Design of a Single Balanced Mixer Using a Quadrature hybrid and a Low Pass Butterworth Filter

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I. INTRODUCTION

A mixer is a three-port device that uses a nonlinear or time-varying element to achieve frequency conversion[1]. An ideal mixer produces an output which has the sum or the difference of the two input frequencies. In this project the non-linearity property of the diode was exploited to build a balanced mixer. A nonlinear component produces harmonics and hence there is a low pass filter after the mixer to select the desired IF(Intermediate Frequency) output. The architecture of the mixer used in the project is given in **Fig. 1**. We would start with the design of a matched 90° hybrid coupler and build on to achieve the desired down converted IF of 25MHz from a LO(Local oscillator) frequency of 900MHz and RF(Radio Frequency) input of 925MHz.

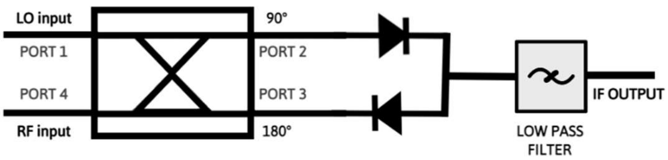

Fig. 1. Architecture of balanced mixer using a 90° hybrid coupler

II. 90° HYBRID COUPLER

Hybrid couplers are a special case of directional couplers where the coupling factor is 3dB, which means that half of the power from the input port reaches port 2 and the other half to port 3. Here we will discuss about the quadrature hybrid coupler also called branchline coupler which has a 90° phase shift between the ports 2 and 3.

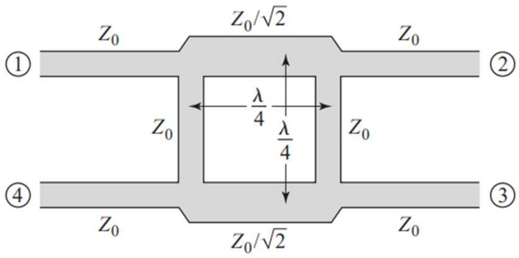

Fig. 2. Geometry of a branchline coupler[1].

The branchline coupler as shown in **Fig. 2.** is made up of four quarter-wavelength long transmission lines at target frequency. Two of the transmission lines have the characteristic impedance of the termination(Z_0) and the other two have $1/\sqrt{2}$ times Z_0. Here in our project, Z_0 is taken as 50Ω. Hence $Z_0/\sqrt{2}$ can be calculated as 35.35Ω. Looking at the geometry of the device we can easily notice that it is symmetric hence we can easily perform even-odd mode analysis.

A. Even-Odd Mode Analysis

We first normalize all the impedances to 50**Ω.** We assume that a wave of unit amplitude is incident on port 1,4. We consider the ports 1,4 as input ports and the ports 2,3 as the output ports. To start with the even mode analysis, first we divide the scheme into two halves using the dotted lines A-A' as shown in **Fig. 3.**

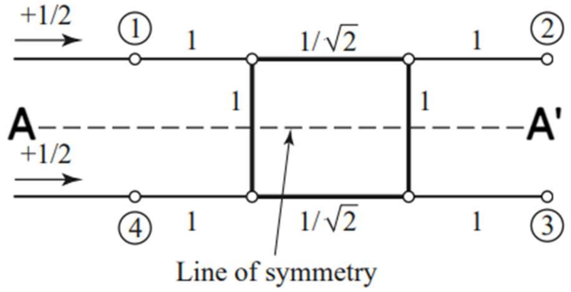

Fig. 3. Symmetric division of the scheme for even-odd mode analysis[1].

In even mode, the input source is divided into two equal halves as per symmetry and an open circuit is observed as shown in **Fig. 4.** On looking at the top half of the circuit and considering the line of symmetry as a virtual open circuit, we notice that now the transmission line that is divided has becomes λ/8. Also, the input source is divided into two equal halves. We can notice that no current flows through the transmission lines and the signals are in phase. The even mode can also be thought of to be sourced using a common mode signal.

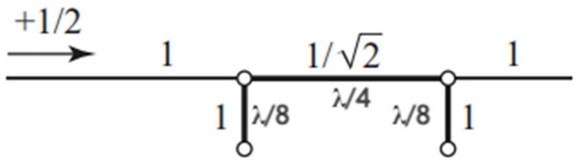

Fig. 4. Half circuit analysis for even mode[1].

Similarly, in odd mode, the input source is divided into two equal halves as per symmetry and a virtual ground is observed as shown in **Fig. 6.** We can notice that maximum current flows through the transmission lines and the signals are out of phase. The odd mode can also be thought of to be sourced using a differential mode signal as shown in **Fig. 5**.

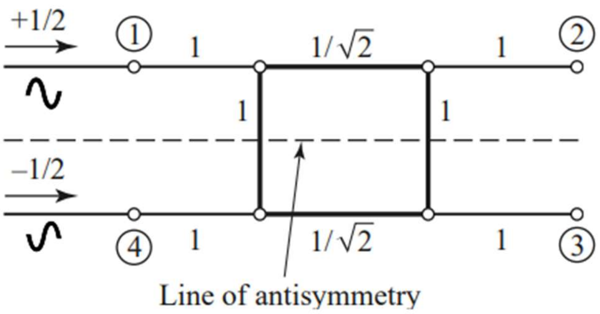

Fig. 5. Antisymmetric division[1].

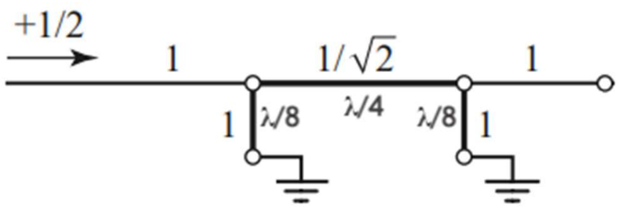

Fig. 6. Half circuit analysis for even mode [1].

The analytical expressions for each of the analysis could be found in [1] & [2]. The purpose of the even odd analysis is to simplify the analysis of the scheme using superposition and hence get the S-parameters by solving for the ABCD parameters of the simplified lines. The corresponding S-Parameters matrix after the analysis is given below.

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

We can see from the S-parameters matrix that the device is ideally lossless and reciprocal. Also, we notice that each row can be easily obtained by transpose of the first.

B. Ideal transmission line implementation

The design of an ideal branchline coupler starts with the choice and calculation of the following parameters:

- **Z_H :** Horizontal quarter wave line impedance;
- **Z_V :** Vertical quarter wave line impedance;
- **E_L**: Electrical length of transmission line (βℓ);
- **Z_o**: Impedance of the 4 ports;
- **f_0**: Operating frequency;

We start our design for **f_0 = 950MHz**. We know, that the quarter wave transmission line has **E_L or βℓ = π/2 or 90°.** We choose the impedance of ports as **Z_o= 50Ω** which gives us **Z_V = Z_o= 50Ω; $Z_H = \frac{Zo}{\sqrt{2}}$ = 35.35Ω;** The coupler was implemented in AWR microwave office using TLIN as shown in **Fig. 7.**

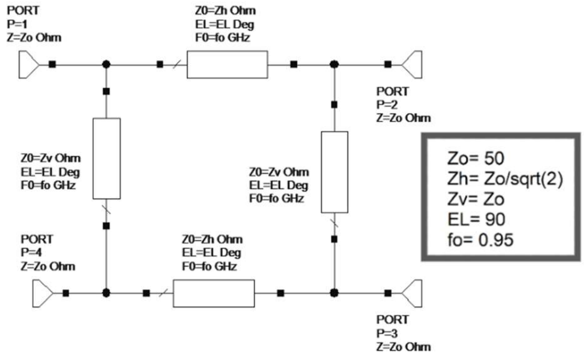

Fig. 7. Ideal transmission line implementation in AWR

On plotting the S-parameters (S11, S21, S31, S41) for the ideal branchline coupler shown in **Fig. 7.** we notice that, at our design frequency of 950MHz, the S11 has an extremely low value < - 75dB, as expected. The S21 and S31 are -3dB as expected for a branchline coupler. This makes it possible for the input power to be routed equally between output port and the coupled port. Also, we notice that at each odd multiples of quarter wavelength we have an identical behavior of our device, here in the **Fig. 8.** we see the behavior at $3xf_0$ i.e., 2.85GHz.

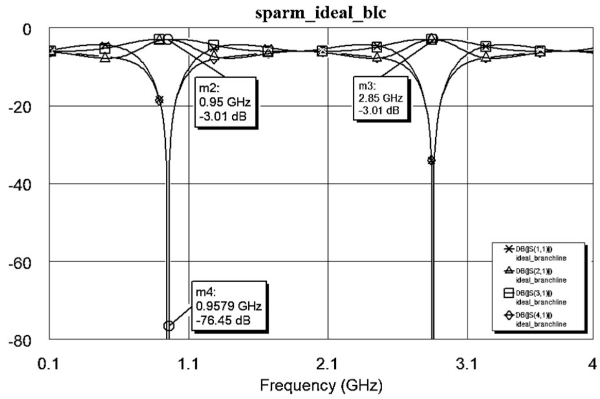

Fig. 8. S-parameters simulation for ideal branchline coupler.

We also see that the output ports 2 & 3 have a 90° phase difference as shown in **Fig. 9.** owing to the property of a quarter wave transformer that creates a 90° phase change from input to output.

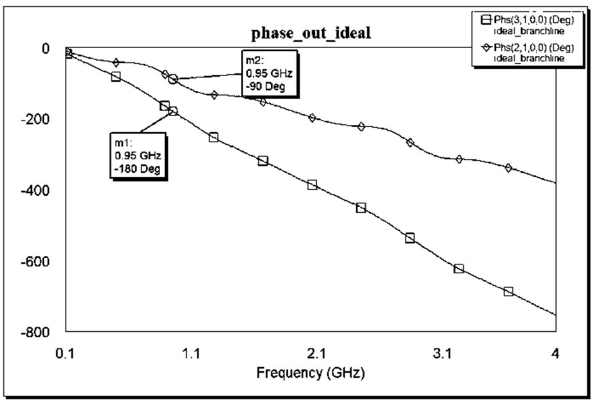

Fig. 9. Phase plot at the output ports with respect to the input port.

C. Microstrip implementation

The design of a real branchline coupler using a microstrip line starts with the choice and calculation of the following parameters:

- **ε_r :** Dielectric constant;
- **T :** Thickness of the substrate;
- **H:** Height of the substrate;
- **σ:** Conductivity of material used;
- **Conductor:** Cu
- **Loss tangent:** 0.0001

We start with finding the length and width of the microstrip line using TX-Line tool [3]. We used the $\boldsymbol{\varepsilon_r = 4.5}$ for our selected FR4 copper clad sheet which was taken as the substrate for the design. The H and T was taken as 1.53mm and 18um respectively. Copper was selected to be the conductor. At the design frequency of 950MHz and E_L of 90°, we use the tool to calculate the length and width of the lines for the required characteristic impedances. The extracted horizontal line length(Lh) and width(Wh) corresponding to Z_H was 41.72mm & 4.89mm respectively. Similarly, the extracted vertical line length(Lv) and width(Wv) corresponding to Z_V was 42.8mm & 2.85mm respectively. The layout for the designed coupler implementation is given in **Fig. 10.**

Fig. 10. Layout of the microstrip implementation

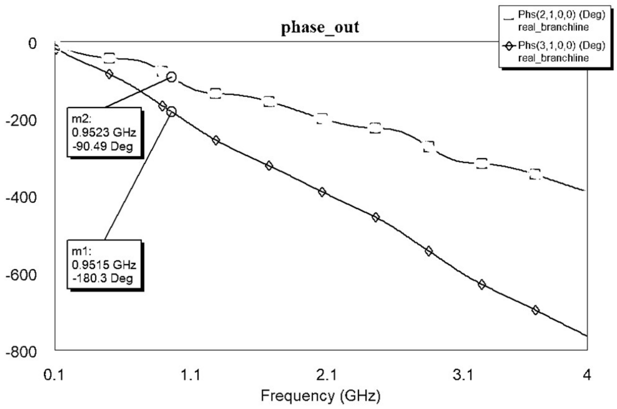

Fig. 11. Phase plot for the real transmission line implementation.

Since, the horizontal lines have lower characteristic impedance, they are wider compared to the vertical lines. The S-parameters and the phase for the microstrip implementation are seen to be comparable to the ideal transmission line one as shown in **Fig. 11.**

III. LOW PASS FILTER DESIGN

We start the design of a 3-element prototype LPF as shown in **Fig. 12.** We proceed to design our filter having maximally flat or Butterworth response.

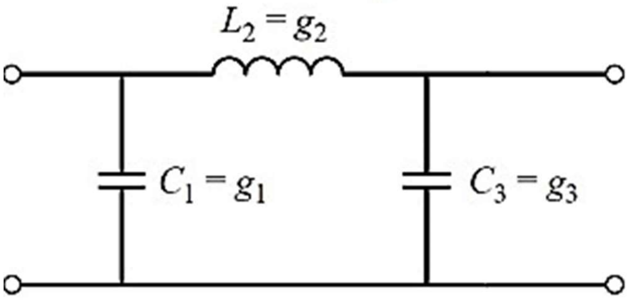

Fig. 12. Prototype LPF in 3 element pi model [1].

From the table)[7] in **Fig. 13.** we find that the normalized design elements have values as g0=g4=1; g1=1; g2=2; g3=1. This filter assumes a normalized load g4 which is equal to the source of g0=1Ω with a cutoff angular frequency of ω_0 = 1 rad/sec.

N	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}	g_{11}
1	2.0000	1.0000									
2	1.4142	1.4142	1.0000								
3	1.0000	2.0000	1.0000	1.0000							
4	0.7654	1.8478	1.8478	0.7654	1.0000						
5	0.6180	1.6180	2.0000	1.6180	0.6180	1.0000					
6	0.5176	1.4142	1.9318	1.9318	1.4142	0.5176	1.0000				
7	0.4450	1.2470	1.8019	2.0000	1.8019	1.2470	0.4450	1.0000			
8	0.3902	1.1111	1.6629	1.9615	1.9615	1.6629	1.1111	0.3902	1.0000		
9	0.3473	1.0000	1.5321	1.8794	2.0000	1.8794	1.5321	1.0000	0.3473	1.0000	
10	0.3129	0.9080	1.4142	1.7820	1.9754	1.9754	1.7820	1.4142	0.9080	0.3129	1.0000

Fig. 13. Element Values for Maximally Flat Low-Pass Filter Prototypes (g0 = 1, ωc = 1, N = 1 to 10)[7]

We would like to design our filter to have a cutoff frequency fc = 25MHz which is the IF frequency for our mixer. The input and output ports of the LPF will have an impedance of 50Ω. Consider the Formulas for the elements as per the table in **Fig. 14.**

$R = \left(\frac{R_0}{R_0'}\right) R'$	$L = \left(\frac{R_0}{R_0'}\right)\left(\frac{\omega_1'}{\omega_1}\right) L'$	$C = \left(\frac{R_0'}{R_0}\right)\left(\frac{\omega_1'}{\omega_1}\right) C'$

Fig. 14. Expressions for values of elements for prototype LPF[7]

We use these formulas given in [7], to derive the values for our elements as given in **Fig. 12.** implicitly in the AWR Microwave Office. We find that for our desired fc, we have (C1)g1= (C2) g3= 127.3pF and (L2)g2= 636nH. We further simulate for the S21 and S11 of the filter as given in **Fig. 15.** As expected, the Insertion Loss or S21 is around 0dB in the passband and gets attenuated to low levels of around -40dB in the stopband.

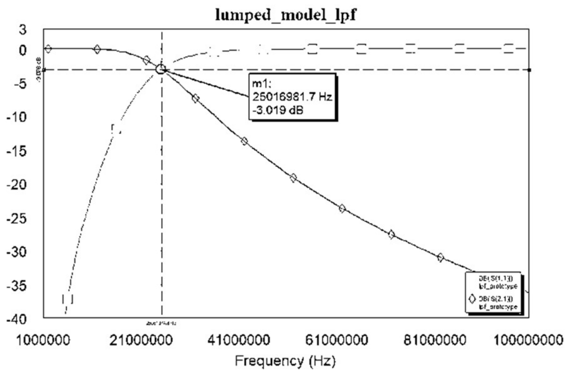

Fig. 15. Insertion and return loss plots for prototype LPF

IV. BALANCED MIXER

We design an RF down converting mixer that uses the quadrature phase output property of the branchline coupler, a non-linear voltage to current converter and a low pass filter as shown in **Fig. 1.** The input LO and RF are provided at port 1&4 respectively. The RF input is split as sine and cosine versions in ports 2&3, similarly LO is split in the same way. Hence, at each terminal of the diode we have sin(RF)+cos(LO) and sin(LO)+ cos(RF).

The voltage to current conversion function is performed by the diode using the non-linear representative equation of the diode current.

$$I_o = e^{\frac{Vin}{\eta V_T}}$$

Where V_T is the thermal voltage under ambient conditions and Vin is the voltage input seen by the diodes. This equation can be expanded in Taylor series form and can be written like:

$$e^{A\,Vin} = A0 + \frac{A1\,Vin}{1!} + \frac{A2\,Vin^2}{2!} + \frac{A3\,Vin^3}{3!} + \cdots$$

Where Vin ~ [$\boldsymbol{sin(\omega_{RF}\, t) + cos(\omega_{LO}\, t)}$]

We can see that the Taylor expansion gives us a lot of harmonic components of Vin. The first component is the largest component, and we try to cancel it out using a combination of two oppositely connected diodes or balanced diodes. Then we use the **Vin²** term because we can write it in the form:

$$\begin{aligned} Vin^2 &= [sin\,(\omega_{RF}t) + cos\,(\omega_{LO}t)]^2 \\ &= sin^2\,(\omega_{RF}t) + cos^2\,(\omega_{LO}t) + 2sin\,(\omega_{RF}t)\,cos\,(\omega_{LO}t) \\ &= \frac{1}{2}[1 - cos\,(\omega_{RF}t) + 1 + cos\,(\omega_{RF}t) + sin\,(\omega_{RF}t + \omega_{LO}t) + \boxed{sin\,(\omega_{RF}t - \omega_{LO}t)} \end{aligned}$$

We would use the last part of the **Vin²** equation to help us get our down converted IF frequency. Other harmonic components are filtered out using the low pass filter. The architecture shown in **Fig. 1** was implemented in AWR Microwave Studio. The schematic for the same is given in **Fig. 16.**

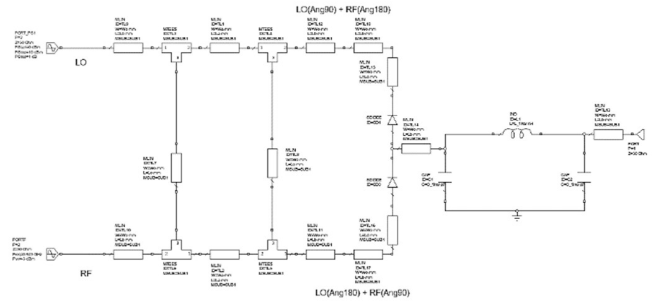

Fig. 16. Schematic for Balanced Mixer.

The layout as shown in **Fig. 17.** was done in microstrip line implementation was generated and we have a few discrete components used for the LPF in 805 SMD package.

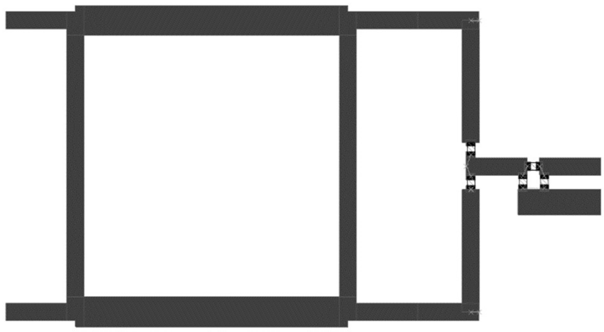

Fig. 17. Layout for Balanced Mixer.

After performing a LSSnm simulation in AWR Microwave studio for the microstrip implementation of our balanced mixer we found that the conversion loss of the mixer was around -7.9dBm as shown in **Fig. 18.**

As the RF frequency used in the project is 925MHz and the LO frequency is 900MHz, we expect our IF to be around 25MHz. The frequency domain power simulation was done on the implementation of the circuit without a Low Pass Filter(LPF). This simulation gave out the expected IF tone of about -15dBm at 25MHz. The problem was that, the LO tone of around -9dBm which was stronger than our desired IF tone was seen at the output. This effect is shown in **Fig. 19.**

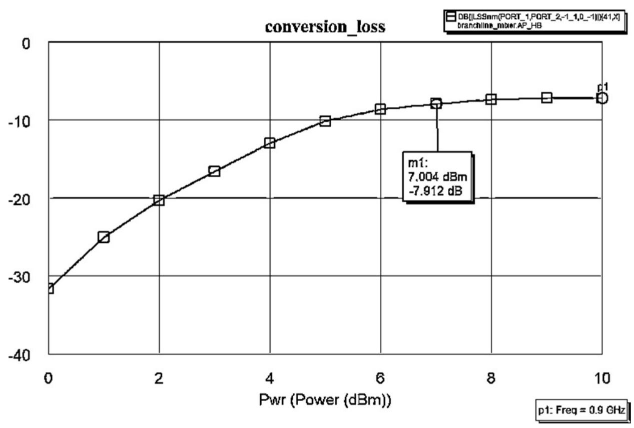

Fig. 18. Conversion loss simulation.

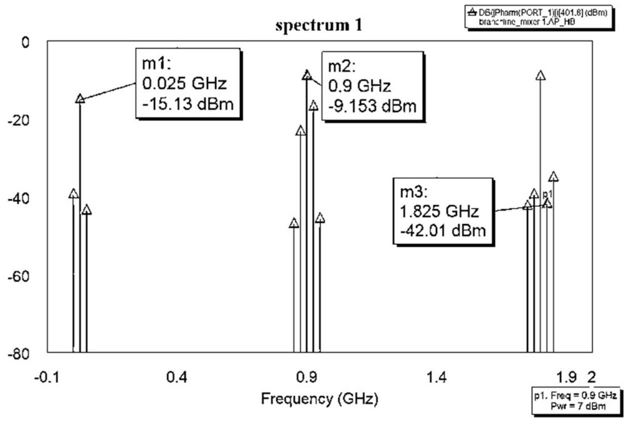

Fig. 19. Output spectrum without LPF

The LPF designed in Section III was used as per the architecture in **Fig.1.** We see that the spectrum looks a lot better. The LO tone is significantly attenuated by our LPF to about -101dBm. At our preferred IF frequency we get a tone of about 14dBm. The output spectrum of the final design is shown in **Fig. 20.**

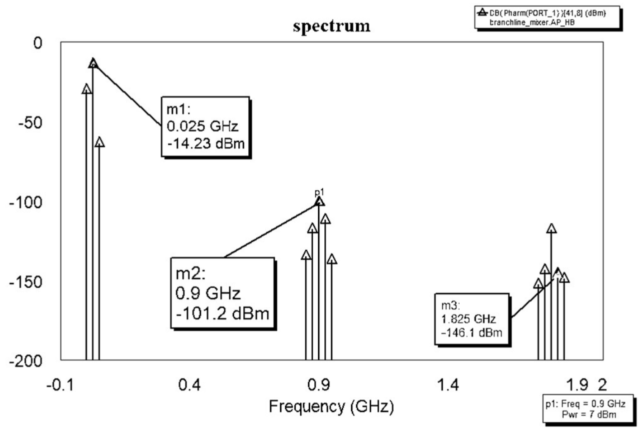

Fig. 20. Final output spectrum of the balanced mixer

V. CONCLUSION

In this project we were able to build a balanced mixer using a branchline coupler and a 3-element Low Pass Butterworth Filter. We achieved a conversion loss of around -7.9dBm. We achieved our desired IF of 25MHz having a tone of around 14dBm. The spurious LO signals at 900MHz were attenuated to a tone of about -101dBm.

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