Limitation of a Five-Port Reflectometer Using Planar Elliptic Couplers for UWB Applications

Imen Sfar
UR - CSEHF
University of Tunis El Manar
Tunisia
imen.sfar@voila.fr

Lotfi Osman
UR - CSEHF
University of Tunis El Manar
Tunisia
Lotfi.osman@supcom.rnu.tn

Ali Gharsallah
UR - CSEHF
University of Tunis El Manar
Tunisia
ali.gharsallah@fst.rnu.tn

Marjorie Grzeskowiak
ESYCOM EA2552
Université Paris-Est, Marne-la-Vallée
France
grzeskow@univ-mlv.fr

Odile Picon
ESYCOM EA2552
Université Paris-Est, Marne-la-Vallée
France
odile.picon@univ-mlv.fr

***Abstract*— In this paper, we present the design of ultra wideband (UWB) planar Five-Port Reflectometer using the elliptic coupler which has the advantage to offer a wide bandwidth compared to the hybrid one. The elliptic coupler and the five-port circuit were designed and simulated in the 1.5 to 3.5 GHz band using ADS software of Agilent. An analytical calculus and simulation results of a five-port circuit using elliptic coupler demonstrate limitations of this structure.**

Keywords- Elliptic coupler, six-port circuit, five-port reflectometer, ultra wideband applications.

I. INTRODUCTION

During recent years, there has been a great interest in six-port circuits due to their broad range of applications. Initially, a six-port was applied for flexible measurements of complex reflection coefficients in microwave network analysis [1]. Many new applications using the six-port technology have been proposed over the last ten years, such as direct conversion receivers [2]-[3], which reduce circuit complexity and allow a higher level of integration than conventional heterodyne receivers [4], phase and frequency discrimination in a radar system [5] and beam-direction-finding circuits [6].

A typically six-port circuit is composed of a passive linear device of four 90° hybrid couplers, it has two inputs and four outputs connected to power detectors in order to obtain the baseband signals. One of the four power detectors is used as a reference to normalize any variation in the source power.

In the same perspective, a multiband six-port circuit using micro-strip rounded couplers has been built [7]. However, these designs are characterized by low bandwidth, which is not sufficient for the high speed communication system. Recently, a new six-port based on elliptic coupler is proposed to overcome this inconvenience [8]. The circuit allows performing low losses and wide bandwidth. During the development of the six-port technique, the reference detector may not be required and the six- port circuit can be replaced by simplified five-port architecture, thus reducing the number of analog-digital converters and low-pass filters. The possibility of using a relatively simple five-port circuit like a vectorial reflectometer had been discussed by Engen [1].

Homodyne five-port receiver configurations were proposed in [9] and [10], and some experimental five-port reflectometers have been published [11]-[12]. This structure is very suitable for software-defined radio applications [13]. Lately, some direct-conversion receivers with different architectures, based on five-port technology, were proposed and studied [14]-[15]. This technique was particularly interesting due to the use of power detectors instead of mixers and hence provided simpler circuits in comparison with its heterodyne counterpart. The approach has many advantages. First of all, the broadband specifications can be easily obtained using passive elements. Also, they have very low power consumption. There are various methods to implement the five-port architecture but it is important to choose an efficient one to provide suitable performance. In this paper, we present the general theory and prototype of five-port reflectometer, the quadrature hybrid using the elliptic disk is designed. The

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ambiguity problem and design consideration of the new five-port reflectometer will be discussed and some experimental results regarding low losses and wide bandwidth will be reported.

II. THE FIVE-PORT REFLECTOMETER

The conventional five-port reflectometer is a linear device with two inputs corresponding respectively to the radio-frequency (RF) and local oscillator and three outputs. Power detectors are connected to these outputs in order to obtain the baseband signals. Usually, the power detector consists of a Schottky diode and a low-pass filter. The voltage measured at the filter output is linearly related to the power of the RF signal present at the diode's input terminal only for small values. This condition cannot always be guaranteed and a linearization procedure must be performed on the measured voltage values [16].

The five-port reflectometer performs a direct conversion but using a redundant mixer and a 120 degrees basis [17] instead of a cartesian basis. This redundancy makes the five-port system more robust to the phase and amplitude.

The relation between the incident (a_i) and reflected (b_i) pseudo-waves at the interferometer ports are completely described by the junction's scattering parameters S_{ij} :

$$b_i = \sum_{j=1}^{5} S_{ij}\, a_j \quad \text{with} \quad i \in \{1,2,3,4,5\} \tag{1}$$

Assuming the power detectors permanently connected at the reflectometer's output ports, one may write:

$$a_j = b_j\, \Gamma_j \quad \text{with} \quad j \in \{3,4,5\} \tag{2}$$

where Γ_j is the reflection coefficient of the power detectors at port j. The eight-equation linear system formed by (1) and (2) may be rewritten in function of two of the ten variables. It can be solved as function of two free variables a_1 and a_2 and we may write:

$$b_i = A_i\, a_1 + B_i\, a_2 \quad \text{with} \quad i \in \{3,4,5\} \tag{3}$$

where A_i and B_i are constants of the reflectometer. These constants are determined by a calibration [18]. It is important to notice that, contrary to a six-port system, the power level at the reference input port of a five-port system must be the same for system calibration**.** The power level detected at the reflectometer outputs may be expressed by:

$$P_i = |b_i|^2 \left(1 - |\Gamma_i|^2\right), \quad i \in \{3,4,5\} \tag{4}$$

It is possible to write:

$$|\rho - q_i|^2 = \frac{\alpha_i}{|a_1|^2} P_i \ , \ \ i \in \{3,4,5\} \tag{5}$$

where $\rho = \frac{a_2}{a_1}$, q_i and α_i are constants of the junction.

By means of a straightforward mathematical development equation (5), it is possible to write the complex ratio between the input pseudo power waves as a linear combination of the voltages v_{out-3} , v_{out-4} and v_{out-5} measured at the low-pass filter outputs after dc-offset cancellation, as shown below:

$$\rho = \alpha . v_{out-3} + \beta . v_{out-4} + \gamma . v_{out-5} \tag{6}$$

where the complex values α, β and γ are three calibration constants and v_{out-3} , v_{out-4} and v_{out-5} are proportional to P_3, P_4 and P_5. The calibration procedure [19] determines three complex constants α, β and γ to regenerate the complex envelope from the three output voltages using equation (6).

We use the same kit for calibrating both six-port and five-port reflectometers. A new adapted procedure for the diode linearization is suggested. The calibration of the five-port is done using a similar approach as in six-port case where an intermediate variable *w* is introduced [20].

The used system is a part of a six-port reflectometer where the reference power data is not used. This simplification provides a considerable reduction in the time processing usually required by six-port based systems. In comparison to the six-port, the five-port reduces the circuit size.

A photograph of the implemented circuit is shown in Fig. 1. This circuit, inscribed in a rectangle of 10 cm in length and 8 cm in width, was designed to operate at 2.4 GHz with a bandwidth of 1 GHz. The five-port reflectometer is fabricated on TMM4 substrate with ε_r = 4.5 and thickness of 1.524 mm.

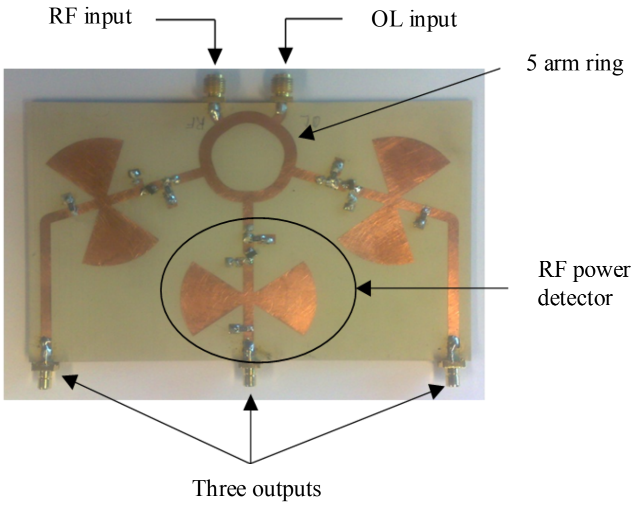

Figure 1. Microstrip five-port reflectometer.

III. ELLIPTIC COUPLER DESIGN

The design of the directional coupler can follow different implementations among which, lumped elements [21], micro-strip branch-line couplers [22]**,** multilayered structures [23], ring couplers [24]**,** and circular or elliptical couplers [25] are the most used.

The elliptic patch design method is based on [26], where short-circuit stubs are introduced at the periphery of a metallic printed disk to function as dummy ports, in addition to the matching networks at the physical ports. The approach in [26] allows changing the overall impedance matrix of the structure by a suitable choice of the stub parameters.

The Green's function of an elliptic patch consists of even and odd parts. So from [27], the even and odd elements of the impedance matrix Z for an elliptic patch with ports at the periphery, as shown in Fig. 2, are given as illustrated in the following equations:

$$Z_{ij}^{e} = \frac{jw\mu dl^2}{W_i W_j} \sum_{n=0}^{\infty} \frac{Je_n(h, \cosh u_1) ISe_n(h, v_i, \Delta_i, u_1) ISe_n(h, v_j, \Delta_j, u_1)}{Me_n(h) Je_n'(h, \cosh u_1)} \quad (7)$$

$$Z_{ij}^{o} = \frac{jw\mu dl^2}{W_i W_j} \sum_{n=0}^{\infty} \frac{Jo_n(h, \cosh u_1) ISo_n(h, v_i, \Delta_i, u_1) ISo_n(h, v_j, \Delta_j, u_1)}{Mo_n(h) Jo_n'(h, \cosh u_1)} \quad (8)$$

where $je_n(h, \cosh u)$ and $jo_n(h, \cosh u)$ are even and odd first-kind radial Mathieu functions, $Se_n(h, \cosh v)$, $So_n(h, \cosh v)$ are the even and odd first-kind circumferential Mathieu functions, respectively.

By the introduction of the effective parameters, the fringing fields are accounted. The effective relative permittivity [28] of the elliptic patch can be taken as:

$$\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left\{1 + \frac{10d}{2b}\right\}^{-1/2} \quad (9)$$

The effective semiminor axis *be* of the elliptic patch can be taken as [29]:

$$b_e = b\left[1 + \frac{2d}{\varepsilon_r \pi b}\left\{\ln\left(\frac{b}{2d}\right) + (1.41\varepsilon_r + 1.77) + \frac{d}{b}(0.268\varepsilon_r + 1.65)\right\}\right]^{1/2} \quad (10)$$

b represents the semiminor axis, given by *b*=*l*.sinh *u* and *d* is the thickness of the substrate.

The S-parameters of the quadrature hybrid are evaluated using the expression $S = (Z - Z_0)(Z - Z_0)^{-1}$, where Z_0 is the characteristic impedance of the transmission lines connected to the device.

The elliptic directional coupler was proposed by [27] to enhance the bandwidth of hybrid coupler. While introducing modifications on the structure original, a scaling procedure based on the following relation is performed:

$$f_1 a_1 \sqrt{\varepsilon_{r1}} = f_2 a_2 \sqrt{\varepsilon_{r2}} \quad (11)$$

where f_1 and f_2 are the central frequencies of the two bands (existed and desired), a_1 and a_2 are the corresponding characteristic dimensions of the patch (semi-major elliptic axis), ε_{r1} and ε_{r2} are the dielectric constants of the substrates.

The initial parameters are found after scaling and the final dimensions are obtained by using Momentum which is part of Advanced Design System (ADS). An optimization was carried out to determine the optimal values of the elliptic coupler dimensions.

The procedure of optimization is based on the scanning of a four-dimensional space by varying the length of the semimajor elliptic axis, the eccentricity, and the width and length of the impedance steps. The port angle is kept constant and in each circle of optimization, the flatness and the bandwidth of the coupler are evaluated. In the final step, the resultant geometry matches the desired characteristics. This geometry is given in the third column of Table I.

The actual implementation of the coupler is fabricated on a TMM4 substrate with relative dielectric constant ε_r= 4.5 and thickness h=1.524 mm. The implemented elliptic coupler is shown in Fig.2, while the simulated and measured responses of the reflection and transmission scattering S parameters are compared in Fig. 3, 4, 5 and 6. The elliptic coupler has been measured on network analyzer (E8361C).

TABLE I. DESIGN OPTIMIZATION OF THE PROPOSED ELLIPTIC COUPLER

Parameter	Initial Values	Optimum Values
Semimajor elliptic axis	18 mm	19.5 mm
Eccentricity	0.58	0.64
W (width of the impedance steps)	2.8 mm	3.8 mm
Ds (length of the impedance step ~ λ/4 transformers)	16.9 mm	17 mm
Φ (the port angle)	51°	51°

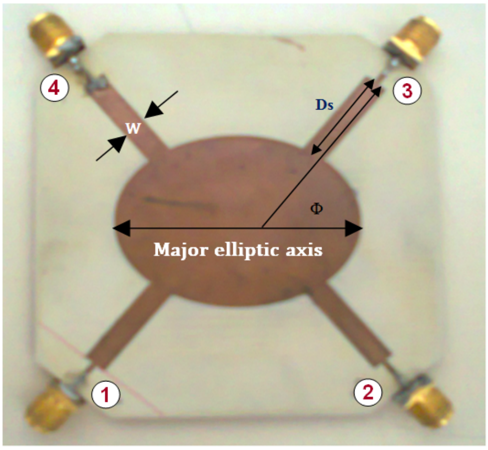

Figure 2. Implemented elliptic coupler.

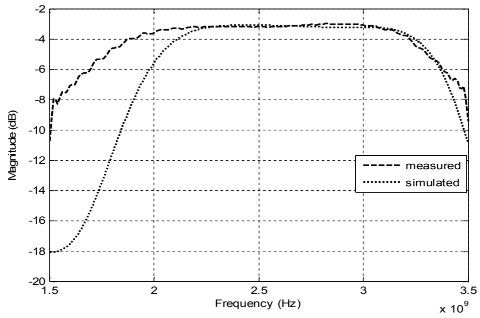

Figure 3. Simulated and experimental results of the coupling port S_{13} of the proposed elliptic coupler.

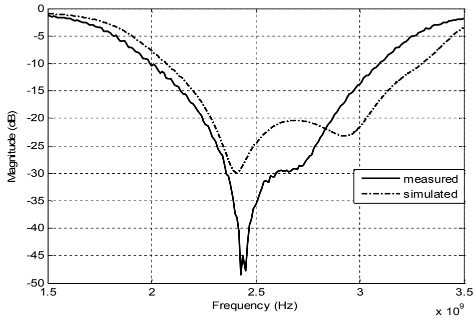

Figure 4. Simulated and experimental responses S_{11} of the elliptic coupler.

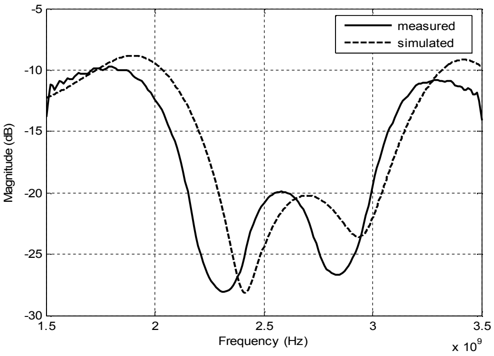

Figure 5. Simulated and experimental responses S_{12} of the proposed elliptic coupler.

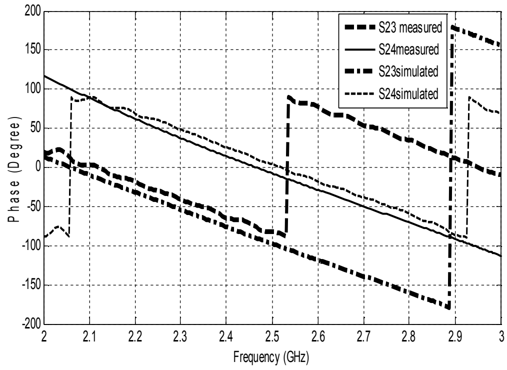

Figure 6. Phase difference between S_{23} and S_{24}.

The power of the coupled ports, S_{13} and S_{14}, is equal to - (3 ± 0.1) dB and presented in Fig .3. The simulated return loss, S_{11}, and the insertion loss, S_{12}, of the elliptic coupler are shown in Fig. 4 and 5.

The measured responses confirm the coupler's compliance with the bandwidth requirements, both the return loss and the isolation of better than 15 dB (at the center frequency). In terms of phase shift, Fig. 6 shows that the coupler produces a phase shift of about 90°.

IV. CONFIGURATION OF THE PROPOSED FIVE-PORT CIRCUIT

The standard six-port junction consists of four 90° -3dB hybrid couplers and one phase shifter. This circuit doesn't provide any high performances in terms of low losses and wide bandwidth. To overcome these problems, a new six-port based on elliptic couplers is achieved [8]. Performances concerning magnitude and phase shift of this coupler are better than the hybrid one. With these features, the six-port circuit based on elliptic couplers is suitable for wideband wireless applications system.

We have thought that we can adapt this optimization to achieve a UWB five-port circuit. Fig.7 shows the desired layout of the wideband five-port configuration based on three elliptic couplers.

The five-port junction presented in this section is designed and simulated at a center frequency of 2.4 GHz. The proposed five-port circuit was simulated using Momentum of ADS [30].

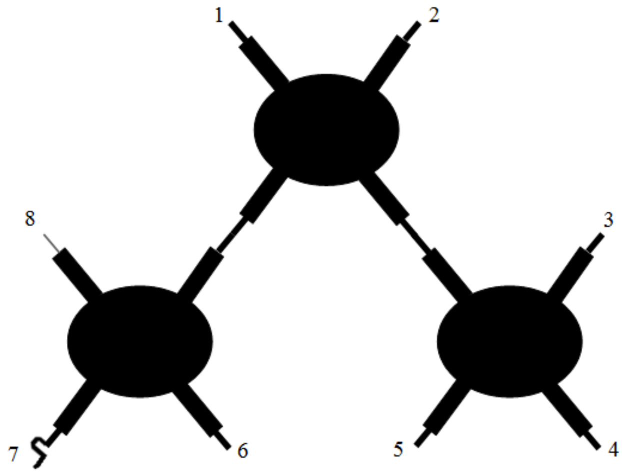

Figure 7. Layout of the five-port circuit.

It is a passive circuit composed of 3 four-port microstrip elliptic couplers and a phase shifter connected in port 7. The five-port circuit is designed with TMM4 having a thickness h of 1.524 mm and a relative dielectric constant $\varepsilon_r = 4.5$.

In this case, the size of the proposed five-port (159.6 mm*118.7 mm) is bigger compared with classic structure of the five-port. As perspective, a multi layer structure is considered to reduce the size. This technology needs an accurate fabrication process and alignment between the layers.

Figures 8 and 9 show simulation results of scattering parameters of the proposed five-port as well as those the classic five-port.

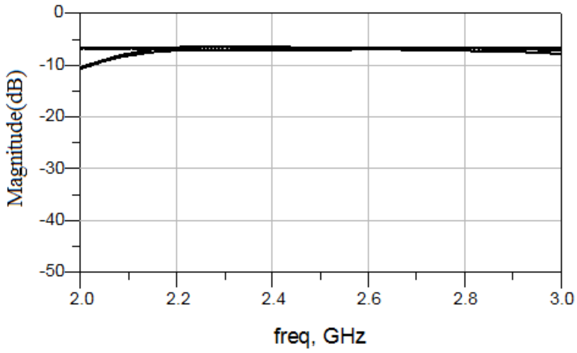

Figure 8. Scattering parameters of the proposed five-port magnitude.

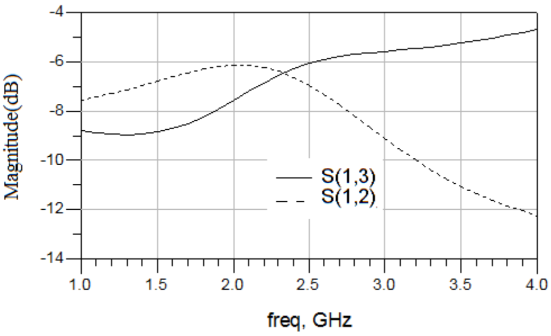

Figure 9. Scattering parameters of the five-port ring magnitude.

From these results, it can be shown that the coupling is almost constant over the frequency band of (1-3 GHz) with a value of 6 dB. Contrary to the classic five-port, it can be seen that the coupling is about 6 dB for the central frequency to 2.4 GHz, which is suitable to build the proposed circuit or other microwave networks. In term of phase shift as illustrated in figure 10, the five-port ring provides a phase of approximately 120 ° between two ports.

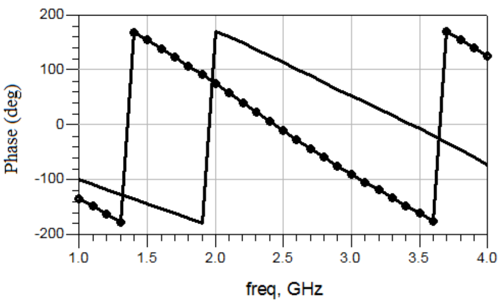

Figure 10. Scattering parameters of the five-port ring phase difference.

During simulation, we have noticed that the phase differences of the novel structure are not consistent for all ports for which the theoretical values are 120°. A mathematical treatment has been developed to justify the insufficiency of this technique, which allowed us to conclude that this structure had limits since we did not get good phase differences of the classic five-port.

V. CONCLUSION

In this paper, a new wideband five-port circuit based on the use of elliptic coupler has been presented. This elliptic coupler, which is also a part of the directional coupler family, is particularly suitable for low cost and UWB applications.

The performances of the proposed five-port in terms of magnitude are very promising, but this structure has limits in the phase shift. A five-port circuit composed of four elliptic couplers, which lead to sufficient degrees of liberty to optimize the 5-port circuit, is being studied, in despite of the size of this system. We have the possibility to realize a multilayer system in order to reduce the size.

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