickness is considered finite, conformal transformation analysis is needed. Due to the complementary natures of CPS and CPW, the same expression can be used for the effective dielectric constant:

$$\boxed{e_{re} = 1 + \frac{e_r - 1}{2} \frac{K(k_2)}{K'(k_2)} \frac{K'(k_1)}{K(k_1)}}$$

where

$$k_2 = \frac{t_1}{t_2} = \frac{\sinh\left(\frac{p \cdot a}{2h}\right)}{\sinh\left(\frac{p \cdot b}{2h}\right)}$$

And the expression for the characteristic impedance of a CPS is the same, but with the new e_{re} calculated:

$$Z_0 = \frac{120p}{\sqrt{e_{re}}} \frac{K(k_1)}{K'(k_1)}$$

For coplanar strips, the effect of strip thickness is similar to that in CPW. The dielectric constant can be calculated by adding a term due to metal thickness to the expression obtained earlier:

$$e_{re}^{t} = e_{re} - \frac{1.4(e_{re} - 1) \cdot t / s}{\left[K'(k)/K(k)\right] + 1.4t / s} \qquad \begin{cases} e_{re} \geq 9 \\ t / w < 0.1 \end{cases}$$

Effective values of strip width and slot width can be defined:

$$\begin{cases} s_e = s - \Delta \\ w_e = w + \Delta \end{cases}$$

where

$$\Delta = (1.25t / p)\left[1 + \ln(4p \cdot w / t)\right]$$

Effective values of geometric factors can be defined as:

$$\begin{cases} a_e = \frac{s_e}{2} \\ b_e = \frac{s_e}{2} + w_e \end{cases}$$

From them, a new effective factor k_e can be derived, that takes into account metallization thickness:

$$k_e = \frac{a_e}{b_e} = \frac{s_e}{s_e + 2w_e}$$

And the characteristic impedance is found as

$$Z_0 = \frac{120p}{\sqrt{e_{re}^{t}}} \frac{K(k_e)}{K'(k_e)}$$

3 Balun

It's a transformer used to connect balanced transmission line circuits (two conductors with equal potential with 180 degrees phase difference, no current flows through a grounded shield) to unbalanced transmission line (a current flows through the ground).

The planar transmission line baluns consist of two sections:

- the first section divides the signal into two signals having equal magnitude and phase
- the second section provides -90 degrees and +90 degrees phase shift for these signals

3.1 CPW to CPS transitions

These structures are employed to transform electromagnetic energy between two different types of transmission lines; this requires impedance match and also field match. We are looking for a coplanar waveguide (CPW) to coplanar stripline (CPS) transition that maximize the bandwidth and minimize the insertion loss.

CPW and CPS lines are chosen because of its small dispersion, small sensitivity to substrate thickness, simple realization and easy integration, as they don't need via holes.

The structure that has been studied is:

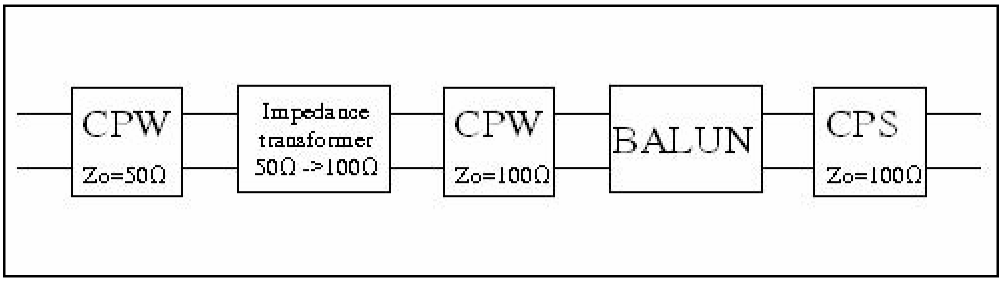


Fig. 5 – Structure been studied

First block is a coplanar waveguide, with 50O as characteristic impedance; this line will carry the signal to feed the antenna, or that it receives.

Then an impedance adapter is used to transform a CPW line with Z_0 = 50 O to a CPW line with Z_0 = 100O.

After this, a balun structure is used to transform an unbalanced line (CPW) to a balanced line (CPS).

Finally, a coplanar stripline with characteristic impedance of 100O will be used to feed the antenna.

Dimensions designed for these lines are:

	CPW2	CPW1	CPS

Strip width	s = 2.6 mm	s = 1.2 mm	w = 2.5 mm
Gap width	w = 0.2 mm	w = 0.9 mm	s = 0.25 mm
Calculated Z_0	49.95 O	104.44 O	98.17 O
Simulated Z_0	51 O	100 O	98 O

3.1.1 Impedance transformer

The impedance transformer used is Chebyshev multi-section matching transformer. For this structure, some possibilities of number and length of sections are possible, and have been studied.

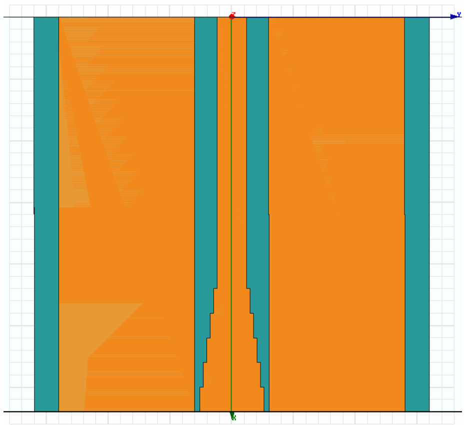


Fig. 6 – Chebyshev 4 sections impedance transformer (50O to 100O)

When simulate this structure, S_{11} and S_{21} parameters present maximums and minimums in frequency. They occur at those frequencies which corresponding wavelength is such that any multiple of its quarter is the same as the length of the final CPW line, but with influence of length of some sections of the Chebyshev transformer.

$$\left.\begin{array}{l} length = K\dfrac{l}{4} \\ l = \dfrac{c}{f} = \dfrac{c_0}{f \cdot \sqrt{e_{re}}} \end{array}\right\} \Rightarrow f = \frac{c_0}{l \cdot \sqrt{e_{re}}} = K \cdot \frac{c_0}{4 \cdot length \cdot \sqrt{e_{re}}}$$

The substrate chosen for simulations is Taconic TLX, with relative permittivity of $e_r = 2.55$. Effective relative permittivity of CPW lines of the structure (including those that are part of the impedance transformer) is $e_{re} = 1.5$.

Results of simulations that have been performed show how the bandwidth is influenced by the length of the sections that form the transformer. As observed in Fig. 7, peak occurs in lower frequency as the length of sections increases.

Even if maximum value reached in all cases is the same, in order to keep good values at low frequencies, it's better to use sections as short as possible. Consequently, peak will occur at higher frequencies, and S_{11} values will be lower in low frequencies band.

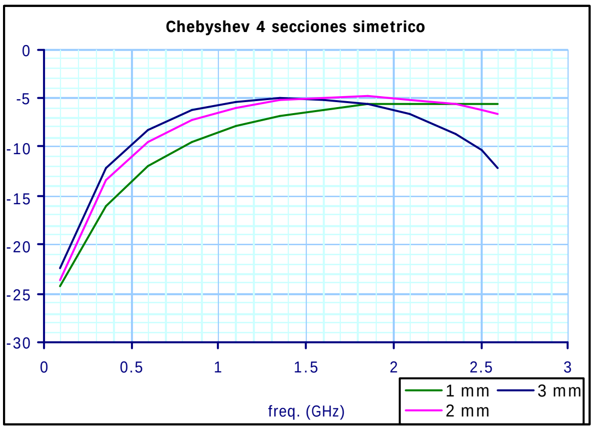


Fig. 7 – S_{11} parameter for back2back 4 sections Chebyshev transformers, connected by a CPW line 10 mm long

We can observe the influence of the length of the line that connects both transformers in back2back configuration in Fig. 8. For longer lines, the peak appears at lower frequencies; consequently, it's necessary to use a short CPS line to connect balun and antenna, for obtaining a good bandwidth.

Obviously, maximum losses value is worse for long lines. Using short lines, maximum losses are lower and at higher frequencies, so up to 1 GHz better values will be obtained.

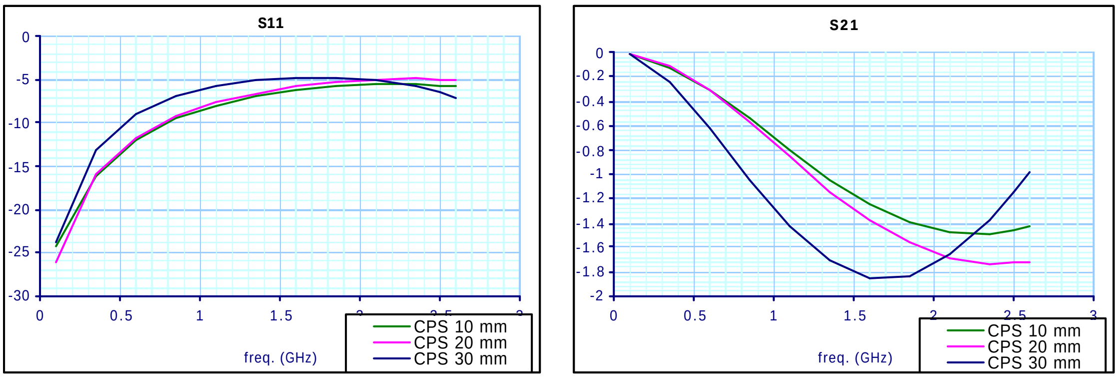


Fig. 8 – Parameters of 4 sections Chebyshev transformers, in back2back configuration, depending on the length of the CPW line connecting them

Our conversion is for CPW lines, from 50O to 100O, so dimensions mentioned above are used:

	CPW2 (50O)	**CPW1 (100O)**
Strip width	s = 2.6 mm	s = 1.2 mm
Gap width	w = 0.2 mm	w = 0.9 mm

In this case, the best choice is a Chebyshev transformer with 6 sections 1 mm long, with a step distance of 0.1 mm between them. Therefore, next table shows the dimensions and impedances for every section:

6 sections Chebyshev transformer $Z_L/Z_{IN} = 2$ *? = 0.05*	CPW							
	Z_{IN}	section 1 Z_1	section 2 Z_2	section 3 Z_3	section 4 Z_4	section 5 Z_5	section 6 Z_6	Z_L
designed Z_0 [O]	50 O	60 O	67.8 O	75 O	82 O	89.17 O	96.62 O	104.65 O
simulated Z_0 [O]	55.5 O	60 O	65.5 O	73.5 O	81 O	88 O	64 O	102 O
CPW Physical parameters:								
Gap width w [mm]	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9
Strip width s [mm]	2.6	2.4	2.2	2	1.8	1.6	1.4	1.2

Ground strips width is 6 mm in all cases; they are recommended to be 2.5 times wider than central strip. Maximum strip width is 2.6 mm at the input of the structure, so ground strips width should be at least 6.5 mm; anyway, 6 mm is enough, as it doesn't have a big influence in results.

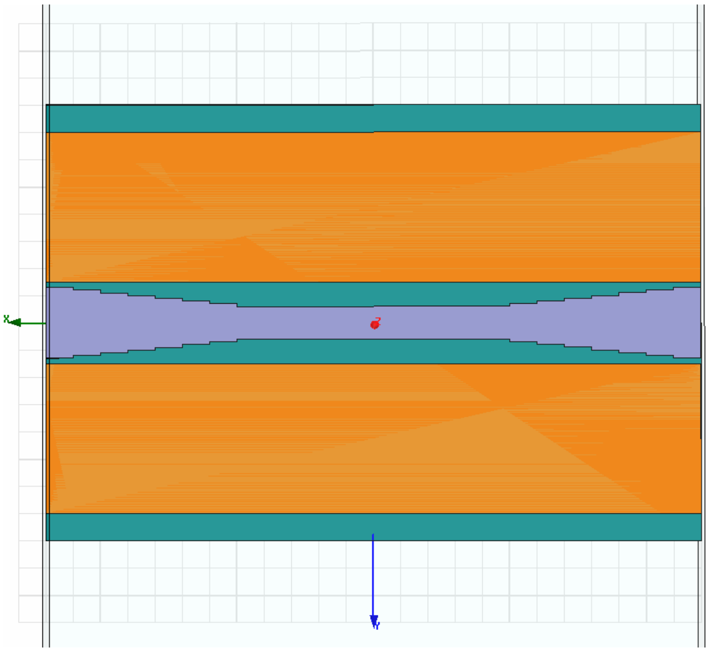


Fig. 9 – Back to back configuration. Two Chebyshev transformers connected by a CPW line

These are the resulting S parameters obtained for this structure, in back-to-back configuration with a 20 mm length line between them:

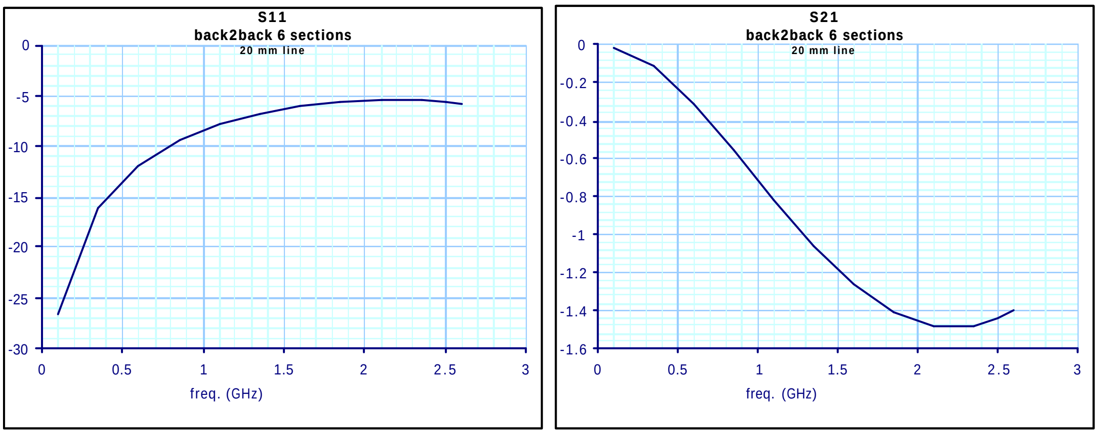


Fig. 10 - S parameters for 6 sections, 1 mm long each, Chebyshev transformer, in back2back configuration and 20 mm long CPW line between

As observed in Fig. 10, up to 1 GHz, reflection coefficient is better than 8 dB, and losses are 0.7 dB (0.35 dB every Chebyshev structure, as this is back to back configuration).

If we consider specifications for wideband S_{11}<-10 dB and S_{21}<0.2 dB, designed transformer can be used up to 800 MHz.

Results are better if both transformers are connected by a shorter CPW line. In Fig. 11 a comparison is shown between a 10 mm and a 20 mm long CPW line connection. Using a 10 mm long CPW line, reflection coefficient is better than -10 dB and insertion loss are less than 0.2 dB (0.4 for back to back configuration) for frequencies up to 1 GHz.

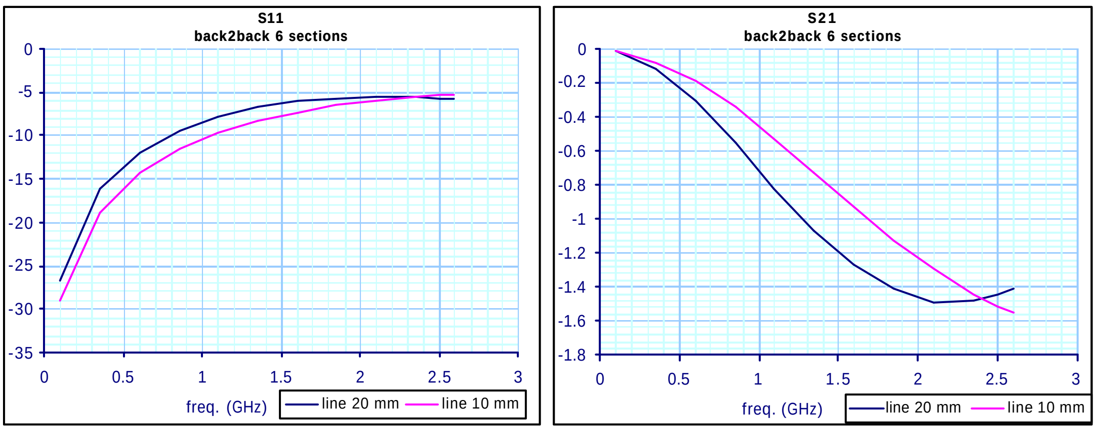


Fig. 11 – Comparison between results obtained connecting Chebyshev tranformers with 10 mm and 20 mm long CPW lines

3.1.2 Balun

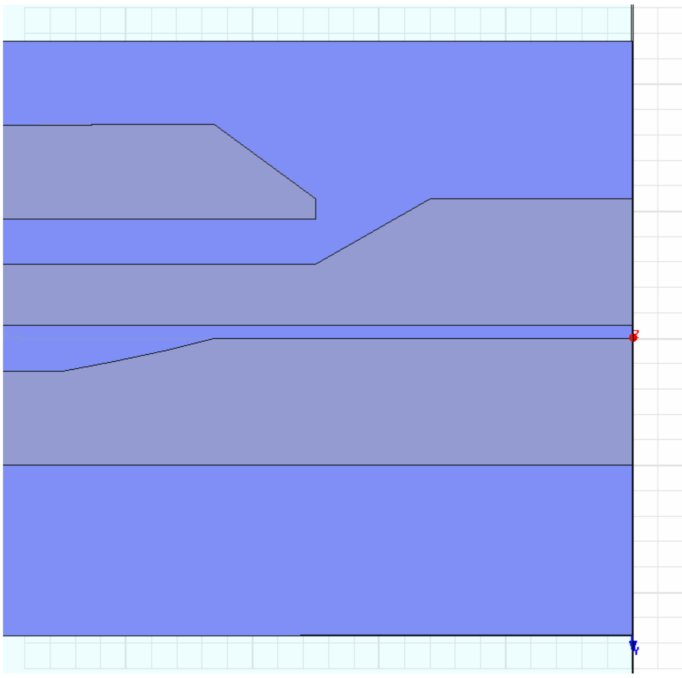


Fig. 12 – Balun. Wideband CPW to CPS transition

The structure studied [1] is a wideband transition (Fig. 12), that can be decomposed into 6 parts (Fig. 13) for theoretical modelling:

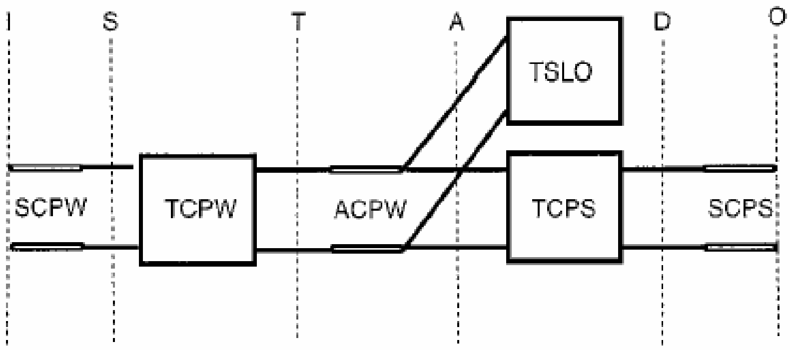


Fig. 13 – Equivalent-circuit model

- SCPW – Symmetric coplanar waveguide
- TCPW – Coplanar waveguide tapered linearly in the lower slot
- ACPW – Asymmetric coplanar waveguide tapered linearly in the upper ground strip
- TSLO – Unterminated slotline open
- TCPS – Asymetric coplanar stripline tapered linearly in the upper strip
- SCPS – Symmetric coplanar stripline

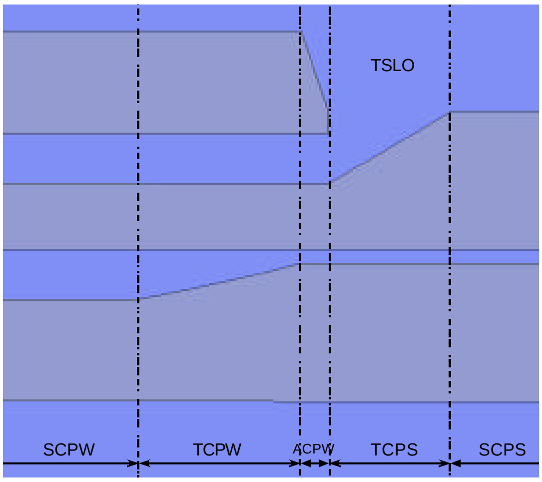
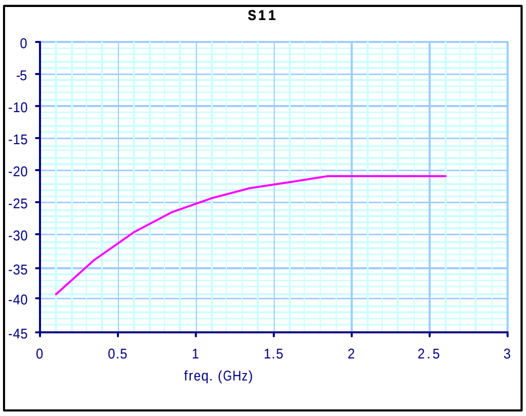


Fig. 14 – Relation between balun structure and parts of equivalent-circuit model

Fig. 15 represents S parameters of back to back configuration for balun structure. Reflection coefficient obtained is better than -20 dB for all frequencies, and losses are less than 0.3 dB (0.15 each balun). In this case, baluns are connected by an 8mm long CPS line.



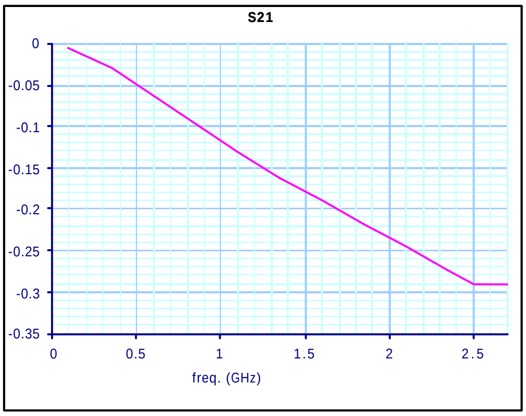


Fig. 15 – Simulated S parameters of back2back balun structure

Some geometric parameters were varied to study their influence in the structure response, and with the objective to improve its performance.

Adding bond wires to the model at the discontinuities, non-CPW modes are suppressed. It results in a wider band of frequencies as for both insertion losses and reflection coefficient, as can be observed in Fig. 17. Three bond wires are used, located at the interface between SCPW and TCPW, at the interface between TCPW and ACPW, and at the end of ACPW.

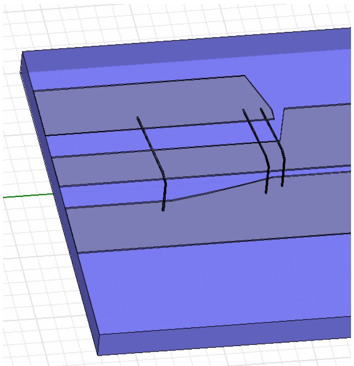


Fig. 16 – Bond wires location at balun structure

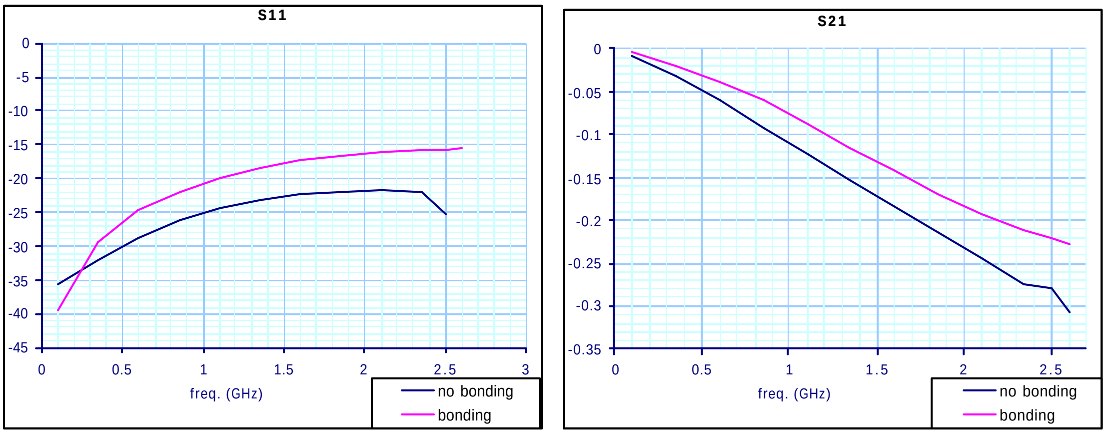


Fig. 17 – Comparison of simulated S parameters of back to back balun structure with and without bonding.

The circuit performance is degraded by the discontinuity mismatch in transition configurations. To reduce this mismatch effect, the inclined angle ? associated with the tapered structures TCPS and TSLO is varied to study his influence in the whole circuit response. Fig. 18 shows this effect; narrower angles provide better $|S_{21}|$ and $|S_{11}|$ values. Therefore, a 15º angle seems to be the most suitable for the structure.

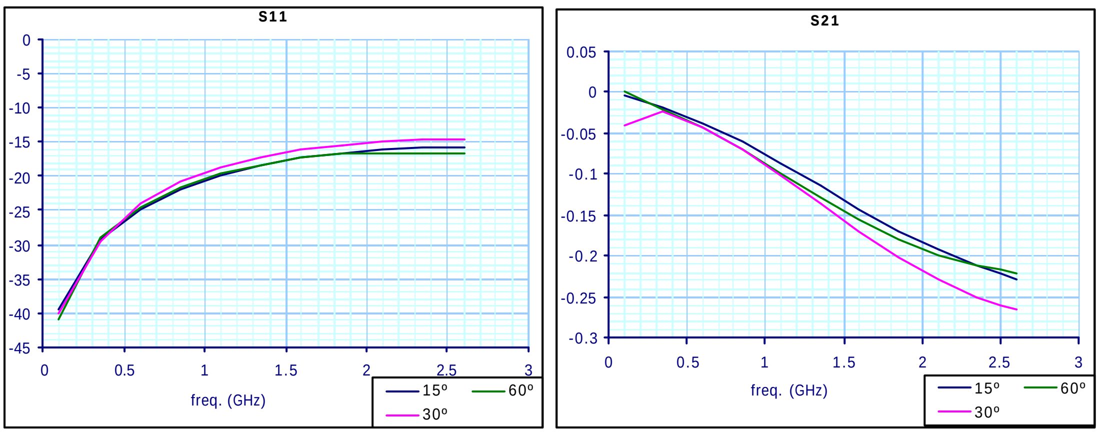


Fig. 18 – Simulated S parameters of back to back balun with inclined angle ? as parameter

Another parameter to be studied is the length of ACPW section; Fig. 19 shows that shorter lengths provide better resulting values of $|S_{21}|$ and $|S_{11}|$. A value of 0.5 mm is chosen for the balun being designed.

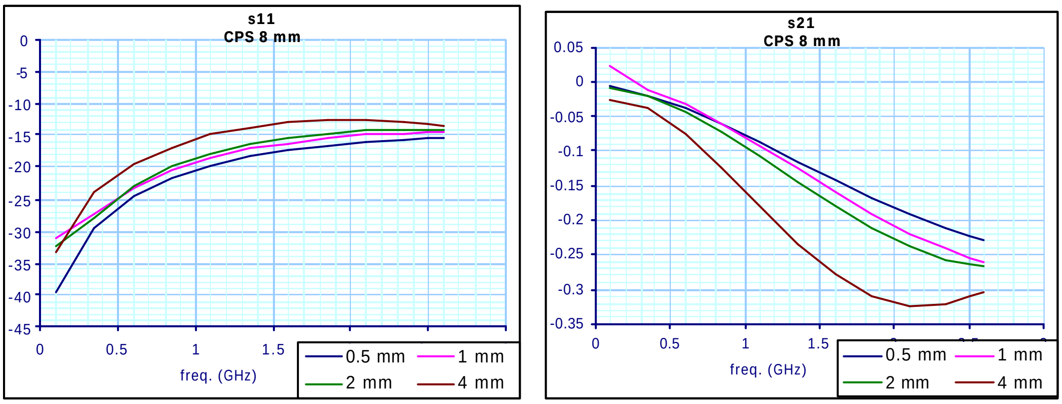


Fig. 19 - Simulated S parameters of back to back balun with ACPW length as parameter

Similarly, TCPW section is probed to provide better results when it's shorter, as shown in Fig. 20. The difference is significant specially for reflection coefficient, whereas insertion loss doesn't practically vary. Hence, the TCPW section is shortened in the design for getting better performance.

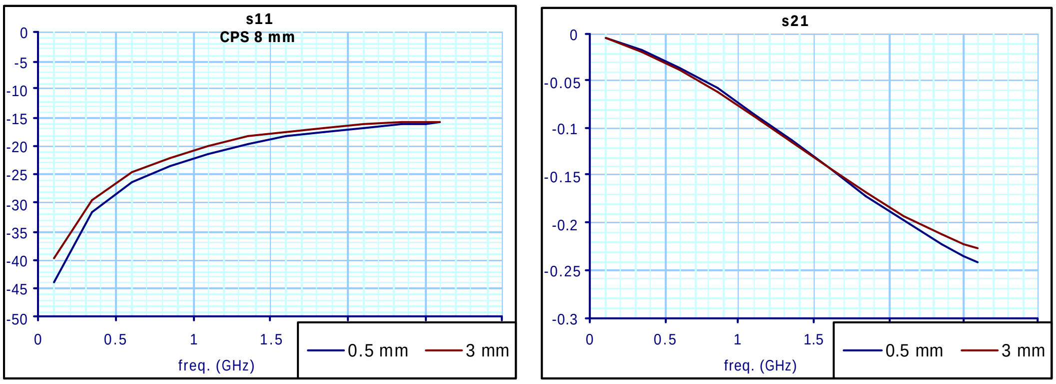


Fig. 20 – Simulated S parameters of back to back balun with TCPW section length as parameter

As occurs with Chebyshev transformer, the length of the interconnecting line is very important in the results obtained. Resonances happen at frequencies related to the length of TCPS + SCPS, so it's worth using lines as short as possible to connect the balun with the antenna.

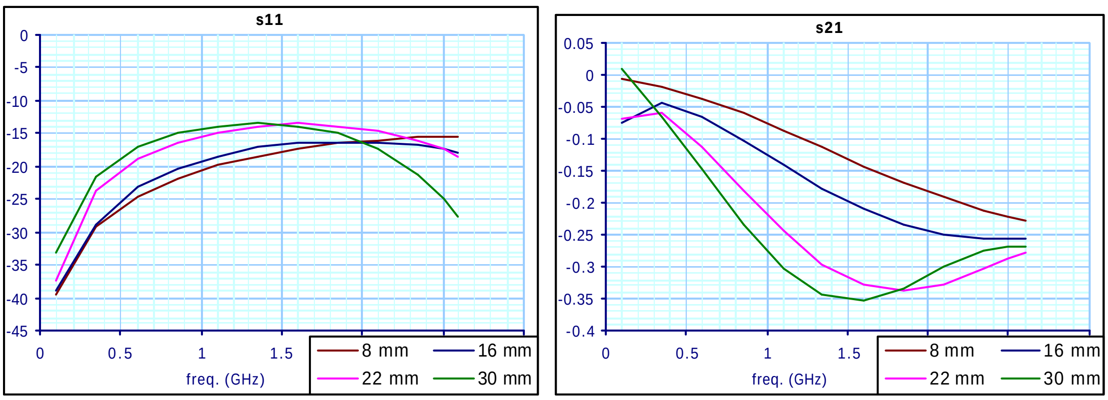


Fig. 21 – Simulated S parameters of back to back balun with CPS connecting line length as parameter

Fig. 21 shows the strong dependency of S parameters with the length of the SCPS line that connects both balun structures. As the distance between baluns increases, the frequency at which the peak occurs is lower, so to have good values of S parameters at low frequencies, it's better to use short lines.

And finally, concerning the length of the first section of the structure (SCPW), a value of 3 mm provides better results than a shorter one of 0.5 mm. But when increasing its length, performance is not improving, but it even gets worse.

This effect can be observed in Fig. 22; the reason for this is that wave ports require a length of uniform cross section; HFSS assumes that each port is connected to a semi-infinitely long waveguide that has the same cross section as the wave port, and it is excited by the natural field modes associated with it.

The length of the uniform cross section must be long enough to allow non-propagating modes to die out, and thus to ensure accurate simulation results. If this length is not enough, or discontinuities are too close to the port, non-propagating modes can reach the port, and its energy will affect the apparent energy in the dominate mode; therefore, it will produce erroneous results.

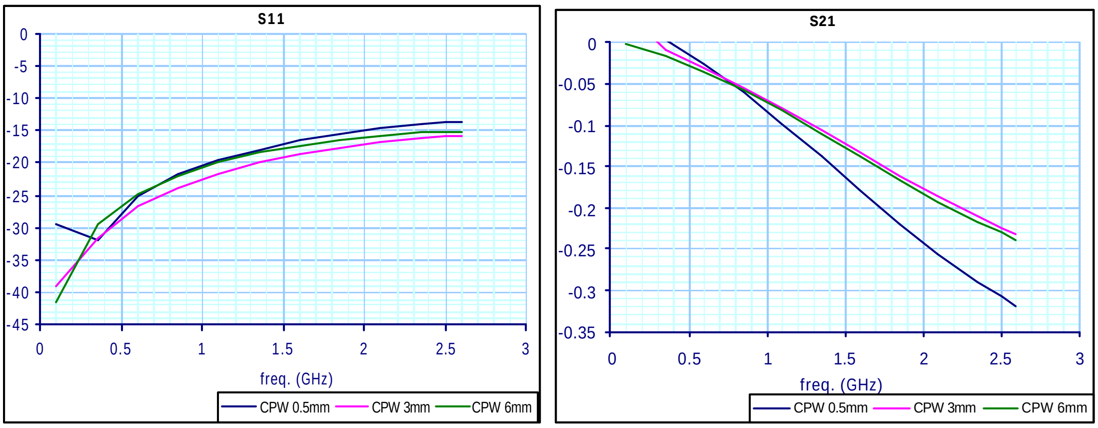


Fig. 22 – Simulated S parameters of back-to-back balun with SCPW section as parameter

Error produced will be considerable at low frequencies specially, since associated wavelength is bigger, and therefore uniform cross section should be longer to allow non-propagating modes to die.

Conclusions

For improving balun performance, the following design guidelines can be deduced from this study:

- Use of bond wires between ground strips
- Narrow inclined angle ? associated with TCPS and TSLO
- Short ACPW section
- Short TCPW section
- Short SCPS interconnecting line
- Mid-length SCPW section

Final geometric dimensions chosen for balun structure are specified in Table 1, and simulated results of back-to-back configuration are shown in Fig. 23:

Geometric dimension	Designed value
Inclined angle ?	15º
ACPW length	0.5 mm
TCPW length	0.5 mm
SCPS length	8 mm
SCPW length	3 mm

Table 1 - Geometric dimensions of designed balun

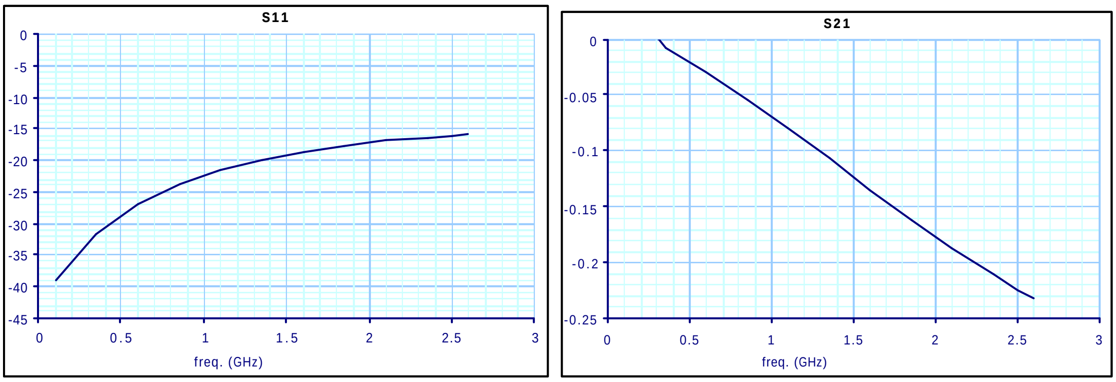


Fig. 23 – Simulated S parameters of final designed balun structure

Obtained results are excelent (S_{11} parameter is always lower than -15 dB, and losses are quite less than 0.1 dB for frequencies up to 1 GHz), but they are concerning only balun structure. Its simulated port Z_0 value is 102O, so an impedance transformer is necessary in order to connect a feeder via a 50O CPW line.

3.1.3 Impedance transformer and balun

Both structures were modelled together, resulting in a 50 O CPW to 100 O CPS transformer (Fig. 24).

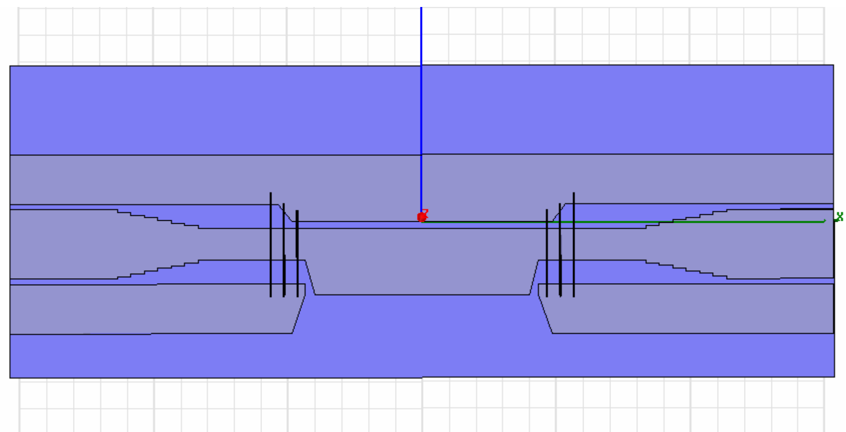


Fig. 24 –Balun structure with impedance transformer

This back to back configuration was simulated with HFSS, and conclusions extracted are resumed as follows.

Laboratory measurements, such as those from a vector network analyzer, use constant reference impedance. To obtain results consistent with measurements, the S parameters calculated by HFSS are renormalized to a constant Z_0 of 50O. Simulated ports Z_0 is about 53O, so this renormalization wosn't mean big mismatch.

Chosen physical dimensions are grounded on previous studies of both structures, separately.

For Chebyshev impedance transformer, these dimensions are:

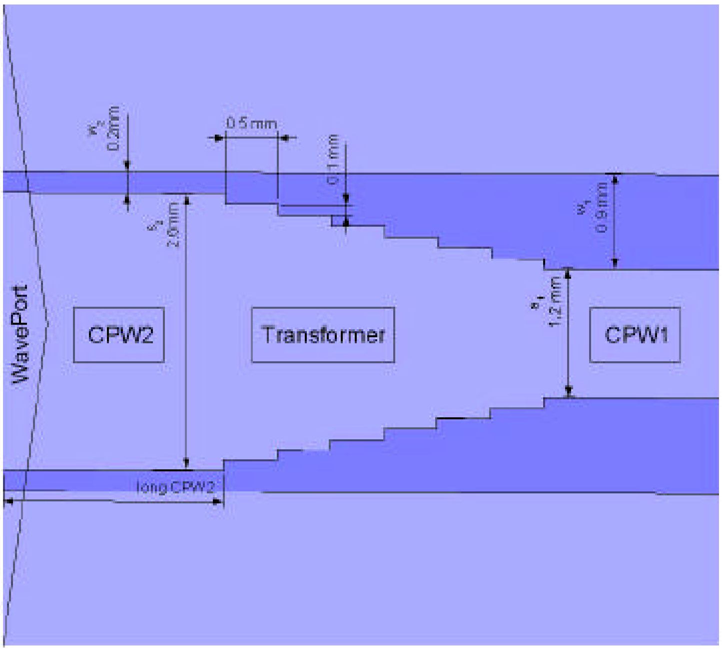


Fig. 25 – Physical dimensions of Chebyshev impedance transformer

CPW 2	*central strip width (s_2)*	2.6 mm
	gap width (w_2)	0.2 mm
	CPW2 length	7 mm
Chebyshev Transformer	*number of sections*	6
	sections length	0.5 mm
	lateral difference between sections	0.1 mm
CPW 1	*central strip width (s_1)*	1.2 mm
	gap width (w_1)	0.9 mm
	CPW1 length	1 mm

Table 2 – Physical dimensions of Chebyshev impedance transformer

And physical dimensions for balun structure are:

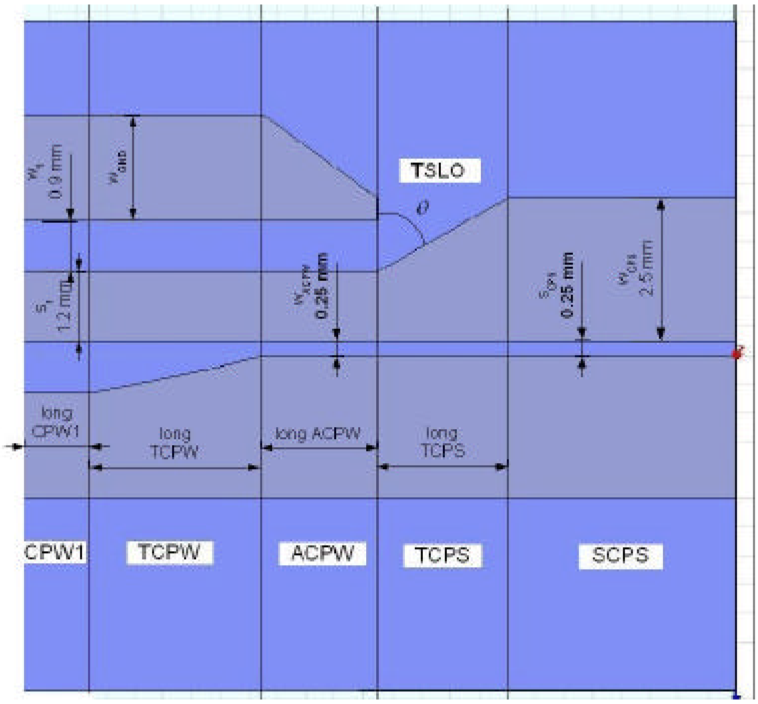


Fig. 26 – Physical dimensions of balun structure

CPW 1	*central strip width (s_1)*	1.2 mm
	gap width (w_1)	0.9 mm
	ground strip width (w_{GND})	1.8 mm
	length	1 mm
TCPW	*length*	0.5 mm
ACPW	*tapered gap width (w_{ACPW})*	0.25 mm
	length	0.5 mm
TCPS TSLO	*inclined angle (?)*	15º
	length	0.35 mm
CPS	*central gap width (s_{CPS})*	0.25 mm
	strip width (w_{CPS})	2.5 mm
	length	8 mm

Table 3 – Physical dimensions of Chebyshev impedance transformer

Bond wires in this structure have a notable influence in the final performance, as can be observed in Fig. 27. Bonding provides wider bandwidth and better maximum values.

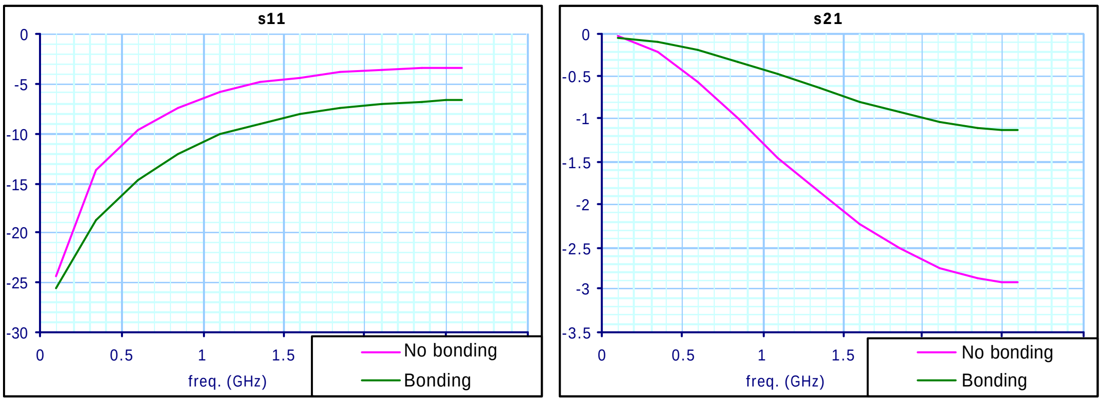


Fig. 27 – Simulated S parameters of complete balun structure back to back configuration

Between the Chebyshev transformer and the balun structure there is a CPW line that in this design is called CPW1. Fig. 28 shows performance dependence with this section's length.

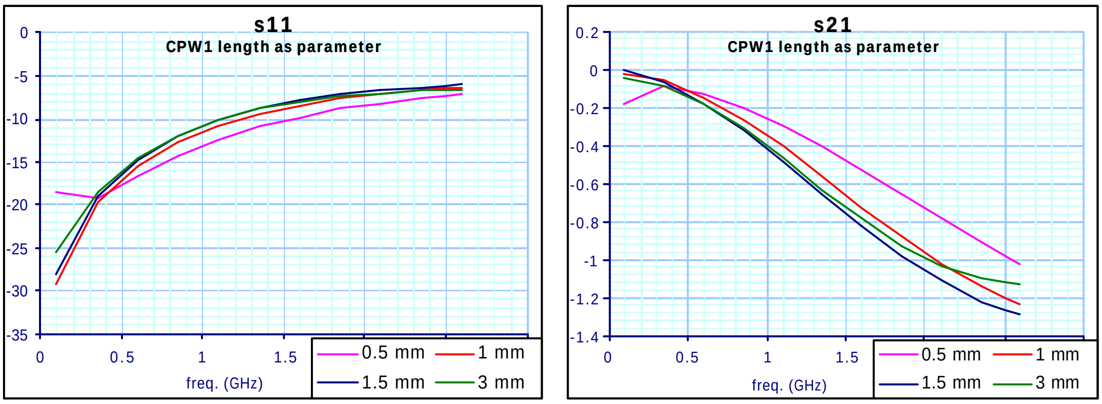


Fig. 28 – Simulated S parameters of complete balun structure in back to back configuration, with CPW1 length as parameter

In addition, first section of the complete structure is another CPW line that in this model is called CPW2. Fig. 29 shows performance dependence with this section's length:

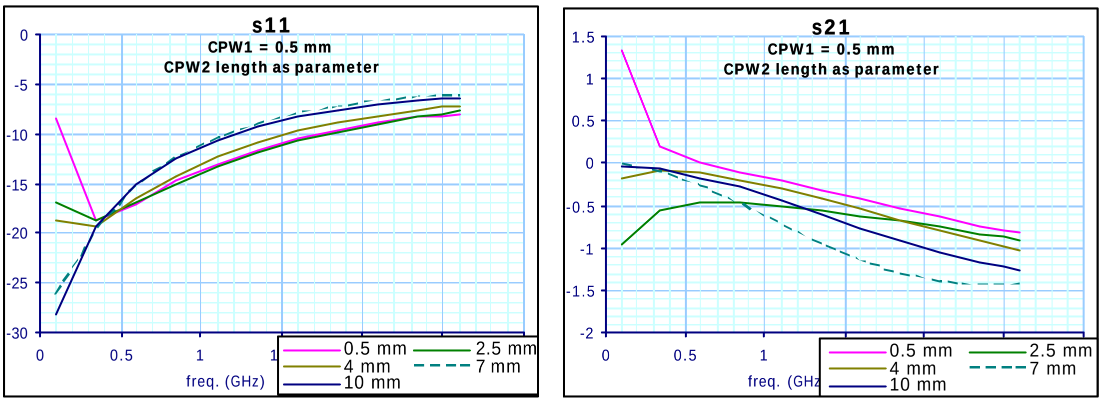


Fig. 29 – Simulated S parameters of complete balun structure in back to back configuration, with CPW2 length as parameter

Again, short CPW2 section gives better results.

But it can be observed that when combining short CPW1 and CPW2 lines, results are not coherent at low frequencies. This is due to discontinuities placed close to the port, causing non-propagating modes to reach the port. At low frequencies, wavelength is longer, so this short distance's effect is more important.

Consequently, minimum lengths to insure non-propagating modes attenuation have been chosen: 1 mm for CPW1 and 4mm for CPW2. Resulting performance of the structure is shown in

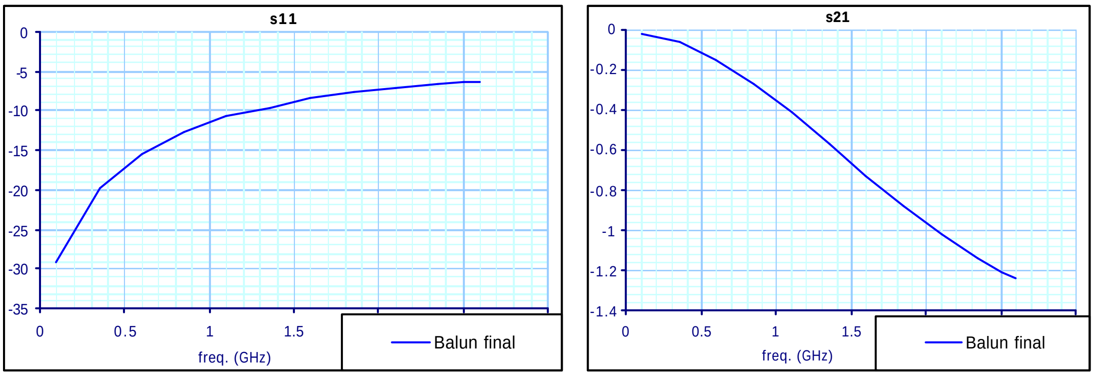


Fig. 30 - Simulated S parameters of final desiged balun structure in back to back configuration

This design could be used up to 1 GHz which resulting reflection coefficient is better than -10 dB, and losses are less than 0.4 dB (in back to back configuration, what means 0.2 dB for the single structure).

4 Construction

This structure has been constructed with the LPKF in CAY and the results obtained are exposed next.

The substrate used is Taconic TLX-8 C2/C2. Its dielectric constant is 2.55 (tolerance of ±0.04), with a thickness of 0.8 mm and up. It is clad on both sides with electrodeposited copper, but it was eliminated in one of them to make coplanar structures.

Fig. 31 shows a comparison between simulated and constructed balun with no bond wires:

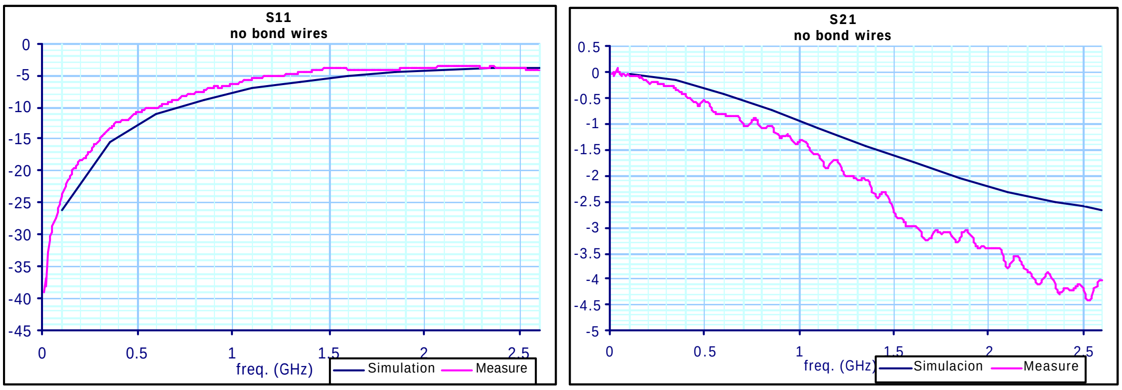


Fig. 31 – Comparison between simulated and measured balun with no bond wires

Results in reflection coefficient are very similar, but losses are higher for measured real balun. As showed in later measurements, this happens in every case. It can be due to connectors used, or to any problem with ground conections.

Several configurations for number and situation of bond wires were tested, and all measures made were compared to the simulation, which uses 3 bond wires in the positions described in 3.1.2. The ideal simulated position is too accurate to be achieved in electronics laboratory; better bonding techniques could approach better performance.

In practice, the characteristics of these wires and their effect have been found to have more importance than in simulations.

Next figures show the effect of the number of bond wires (1, 2 or 3 wires):

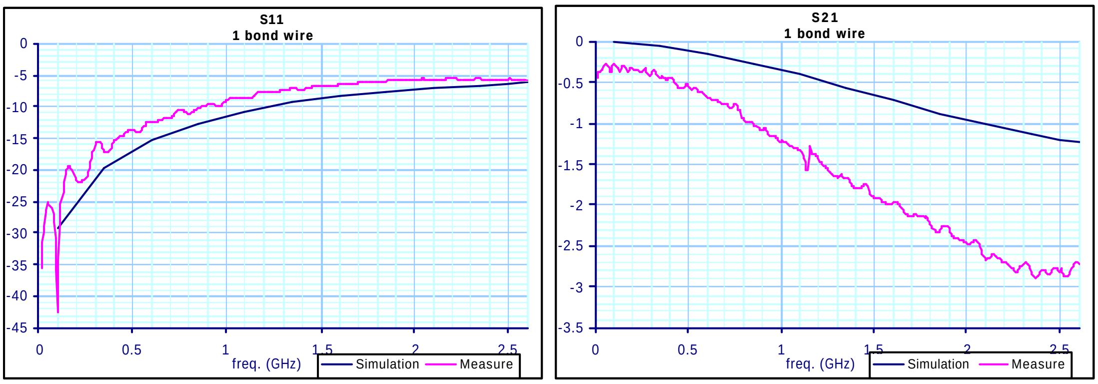


Fig. 32 – Comparison between simulated 3 wires balun, and measured balun with 1 bond wire

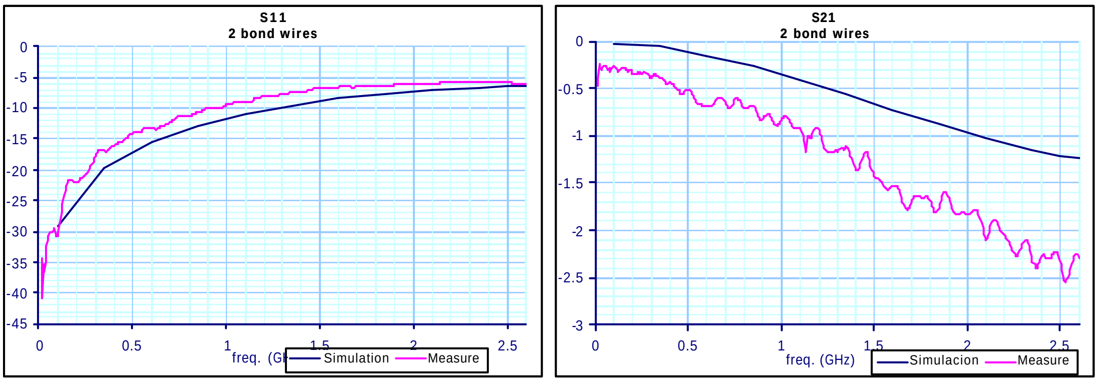


Fig. 33 – Comparison between simulated 3 wires balun, and measured balun with 2 bond wires

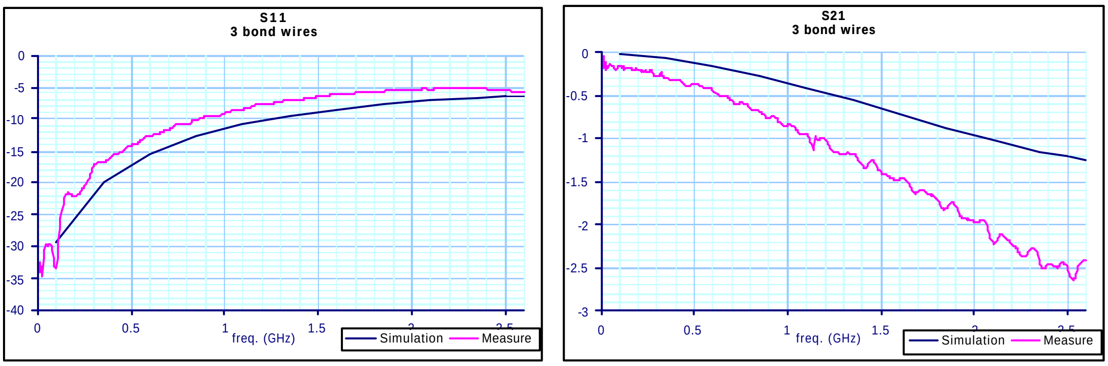
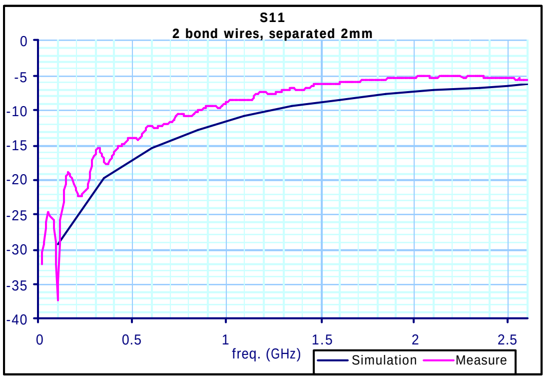


Fig. 34 – Comparison between simulated and measured 3 wires balun

Reflection coefficient is very similar for all cases, and close to ideal simulated behaviour. With only one wire, losses are quite worse than with 2 or 3, but in all cases much higher than values obtained in simulations.

All these bond wires were placed in the last zone of the balun structure, and very close to each other. When increasing this distance, results don't improve, as can be observed in , where distance between two balun were about 2 mm (one at the end of the structure and the other about the second section of the impedance transformer):



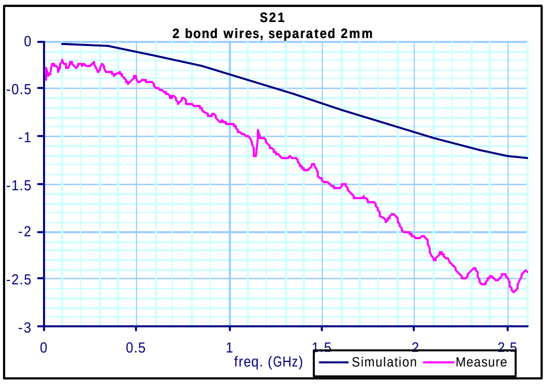


Fig. 35 – Comparison between simulated 3 wires balun, and measured balun with 2 wires 2mm separated

Next figures compare all configurations under test:

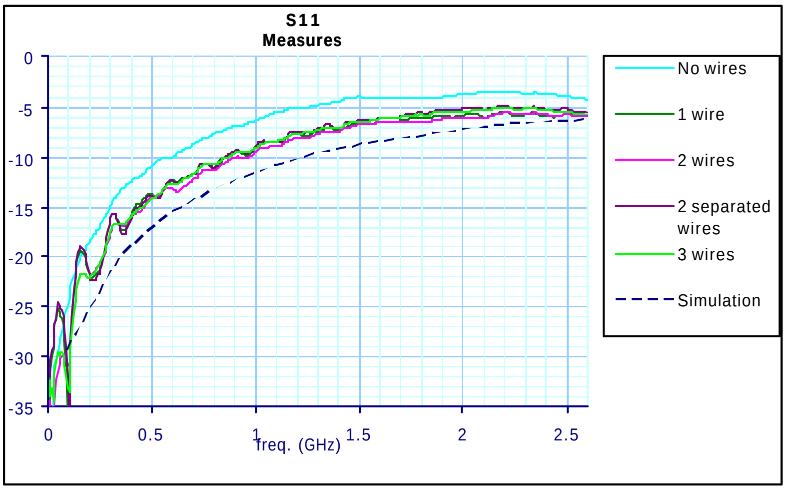


Fig. 36 – s11 comparison between all measured configurations for balun's bond wires

Reflection coefficient is very similar in all wires configuration. It is about 3 dB better than when no bond wires are used, and about 3 dB worse than simulation results. The case of 2 wires close to each other seems to be the best choice.

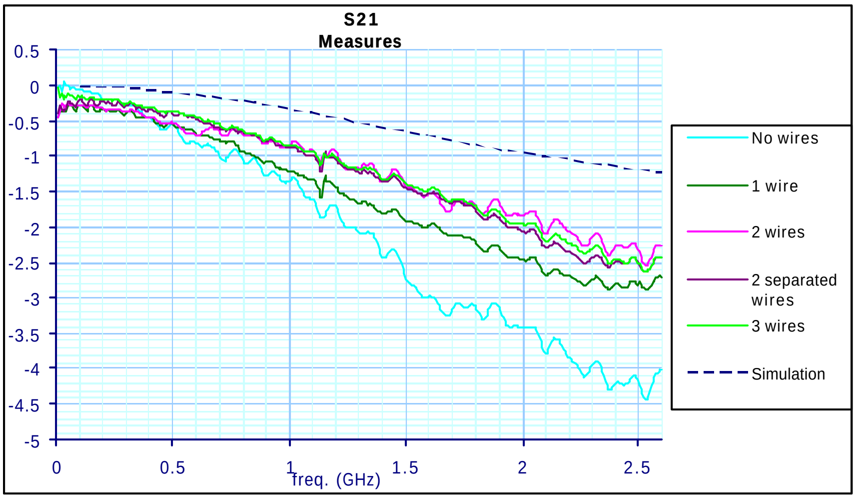


Fig. 37 – s21 comparison between all measured configurations for balun's bond wires

Losses are more dependent on wires configuration, at high frequencies specially. Highest losses are obtained when no wires are bonded, and then for 1 wire configuration. When using 2 or 3 wires, results are very similar. All measurements are much worse than simulation's results.

Finally, it was chosen the 2 wires configuration (), very near each other. shows this structure's performance.

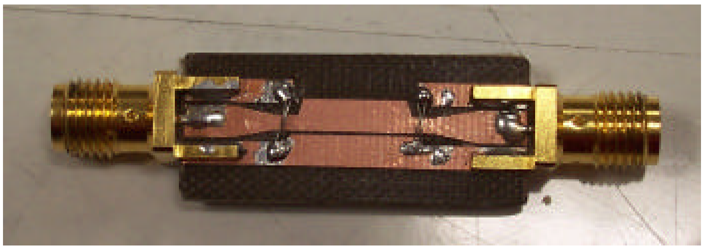


Fig. 38 –Balun constructed (back to back configuration)

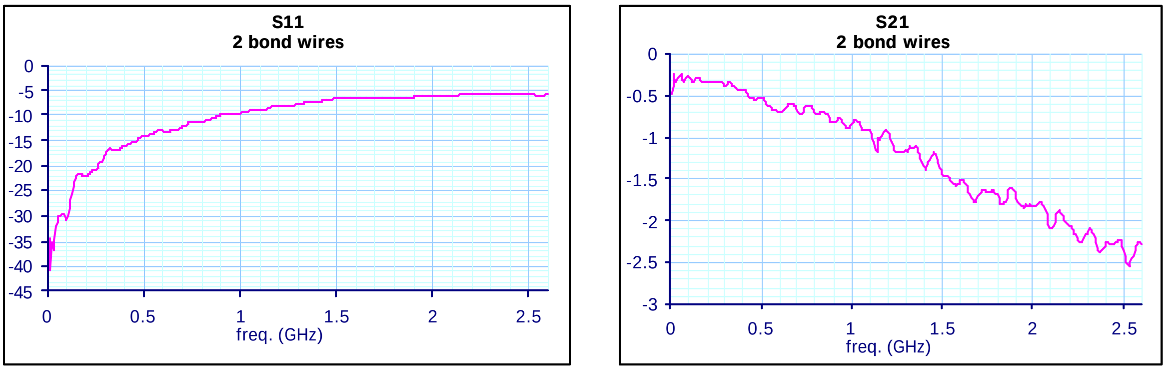


Fig. 39 – Measured S parameters of balun constructed

Up to 1 GHz, reflection coefficient is better than -10 dB. Losses are higher than supposed to be, with a maximum value at this frequency of 0.9 dB; as this is back to back configuration, each balun's losses would be 0.45 dB.

5 Conclusions

In order to improve the designed balun's performance, next guidelines are proposed:

- Change connectors used and try to find some better models. The ones than have been use could cause higher losses in the structure.

- Insert the circuit in a conductor box, that could improve ground connection

- Try any other bonding technique that could provide better results.

References

[1] *Analysis of Coplanar Waveguide-to-Coplanar Stripline Transitions*.
Shau-Gang Mao, Chieh-Tsao Hwang, Ruey-Beei Wu, Chun Hsiung.
IEEE Transactions on Microwave Theory and Techniques, January 2000

[2] *Dipolo de Banda Ancha con Balanceador de Corriente integrado*
E. de Lera, E. García, J. Vázquez, E. Rajo, J. M. Serna, M. Azuaga, J. A. López Fernández
URSI Oviedo 2006

[3] *Microstrip Lines and Slotlines*
K. C. Gupta, Ramesh Garg, Inder Bahl, Prakash Bhartia
Artech House Publishers, second edition 1996

[4] *Ansoft HFSS User's Guide*

Apendix A - Formulas

- *Skin depth of a conductor*. Distance at which current density drops to 37%.

$$d_s = \sqrt{\frac{2}{wms}} \qquad (\ \mu_0 = 4p \cdot 10^{-7}\ \text{H/m}\ )$$

- *Surface impedance of a conductor*. It’s the ratio of tangential electric and magnetic fields at the surface:

$$Z_S = R_S + jwL_S = \frac{1+j}{s d_S}$$

For example, for copper at 10GHz it is 0.026+j0.026.

Apendix B – HFSS model

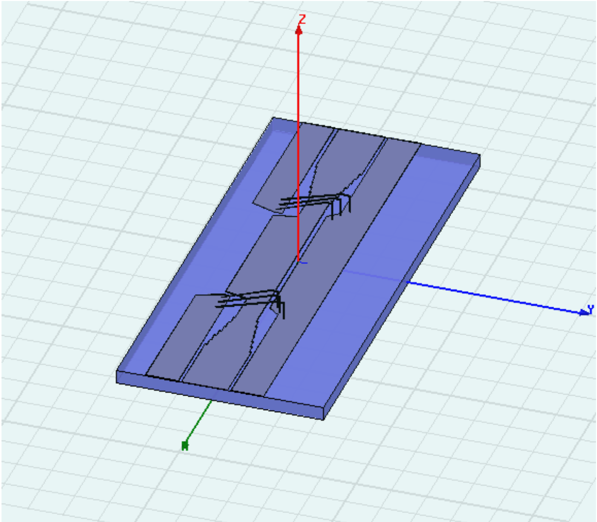


Fig. 40 – HFSS model of designed balun in back to back configuration

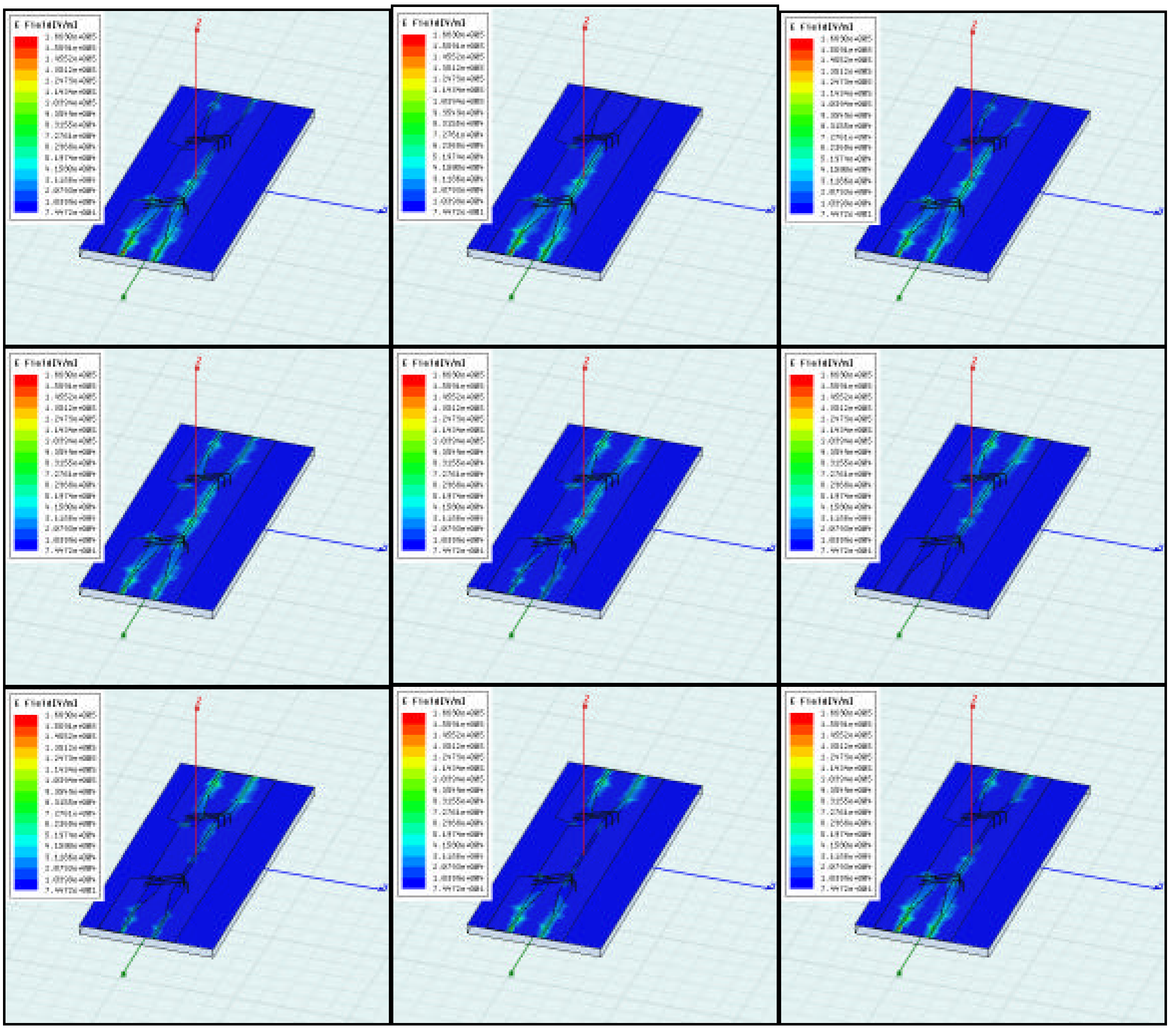


Fig. 41 – Plot of E field evolution in phase, simulated by HFSS

Apendix C – CADStar model

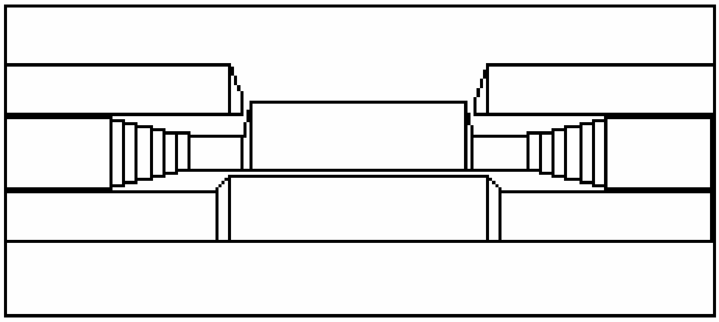


Fig. 42 – CADStar model for construction

Apendix D – Substrate datasheet

TLX

Excellent Mechanical & Thermal Properties
Low Dissipation Factor
Low & Stable Dielectric Constant
UL 94 V-O Rating
Tightly Controlled Dielectric Constant

TACONIC

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TOLL FREE: 800-833-1805 • FAX: 800-272-2503

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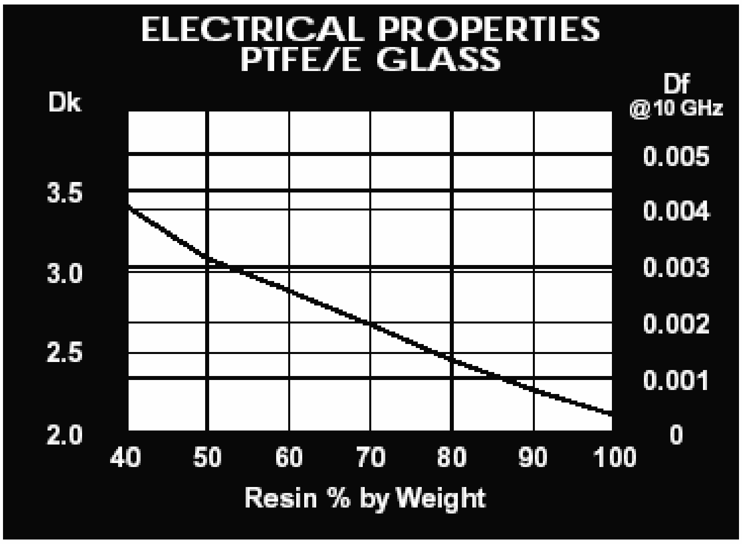
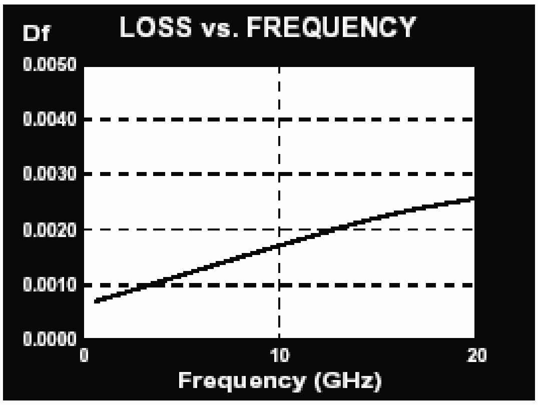
366-4 Yatap-dong Bundang-ku
Sungnam-si, Kyungki-do
Republic of Korea
TEL: 82-31-704-1858/9
FAX: 82-31-704-1857

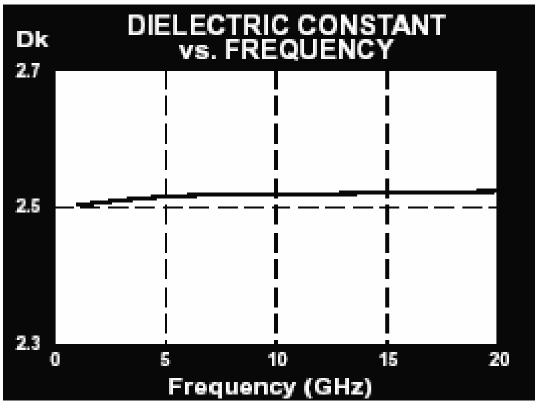
TLX-9 TYPICAL VALUES

Property	Test Method	Units	Value	Units	Value
Dielectric Constant @ 10 GHz	IPC-TM 650 2.5.5.5		2.50		2.50
Dissipation Factor @ 10 GHz	IPC-TM 650 2.5.5.5		0.0019		0.0019
Moisture Absorption	IPC-TM 650 2.6.2.1	%	<0.02	%	<0.02
Dielectric Breakdown	IPC-TM 650 2.5.6	kV	>60	kV	>60
Volume Resistivity	IPC-TM 650 2.5.17.1	Mohm/cm	10^7	Mohm/cm	10^7
Surface Resistivity	IPC-TM 650 2.5.17.1	Mohm	10^7	Mohm	10^7
Arc Resistance	IPC-TM 650 2.5.1	seconds	>180	seconds	>180
Flexural Strength Lengthwise	IPC-TM 650 2.4.4	lbs./in.	>23,000	N/mm^2	>159
Flexural Strength Crosswise	IPC-TM 650 2.4.4	lbs./in.	>19,000	N/mm^2	>131
Peel Strength (1oz copper)	IPC-TM 650 2.4.8	lbs./linear in.	12.0	N/mm	2.1
Thermal Conductivity	ASTM F 433	W/m/K	0.19	W/m/K	0.19
x-y CTE	ASTM D 3386 (TMA)	ppm/°C	9-12	ppm/°C	9-12
z CTE	ASTM D 3386 (TMA)	ppm/°C	130-145	ppm/°C	130-145
UL-94 Flammability Rating	UL-94		V-0		V-0

Type		Dk
TLY-5A		2.17
TLY-5		2.20
TLY-3		2.33
TLT-0	TLX-0	2.45
TLT-9	TLX-9	2.50
TLT-8	TLX-8	2.55
TLT-7	TLX-7	2.60
TLT-6	TLX-6	2.65
TLE-95		2.95
TLC-27		2.75
TLC-30		3.00
TLC-32		3.20
RF-30		3.00
RF-35		3.50
RF-60		6.15
CER-10		10







How to Order

Designation	Dielectric Constant	Dielectric Thickness	Dielectric Thickness
TLX - 0	2.45 +/– .04	.0050" - .0190"	0.13mm – 0.48mm
		.0200" - .0300"	0.50mm – 0.76mm
		≥ .0310"	≥ 0.78mm
TLX - 9	2.50 +/– .04	.0050" - .0190"	0.13mm – 0.48mm
		.0200" - .0300"	0.50mm – 0.76mm
		≥ .0310"	≥ 0.78mm
TLX - 8	2.55 +/– .04	.0050" - .0190"	0.13mm – 0.48mm
		.0200" - .0300"	0.50mm – 0.76mm
		≥ .0310"	≥ 0.78mm
TLX - 7	2.60 +/– .04	.0050" - .0190"	0.13mm – 0.48mm
		.0200" - .0300"	0.50mm – 0.76mm
		≥ .0310"	≥ 0.78mm
TLX - 6	2.65 +/– .04	.0050" - .0190"	0.13mm – 0.48mm
		.0200" - .0300"	0.50mm – 0.76mm
		≥ .0310"	≥ 0.78mm

Standard sheet size is 36" x 48" (914mm x 1220mm). Please contact our Customer Service Department for the availability of other sizes and claddings.

TLX can be ordered with the following electrodeposited copper:

Designation	Weight	Copper Thickness	Copper Thickness
CH	1/2 oz./sq. ft.	~ .0007"	~18 µm
C1	1 oz./sq. ft.	~ .0014"	~ 35 µm
C2	2 oz./sq. ft.	~ .0028"	~ 70 µm

Panels may be ordered cut to size

Typical Panel Sizes	
12" x 18"	304mm x 457mm
16" x 18"	406mm x 457mm
18" x 24"	457mm x 610mm
16" x 36"	406mm x 914mm
24" x 36"	610mm x 914mm
18" x 48"	457mm x 1220mm

An example of our part number is: TLX-9-0310-CH/CH-18" x 24" (TLX-9-0310-CH/CH-457mm x 610mm)

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