

PROYECTO FIN DE GRADO

TITLE: Design of high isolation satellite feeding networks

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TITULACIÓN: Grado en Ingeniería de Sistemas de Telecomunicación

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Fecha de lectura: 10 de julio de 2024

Calificación:

El Secretario/La Secretaria,

Dedicatoria

Dedicado a

...

esos grandes profesores y técnicos de laboratorio, por ayudarme a adentrarme en el maravilloso mundo de la ingeniería,

mis amigos, y los compañeros de APIDE y AETEL, sin el apoyo de los cuales dudo mucho que hubiera llegado a donde estoy ahora mismo,

mi novia Carolina, por estar ahí en todos esos malos momentos y aguantar todas esas chapas sobre “antenitas”,

mi familia, por su apoyo incondicional y fe en mí desde el primer momento.

Resumen

El objetivo de este proyecto, denominado “Diseño de redes de alimentación satelitales de alto aislamiento”, es el de aportar una solución para el UPMSat-3, proyecto del Instituto Universitario de Microgravedad “Ignacio de Riva”, el cual embarca dos transceptores de RF (radio frecuencia) que operan en la misma banda de frecuencia, y cuya operación sería detrimental para el otro.

Esto se hace mediante la creación de una red de alimentación hecha a medida usando tecnologías microstrip sobre un sustrato de uso espacial. Conteniendo un acoplador híbrido de 90° y dos balunes para aislar ambos dispositivos, la red de alimentación permite el uso de doble polarización circular para prestar también aislamiento en radiación. Siempre buscando minimizar el tamaño del dispositivo.

El proyecto comienza analizando el actual problema presentado por el UPMSat-3, así como estableciendo los parámetros de interés concretos de cada dispositivo contribuyente a ello.

Continúa aportando una base teórica de los diferentes dispositivos y procesos involucrados en la solución, analizando requisitos especiales para aplicaciones espaciales como la baja desgasificación antes de mostrar un diagrama conceptual de alto nivel de la solución.

Sigue con una detallada explicación de los pasos seguidos para concebir la solución, incluyendo el modelado y la simulación, hechas mediante el uso de CAD (diseño asistido por ordenador) como Cadence AWR Microwave Office, Simulia CST Studio Suite, Autodesk Fusion360 y Altium Designer, y fabricación, donde se pone en contacto con quince fabricantes de PCB (tarjeta de circuito impreso).

El proyecto también analiza el actual sistema de antenas empleado en el UPMSat-3 y propone un diseño nuevo y mejorado junto con su propia red de alimentación de cara a futuro desarrollo.

Para finalizar, un juego de prototipos es medido mostrando buena correlación con los requisitos listados, aunque los prototipos no son lo suficientemente fieles al diseño concebido para caracterizar de manera completa el funcionamiento de la solución.

Abstract

The aim of this project, titled "Design of high isolation satellite feeding networks", is to provide a solution to the UPMSat-3, a project led by the Instituto Universitario de Microgravedad "Ignacio de Riva", which embarks two RF (radio frequency) transceivers operating on the same frequency band and whose operation would be detrimental to the performance of the other.

This is done by creating a tailored feeding network using microstrip technologies on a space rated substrate. Containing a 90° hybrid coupler and two baluns to isolate both devices, the feeding network permits the use of dual circular polarizations in order to also provide radiation isolation. Always attempting to minimize the surface area of the device as much as possible.

The project begins by analyzing the current issue presented in the UPMSat-3, as well as laying out the concrete parameters of interest of either device contributing to it.

It continues by providing a theoretical background of the different devices and processes involved in the solution, analyzing unique requirements for space applications such as outgassing before displaying a conceptual high-level model of the solution.

It follows with a detailed explanation of the steps taken to conceive the solution, including the modelling and simulation, done through the use of modern CAD (Computer Aided Design) software such as Cadence AWR Microwave Office, Simulia CST Studio Suite, Autodesk Fusion360 and Altium Designer, and manufacturing, where fifteen different PCB (printed circuit board) manufacturing companies are contacted.

The project also analyzes the current antenna system employed on the UPMSat-3 and proposes a new and improved design along its own matching network intended for further development.

To finalize, a set of prototypes are measured showing good correlation with the listed requirements, although the prototypes are not sufficiently accurate to the conceived design to fully characterize the performance of the solution.

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Acronym List

3D	three dimension
BOM	bill of materials
CNC	computer numerical control
COTS	commercial of the shelf
dB	decibel
EM	electromagnetic
FDM	fused deposition modelling
IVA	value added tax (<i>impuesto de valor añadido</i>)
LC	inductor-capacitor
MLCC	multilayer ceramic capacitor
PLA	polylactic acid
RX	reception
SMD	surface mounted device
TDM	time domain multiplexing
TVAC	thermal vacuum chamber
TX	transmission
UHF	ultra-high frequency
VAT	value added tax
VHF	very-high frequency

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1. Introduction

1.1 Frame and motivation of the project

UPMSat-3 is a university microsatellite project led by the Instituto Universitario de Microgravedad “Ignacio de Riva” in collaboration with the STRATST group (Real-Time Systems and Architecture of Telematic Services) of the UPM; thought as a logic continuation to the previous UPM-Sat1 and UPM-Sat2 satellites.

The objective of the UPMSat-3 is to design, fabricate, qualify, launch, and operate a low-cost satellite usable as a technology demonstrator platform with educational and scientific purposes.

The satellite embarks on two radiofrequency transceivers, a Gomspace AX100 and a transceiver developed by HYDRA Space Systems and AMSAT-EA. Both being required to operate without being detrimental to the other, sharing the same frequency band.

The consequence of embarking on two transponders which share the working frequency band and operating them simultaneously may become a problem. The emissions of one transceiver may desensitize the receiver of the other, and vice versa, interrupting its service and even being able to degrade its performance irreversibly. This is due to the output power of one device being higher than the maximum input power of the other. Being more concrete:

Both devices share a common radiating element.

The TX output power of either device is higher than the absolute maximum rated RX power of the other.

There is no control over the operation of either device once on mission.

The two embarked devices, referenced as TTL1 and TTL2, are a Gomspace AX100 transceiver and a Hydra Space/AMSAT Genesis transceiver. The characteristics pertinent to this issue of both devices are displayed in Table 1.

	TTL1	TTL2
TX Band	UHF 437MHz	UHF 437MHz
TX Power	23dBm	27dBm
RX Band	UHF 437MHz	VHF
Maximum RX Power	-5dBm	N/A
Ports	Single TX/RX	Dual TX and RX

Table 1: Characteristics of interest of the transceivers.

From these characteristics we can assume a minimum 32dB of isolation is required between both devices in the UHF band. The VHF band requires no additional feeding network since the aforementioned issue is not present.

Both devices operate around the 437MHz frequency with less than 500kHz deviation, so the assumed center frequency is 437MHz.

This project main goal is to develop a feeding network which allows the use of both transceivers simultaneously without permanently damaging each other and, to the maximum extent possible, reduce the sensibilization.

As a secondary goal, the project analyzes the current antenna design used by UPMSat-3 and proposes a new antenna design and feeding network for further development.

1.2 Technical and academical objectives

The objectives of this end of bachelor's degree project are, from a technical standpoint:

- Provide a solution to an existing problem in the UPMSat-3.

- Design, simulate and fabricate a bespoke circuit for the UPMSat-3.

- Introduce further upgrades in the radio frequency segment of the UPMSat-3.

From the academical standpoint, the writer acquires competencies and abilities in the following:

- Proficiency in the use of CAD and simulation platforms applied to radio frequency and circuit design applications.

- Experience in the radio frequency realm of the aerospace segment.

- Experience in writing for academia and research purposes.

1.3 Structure of the rest of the memory

In the following chapters, an overview of how couplers work and what some of their main figures of merit are, as well as a more in-depth description of 90° hybrid couplers is presented. This is followed by an overview on baluns including some of their main figures of merit, and a substrate selection description, including an overview of outgassing criteria.

This is followed by a generation of the initial design constraints and specifications, and the development of the solution itself. The development of the solution is separated into 8 chapters which include the design of the 90° hybrid coupler, the simulation results of the 90° hybrid coupler, the design of the balun, the simulation results of the balun, the design and simulation results of the whole feeding network, the PCB design and fabrication, the analysis of the current antennas and a proposal for a future antenna and feeding network design.

2. Technological frame

2.1 Couplers

2.1.1 Description of a coupler

Couplers are RF devices with at least one input and two outputs. These ports are typically referred to as the *input* port, *output* port, and *coupled* port. As the name implies, both output ports must be electrically coupled in order to function as intended. From this coupling, we extract three characteristic figures of merit:

The first one is the *coupling factor*, which indicates the power relationship between the input port and the coupled port, defined with the formula:

$$C[dB] = -10\log_{10}\left(\frac{P_{coupled}}{P_{input}}\right) \quad [1]$$

The second is the *insertion loss*, which indicates the power relationship between the input port and the output port, defined with the formula:

$$L_i[dB] = -10\log_{10}\left(\frac{P_{output}}{P_{input}}\right) \quad [2]$$

There are primarily three types of couplers: *directional couplers*, *bidirectional couplers*, and *hybrid couplers*.

We consider the ports from now on with the following numbering: input port P1, output port P2, coupled port P3.

Directional couplers consist of the three aforementioned ports, although internally they contain a matched load terminating the opposite end of the transmission line leading to the coupled port. An illustration of the device is shown in Figure 1.

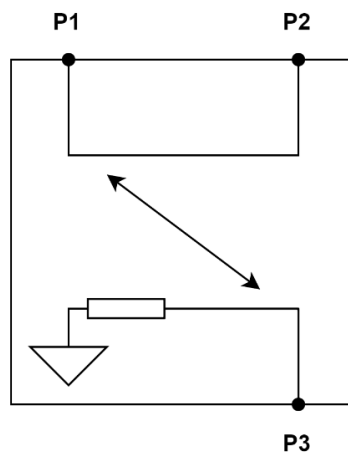


Figure 1: Directional coupler schematic.

Bidirectional couplers consist of two directional couplers connected back-to-back. The matched load is omitted, joining both coupled ports where it was previously found, generating a fourth port called *isolated* port.

We will consider the isolated port P4 from now on.

This coupler introduces a new concept and characteristic figure of merit called *isolation*, which indicates the power relationship between the input port and isolated port, defined with the formula:

$$I[dB] = -10\log_{10}\left(\frac{P_{isolated}}{P_{input}}\right) \quad [3]$$

Additionally, these devices present port symmetry under ideal conditions, being able to consider the coupler parameter equal in the following case:

Swap between input port and output port, swap between coupled port and isolated port.

An illustration of the device is shown in Figure 2.

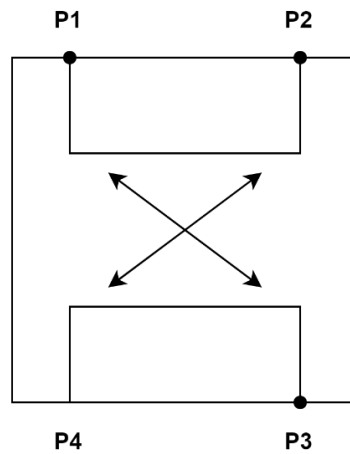


Figure 2: Bidirectional coupler schematic.

Lastly, hybrid couplers are a specific version of bidirectional couplers, characterized by presenting, under ideal conditions, a coupling factor and insertion loss of 3dB and infinite isolation.

Within the hybrid couplers, we can make distinctions between two possible variants: 90° hybrid couplers and 180° hybrid couplers. Being their only difference, the phase shift induced between output ports and coupled ports: 90° in the first case and 180° in the second case.

These couplers introduce two new characteristic figures of merit:

The first one is the *amplitude unbalance*, which indicates the difference between the coupling factor and insertion loss, defined with the formula:

$$\Delta A[dB] = |C - L_i| \quad [4]$$

The second one is the *phase unbalance*, which indicates the difference in phase between the output and coupled port when compared to the 90° or 180° characteristic of the hybrid coupler, defined with the formula:

$$\Delta\theta[^\circ] = |\theta_{coupled} - \theta_{output} - \theta_{hybrid}| \quad [5]$$

An additional characteristic figure of merit common to all circuits is the *insertion loss*, which indicates the power relationship between the incident and reflected signals, due to impedance mismatching, at a given port, defined with the formula:

$$RL[dB] = -10\log_{10}\left(\frac{P_{incident}}{P_{reflected}}\right) \quad [6]$$

For this project we make use of a 90° hybrid coupler.

2.1.2 Topology of a 90° hybrid coupler

In order to achieve the aforementioned characteristics, we may use a variety of topologies. Bellow, the most common one and the one employed is further described:

The *branchline*, in its most basic form, would be made up of four transmission lines with an electrical length of 90° each, making up squares whose vertices would be the different ports in order. In a system with a characteristic impedance of $Z_0\Omega$, the transmission lines would have the impedances shown in the figure in Figure 3.

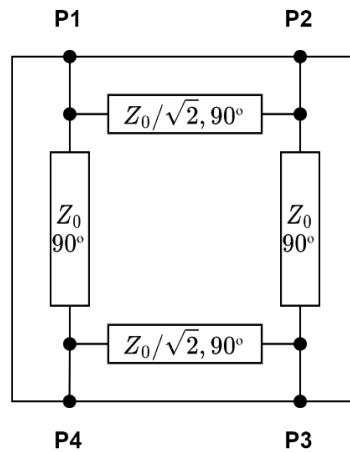


Figure 3: Branchline 90° hybrid coupler schematic.

The operational bandwidth of the device in this basic form is narrow.

The bandwidth of a device may be defined in many ways, one of the most common ones is using the *fractional bandwidth*, which indicates the relationship between the working

frequency range, defined as the range between superior and inferior frequencies that intersect a given amplitude, and center working frequency. It is defined with the formula:

$$B_F[\%] = \frac{\Delta_f}{f_c}; \Delta_f[Hz] = f_H - f_L \quad [7]$$

In order to increase the bandwidth of the device, couplers may be cascaded as shown in Figure 4:

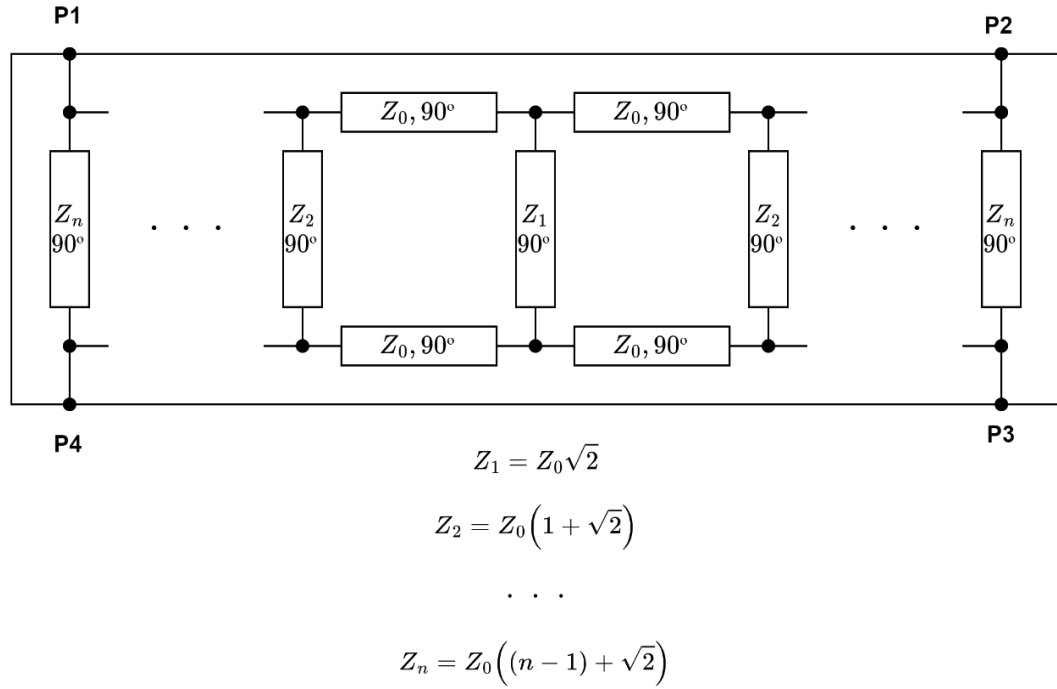


Figure 4: Tandem branchline 90° hybrid coupler schematic.

The branchline topology may be developed using *lumped elements* as well as *microstrip lines* and others. Although the use lumped elements may, in some cases, result in a circuit with a reduced footprint, fabrication tolerances and parasitic effects, intrinsic of their packages, may lead to lower performance characteristics and replicability when compared to those made using microstrip technology. These effects are directly accentuated with higher working frequencies.

For this project we use microstrip lines in order to form a single stage branchline circuit, due to the required bandwidth being narrow and isolation being remarkably high.

Additionally, there are other possible topologies such as *tandem 8.34dB couplers* or *lange couplers*, which may, using microstrip technologies, offer wider bandwidth at the cost of a more complex design and stackup. They are deemed unnecessary due to the aforementioned narrow bandwidth.

2.2 Baluns

2.2.1 Description of a balun

Baluns are RF devices with at least one input and two outputs. When the reference signal is unbalanced, these ports are referred to as *input* port, as *input* port, *output positive* port, and *output negative* port. As the name implies, baluns convert balanced to unbalanced signals, and vice-versa. The concept of balanced and unbalanced lines stems from the notion impedance referencing, balanced lines are characterized by impedance between two lines, while unbalanced lines are characterized from a line to ground.

The process of extracting a balanced signal from an unbalanced signal consists of splitting the signal between two output ports and phase shifting it so that the signal at output negative port is 180° out of phase with one at the output positive port. This splitting induces a 3dB reduction in power between any single balanced line and the unbalanced line.

From this, we extract three figures of merit:

The first one is the *insertion loss*, which indicates the power relationship between the input port and any the output ports, defined with the formula:

$$L_i[dB] = -10\log_{10}\left(\frac{P_{input}}{P_{output+/-}}\right) \quad [8]$$

The second one is the *amplitude unbalance*, which indicates the power difference between the output positive port and the output negative port, defined with the formula:

$$\Delta A[dB] = |P_{output+} - P_{output-}| \quad [9]$$

The third one is the *phase unbalance*, which indicates the difference in phase between the output positive port and output negative port compared to the 180° characteristic of the balun, defined with the formula:

$$\Delta\theta[^\circ] = |\theta_{output+} - \theta_{output-} - 180^\circ| \quad [10]$$

Baluns may be used as signal splitters due to them dividing the signal between two ports, but they may additionally be used as impedance converters due to the impedance converting nature of transformers. This provides us with the characteristic of *impedance ratio* which indicates the relationship between the balanced line-to-line impedance and the unbalanced line-to-ground impedance, defined with the formula:

$$N = \frac{Z_{balanced}}{Z_{unbalanced}} \quad [11]$$

There are primarily four types of baluns: *conventional transformer baluns*, *auto transformer baluns*, *transmission line baluns* and *Marchand baluns*.

We consider the ports from now on with the following numbering: input port P1, output positive port P2, output negative port P3.

Conventional transformer baluns, oftentimes called voltage baluns, force equal voltage on both balanced lines. They are composed of a transformer where the input port is connected to one primary terminal and ground to the other, the output positive port is connected to the secondary terminal opposite to the input port and the output negative port to the other.

An illustration of the device is shown in Figure 5.

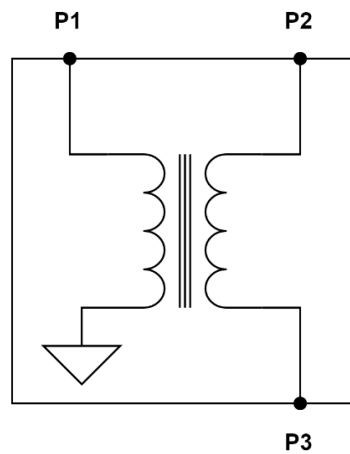


Figure 5: Conventional transformer balun schematic.

Autotransformers are composed by a single winding with a center tap connected to ground. The input port is connected to one terminal of the transformer, so is the output positive port. The output negative port is connected to the other terminal of the winding.

An illustration of the device is shown in Figure 6.

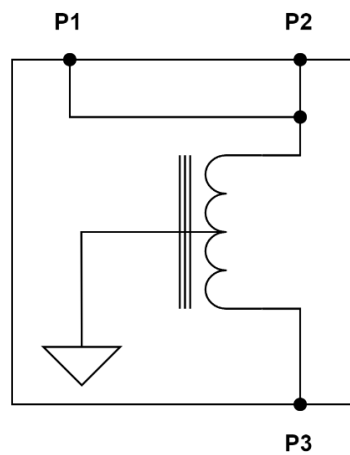


Figure 6: Autotransformer balun schematic.

Transmission line baluns, oftentimes called current baluns, force equal current on both balanced lines. They are composed of a transformer where the input port is connected to one primary terminal and the output positive port to the other, ground is connected to the secondary terminal opposite to the input port and the output negative port to the other.

An illustration of the device is shown in Figure 7.

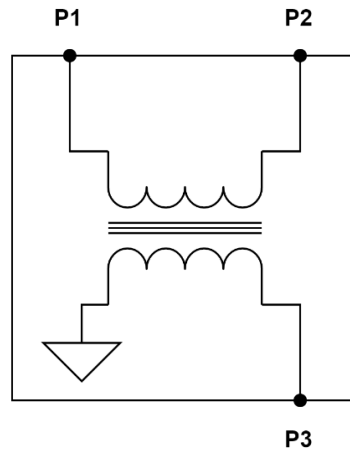


Figure 7: Transmission line transformer balun schematic.

Marchand baluns are composed by cascading two transmission line baluns back-to-back and connecting what would be the output positive ports of both baluns together, leaving the second balun primary port floating. The output positive port and output negative port would be the former output negative ports of either balun, the one connected to the first balun terminal the positive.

An illustration of the device is shown in Figure 8.

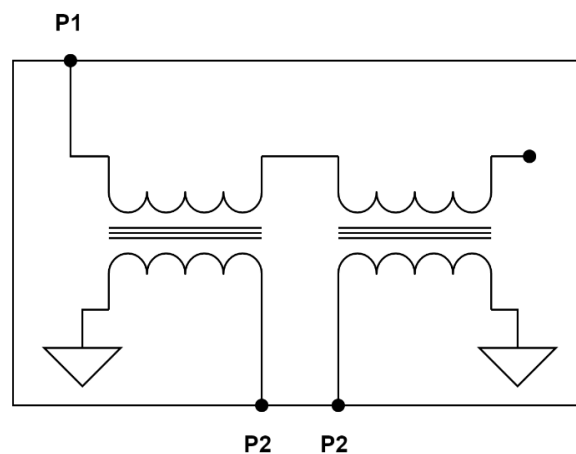


Figure 8: Marchand balun schematic.

2.3 Substrates

2.3.1 Adequate substrate selection

When working with microstrip technologies, selecting an appropriate substrate is crucial. There are a series of characteristics which must be considered to attend all electromagnetic, thermal, mechanical, and chemical requirements and necessities.

The main characteristics considered for the substrate selection include:

The first one is the *dielectric constant*, which indicates the relationship between the permittivity of a substrate material and the permittivity of vacuum, defined with the formula:

$$\varepsilon_r = \frac{\varepsilon}{\varepsilon_0} \quad [12]$$

And allows us to calculate the wavelength of a given frequency for the substrate with the formula:

$$\lambda[m] = \frac{c_0}{f\sqrt{\varepsilon_r}} \quad [13]$$

Consequently, the dielectric constant is going to define the size of the circuit elements in an inversely proportional manner. High dielectric constants are used when miniaturization is required, when working with lower frequencies for example, and low dielectric constants are preferred when fabrication tolerances are more significant in relationship with the frequency wavelength in vacuum, when working with higher frequencies for example.

The second characteristic is the *tangent loss*, which indicates the relationship between the real and imaginary parts of the displaced currents in a substrate material, defined with the formula:

$$\tan\delta = \frac{\omega\varepsilon'' + \sigma}{\omega\varepsilon'} \quad [14]$$

And allows us to estimate the quality of the transmission properties of a substrate and its associated losses. The loss tangent varies depending on other substrate material properties and working frequency.

The third characteristic is the *thermal coefficient of dielectric constant*, which indicates the relationship between the variation in dielectric constant of a substrate material and temperature, defined with the formula:

$$TC_{\varepsilon_r} \left[\frac{ppm}{^\circ K} \right] = \frac{\Delta_{\varepsilon_r}}{\Delta_T} \cdot 10^6 \quad [15]$$

A lower thermal coefficient of dielectric constant indicates higher thermal stability of a circuit and, consequently, higher operating temperature ranges.

The fourth characteristic is the *coefficient of thermal expansion*, which indicates the relationship between the variation in size of a substrate material and temperature, defined with the formula:

$$CTE \left[\frac{ppm}{^{\circ}K} \right] = \frac{\Delta_L}{\Delta_T} \cdot 10^6 \quad [16]$$

When selecting substrates, especially those which may be exposed to an extended temperature range, it is important to attempt to match the coefficient of thermal expansion of the substrate material, to that of pure copper, $16.5 \frac{ppm}{^{\circ}K}$. This is due to dissimilar expansion in the X, Y axes being able to damage or even delaminate the copper traces deposited on the substrate material; while dissimilar expansion on the Z axes may lead to broken vias or stackup separation, among other issues.

The fifth characteristic is the compliance with *low outgassing* requirements as per the ASTM E595 [1] or ECSS-Q-70-02 [2] standards, followed by NASA and ESA, respectively. These standards lay a series of indicators and conditions:

Recovered mass loss (*RML*), defined with the formula:

$$RML = TML - WVR \quad [17]$$

TML (total mass loss), defined as the total outgassed mass loss of a material sample at a constant specific temperature and working pressure over a specific time period.

WVR (water vapor recovered), defined as the mass of the water vapor recovered by the sample after an optional reconditioning step.

CVCM (collected volatile condensable materials), defined as the amount of outgassed material of a test sample condensed on a collector kept at a specific temperature over a specific time period.

ASTM E595:

Requires a $TML < 1.0\%$ and a $CVCM < 0.1\%$.

ECSS-Q-70-02:

Requires a $RML < 1.0\%$ and a $CVCM < 0.1\%$.

Outgassing occurs in environments exposed to high vacuum and/or extremely elevated temperatures. The vestigial gases may hinder the performance of other embarked systems and provide a medium for corona discharges. A *bake-out*, which is a procedure

whereby running the material through high vacuum and elevated temperatures an attempt is made to outgas all of the aforementioned materials prior to operational deployment, shall be performed, minimizing the chances of it occurring while on mission.

For this project, the selected substrate material is TMM6 [3], manufactured by Rogers Corporation. It is a “ceramic, hydrocarbon, thermoset polymer composite material designed for high plated-thru-hole reliability strapline and microstrip applications.” The substrate material “combines many of the benefits of both ceramic and traditional PTFE microwave circuit materials, while offering an exceptionally low thermal coefficient of dielectric constant and nearly isotropic coefficients of thermal expansion, very closely matching those of copper.”

In Table 2, a comparison between TMM6, generic FR4, a high PTFE content substrate such as RT5880 and a high ceramic content substrate such as RT6010.2LM is made.

Characteristic	TMM6	FR4	RT5880	RT6010.2LM	Unit
ϵ_r	6.3	4.5	2.2	10.7	N/A
$\tan\delta$	0.0023	0.02	0.0009	0.0023	N/A
$TC\epsilon_r$	-11	200	-125	-425	$\left[\frac{ppm}{^\circ K}\right]$
$CTE - x, y$	18,18	14-17, 17-17	31,48	24,24	$\left[\frac{ppm}{^\circ K}\right]$
$CTE - z$	26	~70	237	47	$\left[\frac{ppm}{^\circ K}\right]$
TML	*0.04-0.07	0.18-0.62	0.01	0.02	%
WVR	*0.02-0.04	0.05-0.20	0.02	<0.01	%
CVCM	*0.00	0.00-0.01	<0.01	<0.01	%

Table 2: Substrate material comparison, data extracted from different Rogers datasheets, [4] and [5].

*Although the TMM6 material is not characterized in the aforementioned document or in the NASA database, the results of the TMM family of materials are consistent with the compliance of established low outgassing requirements. This way, it is understood that the substrate material, fabricated with the same process as the rest of the family, also complies with the established low outgassing requirements.

As can be seen in the table, both TMM6 and RT6010.2LM offer a higher degree of miniaturization when compared with the standard, FR4; greatly appreciated due to the low working frequency used in this project. Even though the loss tangent of TMM6 is not as low as the one of RT5880, it performs better than the standard, FR4, with acceptable loss performance. The thermal coefficient of dielectric constant of TMM6 is by far the lowest of them all, and the thermal coefficients are, overall, the ones that more closely match those of pure copper. All the substrates pass outgassing requirements, even though FR4 is clearly the worst performer in this aspect.

3. Specifications and design restrictions

To design the 90° hybrid, we use AWR Microwave Office by Cadence. A progressive approach is used throughout the design process, starting with ideal element models, and ending with EM models of the final circuit, iterating through different versions and improvements.

The 90° hybrid design adheres to the following premises:

The circuit has a Z_0 of 50Ω.

The circuit has a Δ_f of 4MHz, being f_c 437MHz.

The hybrid has an I equal or superior to 32dB within the working frequency.

The substrate material has a ϵ_r of 6.3 and a $\tan\delta$ of 0.023.

The substrate is 0.635mm* thick and the copper 0.018mm thick.

And the following parameters are set in AWR:

Start simulation frequency of 430MHz.

Stop simulation frequency of 444MHz.

Step of simulation samples of 0.1MHz.

The steps taken and different measurements are detailed below.

**The substrate initially selected is 0.762mm, thick, this thickness is changed to 0.635mm for the definitive version of the circuit due to the availability of samples offered by Rogers. The intermediate design steps are covered using the 0.762mm thick substrate.*

To design the balun two approaches are taken and one of them is selected to be used on the final design: a design using lumped elements consisting of an LC transformer, and a design using a COTS transformer. Both are designed and simulated using AWR Microwave Office by Cadence.

Throughout the design process a progressive approach is used, beginning with ideal element models, and ending with EM models of the final circuit, iterating through different versions and improvements.

In this case, the steps taken to reach the final result of either circuit are much more streamlined due to the circuit simulation being significantly more similar to the electromagnetic one, and the computing times being lower.

The balun designs adhere to the following premises:

The circuit has a Z_0 of 50Ω.

The circuit has a Δ_f of 4MHz, being f_c 437MHz.

The balun has an RL equal or superior to 29dB within the working frequency.

The substrate material has a ϵ_r of 6.3 and a $\tan\delta$ of 0.023.

The substrate is 0.635mm* thick and the copper 0.018mm thick.

And the following parameters are set in AWR:

Start simulation frequency of 430MHz.

Stop simulation frequency of 444MHz.

Step of simulation samples of 0.1MHz.

4. Description of the proposed solution

The solution presented is a passive PCB circuit capable of isolating the problematic ports using a 90° hybrid and two baluns as shown in Figure 9.

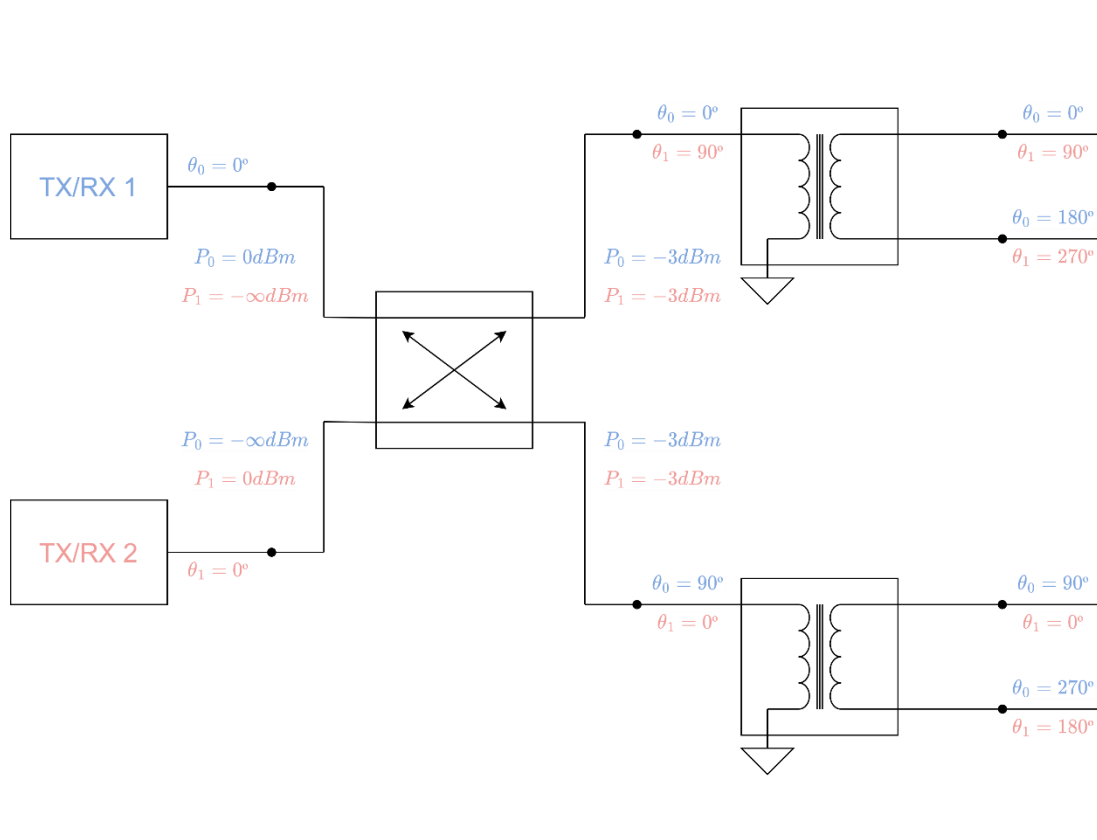


Figure 9: Proposed feeding network schematic.

Other possibilities may include TDM using RF switching or the use of power dividers on a PCB circuit. The 90° hybrid + baluns option is preferred due to it allowing the uninterrupted use of both devices via a dual circular polarization separation using 4 antennas, maintaining isolation in radiation as shown in Table 3. Additionally, the fact that the device is passive implies that no power buses need to be derived to the feeding network.

	LHCP	RHCP	Slant -45°	Slant 45°	Horizontal	Vertical
LHCP	0dB	∞	3dB	3dB	3dB	3dB
RHCP	∞	0dB	3dB	3dB	3dB	3dB
Slant -45°	3dB	3dB	0dB	∞	3dB	3dB
Slant 45°	3dB	3dB	∞	0dB	3dB	3dB
Horizontal	3dB	3dB	3dB	3dB	0dB	∞
Vertical	3dB	3dB	3dB	3dB	∞	0dB

Table 3: Theoretical polarization loss comparison, where blue indicates no loss and orange indicates complete loss.

4.1 Design and simulation of the 90° hybrid coupler

4.1.1 Ideal transmission line circuit

We start designing an ideal branchline circuit using TLIN elements, as shown in Figure 11. These elements model a lossless transmission line with a specified impedance, electrical length and working frequency.

In this case the elements follow the specifications of an ideal basic branchline such as the one described previously, being $f_c = 437\text{MHz}$, $Z_0 = 50\Omega$ y $Z_1 = Z_0/\sqrt{2}$.

In the measurements in Figure 12 we can appreciate that at the working frequency:

The reflection coefficient of the input port, S_{11} , is null.

The isolation (I), S_{41} , is infinite.

The power is evenly distributed between the output (L_i), S_{21} , and coupled (C), S_{31} , ports. Indicating a null amplitude unbalance (ΔA).

The phase shift between the output and coupled ports is 90° (discontinuous lines), indicating a null phase unbalance ($\Delta\theta$).

The bandwidth (Δ_f) at -32dB is $\sim 10\text{MHz}$.

Even though it may not be discernable, the S_{11} and S_{41} traces fully overlap, the same occurs between the S_{21} and S_{31} traces.

Although seemingly obvious, the ideal circuit simulation allows us to check whether, even though ideally, our topology is capable of fulfilling specifications. In the case where we might, for example, require a bandwidth at -32dB of 20MHz, we would need to switch topologies or reconsider the design.

Provided the topology is acceptable we continue with the next step.

4.1.2 Basic circuit elements circuit

We continue designing a basic branchline using MLIN microstrip elements as shown in Figure 13. These elements model a lossy transmission line with a substrate of defined physical properties. Below an illustration shows the definition of the MSUB microstrip substrate, within the Global Definitions, used for the microstrip circuit simulations as shown in Figure 10.

MSUB
Er=6.3
H=0.762 mm
T=0.018 mm
Rho=1
Tand=0.0023
ErNom=0
Name=TMM6_30

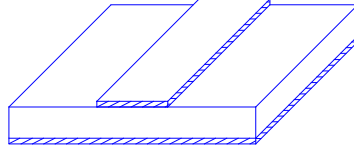


Figure 10: MSUB definition inside AWR.

In this case we make use of AWR TXLINE tool to calculate the initial lengths and widths of the different elements. These are further optimized using both the AWR optimizer and manual optimization in order to adjust the circuit to fulfill the requirements.

Although for determined applications and requirements, circuit simulations may be sufficiently true to reality to be considered valid, in this case they are only used in order to further approximate the parameters of the circuit before performing EM simulations. These are, in all cases, more accurate descriptors of actual circuit performance, with the drawback of requiring significantly more computing time.

Simulation optimized values rarely coincide among both types of simulation, but circuit and electromagnetic parameters are notably much more similar than the ideal or theoretically obtained values. Meaning this stepped optimization model greatly reduces the overall computing time.

The circuit is restimulated, this time utilizing the 2.5D AXIEM electromagnetic simulator and further optimized until the design requirements are met.

Provided the topology and results shown in Figure 14 are acceptable we continue with the next step.

4.1.3 Circuit elements circuit and initial meander

A reduction in physical dimensions via the use of meanders in the transmission lines is pursued as shown in Figure 15. Commencing with the horizontal transmission lines, a section formed by two curves of 60° and 10mm radius is added to the circuit. This way the horizontal size of the circuit is reduced 6.5mm when compared to the initial layout.

The radius of the corner is considered when designing the meander not only by its physical dimensions but also by its potential contribution to additional line coupling, detrimental to the performance of the circuit. In the same way, separation is kept between the different lines to minimize unintended coupling.

The circuit is electromagnetically restimulated corroborating no significant unintended coupling occurs and further optimized until the design requirements are met.

Provided the topology and results shown in Figure 16 are acceptable we continue with the next step.

4.1.4 Circuit elements circuit and final meander

We continue introducing meanders in the vertical lines with a section formed by two curves of 90° and 10mm radius as shown in Figure 17. This way the horizontal size of the circuit is reduced 22.75mm when compared to the initial layout.

The circuit is electromagnetically restimulated corroborating no significant unintended coupling occurs and further optimized until the design requirements are met.

Provided the topology and results shown in Figure 18 are acceptable we continue with the next step.

4.1.5 Circuit elements circuit and final meander, on 0.635mm substrate

Due to the availability of 0.635mm thick TMM6 samples, the previous circuit and substrate definition are modified to match the new substrate thickness. The circuit is then restimulated and optimized, as shown in Figure 19 and Figure 20.

Being this the last iteration of the circuit, the model is exported to Simulia CST Studio Suite as shown in Figure 21, where a 3D fully electromagnetic simulation is performed to corroborate the accuracy of the previously simulated circuits.

CST offers a plethora of solvers closely suited to different applications, being the two most significant ones used for this type of systems the *time domain* solver and *frequency domain* solver, below a brief description is provided:

The time domain solver [6] calculates the development of fields through time at discrete locations and at discrete time samples, using a hexahedral mesh. CST offers two subtypes of time domain solvers: finite integration technique solver, called transient solver inside the program, and transmission line method, called TLM solver inside the program. The time domain solver is remarkably efficient when obtaining the entire broadband frequency behavior of a device over a wide frequency range, converging faster than the time domain solver in these situations.

The frequency domain [7] solver obtains the results by Maxwell's equations being transformed into the frequency domain. It uses either tetrahedral or hexahedral meshes. CST offers two subtypes of frequency domain solvers: general purpose, intended for the widest range of simulation targets, and fast reduced model, intended for broadband calculations. The frequency domain solver is most efficient when calculating a resonant frequency behavior over a small frequency range, converging faster than the time domain solver in these situations.

In this case, being the circuit of narrowband and high-quality factor nature, the general-purpose frequency domain solver is chosen. The simulation results as shown in Figure 22 of both circuits are described in Table 4.

	435Mhz	437Mhz	439Mhz
$I[dB]$	35.21	35.33	33.75
$I[dB]$	35.59	35.08	33.01
phase shift[°]	90.072	90.038	90.001
phase shift[°]	90.159	90.136	90.105
$S_{11}[dB]$	-44.62	-41.52	-36.10
$S_{11}[dB]$	-29.16	-29.55	-29.35
$S_{21}[dB]$	-3.15	-3.16	-3.16
$S_{21}[dB]$	-3.18	-3.18	-3.19
$S_{31}[dB]$	-3.28	-3.28	-3.27
$S_{31}[dB]$	-3.16	-3.16	-3.15

Table 4: AWR simulation results in blue and CST simulation results in orange.

It can be appreciated that both programs show an isolation superior to 32dB for the specified working frequency range and differ less than 0.75dB throughout the entire range.

4.2 Simulation results of the 90° hybrid coupler

4.2.1 Ideal transmission line circuit results

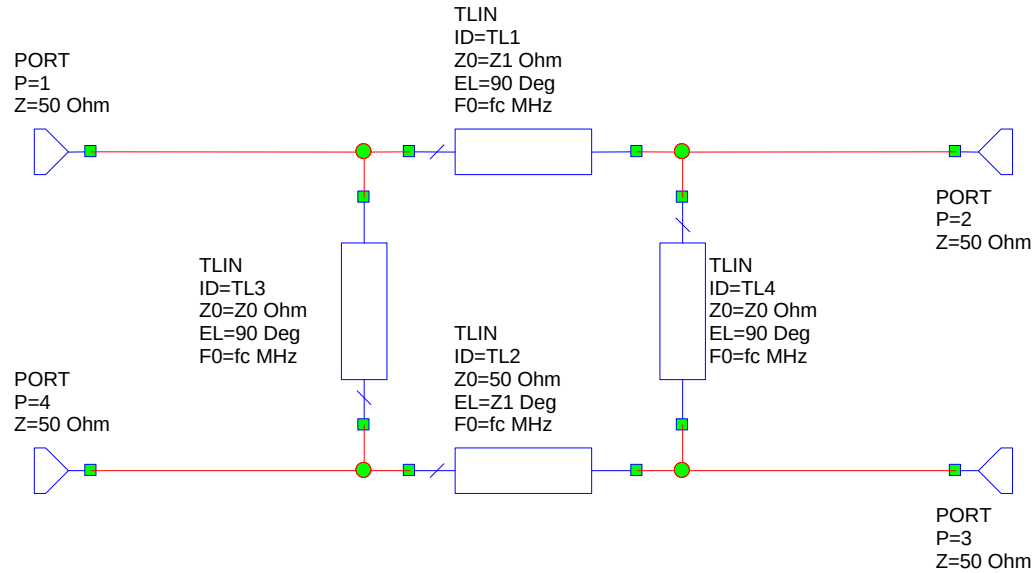


Figure 11: Ideal transmission line 90° hybrid coupler schematic.

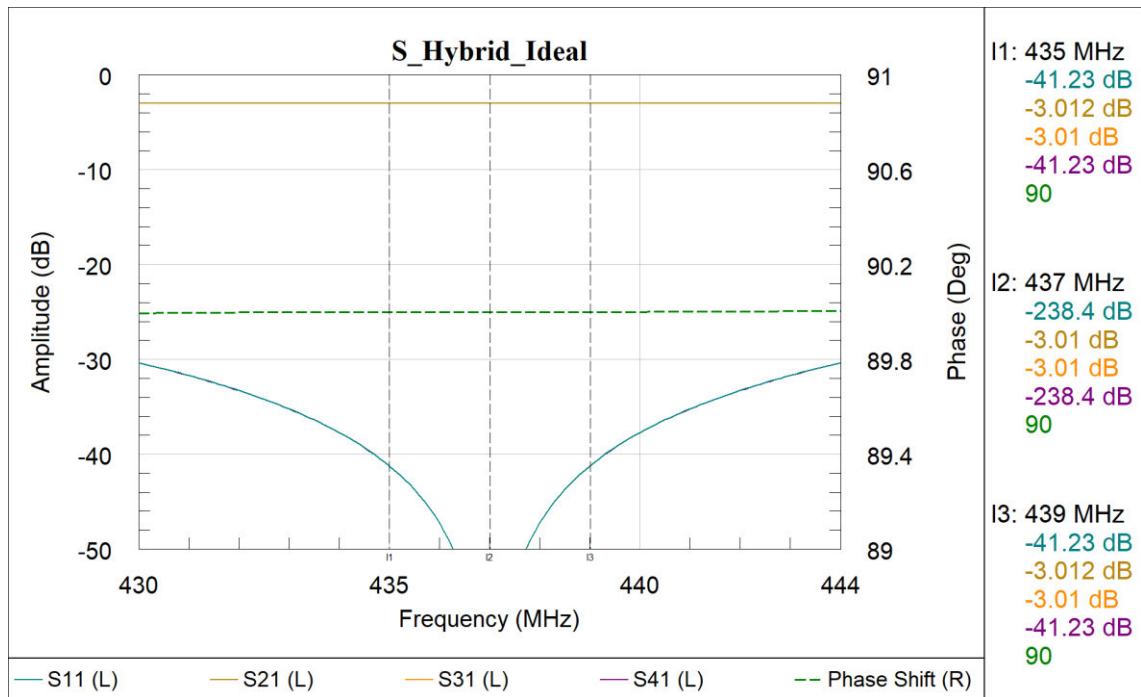


Figure 12: Ideal transmission line 90° hybrid coupler simulation results.

4.2.2 Basic circuit elements circuit result

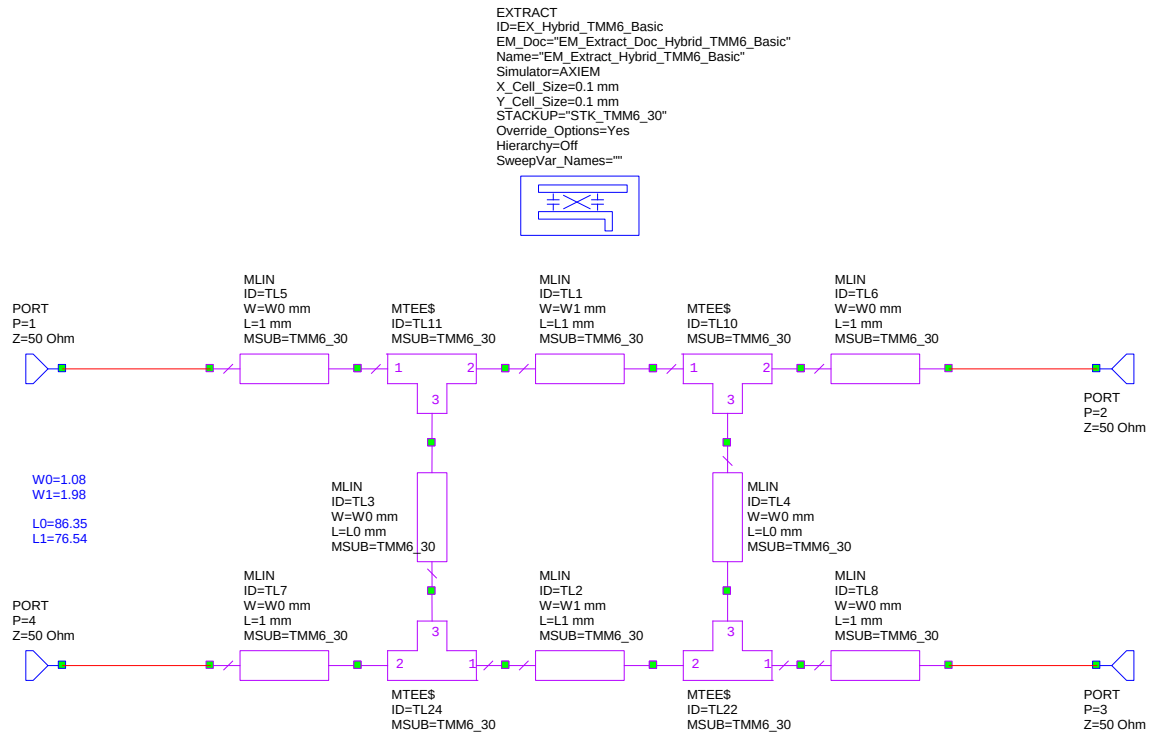


Figure 13: Basic circuit elements 90° hybrid coupler EM schematic.

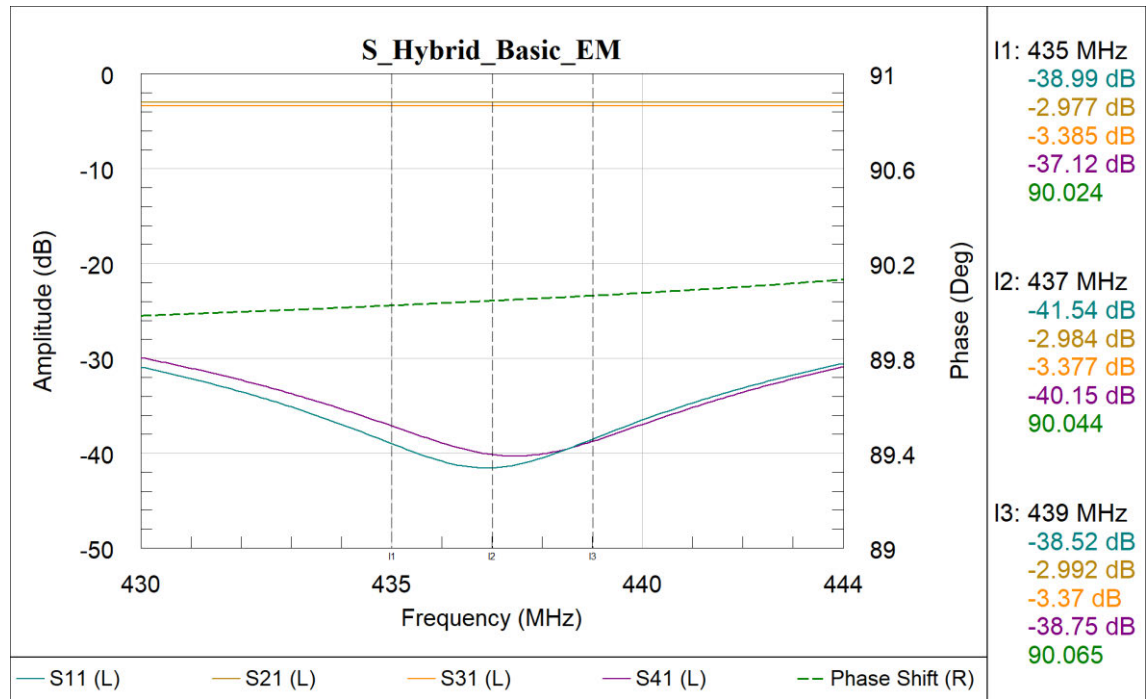


Figure 14: Basic circuit elements 90° hybrid coupler EM simulation results.

4.2.4 Circuit elements circuit and final meander

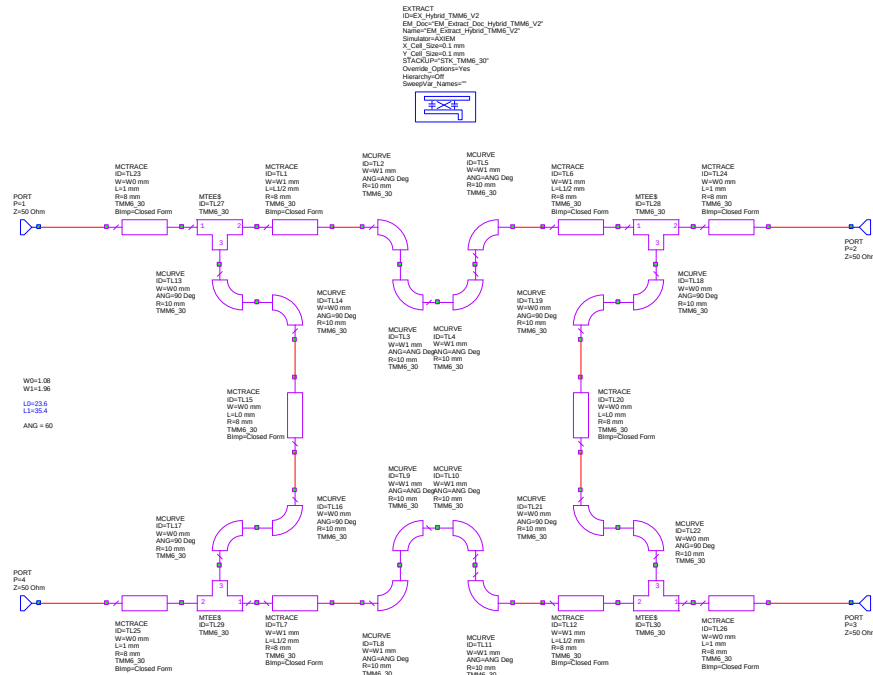


Figure 17: Circuit elements and final meander 90° hybrid coupler EM schematic.

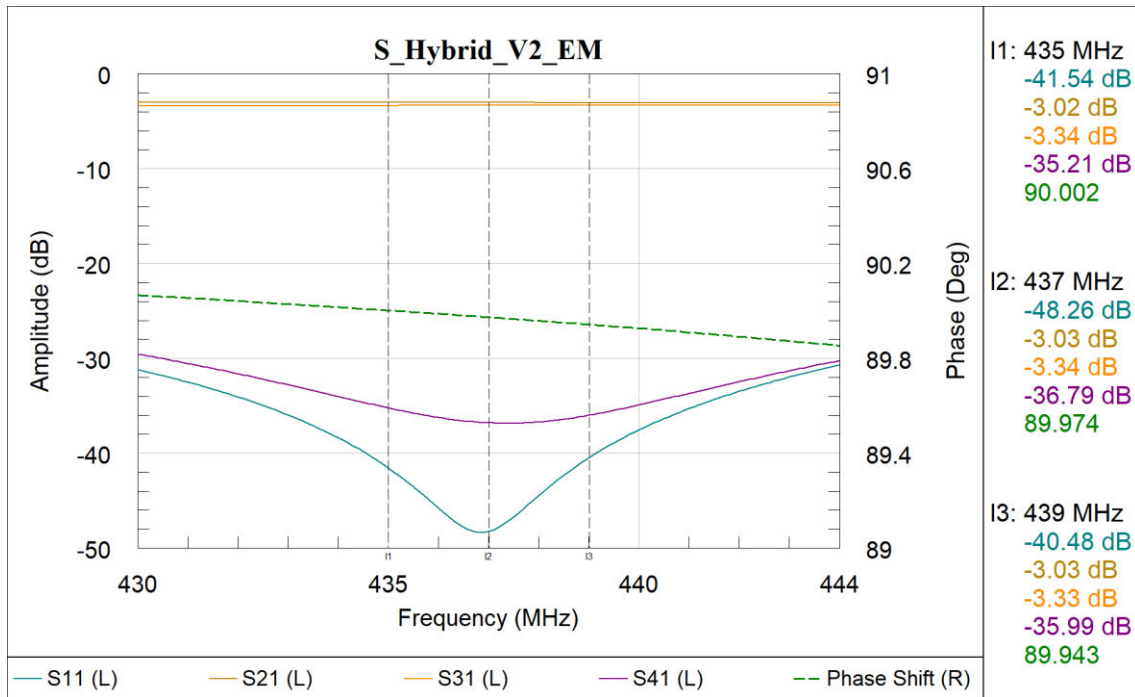


Figure 18: Circuit elements and final meander 90° hybrid coupler EM simulation results.

CST results

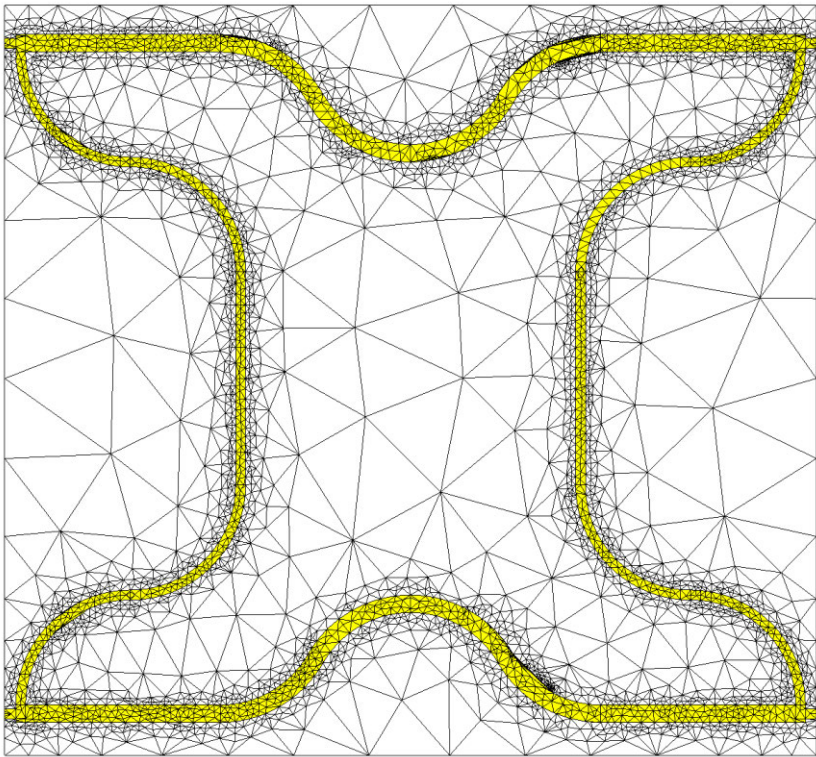


Figure 21: Circuit elements and final meander on 0.635mm substrate 90° hybrid coupler in CST, mesh.

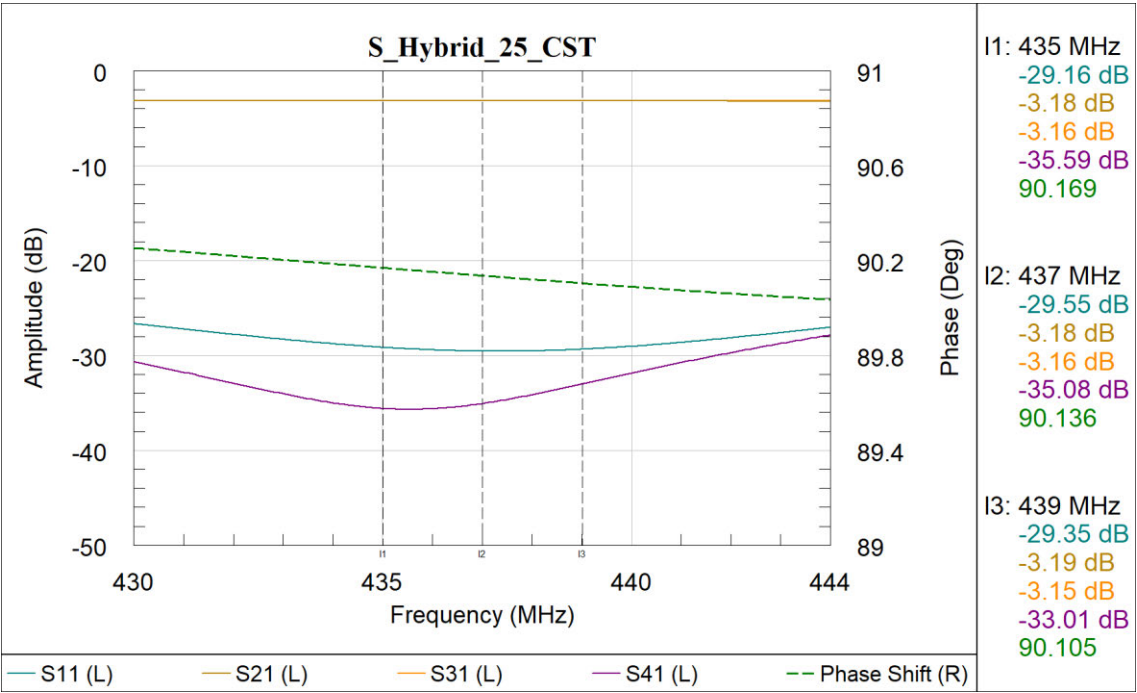


Figure 22: Circuit elements and final meander on 0.635mm substrate 90° hybrid coupler CST simulation results.

4.3 Design and simulation of the balun

4.3.1 Ideal lumped element circuit

We start the design by creating a circuit in AWR using capacitors CAP and inductors IND, whose value we obtain through optimization, as shown in Figure 23 and Figure 24. This design is done primarily to illustrate, with the minimum clutter possible, an LC narrowband voltage transformer, in this case with a N of 2:1.

This basic design is followed throughout the lumped element design process.

4.3.2 Characterized lumped element circuit

We proceed with the initial design of the lumped elements circuit, including characterized LC elements instead of the ideal ones, as shown in Figure 25.

For this design we are using MLCCs [8]. When selecting MLCCs for radiofrequency use, we need to consider a series of basic parameters:

The first one is *packaging*, which indicates the form factor of the device, being in most of the cases SMDs, they are often referenced by their horizontal size in inches. Without exploring more exotic packages, they usually come in sizes ranging from 0201 to 1206. Smaller packages tend to have lower parasitic effects at the cost of capacitance and power rating.

The second one is the *ceramic temperature coefficient*, which indicates the thermal stability of the substrate. There are three classes: *Class I*, *Class II*, and *Class III*, which, in turn, contain an alphanumeric code indicating the concrete substrate within each class. Class I considered “ultra-stable” have low associated capacitance values, whilst Class III substrates are very instable but have extremely high associated capacitance values, Class II falls somewhere in between.

The third one is *Q factor*, which indicates the relationship between reactance and resistance at a given frequency. Higher Q factors indicate less losses while lower Q factors, higher losses.

The fourth one is *tolerance*, which indicates the variation of capacitance values allowed by the capacitor. These are often referenced as capacitance values when working with a low capacitance MLCC, or a percentage of the overall value when working with higher capacitance MLCCs.

For this design we make use of 2 GQM1875C2E5R0BB12 MLCCs. This is a high Q 0603 Class I COG/NPO MLCC capacitor with an overall capacitance of 5.0pF and a tolerance of 0.1pF. Additionally, this device fully complies with the military extended operating temperature range of -55°C to 125°C.

For this design we also use inductors [9]. When selecting inductors for radiofrequency use, we need to consider a series of basic parameters:

The first one is *packaging*, which indicates the form factor of the device, being in most of the cases SMDs, they are often referenced by their horizontal size in inches. Without exploring more exotic packages, they usually come in sizes ranging from 0201 to 1206. Smaller packages tend to have lower parasitic effects at the cost of capacitance and power rating.

The second one is the *lowest self-resonance frequency*, which indicates the first frequency at which the device resonates, changing its inductive properties. Higher values allow for longer stability headroom in the frequency range.

The third one is *Q factor*, which indicates the relationship between inductance and resistance at a given frequency. Higher Q factors indicate less losses while lower Q factors, higher losses.

The fourth one is *tolerance*, which indicates the variation of inductance values allowed by the inductor. These are often referenced as the variation percentage of the overall inductance.

For this design we make use of 2 LQW18AN25NG80MLCC inductors. This is a 0603 inductor with a Q of 40 at 250MHz, lowest self-resonant frequency of 3600MHz, an overall inductance of 25.0nH and a tolerance of 2.0%. Additionally, this device fully complies with the military extended operating temperature range of -55°C to 125°C.

The circuit is simulated using the circuit and, afterwards, the 2.5D AXIEM electromagnetic simulator and further optimized until the design requirements are met.

Provided the topology and results shown in Figure 26 are acceptable we continue with the next step.

4.3.3 Characterized lumped element circuit with meander

In order to connect the device to the output ports, a meander network is added following the previous design, as shown in Figure 27.

This meander network serves two purposes: reach the feeding ports without any introduced significant phase or amplitude unbalance, allow for further tuning in case the feeding network is not optimally impedance matched to either end.

This last is accomplished via the footprint allocation for a PI or L style matching network, only initially populated by a series element, a GQM1875C2E101FB12 MLCC. This is a high Q 0603 Class I COG/NPO MLCC capacitor with an overall capacitance of 100.0pF and a tolerance of 1.0%. Additionally, this device fully complies with the military extended operating temperature range of -55°C to 125°C.

The circuit is simulated using the circuit and, afterwards, the 2.5D AXIEM electromagnetic simulator and further optimized until the design requirements are met.

Provided the topology and results shown in Figure 28 are acceptable we continue with the other design.

4.3.4 COTS transformer circuit

We proceed with the initial design of the COTS transformer circuit, including a characterized transformer and LC elements as shown in Figure 29.

The COTS transformer used in this design is the ADTL1-12+, a core and wire transmission line style transformer with an insertion loss lower than 0.98dB, amplitude unbalance lower than 0.17dB, phase unbalance lower than 1.35° and return loss higher than 9.26dB. A

As the transformer by itself does not provide a sufficiently high return loss, an L style matching network is designed using an LC circuit. Additionally, this device fully complies with the extended commercial operating temperature range of -20°C to 85°C.

This LC circuit is composed of a GQM1875G2E1R0WB12 MLCC. A high Q 0603 Class II X8G MLCC capacitor with an overall capacitance of 1.0pF and a tolerance of 0.05pF. Additionally, this device fully complies with the military extended operating temperature range of -55°C to 125°C.

And an LQW18AN12NG80 inductor. A 0603 inductor with a Q of 37 at 250MHz, lowest self-resonant frequency of 5000MHz, an overall inductance of 12.0nH and a tolerance of 2.0%. Additionally, this device fully complies with the military extended operating temperature range of -55°C to 125°C.

The circuit is simulated using the circuit and, afterwards, the 2.5D AXIEM electromagnetic simulator and further optimized until the design requirements are met.

Provided the topology and results shown in Figure 30 are acceptable we continue with the next step.

4.3.5 COTS transformer circuit and meander

In order to connect the device to the output ports, a meander network is added following the previous design as shown in Figure 31.

This meander network serves two purposes: reach the feeding ports without any introduced significant phase or amplitude unbalance, allow for further tuning in case the feeding network is not optimally impedance matched to either end.

This last is accomplished via the footprint allocation for a PI or L style matching network, only initially populated by a series element, a GQM1875C2E101FB12 MLCC. This is a high Q 0603 Class I COG/NPO MLCC capacitor with an overall capacitance of 100.0pF and a tolerance of

1.0%. Additionally, this device fully complies with the military extended operating temperature range of -55°C to 125°C.

The circuit is simulated using the circuit and, afterwards, the 2.5D AXIEM electromagnetic simulator and further optimized until the design requirements are met.

Provided the topology and results shown in Figure 32 are acceptable, we now proceed to select between both topologies for the use in the final design.

4.4 Simulation results of the balun

4.4.1 Ideal lumped element circuit results

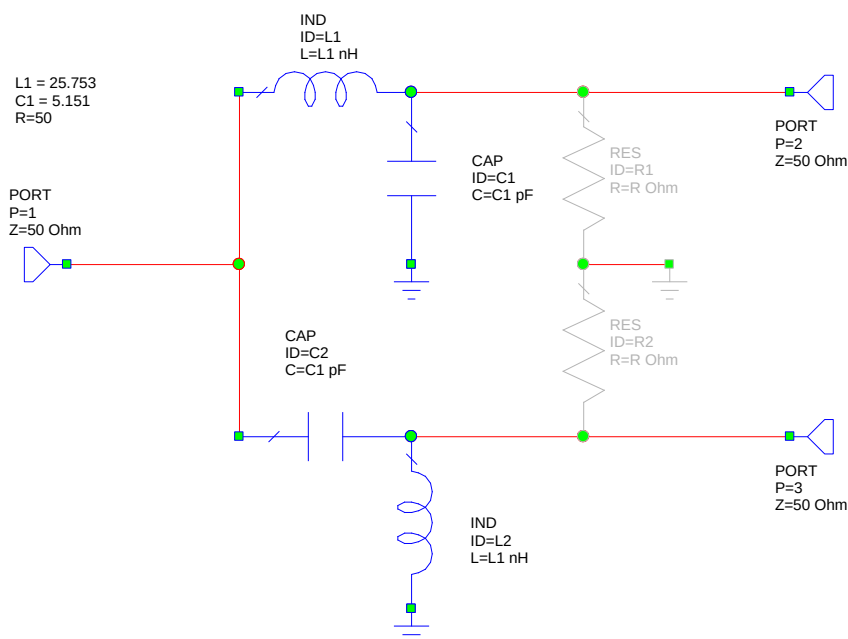


Figure 23: Ideal lumped element balun circuit schematic.

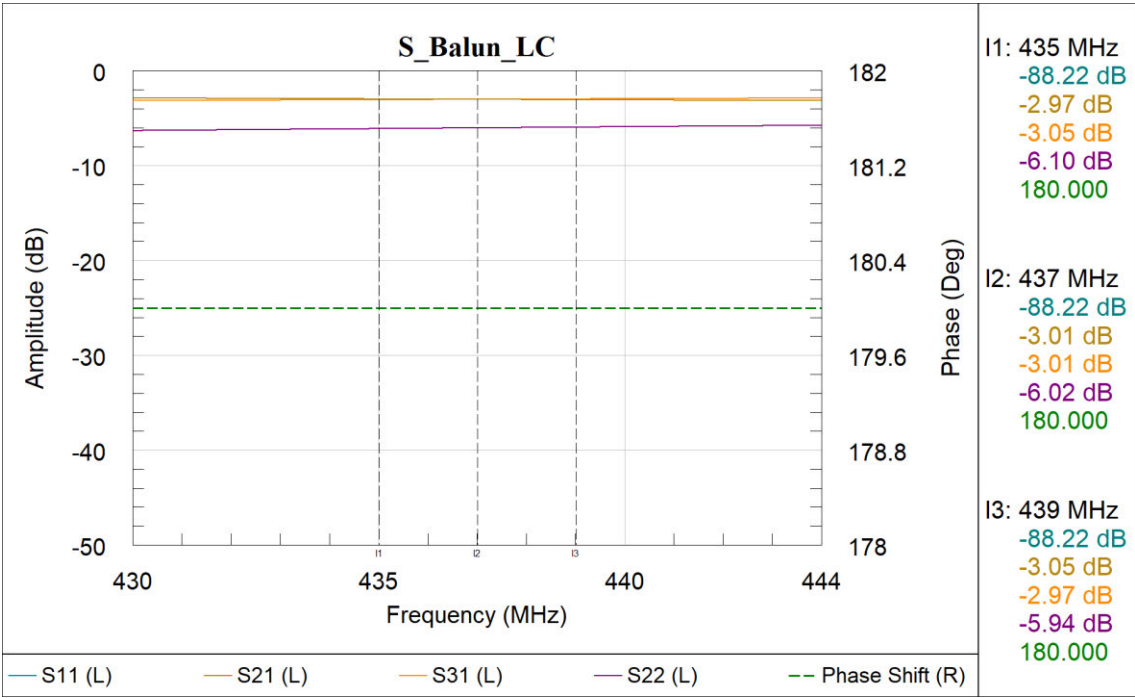


Figure 24: Ideal lumped element balun circuit simulation results.

4.4.2 Characterized lumped element circuit results

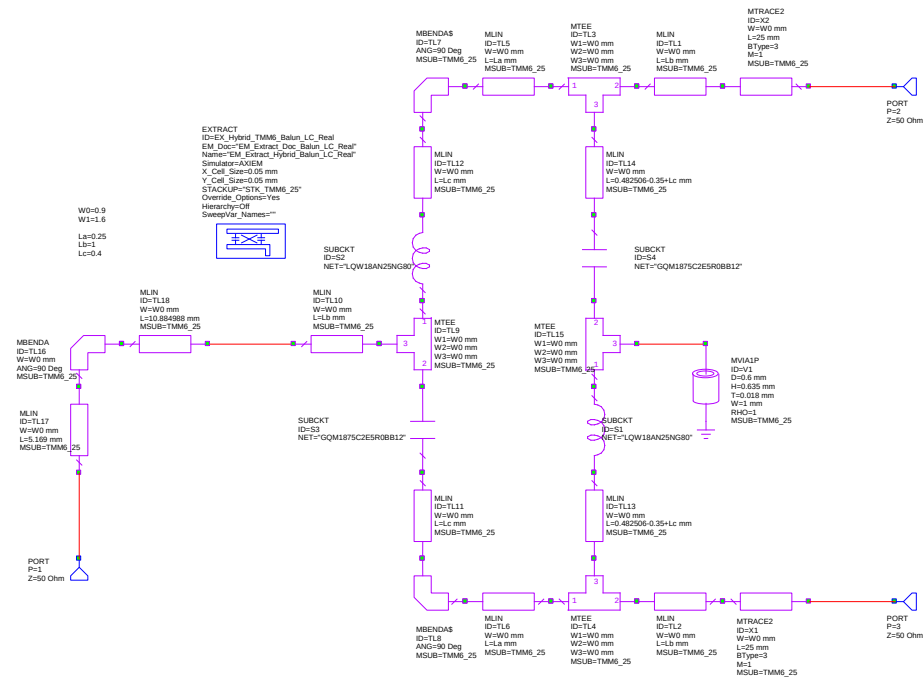


Figure 25: Characterized lumped element balun circuit EM schematic.

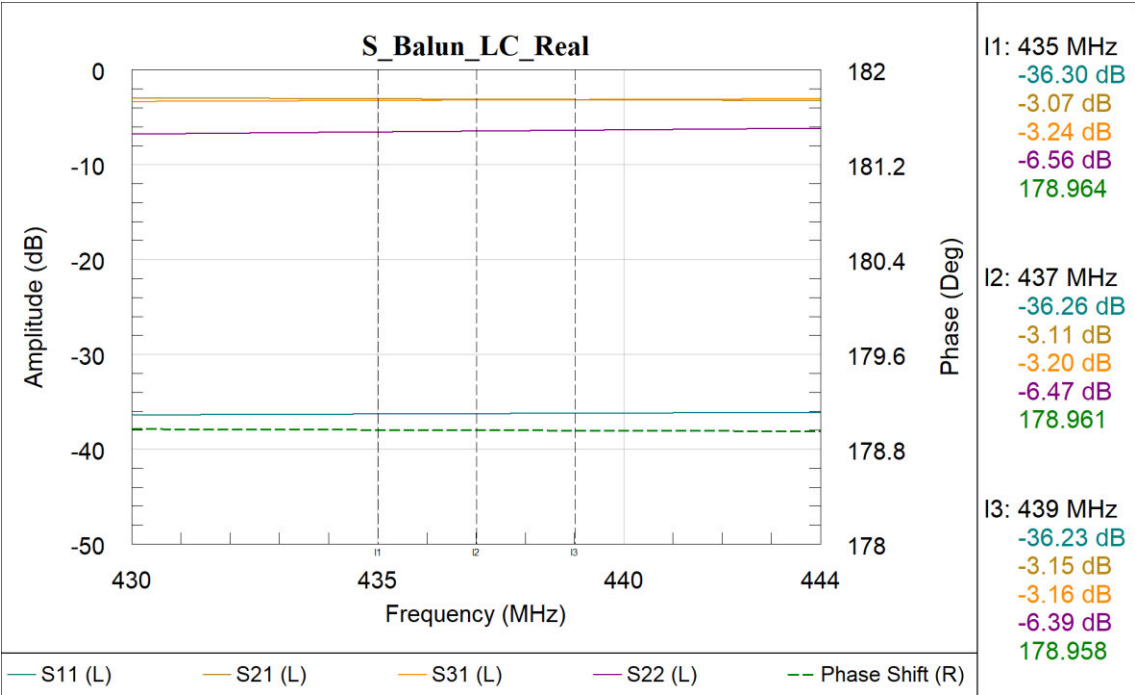


Figure 26: Characterized lumped element balun circuit EM simulation results.

4.4.3 Characterized lumped element circuit with meander results

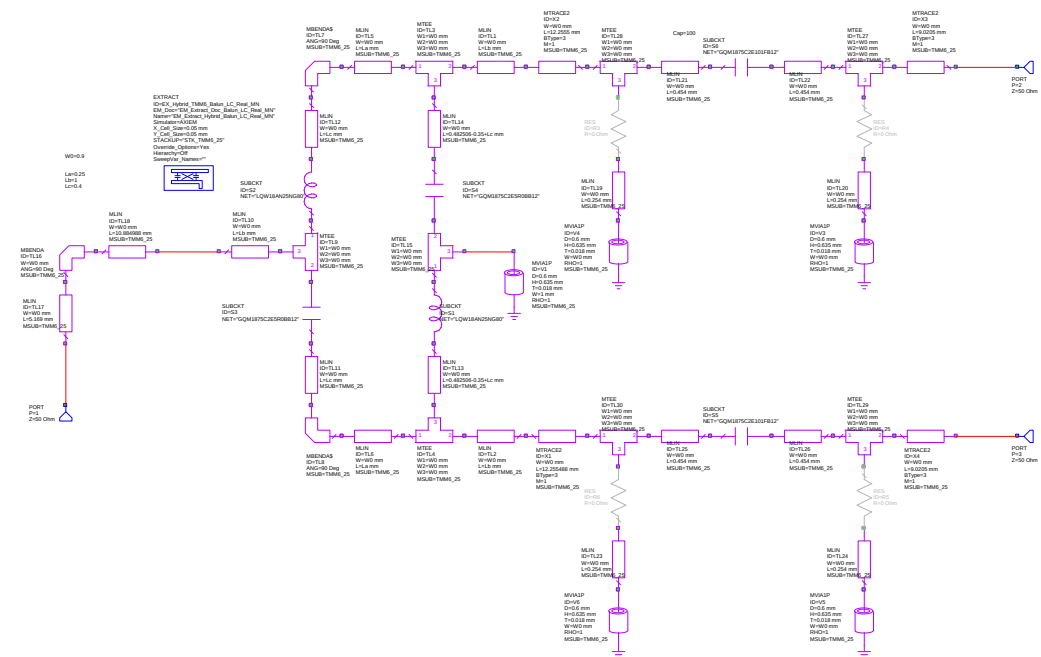


Figure 27: Characterized lumped element balun circuit with meanders EM schematic.

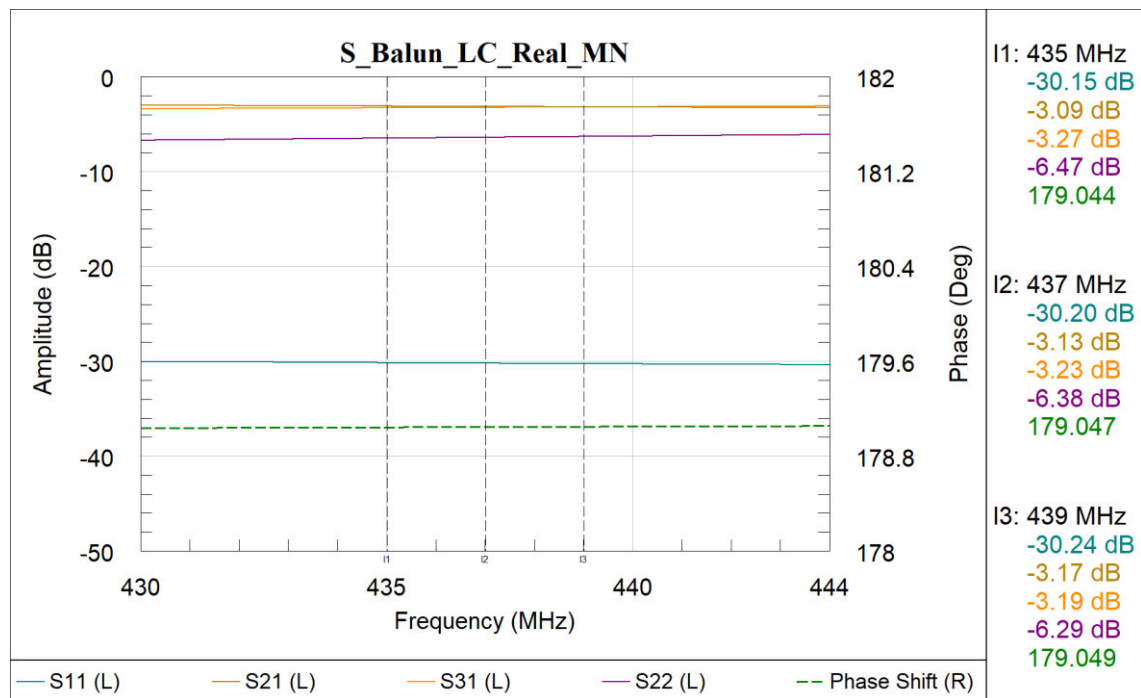


Figure 28: Characterized lumped element balun circuit with meanders EM simulation results.

4.4.4 COTS transformer circuit results

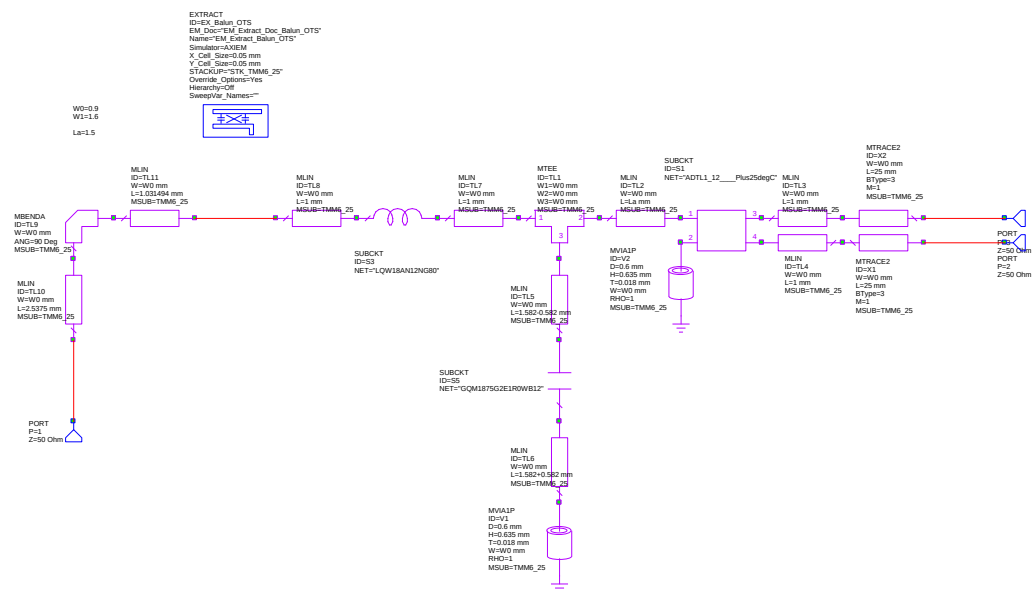


Figure 29: Characterized COTS transformer balun circuit EM schematic.

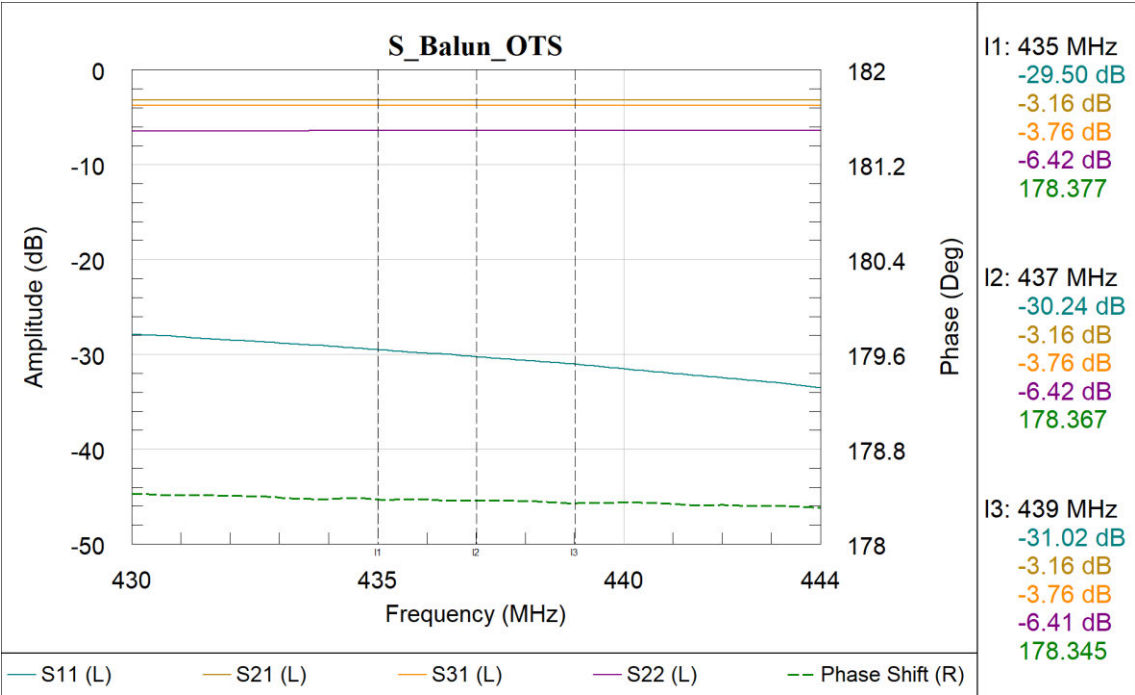


Figure 30: Characterized COTS transformer balun circuit EM simulation results.

4.4.5 COTS transformer circuit and meander results

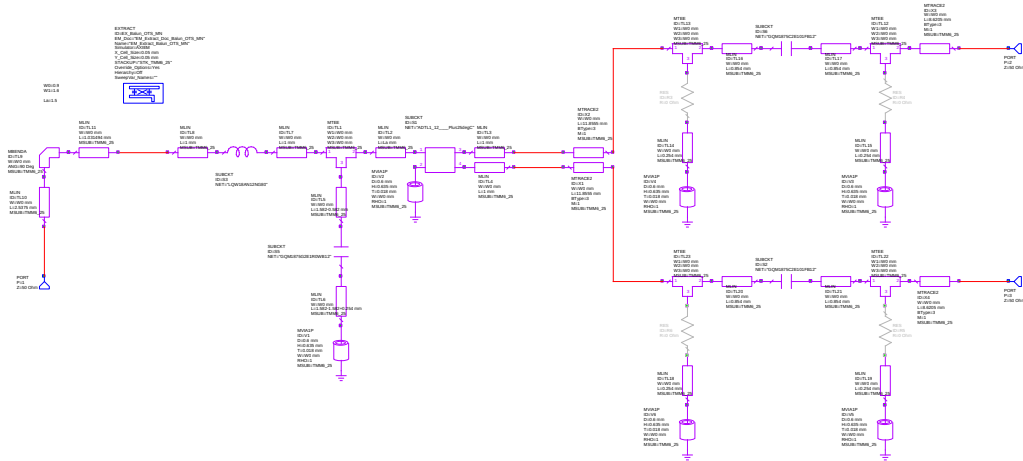


Figure 31: Characterized COTS transformer balun circuit with meanders EM schematic.

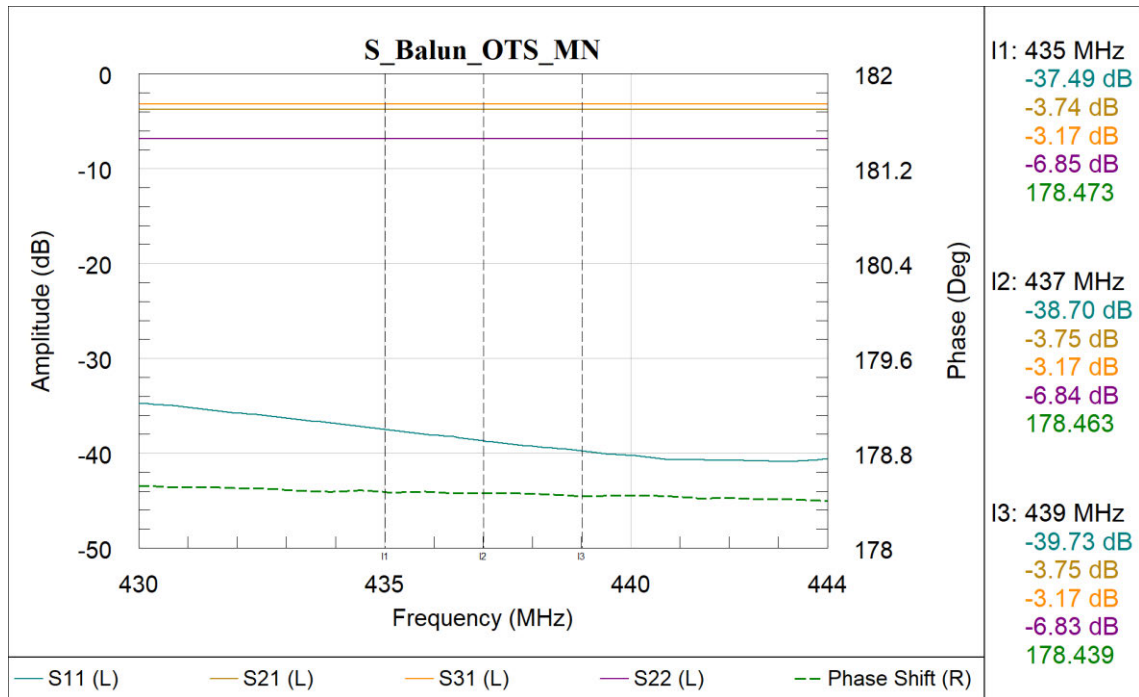


Figure 32: Characterized COTS transformer balun circuit with meanders EM simulation results.

4.5 Design simulation and results of the complete feeding network

In order to proceed with the complete circuit design, a balun topology must be chosen. As a comparison, the simulation results of both topologies are described in the table below:

	435Mhz	437Mhz	439Mhz
phase shift[°]	179.044	179.047	179.049
phase shift[°]	178.463	178.463	178.463
S_{11}[dB]	-30.15	-30.20	-30.24
S_{11}[dB]	-37.49	-38.70	-39.73
S_{21}[dB]	-3.09	-3.13	-3.17
S_{21}[dB]	-3.74	-3.75	-3.75
S_{31}[dB]	-3.27	-3.23	-3.19
S_{31}[dB]	-3.17	-3.17	-3.17

Table 5: Lumped element transformer circuit results in blue and COTS transformer circuit results in orange.

As can be seen, the return loss of the COTS transformer circuit is superior, while the rest of the parameters are similar on either topology, thus, the COTS transformer topology is selected for the complete circuit as shown in Figure 33.

Both the 90° hybrid coupler and balun are connected and simulated together, leading to the creation of the complete feeding network circuit, as shown in Figure 34, Figure 35 and Figure 37.

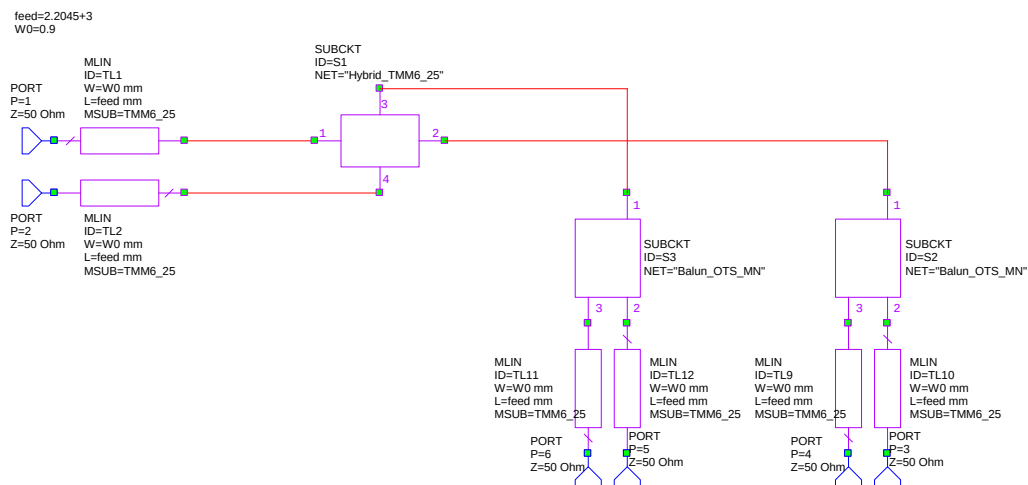


Figure 33: Complete feeding network circuit schematic.

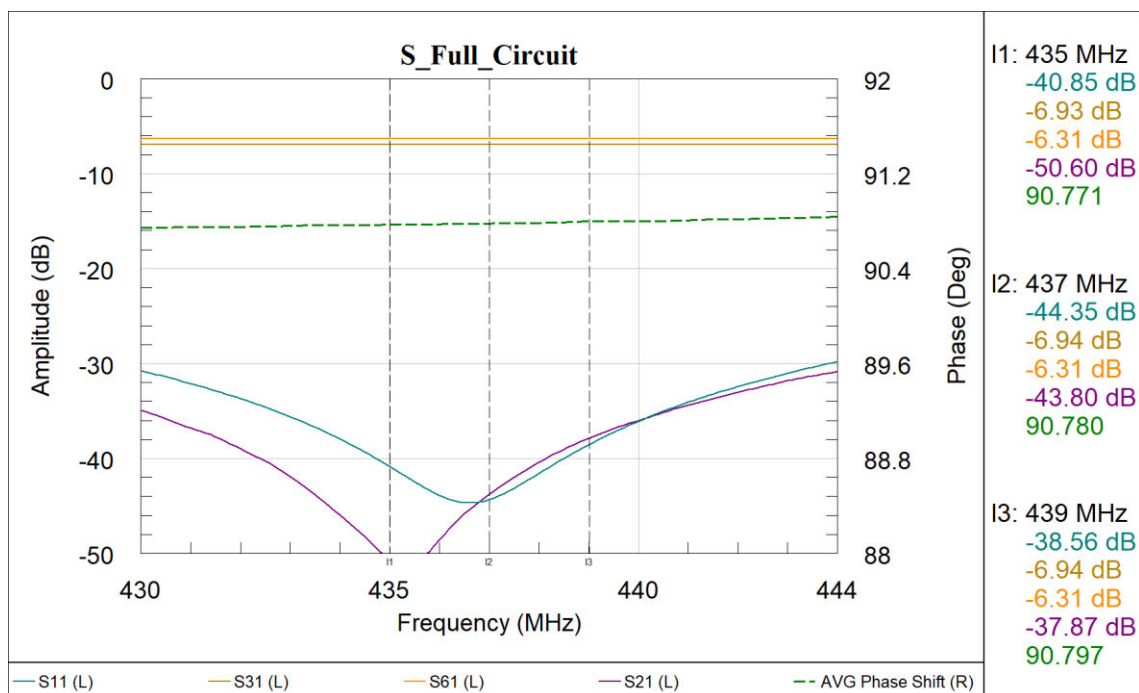


Figure 34: Complete feeding network circuit partial simulation results.

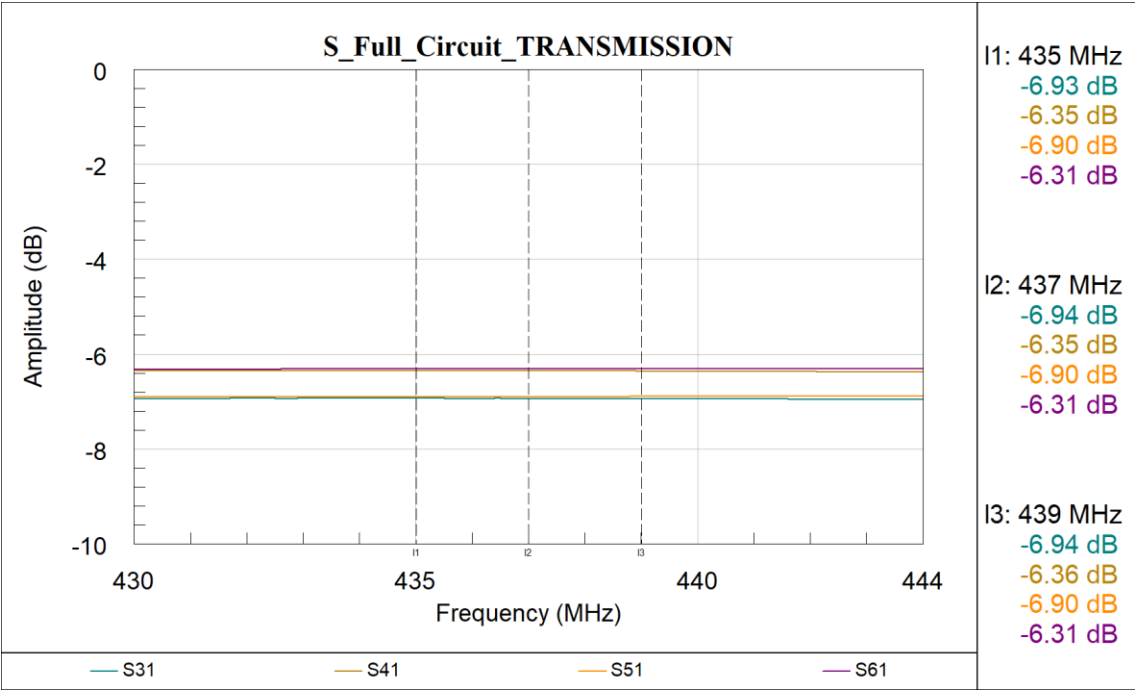


Figure 35: Complete feeding network circuit simulation insertion loss results.

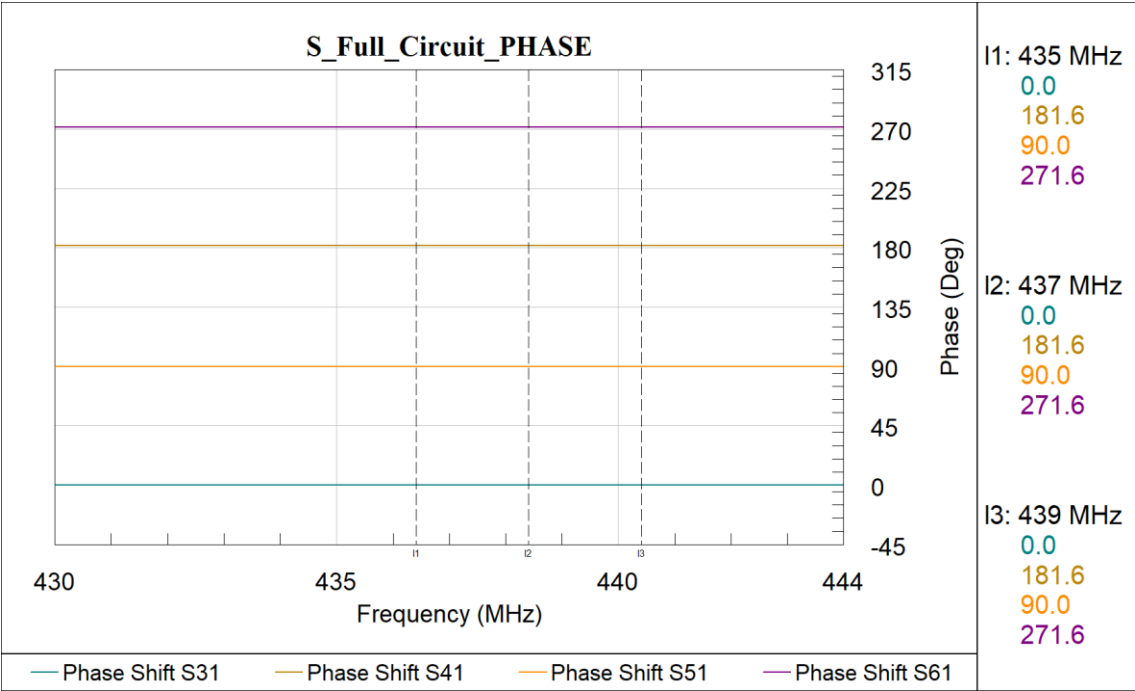


Figure 36: Complete feeding network circuit simulation phase shift results.

4.6 Design and fabrication of the PCB

The previously simulated circuits are exported as DXF. These DXF files are cleaned in Autodesk Fusion360 prior to introducing them in Altium Designer. In Altium Designer the circuits are converted to PCBs where via fences and port footprints are added. These circuits include the 90° hybrid coupler, the balun, and the complete circuit. A total of four PCBs are designed for testing and, eventually, deployment:

The first PCB includes only the 90° hybrid coupler, and the second one includes only the balun, these are intended to evaluate the performance of the devices by themselves. A top facing PCB illustration of either device extracted from Altium is shown in Figure 37.

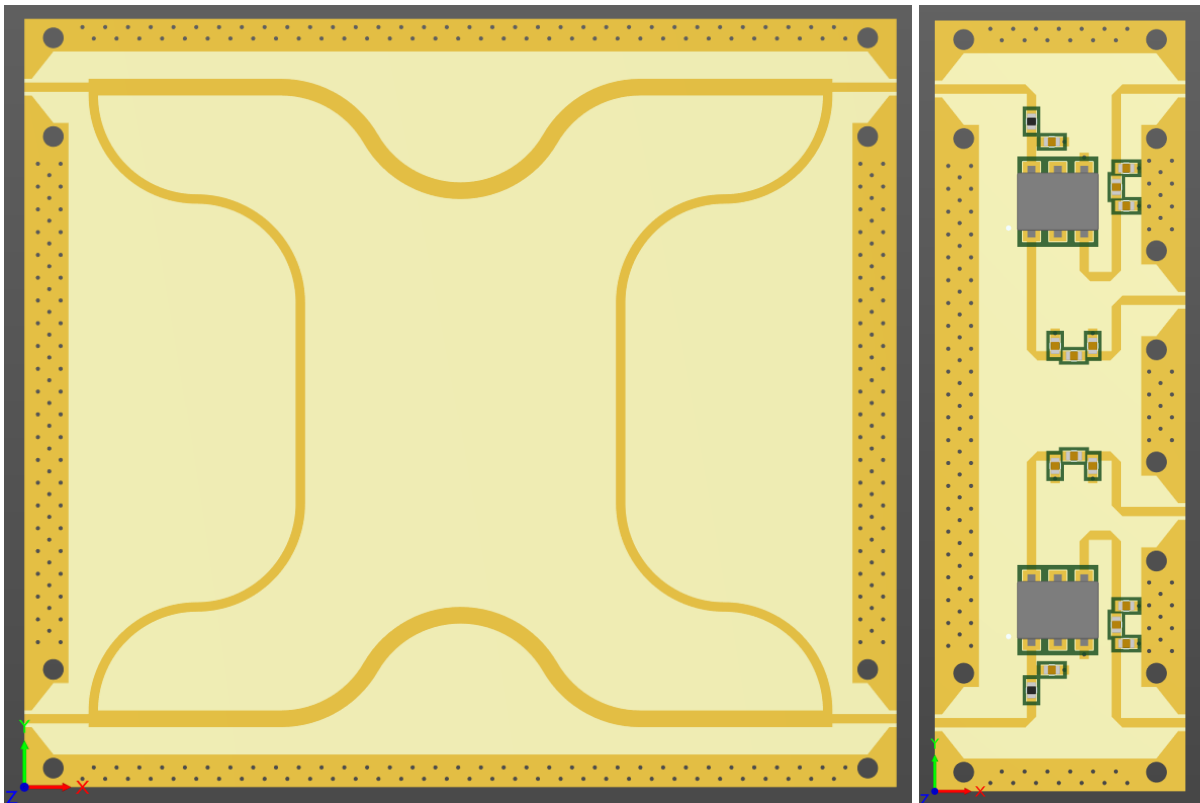


Figure 37: 90° hybrid coupler PCB top (left) and balun PCB top (left) as shown in Altium Designer

Care is taken to minimize the application of solder masks in order to prevent potential outgassing. On the top layer, solder mask is only applied around the pads where component may be soldered, minimizing the chances of solder overflowing outside the intended areas.

A third PCB includes a simple transmission line intended for characterization of the connector and calibration. A top facing PCB illustration of the device extracted from Altium is shown in Figure 38.



Figure 38: transmission line PCB top as shown in Altium Designer

The fourth PCB includes the complete 90° hybrid coupler and balun circuit. This PCB contains four mounting holes with a diameter of 3.2mm as per design requirements. A top and bottom facing PCB illustration of the device extracted from Altium are shown in Figure 39.

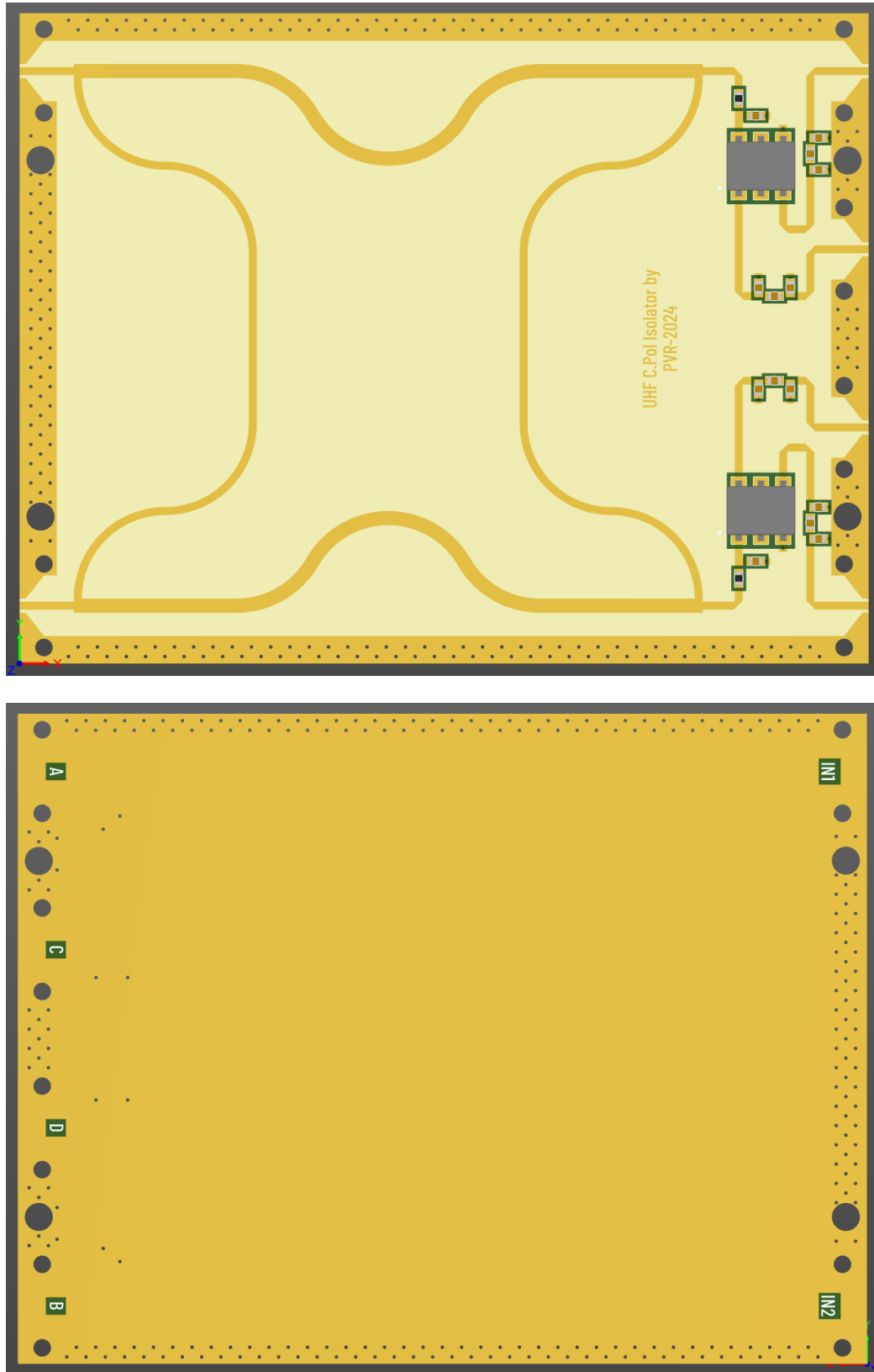


Figure 39: Complete feeding network circuit PCB top (top) and bottom (bottom) as shown in Altium Designer

Due to the negative to using soldered coaxial connectors, screwed port launch connectors are employed. This is itself an issue, since soldered connectors provide in most cases better return loss than screwed ones, and in order to maintain the 32dB of isolation a return loss better than 26dB is required. The required coaxial connector as per specification has to be SMA terminated, so the Southwest Microwave 292-04A-6 connectors are chosen for the design, providing a return loss greater than 26.4dB from DC to past 5000MHz.

In order to find a suitable PCB manufacturer or fabrication house, fifteen are contacted and requested a budget for: 2 90° hybrid coupler circuits, two balun circuits and two complete circuits. All the budgets request specified the following:

TMM-6 substrate material with a thickness of 0.635mm.

Two layers of 18um electrodeposited copper.

ENIG or ENEPIG surface finish.

Green solder mask.

White silkscreen.

IPC 6012 ES Addendum compliant if possible.

Follow IPC-1601 bake-off procedures if possible.

Overall, the material resulted complicated to obtain and significantly more expensive than others, similar in characteristics. Five fabrication houses did not respond to the requests, as summarized in Table 6, none of the indicated prices include IVA or VAT.

Fabrication House	Cost	Remarks
ElectronPCB	N/A	Did not respond
Tedesia	N/A	Did not respond
CISA	N/A	Did not respond
Elate	N/A	Did not respond
PCBWay	14605.92 USD	ENEPIG finish, IPC 6012 ES, Bake-off
IPCB	5888.00 USD	ENEPIG finish, IPC-A-600G Class II
RayMing	4850.00 USD	ENEPIG, IPC-600F Class III
CreativeHiTech	N/A	Did not respond
Canmtech	N/A	Only work RO4003C or RO4350B
Boyunfa Tech	N/A	Only work RO6006
HiTechPCB	6475.00 USD	ENIG, IPC-6012 Class 2
Micron-20	N/A	Only work 4003C
Lab Circuits	4626.24 EUR	ENIG, IPC-6012 Class 3A, Bake-off
McFly	2441.08 EUR	ENIG, IPC-6012 Class 3A, Bake-off

Table 6: List of contacted fabrication houses including provided manufacturing cost and remarks.

Lab Circuits is selected as the manufacturer. They work with standard sized panels, so the cost will be the same as long as the entirety of the PCBs can fit in a standard panel. TMM6 substrates come in panels of 18x12in or 18x24in panels as a standard, so a panelization design for 18x12in is done, as shown in Figure 40.

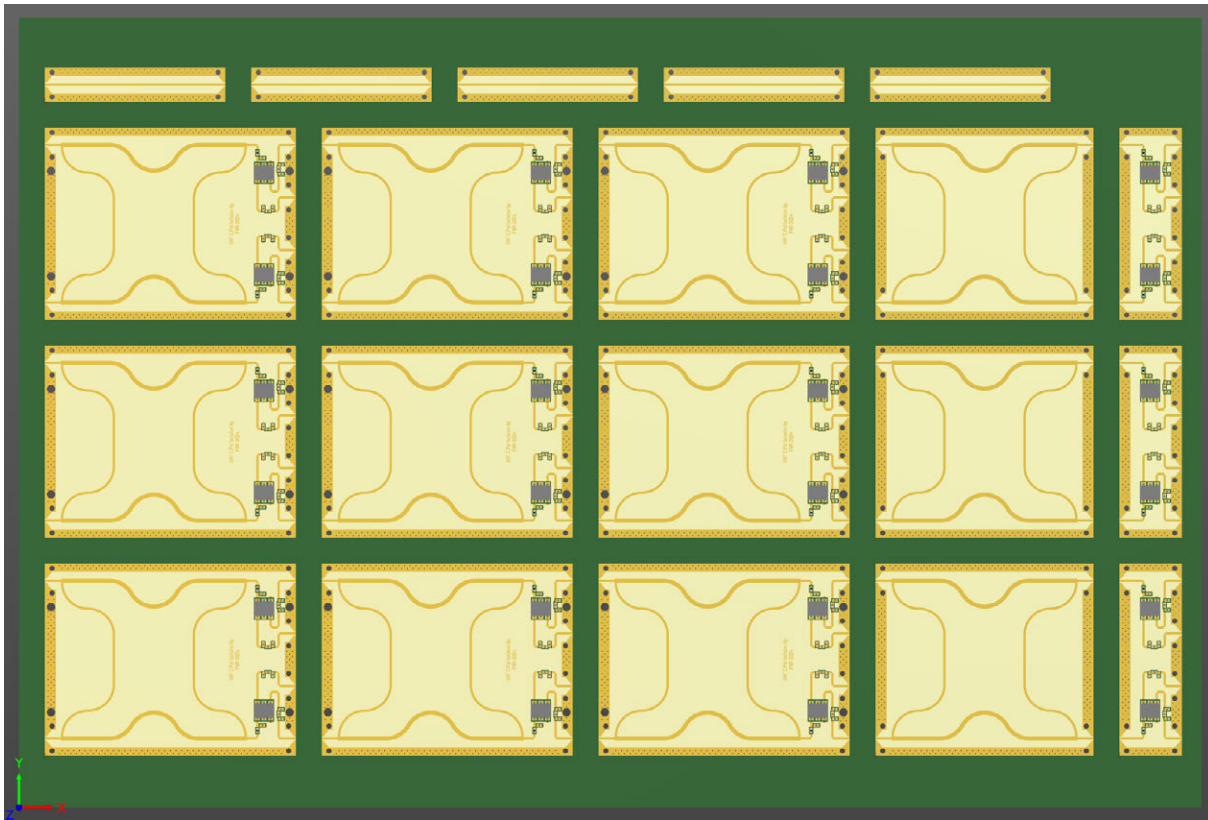


Figure 40: Panelized PCBs in a 12x18in panel as shown in Altium Designer

This includes five transmission lines, 3 90° hybrid couplers, three baluns and nine complete circuits. The specifications sent alongside the gerber to the PCB manufacturer can be seen “as is” in A.1, additionally, the component BOM in A.2.

At least one fully assembled complete circuit will be evaluated for outgassing in a TVAC and mechanical vibrations. At least one of the fully assembled complete circuits will be reserved for flight.

Additionally, circuits intended for initial testing are fabricated using 6x6in panel samples provided by Rogers Corporation. These circuits include a 90° hybrid coupler circuit and a balun circuit. The fabrication of the aforementioned is provided by the GR SSR-UPM via the use of an LPKF CNC routing machine.

4.7 Current antenna system description, simulation and analysis

The UPMSat-3 antenna system is made up of two subsystems, one for the UHF band and another for the VHF band.

The UHF system comprises four identical monopoles aligned with the Y axis protruding the topmost Z-X face of the craft. The monopoles are excited with a clockwise and

counterclockwise phase shift of ninety electrical degrees between elements and are laid out in a rectangular array in the Z-X plane. The antennas are designed to operate at 437MHz and present a quasi-omnidirectional radiation pattern. The length of the antenna element is 310mm. The far fields are shown in Figure 41.

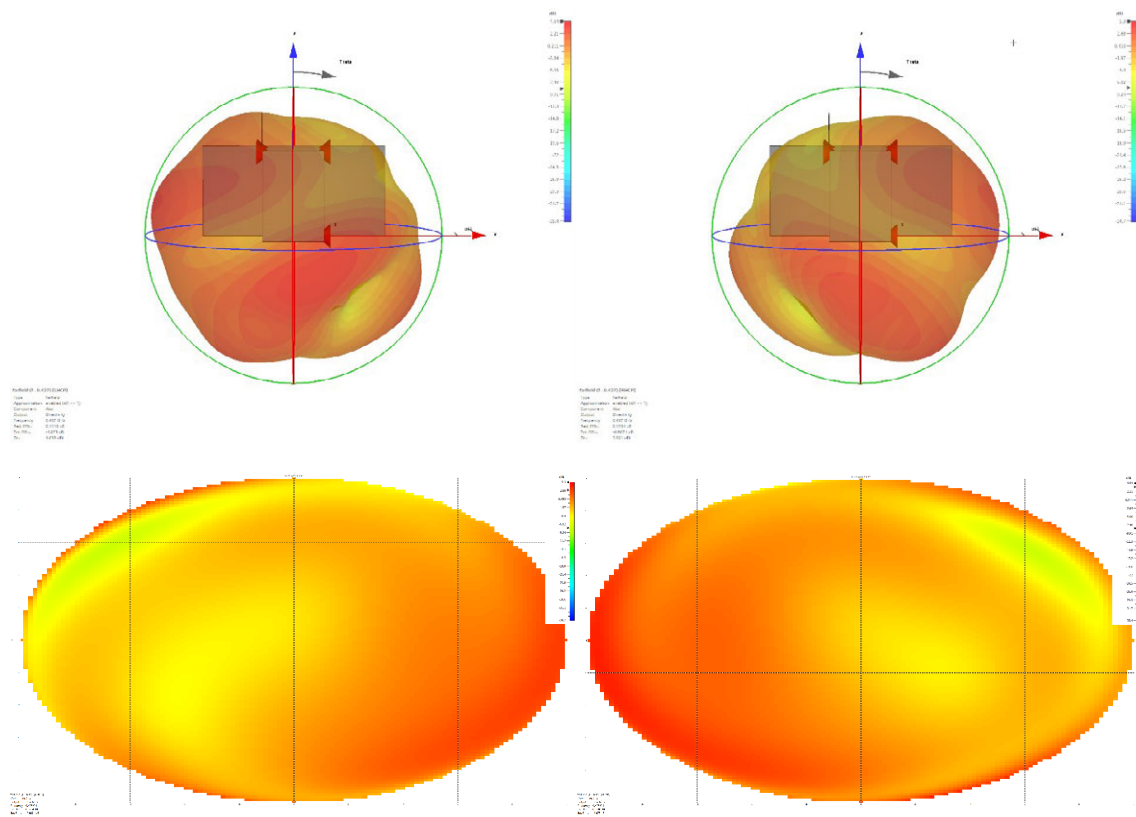
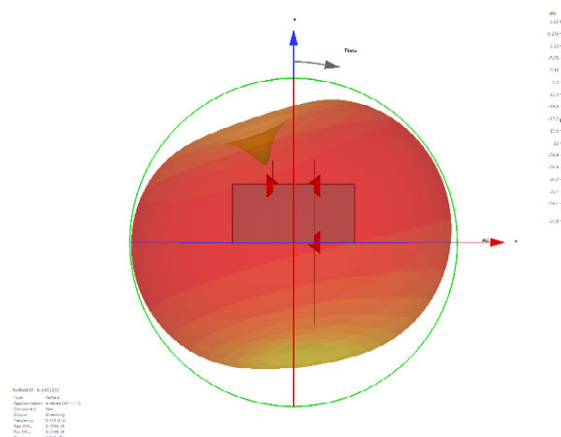


Figure 41: 3D plot (top) and orthographic projection (bottom) of the UHF antennas radiation pattern. Right hand circular polarization (right) and left-hand circular polarization (left).

The VHF system is made up of a single monopole element aligned with the Y axis and protruding the lowermost Z-X face of the craft. The antenna is designed to operate at 145MHz and present a quasi-omnidirectional radiation pattern. The length of the antenna element is 500mm. The far fields are shown in Figure 42.



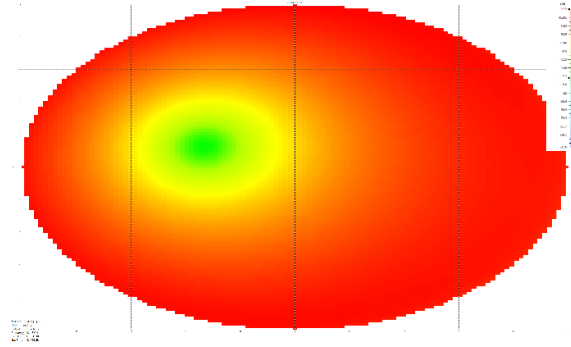


Figure 42: 3D plot (top) and orthographic projection (bottom) of the VHF antenna radiation pattern.

The current UHF antenna layout is inherently incompatible with a circular polarization due to the parallel distribution of the elements when faced from the Y-Z face of the craft. This also leads to strong coupling among the antenna elements.

4.8 Antenna and feeding network upgrades

In order to alleviate this issue, a new UHF antenna design is proposed, simulated and mockups are fabricated. The antenna array itself comprises four identical sequentially rotated IFA antennas laid out on a square perimeter as described by R. Fagnier et al. [10]. The monopoles rest on a conductive ground plane which also mounts the feeding network. The external dimensions of the antenna are 150x150x38mm allowing for its positioning on either face of the UPMSat-3. The far fields are shown in Figure 45.

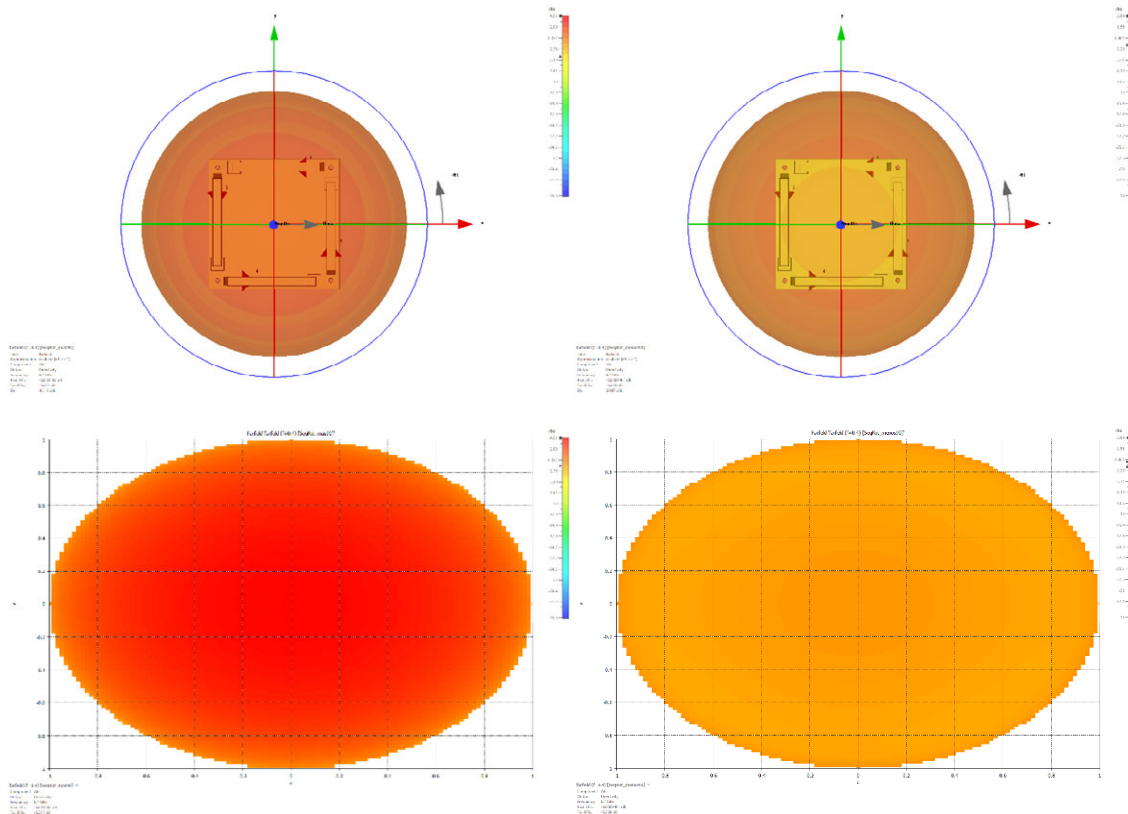


Figure 43: 3D plot (top) and orthographic projection (bottom) of the proposed UHF antennas radiation pattern. Right hand circular polarization (right) and left-hand circular polarization (left).

Two mockups of this antenna design are fabricated, the first one using conventional CNC techniques in 6061 aluminum and the second one using FDM 3D printing and a conductive coating.

The aluminum antenna is fabricated by JLCPCB using a 5-axis CNC machine with a final weight of one hundred ninety-one grams. The 3D printed antenna is fabricated using a Bambu Labs P1S printer acquired by the student association AETEL in Smartfill PLA and coated with a Kontakt Chemie EMI 35 conductive aerosol spray. The final antenna weighs seventy-one grams.

The newly proposed feeding network consists of a 90° hybrid coupler followed by two Wilkinson power dividers, a crossover, and an uneven length transmission line to create a 180° phase shift. All of the above created using microstrip technologies except for the crossover, which would be a surface mounted X2BS crossover manufactured by Xinger; this would be done to minimize the total size of the circuit, as shown in Figure 44.

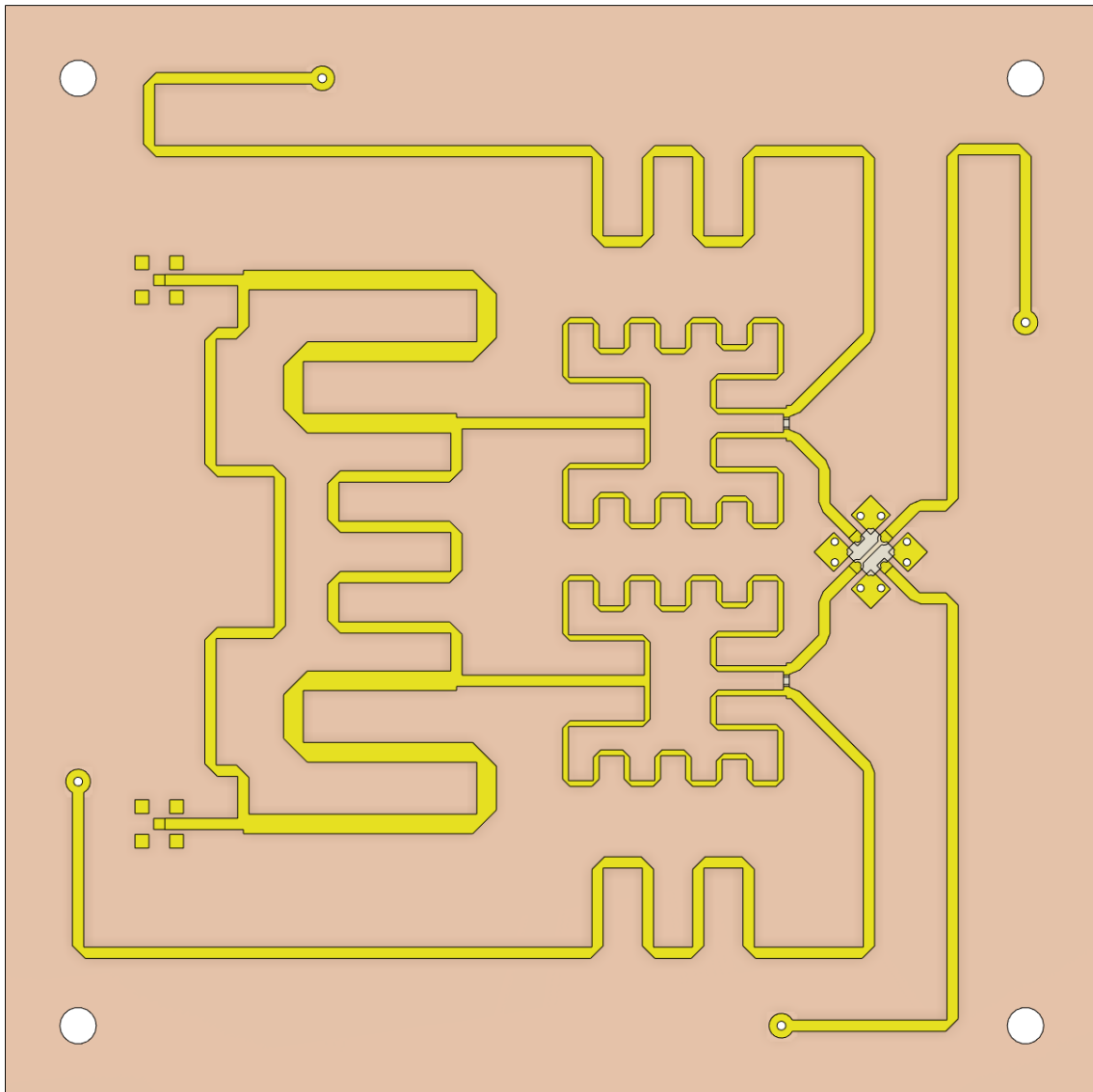


Figure 44: Proposed improved feeding network complete circuit.

5. Results

At the time of this writing the PCBs requested for manufacture at Lab Circuits and specified connectors have not yet been acquired. An accurate characterization of the device is deemed improbable due to:

The only PCBs manufactured are the ones routed by the LPKF CNC which exhibits differences in the manufacturing process when compared with the requested.

The connectors available do not match mounting technique or specifications with the requested.

Nevertheless, the results of the prototypes are considered to be significantly indicative of the performance of the definitive device.

All measurements are performed on a four port Keysight E5080B vector network analyzer.

The prototype 90° hybrid coupler shows correlation between the fabricated and simulated models and complies with the required 32dB of isolation at 437MHz as shown in Figure 45. The measured insertion loss and coupling remain within two hundred of a decibel of the simulated results. The phase shift between the output port and the coupled port increases by about point six of a degree when compared to the simulation. The center frequency shifts about 5MHz. These differences are considered acceptable for a prototype and are attributed to the fabrication process and lack of specified connectors.

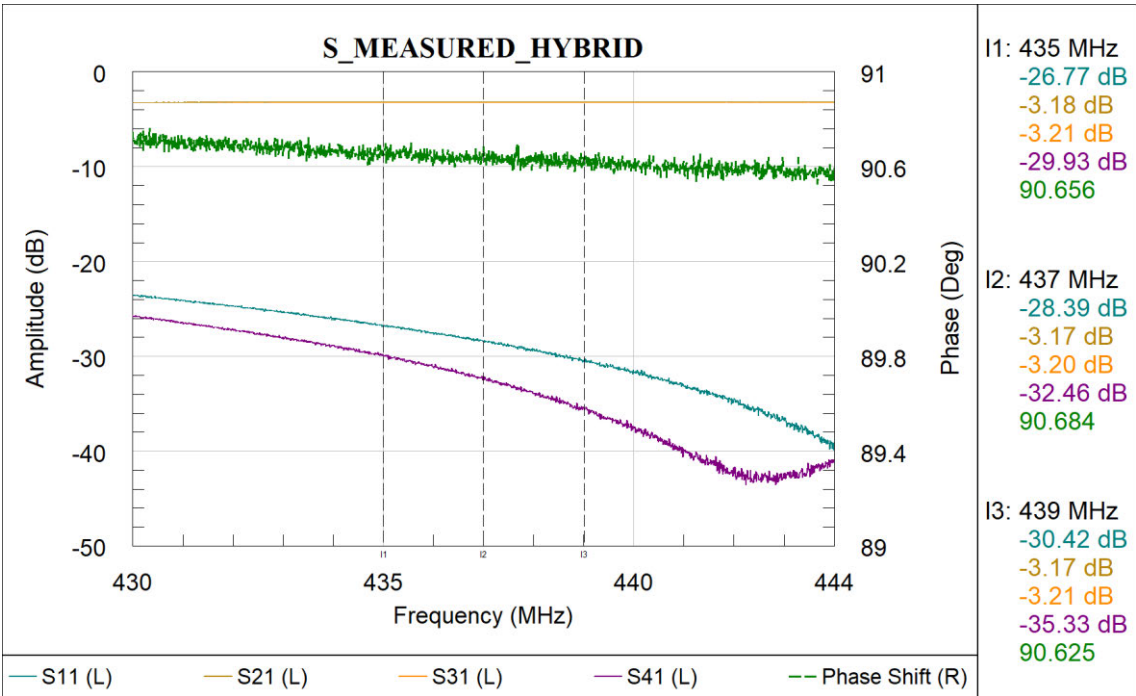


Figure 45: Measurement results of the 90° hybrid coupler.

The complete circuit prototype shows correlation among the fabricated and simulated models but fails to comply with the required 32dB of isolation at 437MHz as shown in Figure 46. This is thought to be a combination of the aforementioned issues and having to solder a wire from

two components to the nearest ground on either balun, since no vias could be placed on the prototype. To measure the complete circuit, two 50Ohm calibration loads are placed on both output ports of either balun in order to obtain coherent measurements from the rest of the ports. Two sets of results are obtained, alternating the terminated baluns as shown in Figure 46.

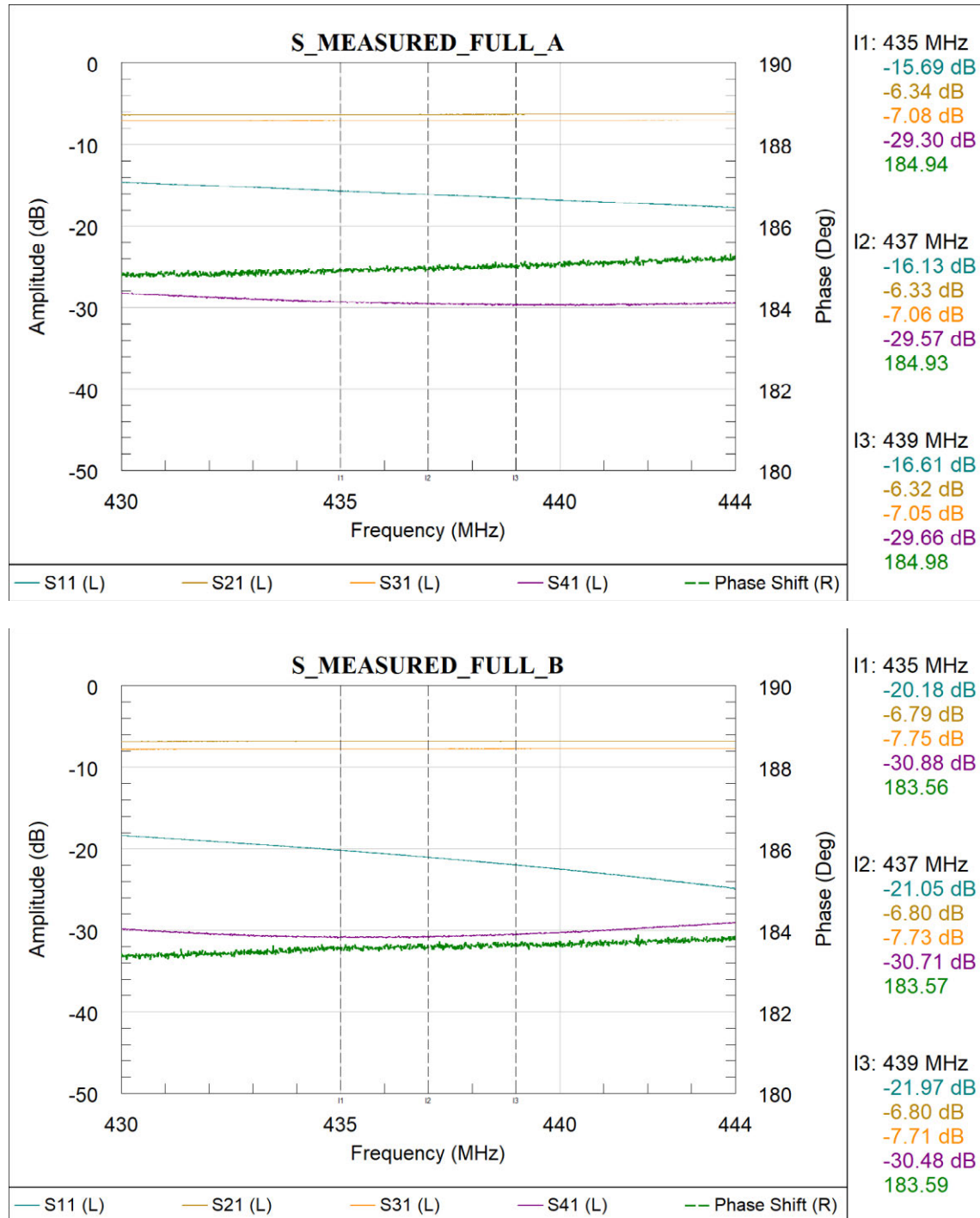


Figure 46: Measurement results of the complete circuit prototype, set A (top), set B (bottom).

The insertion loss remains similar between both samples but is higher than the simulated by over less than 1dB in the worst case as shown in Figure 47. The amplitude imbalance closely matches that of the simulation in both cases. The return loss varies considerably between both cases but remains acceptable at over 16dB, without matching the simulation. The isolation

Results

averages around 30dB which falls 2dB short of the requirements, loosely matching that of the simulation. The phase shift between both output ports increases by about 2-4° when compared to the simulation results as shown in Figure 47.

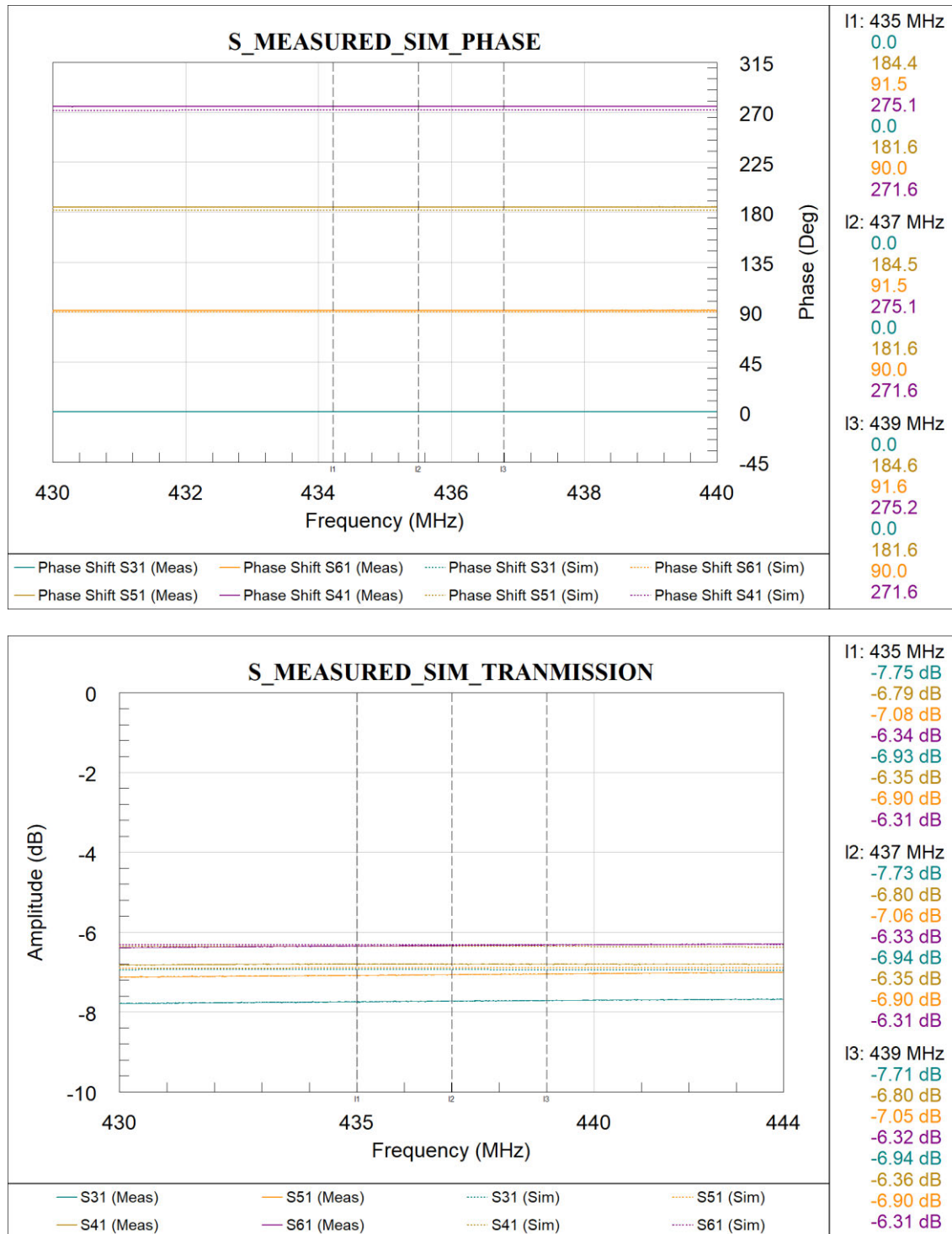


Figure 47: Measurement results of the complete circuit prototype, phase shifts (top), insertion loss (bottom).

Overall, the prototype measurement results are deemed acceptable for this stage and indicate that the design is valid.

6. Budget

The budget of the project can be seen in Table 7. Considering the following:

The student worked three hundred hours at an hourly rate of four comma ninety-seven euros per hour.

The professor worked twenty hours at an hourly rate of one hundred euros per hour.

A AWR Microwave Office license was bought for four months at a cost of fifty thousand USD per year plus IVA.

A CST Studio Suite License was bought for four months at a cost of eighty thousand USD per year plus IVA.

A Fusion360 license was bought for four months at the cost of eighty-five USD per month plus IVA.

An Altium Designer license was bought for four months at the cost of three hundred fifty-eight USD per month plus IVA.

To convert the currency to a single unit, a conversion rate of point ninety-three to one is applied to convert from USD to euro.

	Cost	Time
Student Salary	1491.00€	300h
Professor Salary	2000.00€	20h
Total Salary	3491.00€	

Circuit Components	376.55€	N/A
PCB Manufacturing	4626.24€ + IVA	N/A
Total Materials	5.002,79€ + *IVA	

AWR Microwave Office License	\$16666.67 + IVA	4 months
CST Studio Suite License	\$26666.67 + IVA	4 months
Fusion360 License	\$340.00 + IVA	4 months
Altium Designer License	\$1432.00 + IVA	4 months
Total Licenses	\$45.105,34 + IVA	

Total Project	50.441,76€ + *IVA	*Check chart for applied IVA
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Table 7: Project budget.

7. Project impact

The author considers this project to represent an impact primarily in the technological field. Its impact being limited due to the project's problem-solving nature, rather than a more academic or state-of-the-art research-oriented, which would potentially bear a more significant impact.

The author considers no impact on either the plethora of social implications often referenced or SDGs. The author reflects on the fact that forcefully cramming this type of content on a project without a true impact on these areas may be dishonest and misleading, in addition, it may also shift the focus away from the main topic.

The author also considers the representation of politics or attempt thereof to be utterly ill suited under every circumstance to work like this.

The author finally considers his project to ethically comply with any and every objective laid out in the hopes of a correct development of humankind.

8. Conclusions

8.1 Conclusions

Overall, this work intends to fix the issue found in the UPMSat-3, where two radio frequency transceivers which operate on the same frequency are embarked, and whose operation would be detrimental to the performance of the other.

A tailored feeding network is proposed in order to achieve the isolation required between both devices whilst being able to operate simultaneously. An introduction to the theory behind the design is provided, followed by a conceptual model. Detailed steps of the design procedures, modelling and simulation are provided, along with their simulation results. The current antenna system is analyzed and an improved antenna and feeding network system is displayed. The PCB design and fabrication outputs are displayed, and a suitable fabrication house is selected after comparing over ten of them. Finally, prototypes are measured indicating positive design results.

The results of these early prototypes show good correlation with the requirements although not all of them fully comply with them. A review of the design is considered not to be required.

The compliance of the definitive circuits cannot be known at the moment due to only having characterized early prototypes with different manufacturing properties and connectors.

8.2 Future works

The author considers that a full characterization of the definitive circuit should be performed prior to complete acceptance of the solution. These include electromagnetic characterization, as displayed in this project, as well as outgassing and mechanical characterization.

Further development of the proposed improved antennas should also be pursued in order to achieve a correct radiation performance and polarization isolation.

Alongside the antenna design, an improved feeding network design should follow, intending to minimize the physical dimensions of the feeding network + antenna system and reduce substrate selection related costs and complications.

9. References

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Anex

A.1 Fabrication gerbers specification

El diseño se fabrica con las siguientes especificaciones:

2 capas de cobre electrodepositado de 18um de grosor.

Substrato Rogers TMM6 de 0.635mm de grosor.

Máscara de soldadura verde.

Serigrafía blanca.

Fabricación siguiendo IPC 6012 clase 3A.

El archivo de fabricación con formato Gerber X2 contiene 9 capas que se describen a más adelante.

Se incluyen 3 tipos de taladros en el diseño:

Taladros de 0.4mm de diámetro.

Taladros de 2mm de diámetro.

Taladros de 3.2mm de diámetro.

Todos los taladros son metalizados, sin cubrir, sin relleno y penetran la PCB de lado a lado.

Se genera un panel de dimensiones 457mm X 305mm (tamaño de panel estándar de Rogers) con 4 PCBs diferentes:

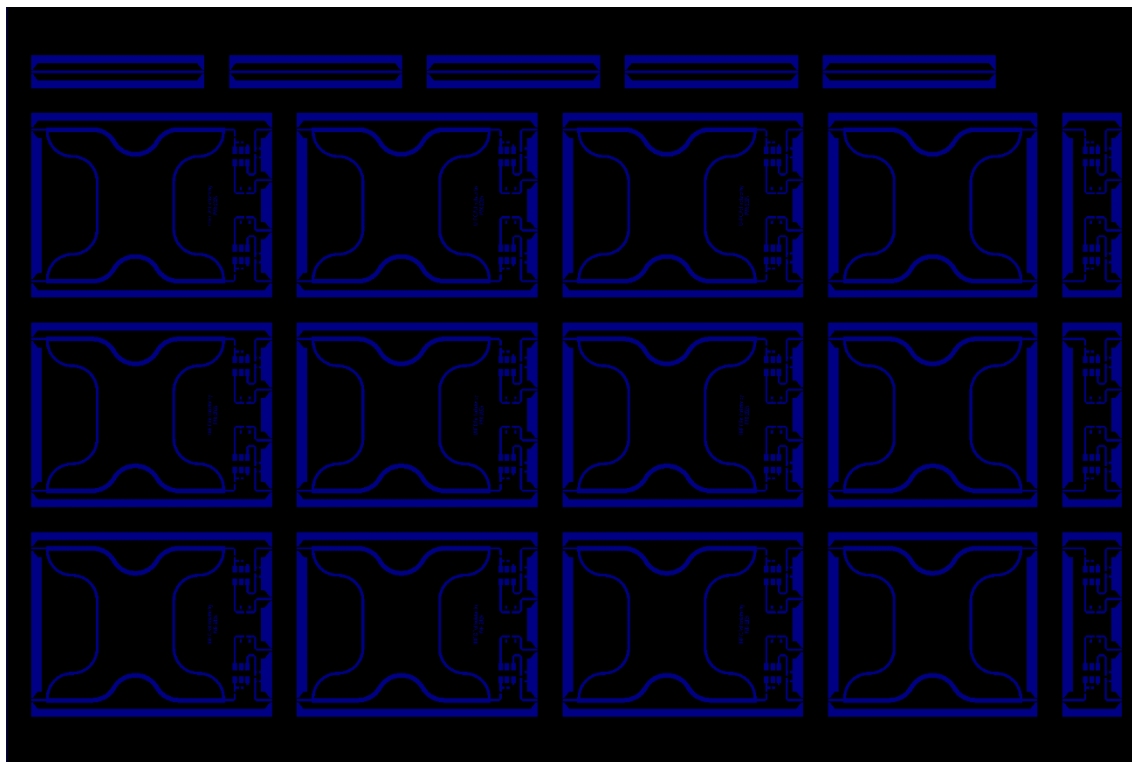
PCB1 – 9 Unidades

PCB2 – 3 Unidades

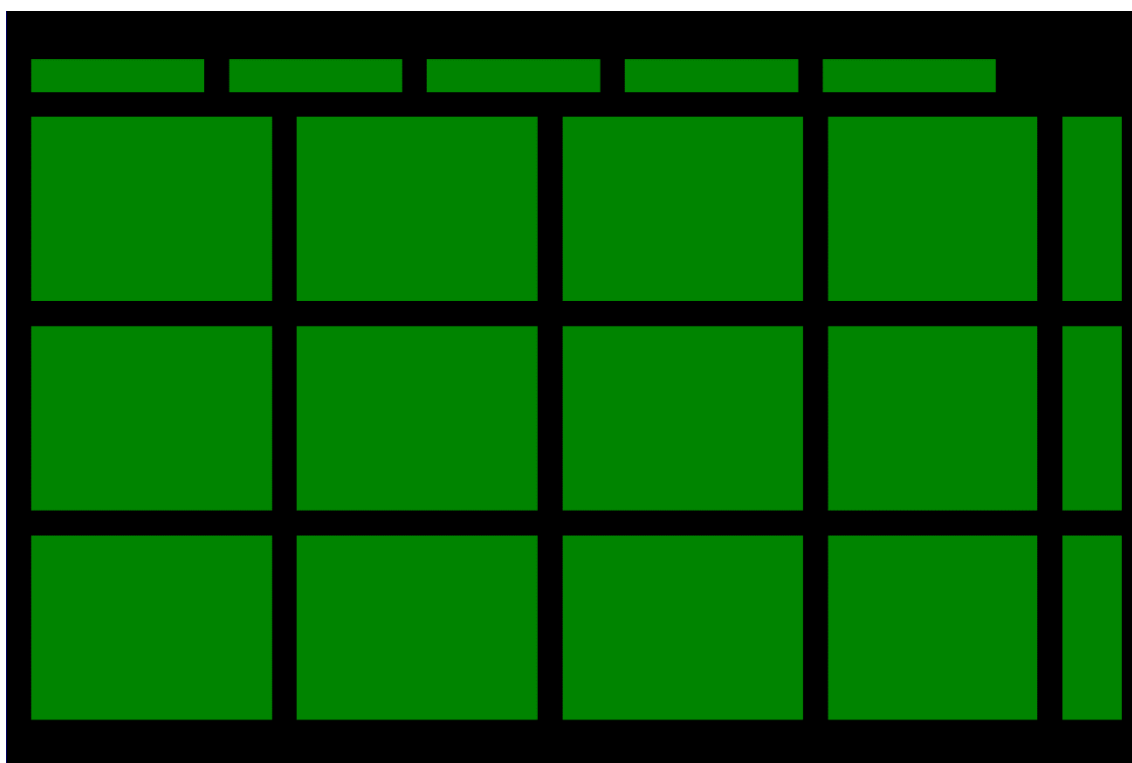
PCB3 – 3 Unidades

PCB 4 – 5 Unidades

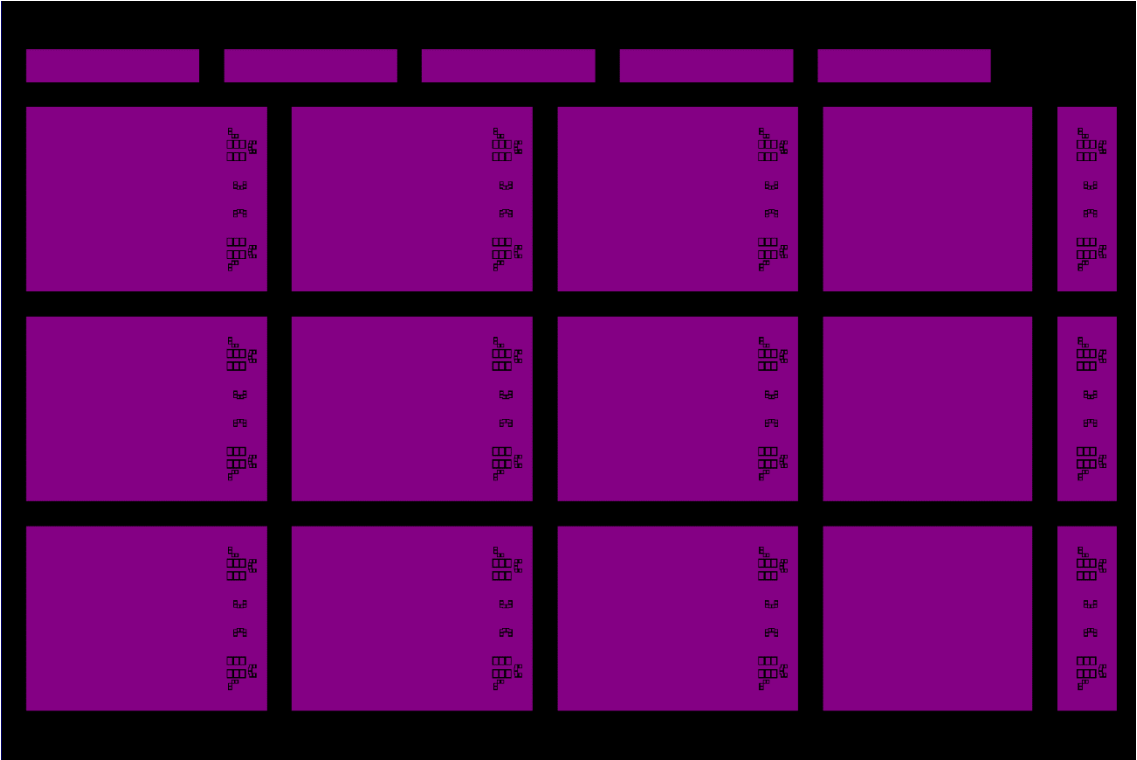
copper_signal_top – Capa superior de cobre



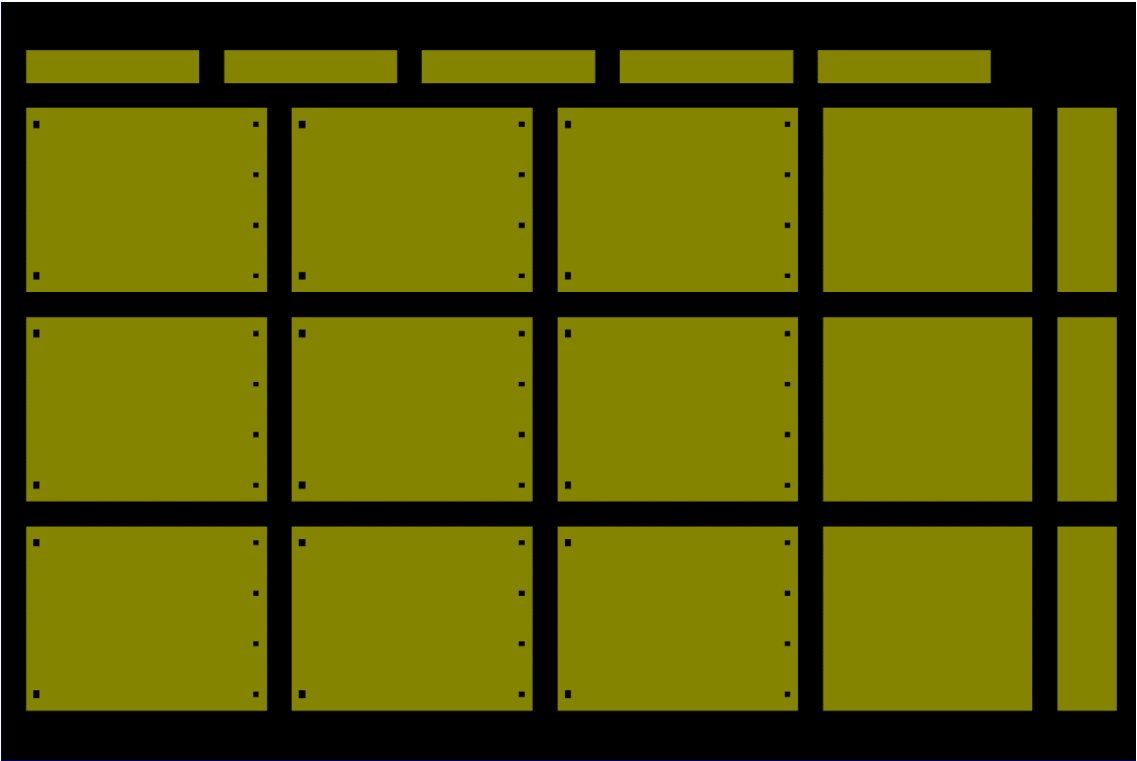
copper_signal_bot – Capa inferior de cobre



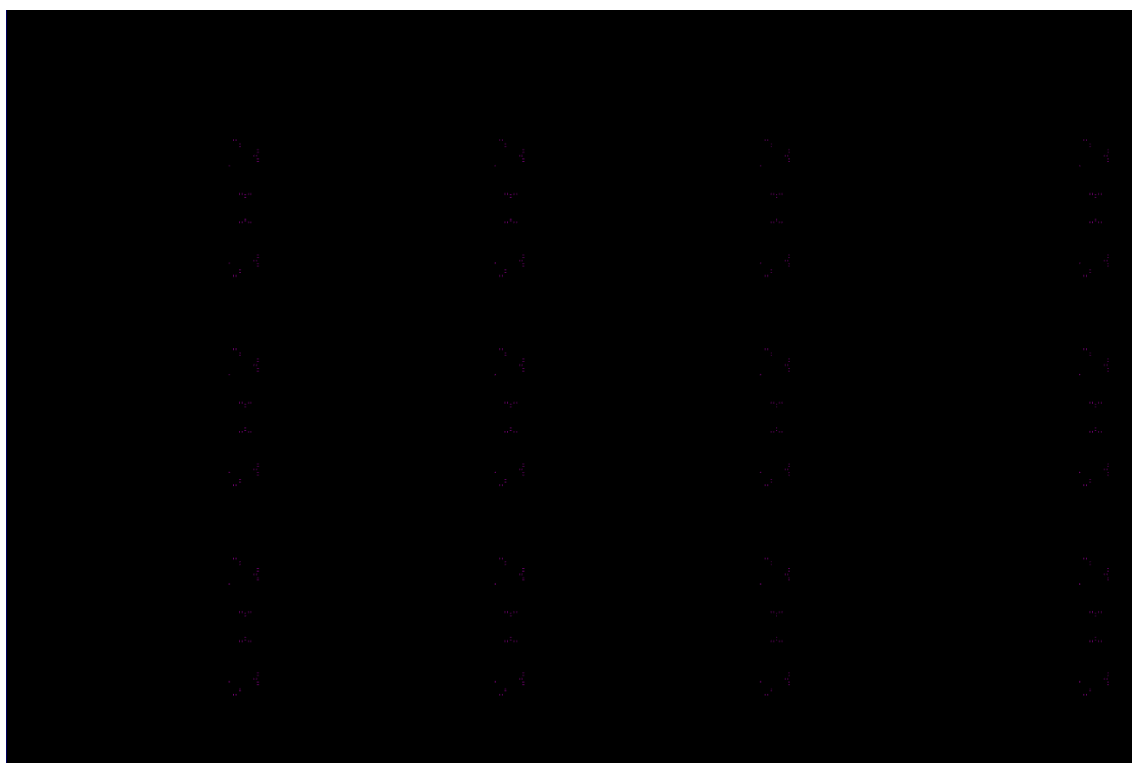
soldermark_top – Máscara de soldadura superior



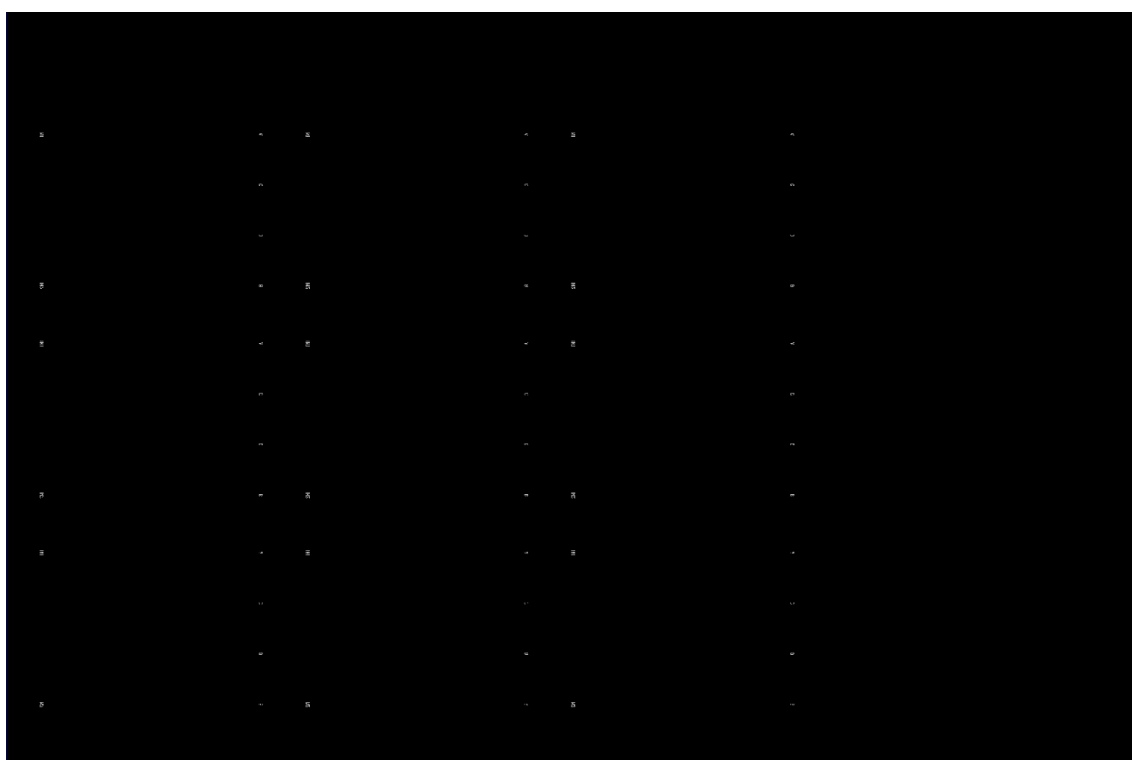
soldermask_bot – Máscara de soldadura inferior



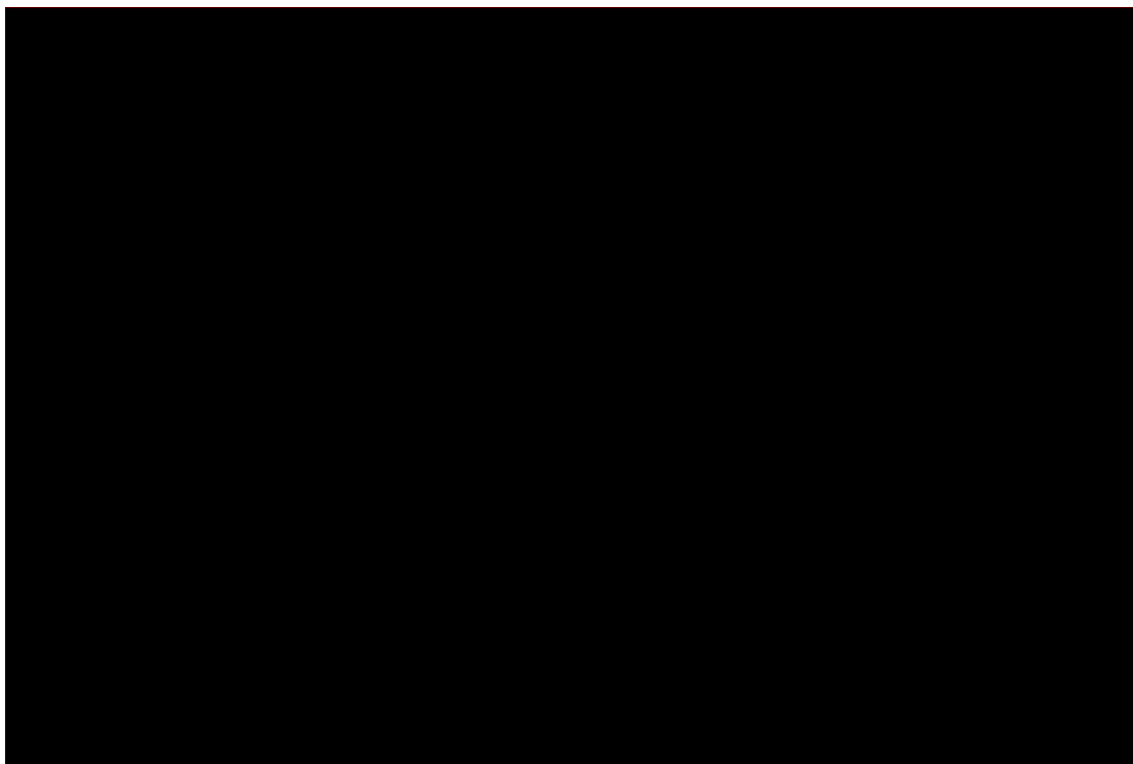
legend_top – Serigrafía superior



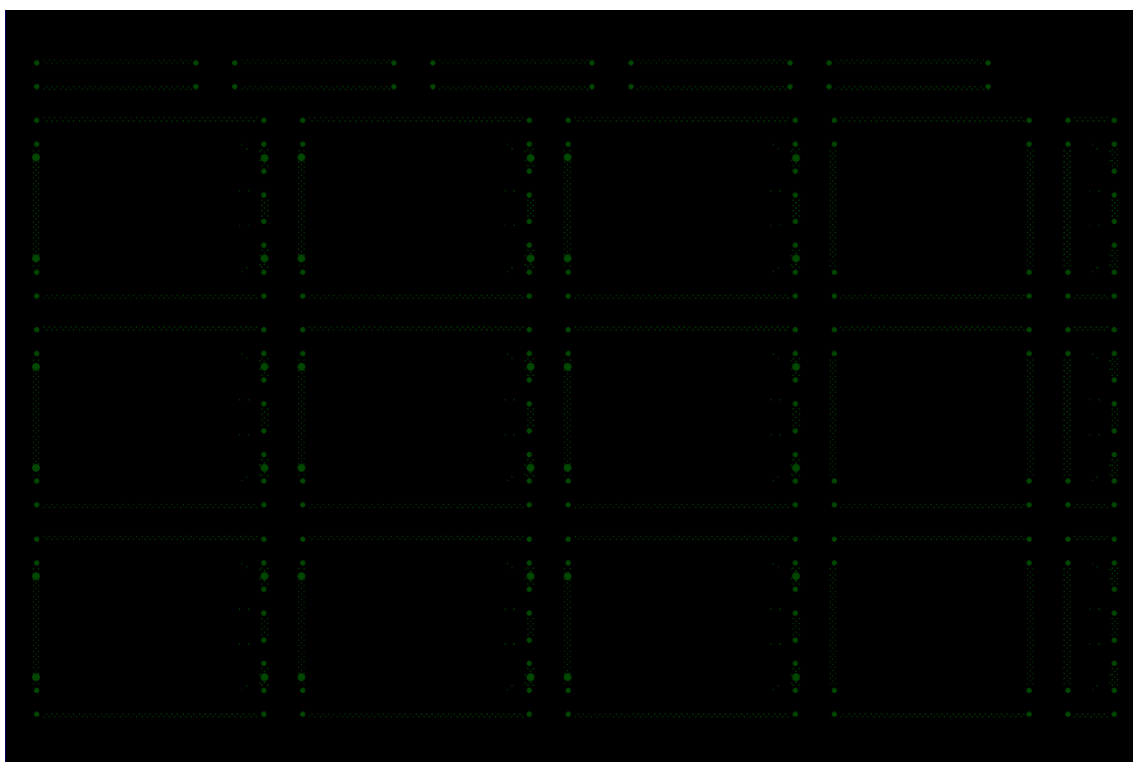
legend_bot – Serigrafía inferior



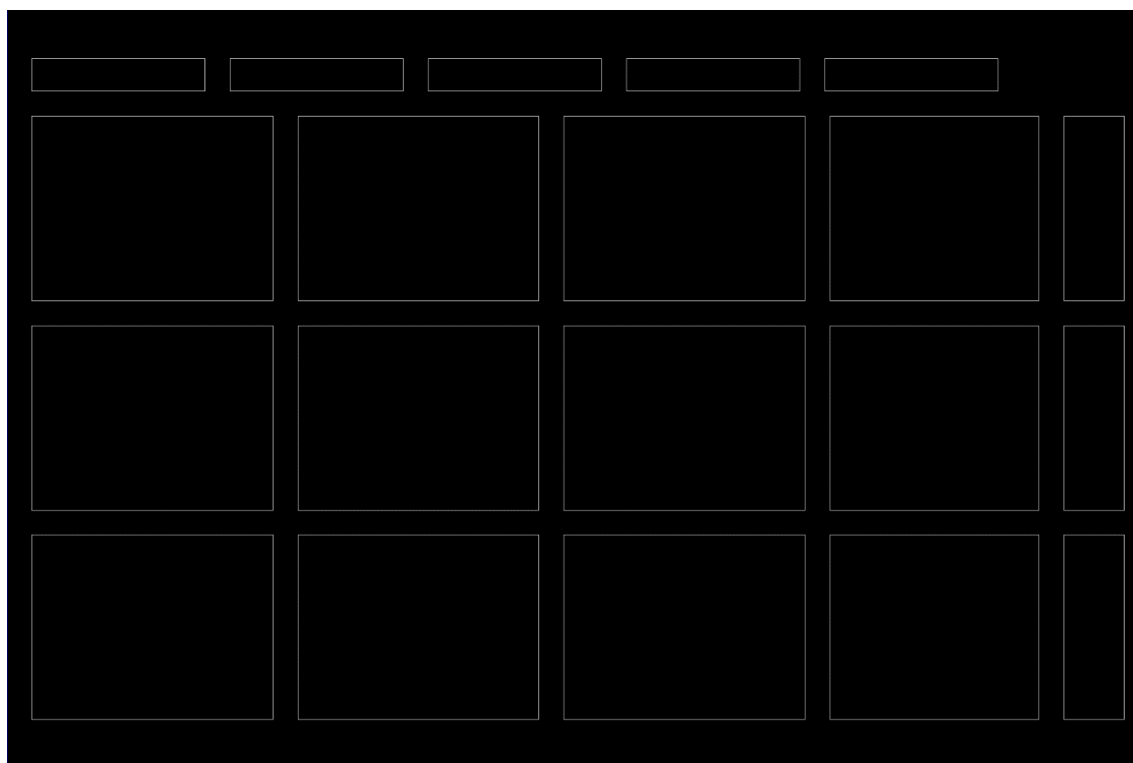
profile – Bordes del panel



pth_drill – Taladros pasantes metalizados



v_cut – Cortes de panel para PCBs individuales



A.2 Component bill of materials

Quantities:		Component	Quantity	ID	Value	Package	Type
2	Full Circuit	LQW18AN12NG80	2	L1,L2	12nH	0603	Inductor
		GQM1875G2E1R0WB12	2	C1,C4	1pF	0603	Capacitor
		GQM1875C2E101FB12	4	C2,C3,C5,C6	100pF	0603	Capacitor
Test+Flight		ADTL1-12+	2	U1,U2	N/A	CD542	Transformer
Minimum 2		292-04A-6	6	N/A	N/A	N/A	Connector
Quantities:		Component	Quantity	ID	Value	Package	Type
1	Balun	LQW18AN12NG80	2	L1,L2	12nH	0603	Inductor
		GQM1875G2E1R0WB12	2	C1,C4	1pF	0603	Capacitor
		GQM1875C2E101FB12	4	C2,C3,C5,C6	100pF	0603	Capacitor
Test		ADTL1-12+	2	U1,U2	N/A	CD542	Transformer
Minimum 1		292-04A-6	0	N/A	N/A	N/A	Connector
Quantities:		Component	Quantity	ID	Value	Package	Type
1	Hybrid	LQW18AN12NG80	0	L1,L2	12nH	0603	Inductor
		GQM1875G2E1R0WB12	0	C1,C4	1pF	0603	Capacitor
		GQM1875C2E101FB12	0	C2,C3,C5,C6	100pF	0603	Capacitor
Test		ADTL1-12+	0	U1,U2	N/A	CD542	Transformer
Minimum 1		292-04A-6	0	N/A	N/A	N/A	Connector
Spares:		Component	Quantity	ID	Value	Package	Type
4	Total + spares	LQW18AN12NG80	10	L1,L2	12nH	0603	Inductor
Reuse connector?		GQM1875G2E1R0WB12	10	C1,C4	1pF	0603	Capacitor
		GQM1875C2E101FB12	16	C2,C3,C5,C6	100pF	0603	Capacitor
0		ADTL1-12+	10	U1,U2	N/A	CD542	Transformer
0 YES; 1 NO		292-04A-6	6	N/A	N/A	N/A	Connector