A WIDEBAND QUADRATURE POWER DIVIDER/COMBINER AND ITS APPLICATION TO AN IMPROVED BALANCED AMPLIFIER

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Abstract—In this paper, a novel broadband quadrature power divider and its robust design method are presented. The QPD consists of a two-section power divider in combination with a 90-degree differential phase shifter. The two-section power divider is calculated to provide equal power split, high output port isolation, and good return loss at all three ports. The differential phase shifter consists of a composed right/left handed transmission line and pure-right handed transmission line named CRLHu-TL and PRHd-TL, respectively. The CRLHu-TL is divided into two parts; one of them consists of a pure-left handed section whose parasitic pad effects are represented by means of a pure right handed section named PRHp. On the other hand, the PRHd-TL is composed by a microstrip transmission line of characteristic impedance $50\,\Omega$ and electrical length $50°$ and two sections equivalents to PRHp. The proposed circuit is applied to develop a broadband balanced amplifier with measured fractional bandwidth (FBW) of 124.4% at the center frequency of $2\,\text{GHz}$.

1. INTRODUCTION

Metamaterials are artificial electromagnetic materials that do not exist in nature and exhibit some unique properties. Successful applications have been proposed in dual-band components [1, 2], enhanced-bandwidth devices [3], super-compact multilayer transmission lines [4],

phase and impedance couplers with arbitrary coupling level [5, 6] and negative/zeroth order resonators [7].

On the other hand, quadrature power dividers are key components for the implementation of balanced amplifiers, image rejection mixers, Doherty amplifiers, circularly polarized antennas, and others. Recently metamaterial quadrature power dividers and baluns have been proposed in [8–10].

In [8], the proposed broadband Wilkinson balun, the authors make use of a one-stage Wilkinson power divider in combination with a $+90°$ composed right/left handed transmission line (CRLH-TL) and a $-90°$ CRLH-TL. The $+90$ CRLH-TL consists of two series capacitor C_{L0}, a shunt inductor L_{L0}, and an initial host transmission line with characteristic impedance of $50\,\Omega$ and length d_{TL1}. The -90 CRLH-TL is composed by two series capacitor C_{L2}, a shunt inductor L_{L2}, and a host transmission line with characteristic impedance of $50\,\Omega$ and length d_{TL2}. The WPD is calculated at the central frequency f_0 and then the metamaterial phase-shifting lines are designed as follows: (a) to design the $+90$ CRLH -TL an initial host transmission line with characteristic impedance of $50\,\Omega$ and length d_{TL1} small enough in order for the effective medium equations presented in [12] remain unchanged is considered. With the purpose of achieve a phase advance for a positively traveling plane wave of the form $\mathrm{e}^{-j\Phi MM+}$, Φ_{MM+} must be chosen to be negative. Thus, letting $\Phi_{MM+} = -\pi/2$ in (1) listed in [8] and imposing the impedance matching condition of (2) listed in [8] the results of C_{L0} and L_{L0} are obtained. (b) The $50\,\Omega$ -90 MM-TL is designed to achieve a phase delay for a positively traveling plane wave of the form $\mathrm{e}^{-j\Phi MM-}$, therefore Φ_{MM-} must be chosen to be negative. Thus C_{L2}, L_{L2}, and d_{TL2} are obtained letting $\Phi_{MM-} = \pi/2$ in (1) listed in [8], imposing the impedance matching condition and enforcing the phase-slope condition of (3) listed in [8]. The circuit was validated with a prototype at $1.5\,\mathrm{GHz}$ and it has achieved a FBW of 77%.

On the other, the modified Wilkinson power divider proposed in [9] makes use of a one-stage Wilkinson power divider in combination with a conventional microstrip transmission line and a CRLH-TL whose unit cell is composed by L_L, C_L, C_R, and L_R. In order to design a broadband quadrature power divider, over the frequency range from f_1 to f_2, the CRLH-TL is chosen to have $-90°$ and $-180°$ at f_1 and f_2, respectively. Solving (1) and (2) listed in [9] the four equivalent circuit parameters L_L, C_L, C_R, and L_R are obtained. On the other hand, the MS-TL is designed to achieve the desired phase difference with the CRLH-TL therefore the microstrip line should be designed to have phase delay of $-180°$ and $-270°$ at f_1 and f_2 respectively.

Proposed circuit has FBW of 94% and exhibits big size due to the electrical length of the microstrip line.

In [10], a broadband quadrature power divider with FBW of 104% was developed. The circuit uses the same circuit as [9] however they differ in the design method. In [10], the quadrature power divider is also designed at the center frequency f_0. After, the length of the microstrip transmission line is properly chosen and simulated to determine the phase Φ_{MS1} and the phase Φ_{MS2} at f_1 and at f_2 respectively. The difference phase between the CRLH-TL and the microstrip transmission line must be $+90$ at f_1 and f_2, therefore the phase Φ_{MM1} and the phase Φ_{MM2} at f_1 and f_2 are obtained by means of (1) and (2) listed in [10]. Subsequently, two values P and Q must be obtained from Equations (6) and (7) listed in [10] and finally, the right-handed section of the CRLH-TL, L_L and C_L are obtained from P, Q and the matching condition. In [10], it is mentioned that the length of the microstrip transmission line is properly chosen however a criteria to choose the length of the microstrip transmission line is not clearly provided. On the other hand, because the solution depends directly of f_1 and f_2 it is worthy to mention that, a criterion to choose f_1 and f_2 is not mentioned.

In this paper, a novel metamaterial broadband quadrature power divider and a simply but robust design methodology based on a closed set of equations are proposed. Also, the proposed quadrature power divider has been utilized to design a balanced amplifier whose measured results show that a balanced amplifier with FBW of 124.4% has been successfully implemented.

2. PROPOSED QUADRATURE POWER DIVIDER AND ITS DESIGN METHOD

The proposed quadrature power divider consists of a two-section power divider in combination with a $90° \pm 10°$ metamaterial differential phase shifter as is shown in Figure 1(a). The two-section power divider has been calculated with Seymour's method [11], to provide equal power split, high output port isolation, and good return losses at all three ports within a fractional bandwidth of 120%. All the arms of the power divider have electrical length of $-90°$ at the center frequency and characteristic impedance Z_1 and Z_2 of $64.77 \, \Omega$ and $77.19 \, \Omega$ respectively. Also R_1 and R_2 were calculated as $165.22 \, \Omega$ and $137.73 \, \Omega$ respectively.

The proposed differential phase shifter consists of a CRLH-TL and PRH-TL named CRLHu-TL and PRHd-TL respectively. The CRLHu-TL is composed by a Pure-Left-Handed section whose unit-cells elements are C_L and L_L and a pure right handed section, named
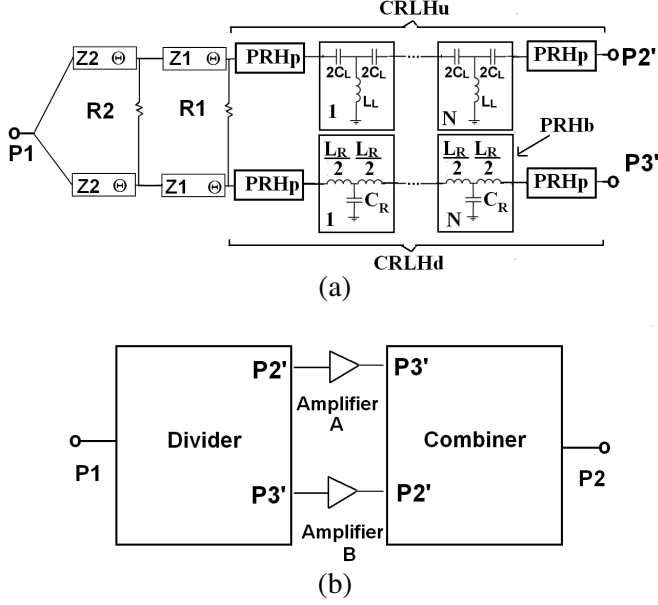
(a)

(b)

Figure 1. (a) Proposed quadrature power divider and (b) balanced amplifier.

PRHp. PRHp is due to the parasitic pad effect of C_L and L_L. The PRHd-TL is composed by a pure right handed section, named PRHb, and two sections equivalents to PRHp. Let $\emptyset_{PLH}$ and $\emptyset_{PRH}$ be the phase shift of the pure left handed section and the phase shift of the PRHb respectively, then the differential phase shift between top branch and the bottom branch of the quadrature power divider shown in Figure 1(a) is given by (1) [12].

$$\Phi(\omega) = \Phi_{PLH}(\omega) - \Phi_{PRH}(\omega) \qquad (1)$$

where

$$\Phi_{PLH}(\omega) = N\frac{1}{\omega\sqrt{L_L C_L}} \qquad (2)$$

$$\Phi_{PRH}(\omega) = -N\omega\sqrt{C_R L_R} \qquad (3)$$

It can be shown that the derivative and second derivate of (1) are given by (4) and (5) respectively. Since $N > 0$, $\omega > 0$, $L_L > 0$ and $C_L > 0$ then $\emptyset'' > 0$ thus the differential phase shift is a convex function because it is a twice differentiable function and its second derivative is positive. In particular, if $\emptyset'(\omega = \omega_{\min}) = 0$, then $\emptyset(\omega_{\min})$ is a global minimum. From this condition we calculated (6) which

represent the frequency in which the differential phase shift is minimal.

$$\Phi'(\omega) = \frac{d}{d\omega}\Phi(\omega) = N\left(\frac{-1}{\omega^2\sqrt{L_L C_L}} + \sqrt{C_R L_R}\right) \tag{4}$$

$$\Phi''(\omega) = N\left(\frac{2}{\omega^3\sqrt{L_L C_L}}\right) \tag{5}$$

$$\omega_{\min}^2 = \frac{1}{\sqrt{L_L C_L C_R L_R}} \tag{6}$$

In this work, the accepted differential phase shift is considered as $90° \pm 10°$ however an analog procedure can be performed for other tolerances. Because the accepted differential phase shift is $90° \pm 10°$ then the minimum accepted differential phase shift is $80°$ ($= 4\pi/9$ radians) and maximum accepted differential phase shift is $100°$ ($= 5\pi/9$ radians). Now, let ω_1 be the frequency in which the differential phase shift is maximum and $\omega_{\min}$ the frequency in which the differential phase shift is minimum, then

$$\Phi(\omega_{\min}) = N\left(\frac{1}{\omega_{\min}\sqrt{L_L C_L}} + \omega_{\min}\sqrt{C_R L_R}\right) = \frac{4\pi}{9} \tag{7}$$

and

$$\Phi(\omega_1) = N\left(\frac{1}{\omega_1\sqrt{L_L C_L}} + \omega_1\sqrt{C_R L_R}\right) = \frac{5\pi}{9} \tag{8}$$

Now, from (7) and (8) and after some straightforward we have obtained the polynomial equation of the second degree given by (9) which solution is given in (10).

$$2\omega_{\min}^2 - 5\omega_1\omega_{\min} + 2\omega_1^2 = 0 \tag{9}$$

$$\omega_1 = \frac{\omega_{\min}}{2} \tag{10}$$

Thus replacing (10) in (7), a new equation in terms of ω_1 can be set as follows:

$$N\left(\frac{1}{2\omega_1\sqrt{L_L C_L}} + 2\omega_1\sqrt{C_R L_R}\right) = \frac{4\pi}{9} \tag{11}$$

Furthermore, if the matching conditions listed in (12) and (13) are imposed in the analysis, then (8), (11), (12) and (13) define a linear system of four equations with the four unknowns (C_L, L_L, C_R and L_R). The solution of this set of equations is given by (14)–(17).

$$Z_C = \sqrt{\frac{L_R}{C_R}} \tag{12}$$

$$Z_C = \sqrt{\frac{L_L}{C_L}} \tag{13}$$

$$C_R = \frac{\pi}{9NZ_C\omega_1} \tag{14}$$

$$L_R = \frac{\pi Z_C}{9N\omega_1} \tag{15}$$

$$C_L = \frac{9N}{4\pi Z_C\omega_1} \tag{16}$$

$$L_L = \frac{9NZ_C}{4\pi\omega_1} \tag{17}$$

Replacing (14)–(15) in (1) it can be found that $\emptyset(\omega = \omega_1) = \emptyset(\omega = 4\omega_1) = 100°$, thus the lowest frequency and the highest frequency where the phase difference is accepted are $\omega = \omega_1$ and $\omega = 4\omega_1$. Therefore the fractional bandwidth of the proposed circuit is 120% and the central frequency is $\omega_0 = 5\omega_1/2$.

On the other hand, replacing $\omega_1 = 2\omega_0/5$ into (14)–(17) one arrives at (18)–(21). These equations are very important because they establish C_L, L_L, C_R and L_R in terms of central frequency ω_0 which facilitates the design; in this context the proposed methodology is simpler to implement than those proposed in [8–10].

$$C_R = \frac{5\pi}{18NZ_C\omega_0} \tag{18}$$

$$L_R = \frac{5\pi Z_C}{18N\omega_0} \tag{19}$$

$$C_L = \frac{45N}{8\pi Z_C\omega_0} \tag{20}$$

$$L_L = \frac{45NZ_C}{8\pi\omega_0} \tag{21}$$

If the PRHb section is considered as a conventional transmission with characteristic impedance Z_{PRH} (= Z_c = 50 Ω) and electrical length Θ_{PRH} then replacing (18) and (19) into (3) ones arrives to the electrical length of the bottom branch is $\Theta_{PRH} = 50°$ at the central frequency. In this context proposed differential phase shifter exhibits smaller size than those proposed in [8–10].

It is important to mention that, if the differential phase shift is considered as $\Phi_{PRH}(\omega) - \Phi_{PLH}(\omega)$ then the differential phase difference is a concave function and the analysis is slightly modified. However, the equivalent circuit proposed in Figure 1(a) is valid for the design.

To validate the schematic of the proposed metamaterial quadrature power divider/combiner, the simulated response of a design at central frequency of 2 GHz with 3 unit cells is shown in Figure 2. This figure shows that from 804 MHz to 3.202 GHz, the differential
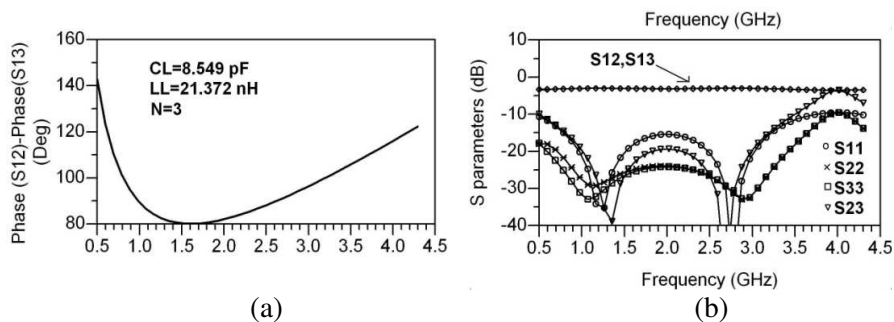
Figure 2. Simulated response of a metamaterial quadrature power divider at 2 GHz with $N = 3$.

phase shift is $90° \pm 10°$, insertion loss are 3 dB, return loss and isolation are better than 10 dB.

3. IMPLEMENTATION AND RESULTS OF A BALANCED AMPLIFIER BY USING THE PROPOSED QUADRATURE POWER DIVIDER

The use of balanced amplifiers is a practical method for implementing the broadband amplifier that has flat gain and good input and output return loss [13]. Figure 1(b) shows the most popular arrangement of a balanced amplifier. It is well known, from the S-parameters of a balanced amplifier, that if the two amplifiers are identical, then the best return loss are achieved ($S_{11} = 0$ and $S_{22} = 0$) and therefore the return loss depend on the combiner/divider [13]. In this context several balanced amplifiers have been proposed in [14–22]. From Table 1, given in [14], it can be found that the maximum FBW of published balanced amplifiers is 97.9%. This FBW was achieved, in [14], by using metamaterial quadrature power dividers/combiners.

Based on the schematic given in Figure 1(b) a balanced amplifier at central frequency of 2 GHz was developed by using two identical ATF-36077 transistors. The balanced amplifier was designed and fabricated on a RT-duroid substrate with $\varepsilon_r = 3.38$ and $h = 1.524$ mm.

The proposed prototype was implemented in three steps. The first step is to develop two identical amplifiers, the second step is to develop a metamaterial differential phase shifter and the third step is to implement the two-stage PD. In the first step, each amplifier is formed by an input bias tee, an ATF-36077 transistor (biased at $V_{ds} = 1.5$ V and $I_{ds} = 10$ mA), and an output bias tee. The response of a single

fabricated amplifier is shown in Figure 3. It can be seen in Figure 3(b) that return losses are very bad. It is due to input matching and output matching networks are not included in the design. However, this was done deliberately to demonstrate the ability of proposed balanced amplifier to improve the return losses.

In the second step, the pure-left handed section of the quadrature power divider is designed by considering one unit cell. The unit-cell values, C_L (= 2.85 pF) and L_L (= 7.124 nH), are calculated from (2) and (3) respectively. Each unit cell is formed by two commercial chip capacitors (ATC-600S5R6) with capacitance of 5.6 pF ($\approx 2C_L$) and a microstrip stub inductor ($w = 0.2$ mm and length $l = 7.555$ mm) shorted to the ground plane by a via is utilized to implement L_L.

On the other hand, the PRHb is designed to have characteristic impedance of 50 Ω and electrical length of 50°. Thus, the PRH-TL has length of 12.688 mm and width of 3.512 mm respectively.

Up to this point, the design of the QPD is almost finished. However when the differential phase shift is measured, it can be found that the differential phase shift is lower than the desired valued due to the parasitic pad effect of C_L and L_L. The problem is easily solved by adding a 50 microstrip transmission line, on the bottom branch, whose length increases from zero degrees until the required differential phase shift is achieved. For our design, a transmission line with length $l = 2.95$ mm and width $w = 3.512$ mm is added to the PRHb section to take into account the parasitic pad effects. Figure 4 shows the differential phase shift before and after compensation. The other parameters are not presented because the changes are only performed on the phase of the pure-right handed transmission line.

In the third step, the dimensions of the two-stage power divider were calculated and arranged as is shown in Figure 5. The width and length of the transmission line with impedance of 64.770 Ω and
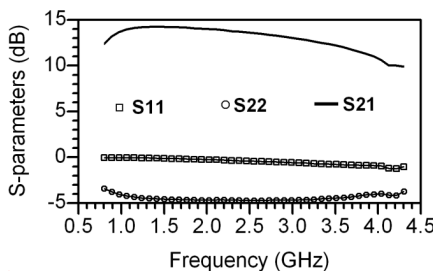

Figure 3. Measured results of a single amplifier.
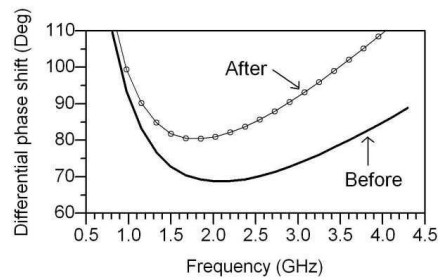

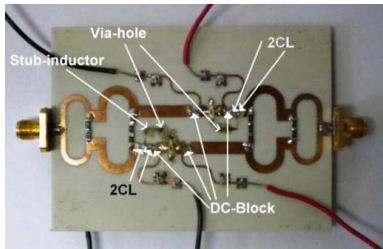**Figure 4.** Compensation of the parasitic effects due to C_L and L_L.

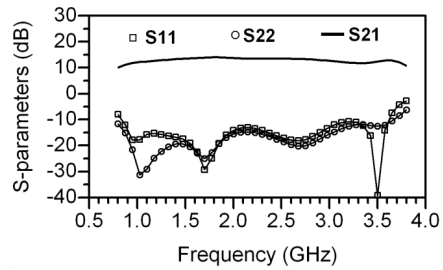**Figure 5.** Manufactured wideband balanced amplifier.

Figure 6. Measured response of a prototype at 2 GHz.

electrical length of $90°$ are $w = 2.259$ mm and $l = 23.247$ mm respectively. Also, the width and length of the transmission line with impedance of $77.190\,\Omega$ and electrical length of $90°$ are $w = 1.607$ mm and $l = 23.530$ mm.

Figure 5 and Figure 6 show the photograph of the fabricated wideband balanced amplifier and its measured results respectively. From Figure 6, it is clear that the return losses are better than 10 dB through the 841.6 MHz to 3.610 GHz frequency range. This frequency range represents a FBW of 124.4% and central frequency at 2.226 GHz. Furthermore the gain of the amplifier is 12.333 dB ± 1.6 dB. From Figure 3 and Figure 6, it can be seen that the balanced amplifier has very good response despite that each individual amplifier has very bad return loss. These results show that with the proposed circuit, we were able to successfully design a broadband balanced amplifier with good return losses and flat gain.

4. CONCLUSIONS

In this work, a novel broadband metamaterial quadrature power divider with fractional bandwidth up to 120% was presented. The design is based on a straightforward, but robust methodology which is based on a set of closed exact equations. The proposed circuit was utilized to develop a broadband balanced amplifier with measured fractional bandwidth of 124.4%.

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