

Design and Comparative Analysis of 90° Hybrids in FR4 and RO4003C Substrates for 5G NR Systems

Roger Metz Meira, André Augusto Mariano and Arthur Liraneto Torres Costa

Abstract—This work presents the design and comparative analysis of two 90° hybrids for phase-shift applications on OFDM signal processing in the context of 5G NR systems operating at 900 MHz (Band n8). Both designs were validated by electrical simulation and measurements demonstrating the expected trade-offs between the low-cost (FR4) and high-precision (RO4003C) solutions. The analysis demonstrates that the physical modeling of the junctions and the fabrication methods shift the simulated 900 MHz FR4 resonance frequency into 917 MHz and the same simulated frequency in RO4003C into only 905 MHz. Nonetheless, in the target frequency of 900 MHz, even with resonance shifts, the FR4 board was able to provide an S_{11} of -33.5 dB, while its S_{21} and S_{31} presented an expected imbalance due to the lack of material homogeneity. In contrast, the RO4003C board presented an excellent S_{21} and S_{31} symmetry, while only an S_{11} of -19.1 dB, due to a reduction in trace thickness near the SMA connector. The differential phase shift was measured to be -92.5° on the RO4003C and -91.5° on the FR4 version. These results confirm the reliability and precision of the RO4003C substrate when compared to FR4, even when manufactured in low-cost CNC machines.

Keywords—phase-shifter, 5G NR, hybrid coupler, OFDM.

I. INTRODUCTION

The transceiver architectures for the fifth generation of cellular networks (5G New Radio - 5G NR) require strict linearity and signal integrity, especially due to the use of multiple carrier modulations such as the Orthogonal Frequency Division Multiplexing (OFDM) [1]. According to the TS 38.101-1 standard specifications of 3GPP [2], the frequency spectrum below 1 GHz plays a fundamental role in long range coverage. In this scenario, the N8 band, which operates at a center frequency of 900 MHz, stands out by requiring an orthogonal treatment of the in phase (I) and quadrature (Q) components to mitigate adjacent channel interference [3].

In laboratory environments, when using vector signal generators to send I and Q, the quadrature component is of utmost importance for signal demodulation on the receiver side of the system. In Figure 1, the 90° hybrid coupler acts as a critical element in both the transmission and reception processes. The severe imbalance in the quadrature signal can make the OFDM signal recovery infeasible and directly compromises the receiver SNR [4].

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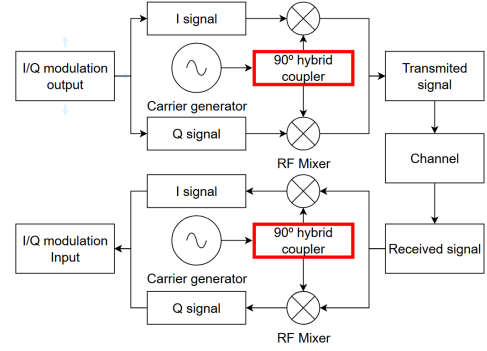


Fig. 1. Application scenario for the phase shifter in the radio frequency chain for 5G NR systems.

In order to quantify the analog imperfections in the signal integrity of the OFDM signal, the Error Vector Magnitude (EVM) parameter is used. In an ideal modulator, the I and Q signals maintain their perfect orthogonality, concentrating all the transmitting power in the desired spectrum component (P_{desired}). However, the amplitude imbalance and quadrature errors introduce an undesired spectral image component (P_{image}), associated to the loss of orthogonality of the I/Q system. The ratio between the desired signal power and the image component power define the image rejection (IR) metric, expressed by:

$$IR = \frac{P_{\text{desired}}}{P_{\text{image}}} \quad (1)$$

In the OFDM context, that image component acts like an additional interference over the received sub-carriers, directly degrading the modulation error vector. Thus, the analytical relation between the physical errors of the hybrid coupler and the EVM degradation can be expressed by: [3], [5]:

$$EVM_{\text{rms}} = \sqrt{\frac{1}{IR}} = \sqrt{\frac{1 - 2(1 + \delta) \cos \theta + (1 + \delta)^2}{1 + 2(1 + \delta) \cos \theta + (1 + \delta)^2}} \quad (2)$$

where δ represents the gain fractional imbalance between the I/Q channels and θ corresponds to the quadrature error in radians.

In order to mitigate the phase imbalance and to comply with the 5G NR specifications, the industry usually adopts high cost microwave substrates. In parallel, the recent literature concentrates on the topological improvement of the design structures, investigating from the analytical behavior of the phase difference [6] and the power division flexibility [7], to

the miniaturization by advanced technologies, such as printed waveguides (Printed-RGW) [8].

Although these advancements solve many bottlenecks in high frequency performance, the literature lacks investigations about practical limits of basic commercial materials operating in the sub-1 GHz range. This paper fills this gap by focusing on the evaluation of a low-cost solution using an FR4 design manufactured with low-cost CNC machines compared to the industrial standard Rogers RO4003C for the same N8 band (900 MHz), measuring its advantages and disadvantages.

II. THEORETICAL FUNDAMENTALS AND ANALYTICAL DESIGN

The chosen hybrid coupler in 90° quadrature is based on the classic hybrid coupler topology of the branch-line type. Due to the geometrical and electrical symmetry of the 4-port circuit, its electromagnetic behavior can be rigorously determined by the even and odd analysis [7].

Consider the coupler fed by a voltage wave V on Port 1, being all the other ports terminated in the reference impedance Z_0 . The structural topology of the hybrid, illustrating each branch characteristics, is shown in Figure 2 [9].

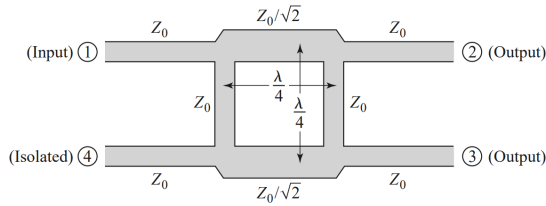


Fig. 2. Theoretical model of the 90° hybrid coupler (branch-line), with ports distribution detailed and their impedances with respect to Z_0 .

The ideal S-matrix ($[S]$) of a lossless 3 dB coupler in quadrature and perfect isolation is expressed by:

$$[S] = -\frac{1}{\sqrt{2}} \begin{bmatrix} 0 & j & 1 & 0 \\ j & 0 & 0 & 1 \\ 1 & 0 & 0 & j \\ 0 & 1 & j & 0 \end{bmatrix} \quad (3)$$

In order to deduce the characteristic impedances for the horizontal (series, Z_{series}) and vertical (parallel, Z_{shunt}) lines, port 1 excitation is decomposed into two symmetrical scenarios along the geometrical center axis of the device:

- 1) **Even mode:** Same amplitude and same phase excitations ($V/2$) applied simultaneously to ports 1 and 4. This symmetry plane behaves like an open circuit.
- 2) **Odd mode:** Same amplitude but opposite phase excitations ($V/2$ and $-V/2$) applied to ports 1 and 4. The symmetry plane acts like a short circuit.

The superposition of both modes results in a single excitation on port 1. From this decomposition, the 4-port circuit is reduced to two equivalent 2-port circuits with two cascaded sections. The reflection (Γ_e , Γ_o) and transmission (T_e , T_o) coefficients for the even (e) and odd (o) modes, obtained from the ABCD transmission matrix of each branch with electrical length $\theta = \pi/2$ ($\lambda/4$), are fundamental deductions of the microwave engineering, being given by [9]:

$$\Gamma_{e,o} = \frac{A_{e,o}Z_0 + B_{e,o} - C_{e,o}Z_0^2 - D_{e,o}}{A_{e,o}Z_0 + B_{e,o} + C_{e,o}Z_0^2 + D_{e,o}} \quad (4)$$

$$T_{e,o} = \frac{2}{A_{e,o}Z_0 + B_{e,o}/Z_0 + C_{e,o}Z_0 + D_{e,o}} \quad (5)$$

The s-parameters of the entire coupler are reconstructed by the linear combination of the modal coefficients:

$$\begin{aligned} S_{11} &= \frac{1}{2}(\Gamma_e + \Gamma_o), & S_{21} &= \frac{1}{2}(T_e + T_o) \\ S_{31} &= \frac{1}{2}(T_e - T_o), & S_{41} &= \frac{1}{2}(\Gamma_e - \Gamma_o) \end{aligned} \quad (6)$$

In order to guarantee the impedance matching on the input port ($S_{11} = 0$) and ideal isolation on the isolated port ($S_{41} = 0$), the condition $\Gamma_e = \Gamma_o = 0$ is imposed. Substituting the ABCD matrix elements under a quarter wave wavelength ($\lambda/4$), the fundamental relation of normalized admittances is obtained:

$$\left(\frac{Y_{series}}{Y_0}\right)^2 - \left(\frac{Y_{shunt}}{Y_0}\right)^2 = 1 \quad (7)$$

In order to obtain an equal power division of -3 dB at the output ports, the magnitudes of the transmission coefficients must be identical ($|S_{21}| = |S_{31}| = 1/\sqrt{2}$), which fixes the relative phase exactly 90°. The solution of the system of linear equations results in the design nominal values of the lines characteristic impedance [9]:

$$Z_{serie} = \frac{Z_0}{\sqrt{2}} \approx 35.35 \Omega, \quad Z_{shunt} = Z_0 = 50 \Omega \quad (8)$$

III. FABRICATION SCENARIOS

The structural analysis comprises two distinct dielectric media:

- **Scenario A (FR4 Low-Cost):** as a commercial laminate focused on low-cost prototyping, its electrical properties are not certified by conventional suppliers. This way, typical empirical values are adopted: $h = 1.6 \pm 0.16$ mm, nominal relative permittivity $\epsilon_r = 4.5 \pm 0.3$ and loss tangent $\tan \delta = 0.02$. It should be noted, however, that the FR4 has a non-homogeneous macrostructure, composed by fiberglass bundles ($\epsilon_r \approx 6$) embedded in an epoxy resin matrix ($\epsilon_r \approx 3$). This physical characteristic provokes local spatial variations in the effective dielectric constant that can surpass $\Delta\epsilon_{r,eff} = 0.2$ [10], impacting the local propagation velocity and, consequently, the quadrature precision of the circuit phase.
- **Scenario B (Rogers RO4003C):** hydrocarbon substrate reinforced with ceramic, ideal to wideband high frequency applications. It was designed with a thickness of $h = 1.52 \pm 0.1$ mm. According to the official manufacturer's datasheet [11], the material ensures a rigorous permittivity control. For design calculations and coupled circuits simulation (*Design Dk*), $\epsilon_r = 3.55$ is adopted, while the loss tangent is maintained in $\tan \delta = 0.0027$. In contrast with the FR4, its composition guarantees an extreme thermal stability and electromagnetic homogeneity of the board.

IV. DESIGN METHOD AND TEST SETUP

The method for the development of this study comprises the transition from the analysis of transmission lines and the schematics created in the simulation environment to the test-bench structuring for the later experiment on the bench.

A. Layout synthesis and Circuit Simulation

The transition from the nominal characteristic impedances calculated in equation (8) to the physical dimensions of the printed circuit required the design of the width (W) of the traces with respect to the dielectric thickness (h). The classical Hammerstad synthesis equations for microstrip lines were used [9]. For narrow lines ($\frac{W}{h} \leq 2$):

$$\frac{W}{h} = \frac{8e^A}{e^{2A} - 2} \quad (9)$$

For wide traces ($\frac{W}{h} \geq 2$), the extended equations are:

$$\frac{W}{h} = \frac{2}{\pi} \left[B - 1 - \ln(2B - 1) + \frac{\varepsilon_r - 1}{2\varepsilon_r} \cdot \left(\ln(B - 1) + 0,39 - \frac{0,61}{\varepsilon_r} \right) \right] \quad (10)$$

where the auxiliary parameters A and B are expressed as:

$$A = \frac{Z_c}{60} \sqrt{\frac{\varepsilon_r + 1}{2}} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1} \left(0,23 + \frac{0,11}{\varepsilon_r} \right), \quad B = \frac{377\pi}{2Z_c\sqrt{\varepsilon_r}} \quad (11)$$

From these equations, the physical lengths (L) of each branch of the coupler were designed analytically to be equivalent of a quarter wave ($\lambda_g/4$) in the frequency of 900 MHz.

In order to extract the s-parameters, the circuit schematic was submitted to a microwave simulation, using the open-source software named uSimmics. Distributed classical models were used to represent the physical behavior of the board (Fig. 3). The model included microstrip transmission lines ($MLIN$) and "T" junctions ($MTEE$) to model the branch intersections of the hybrid coupler.

The transition from the theoretical design to the physical layout required specific geometrical adaptations to the input and output ports. It is important to note that for the Rogers RO4003C laminate, the narrow variation in the dielectric thickness significantly alters the W/h ratio of the 50Ω line. In order to make the trace width compatible with the physical dimensions of the available SMA connectors, a geometrical step was designed to adapt. That transition near the connector was carefully computed in the circuit simulation by means of impedance step blocks ($MSTEP$) to guarantee a correct prediction of parasitic effects.

Fig. 4 illustrates the final physical topology of the structure printed on the FR4 and RO4003C substrates, reflecting the rectilinear modeling of the lines without the need for matching steps, due to its larger dielectric thickness. Additionally, the right side of Fig. 4 shows the adaptation made to prevent coupling effects between the signal and ground lines, using discrete steps that gradually expand the trace width.

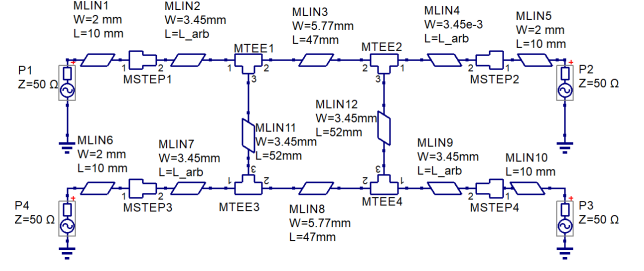


Fig. 3. Schematics model of the coupler using distributed elements for the circuit simulation.

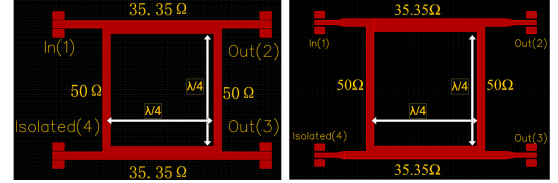


Fig. 4. Physical layout of the circuit topology. FR4 design on the left and RO4003C design on the right.

B. Bench-top electrical characterization

For the full s-parameters measurements of all the 4 ports, a test strategy was developed. Since the coupler has 4 terminals and the Vector Network Analyzer (VNA) available, RIGOL RSA3015N, operates with only 2 ports and 1 path, the complete characterization of the device under test (DUT) was made incrementally.

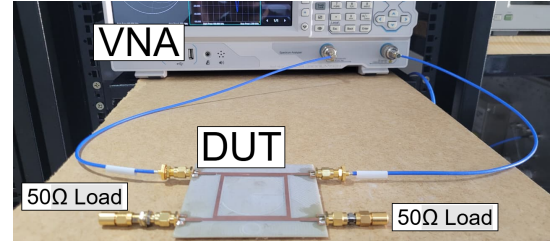


Fig. 5. Experiment setup for the 2-port VNA measurement and 50Ω loads on the terminals not measured.

As shown in the test setup of Figure 5), to obtain the 4-port s-parameter matrix, 16 sequential independent measurements were made. The arrangement considers the simultaneous coupling of 2 ports to the calibrated coaxial terminals of the VNA, while the other remaining 2 ports of the coupler are terminated with 50Ω loads, mitigating external spurious reflections. Before the measurements, an SOLT (Short-Open-Load-Thru) calibration was performed in the RF cables, ensuring the reference plane was well established on the tip of the SMA connectors of the coupler. The calibration kit used was the TBVC-SF6 from Tekbox manufacturer, which provides all the polynomial parameters of each standard to be inserted on the VNA.

V. RESULTS AND DISCUSSIONS

The simulated response of the physical topology exhibits a fundamental characteristic of the microwave design: the

practical insertion of "T"junctions (*MTEE*) and matching steps introduces parasitic elements that affect the effective electrical length of the lines. This phenomenon alters the impedance matching and it is responsible to shift the optimum point of the initial analytical structure operation, independently of the substrate used.

A. Analysis of the Commercial FR4 Substrate

In the FR4 substrate, this phenomenon is evident. As illustrated in Figure 6, the deep resonance of the coupler and the maximum isolation do not occur at 900 MHz, but shifted to the 917 MHz frequency. However, for 5G NR applications, the 900 MHz target band behavior is evaluated. In this context, the hybrid coupler in FR4 presents a simulated reflection coefficient of -33.5 dB, ensuring that less than 0.045% of the energy is reflected to the transmitter.

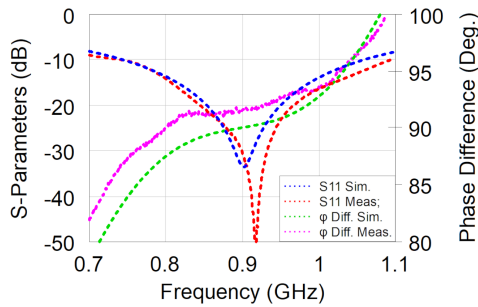


Fig. 6. Reflection coefficient and phase for the FR4 board.

Analyzing the transmission (Fig. 7) and reflection (Fig. 8) parameters, an accentuated imbalance can be noted on the transmitted power at 900 MHz (-3.01 dB and -3.71 dB) and a lack of symmetry in the reflections of each port observed in Figure 8, caused by the combination of frequency shift and losses inherent to epoxi resin and the lack of fabrication method control on the FR4 substrate. The phase behavior follows this pattern (Fig. 6); the quadrature of exact 90° is crossed at 800 MHz, while in 900 MHz the phase falls near the 91.6° , but with a useful bandwidth of more than 100 MHz.

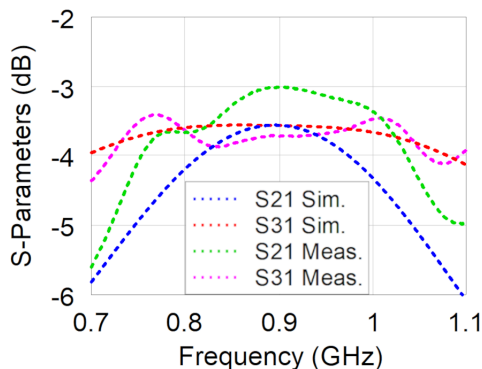


Fig. 7. Simulated transmission coefficients (S_{21} e S_{31}) of the FR4 coupler.

In order to evaluate the results obtained on the VNA, it is of fundamental importance to measure in the time domain as well. Although this is a passive circuit, time domain measurements can ensure that the desired behavior is confronted once

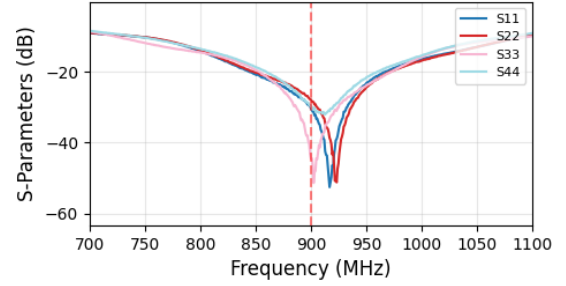


Fig. 8. Measured reflection parameters to the FR4 coupler, evidencing the lack of symmetry in the curves.

more. In Figure 9, it can be observed that there was a 0.5 dB amplitude difference at the output of ports 2 and 3, and a phase difference of -91.5° .

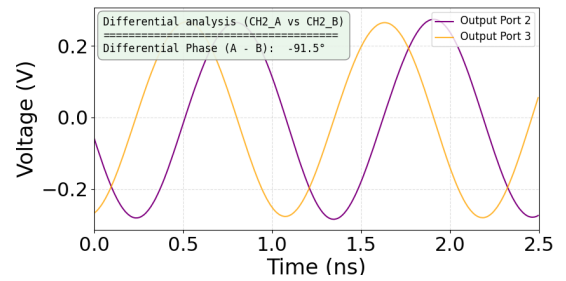


Fig. 9. Time domain response of the FR4 hybrid coupler measured in 900 MHz.

B. Analysis of the Rogers RO4003C Substrate

The design based on the Rogers RO4003C validates the dielectric efficiency in terms of propagation phenomena. Due to the identical impact of the physical junctions, the optimum point of resonance in this substrate established itself in 905 MHz (giving S_{11} of -19.1 dB), which demonstrates a high fidelity to the simulated results (Fig. 10). Besides, the material homogeneity can be observed on the symmetrical results of the reflection parameters obtained in the measurements (Fig. 11).

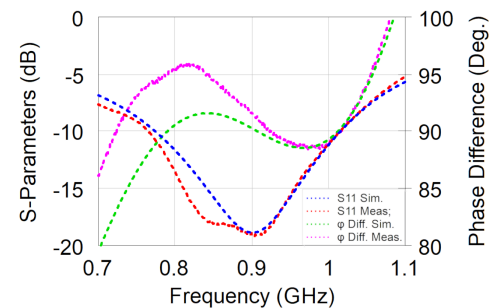


Fig. 10. Reflection coefficients and phase of Rogers RO4003C laminate.

The symmetry of the reflections of the input ports on the RO4003C substrate demonstrates a significative evolution when compared to the FR4. However, in Figure 12, it can be verified that there is still an imbalance between the output ports, however, they are 12% lower than in FR4.

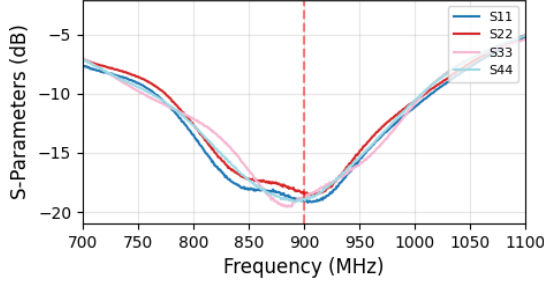


Fig. 11. Reflection parameters of the RO4003C coupler, evidencing the symmetrical results in all ports.

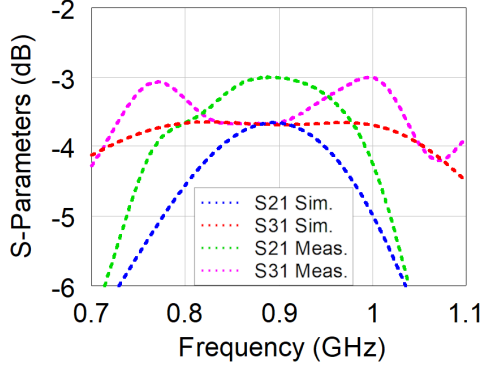


Fig. 12. Simulated transmission coefficients of the hybrid in RO4003C substrate.

The vector difference (Fig. 10), there was a phase difference of 1.6° with respect to what was simulated at 900 MHz, resulting in a frequency performance similar to the FR4 board.

C. Transient analysis

The design using RO4003C (Fig. 13) obtained a time domain performance with a phase of -92.5° , however, the outputs were more well balanced, with a difference of 0.36 dB between ports 2 and 3 at 900 MHz.

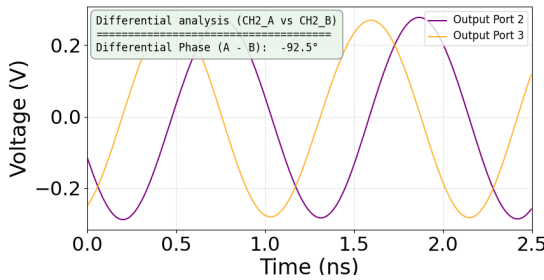


Fig. 13. Time domain response (Rogers): quadrature and waveform analysis at 900 MHz.

To provide a direct evaluation between the two investigated technologies, Table I summarizes the key parameters obtained through both simulation and experimental measurements for the FR4 and RO4003C hybrid couplers operating around the target 900 MHz band.

VI. CONCLUSIONS

This paper showed a technical comparison of 90° hybrid couplers for the 5G NR applications (N8 band). The simulation

TABLE I
PERFORMANCE COMPARISON OF BOTH COUPLERS AT 900 MHz.

Parameter	FR4 (Sim)	FR4 (Meas)	RO4 (Sim)	RO4 (Meas)
Res. Freq. (GHz)	0.900	0.917	0.900	0.905
S_{11} (dB)	-33.5	-33.5	-18.9	-19.1
Amp. Imbal. (dB)	~ 0.0	0.50	~ 0.0	0.36
Phase (deg)	90.0°	-91.5°	90.0°	-92.5°
Substrate Cost	Low		High	

with physical distributed elements revealed that the discontinuities inherent to the layout (junctions and transitions) shift the maximum resonance frequency to higher frequencies in both substrates. However, the performance around 900 MHz dictates the feasibility of the design. In spite of the phase and amplitude losses, the FR4 strategy confirms its usefulness in prototyping for low-cost applications, maintaining design reliability, even with its non-homogeneity of the substrate and fabrication methods. In scenarios that require strict symmetry on all ports and high reliability on the s-parameters, the Rogers RO4003C laminate ensures its relevance as the mandatory technology, shielding the coupler from high energy dispersions and providing a simulated (-18.9 dB) to measured (-19.1 dB) compliance in the desired frequency band of an OFDM signal.

ACKNOWLEDGMENTS

This study was supported by the Conselho Nacional de Desenvolvimento Científico e Tecnológico (CNPq) - Grant number 310783/2023-2 and by the Coordenação de Aperfeiçoamento de Pessoal de Nível Superior - Brasil (CAPES) - Finance Code 001. The authors also acknowledge LiraLab for providing support during device testing and measurement.

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