A Low-Cost Millimeter-Wave Six-Port Double-Balanced Mixer

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Abstract - **A low cost millimeter-wave narrowband six-port double-balanced mixer circuit is proposed. Due to the symmetry and specific properties of the six-port, the mixer exhibits very good suppression of harmonic and spurious products. The mid operating frequency is selected to be 77 GHz, which is dedicated to collision avoidance radar for automobile applications. The simulation results show a conversion loss less than 4 dB and a very good linearity over a wide input power range. The 1 dB compression point and the third-order input intercept point are respectively 7.25 and 17 dBm.**

Index Terms—**Double-balanced mixer, frequency conversion, millimeter wave, six-port.**

I. INTRODUCTION

Mixers are the key components in converters and receivers and it is essential to find solutions which are cost effective in mass production. The most important characteristic of a mixer used in frequency converters or receivers is the spurious performance. Spurious signals generated in the IF frequency band are impossible to be rejected by any filters [1]. Therefore their suppression is achieved by using balanced structures [2].

Balanced mixers result from the symmetric and anti-symmetric pairing of individual basic mixers. These configurations provide an effective means to suppress or, more realistically, to attenuate some unwanted frequency components in the spectra of the input and output signals [2].

As such, the conventional double-balanced mixers provide good isolation between all the ports, as well as all even harmonics of RF and LO signal rejection. They also provide a higher third-order intercept point than a single ended or a single balanced mixer [3].

The symmetric and anti-symmetric pairing of diodes in the double-balanced mixers is achieved traditionally by baluns or hybrids, which significantly affect the bandwidth and the overall performance of the mixer [2]-[7]. The use of baluns in the design of double-balanced mixers involves the use of air bridges. In MHMIC technology, air bridges are workable only at low microwave frequencies. At high microwave and millimeter-wave frequencies their implementation with good performances is difficult to achieve. A solution of avoiding air bridges was proposed in [7]. A double-sided microstrip implementation was done using vias instead of air bridges. Both air bridges and vias are workable only at low microwave frequencies. An alternative way to overcome this problem is the use of six-ports in the design of double-balanced mixers. The applications of the six-port technology in direct conversion receivers with good performance have been demonstrated in [8]-[9].

This paper is focused on the design of six-port doubled-balanced mixers. These mixers avoid the use of air bridges in the RF and LO paths; therefore, they are more suitable for down-conversion at millimetre-wave frequencies.

II. SIX-PORT DOUBLE-BALANCED MIXER

Fig. 1 shows the proposed six-port dedicated to the double-balanced mixer along with the LO and RF phase relationships at the output ports. This circuit is composed by two 180^o and two 90^o hybrids couplers. The 180^o hybrid couplers can be replaced by two 90^o hybrid couplers and two 90^o phase shifters.

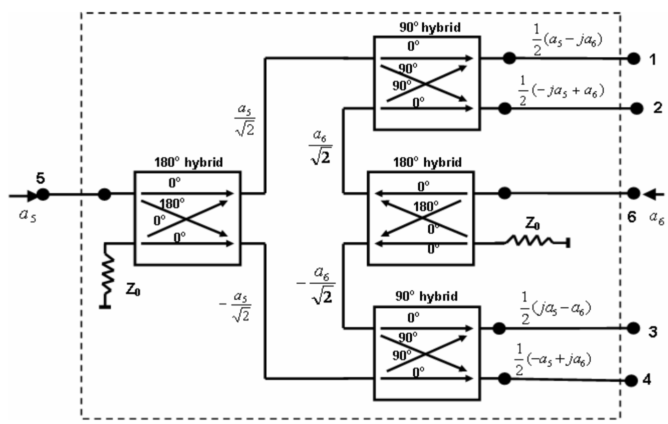

Fig. 1. The six-port for the double balanced mixer

The schematic of the six-port double-balanced mixer is shown in Fig. 2. To provide a ground return at the IF frequencies, four λ/4 short-circuit stubs are used. As known, these stubs are open circuits for RF and LO signal. In addition, a λ/4 open-circuit stub is connected to the mixer output preventing LO and RF leakage in the IF path.

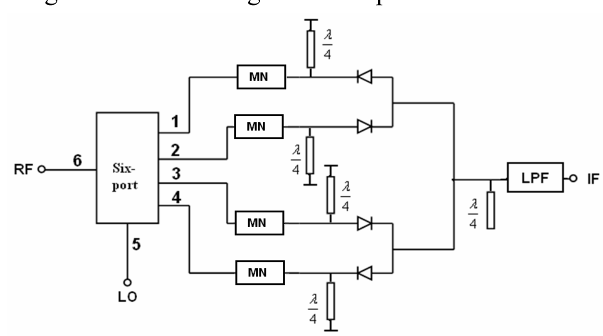

Fig. 2.The schematic of the six-port double-balanced mixer

III. Harmonic and Spurious Product Suppression

Figure 3 shows the diodes connection along with the relationship between incoming and outgoing waves. We assume the ideal case that there is no reflection at the input of each diode, thus $a_1 = a_2 = a_3 = a_4 = 0$. In this case the total voltages at the input of the diodes are given as follows:

$$\mathrm{v}_1 = \sqrt{Z_0}b_1, \quad \mathrm{v}_2 = \sqrt{Z_0}b_2, \quad \mathrm{v}_3 = \sqrt{Z_0}b_3, \quad \mathrm{v}_4 = \sqrt{Z_0}b_4, \tag{1}$$

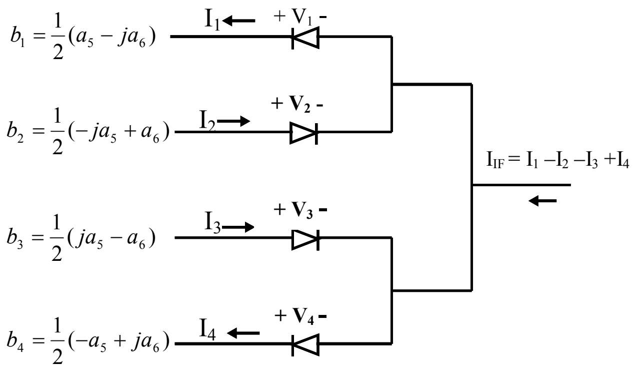

Fig. 3. The diode connections

By injecting the local oscillator signal at port 5, the RF signal at port 6 and neglecting their afferent reflections, we obtain:

$$v_5 = \sqrt{Z_0}a_5 = 2V_{LO}\cos\omega_{\mathrm{LO}}\mathrm{t}, \qquad \mathrm{v}_6 = \sqrt{Z_0}a_6 = 2V_{RF}\cos\omega_{RF}t, \tag{2}$$

We introduce the factor 2 to simplify the calculus. Taking into account the relationships given in Fig. 3, the voltages at the four outputs are obtained as follows:

$$v_1 = V_{LO}\cos\omega_{LO}t + V_{RF}\cos(\omega_{RF}t - 90^o) = V_{LO}\cos\omega_{LO}t + V_{RF}\sin\omega_{RF}t \tag{3a}$$

$$v_2 = V_{LO}\cos(\omega_{LO}t - 90^o) + V_{RF}\cos\omega_{RF}t = V_{LO}\sin\omega_5 t + V_{RF}\cos\omega_{RF}t \tag{3b}$$

$$v_3 = -V_{LO}\cos(\omega_{LO}t - 90^o) - V_{RF}\cos\omega_6 t = -v_2 \tag{3c}$$

$$v_4 = -V_{LO}\cos\omega_{LO}t - V_{RF}\cos(\omega_{RF}t - 90^o) = -v_1 \tag{3d}$$

Assuming the following I/V characteristic of the diodes:

$$I(v) = a_0 + a_1 v + a_2 v^2 + a_3 v^3 + a_4 v^4 + \dots\dots\dots\dots, \tag{4}$$

the currents through the diodes are [2]-[4]:

$$I_1 = I(-v_1) = a_0 - a_1 v_1 + a_2 v_1^2 - \dots. + a_{2n}v_1^{2n} - a_{2n+1}v_1^{2n+1} + \dots. \tag{5a}$$

$$-I_2 = -I(v_2) = -a_0 - a_1 v_2 - a_2 v_2^2 - \dots - a_{2n}v_2^{2n} - a_{2n+1}v_2^{2n+1} - \dots \tag{5b}$$

$$-I_3 = -I(-v_2) = -a_0 + a_1 v_2 - a_2 v_2^2 + \dots - a_{2n}v_2^{2n} + a_{2n+1}v_2^{2n+1} - \dots. \tag{5c}$$

$$I_4 = I(v_1) = a_0 + a_1 v_1 + a_2 v_1^2 + \dots + a_{2n}v_1^{2n} + a_{2n+1}v_1^{2n+1} + \dots. \tag{5d}$$

The total output current of the mixer is given by:

$$I_{IF} = I_1 - I_2 - I_3 + I_4 = 2a_2(v_1^2 - v_2^2) + \dots\dots + 2a_{2n}(v_1^{2n} - v_2^{2n}) + \dots. \tag{6}$$

$$\begin{aligned} I_{IF} &= 2a_2 V_{LO}^2(\cos^2\omega_{LO}t - \sin^2\omega_{LO}t) + \\ &+ 2a_2 V_{RF}^2(\sin^2\omega_{RF}t - \cos^2\omega_{RF}t) + \\ &+ 4a_2 V_{LO}V_{RF}(\cos\omega_{LO}t\sin\omega_{RF}t - \sin\omega_{LO}t\cos\omega_{RF}t) + \dots\dots \end{aligned}$$

$$\begin{aligned} I_{IF} &= 2a_2 V_{LO}^2\cos 2\omega_{LO}t - 2a_2 V_{RF}^2\cos 2\omega_{RF}t - \\ &- 4a_2 V_{LO}V_{RF}\sin(\omega_{LO} - \omega_{RF})t + \dots \end{aligned} \tag{7}$$

This result shows that, the DC current and the odd order harmonics and intermodulation products are suppressed. The product $\omega_{LO} + \omega_{RF}$ is also suppressed. This means that the six-port double-balanced mixer cannot be used for frequency upconversion.

Given that the mixer is narrowband, the RF and LO frequencies are close from one another. Consequently, some of their intermodulation products of the form mf_{LO}-mf_{RF} are in-band and it can be difficult to filter them. For this reason we must see whether these products are suppressed or not.

Let us evaluate the terms $2a_{2n}(v_1^{2n} - v_2^{2n})$ in (6).

$$\begin{aligned} 2a_{2n}v_1^{2n} &= 2a_{2n}(V_{LO}\cos\omega_{LO}t + V_{RF}\sin\omega_{RF}t)^{2n} \\ &= 2a_{2n}\sum_{i=0}^{2n}\binom{2n}{i}V_{LO}^{2n-i}V_{RF}^{i}\cos^{2n-i}\omega_{LO}t\sin^{i}\omega_{RF}t \\ &= 2a_{2\mathrm{n}}(V_{LO}^{2n}\cos^{2n}\omega_{LO}t + V_{RF}^{2n}\sin^{2n}\omega_{RF}t + \dots \\ &\quad + \binom{2n}{n}V_{LO}^{n}V_{RF}^{n}\cos^{n}\omega_{LO}t\sin^{n}\omega_{RF}t + \dots \end{aligned} \tag{8}$$

Where $\binom{2n}{i} = \frac{(2n)!}{i!(2n-i)!}$ are binomial coefficients

$$\begin{aligned} \cos^n\omega_{LO}t\sin^n\omega_{RF}t &= (\cos\omega_{LO}t\sin\omega_{RF}t)^n \\ &= \frac{1}{2^n}\left[\sin(\omega_{LO} + \omega_{RF})t - \sin(\omega_{LO} - \omega_{RF})t\right]^n \\ &= \frac{1}{2^n}\left[\sin n(\omega_{LO} + \omega_{RF})t + (-1)^n\sin n(\omega_{LO} - \omega_{RF})t\right] + \dots. \end{aligned}$$

$$\begin{aligned} 2a_{2n}v_1^{2n} &= 2a_{2\mathrm{n}}\Big(V_{LO}^{2n}\cos^{2n}\omega_{LO}t + V_{RF}^{2n}\sin^{2n}\omega_{RF}t + \dots \\ &\quad + \frac{1}{2^{2n}}\binom{2n}{n}V_{LO}^{n}V_{RF}^{n}\left(\left[\sin n(\omega_{LO} + \omega_{RF})t + (-1)^n\sin n(\omega_{LO} - \omega_{RF})t\right] + \dots\right) \end{aligned} \tag{9}$$

In the same manner, we obtain:

$$\begin{aligned} 2a_{2n}v_2^{2n} &= 2a_{2\mathrm{n}}\Big(V_{LO}^{2n}\sin^{2n}\omega_{LO}t + V_{RF}^{2n}\cos^{2n}\omega_{RF}t + \dots.. \\ &\quad + \frac{1}{2^{2n}}\binom{2n}{n}V_{LO}^{n}V_{RF}^{n}\left(\left[\sin n(\omega_{LO} + \omega_{RF})t + \sin n(\omega_{LO} - \omega_{RF})t\right] + \dots\right) \end{aligned} \tag{10}$$

$$\begin{aligned} 2a_{2n}(v_1^{2n} - v_2^{2n}) &= 2a_{2\mathrm{n}}\Big(V_{LO}^{2n}\cos 2n\omega_{LO}t + V_{RF}^{2n}\cos 2n\omega_{RF}t + \\ &\quad + \frac{1}{2^{2n}}\binom{2n}{n}V_{LO}^{n}V_{RF}^{n}\left[(-1)^n - 1\right]\sin n(\omega_{LO} - \omega_{RF})t + \dots..\Big) \end{aligned} \tag{11}$$

The IF current becomes:

$$\begin{aligned} I_{IF} &= \sum_{n=1}^{\infty}2a_{2\mathrm{n}}\Big(V_{LO}^{2n}\cos 2n\omega_{LO}t + V_{RF}^{2n}\cos 2n\omega_{RF}t + \\ &\quad + \frac{1}{2^{2n}}\binom{2n}{n}V_{LO}^{n}V_{RF}^{n}\left[(-1)^n - 1\right]\sin n(\omega_{LO} - \omega_{RF})\Big) + \dots \end{aligned} \tag{12}$$

Table 1 summarizes the theoretical results of the harmonic and spurious product suppression of the six-port double-balanced mixer in comparison of those of conventional double balanced mixer [3], [4]. The "+" and "-" signs indicate a suppressed or, respectively non-suppressed product. According to these results, the double-balanced six-port mixer is more suitable for down conversion than conventional one, at millimeter-wave frequencies.

TABLE I: HARMONICS AND SPUROUS PRODUCT SUPPRESSION

Harmonic and spurious products	Six port double-balanced mixer	Conventional double-balanced mixer	Comments
DC	+	+	Both suppress DC current
f_{LO}	+	+	
f_{RF}	+	+	
f_{LO}-f_{RF}	-	-	Both are downconverter
f_{LO}+f_{RF}	+	-	The six-port mixer is not an upconverter
Odd order harmonics	+	+	
Even order harmonics	-	+	
Odd order spurious	+	+	
m.f_{LO}+m.f_{RF}, m odd	+	-	
m.f_{LO}+m.f_{RF}, m even	+	+	
m.f_{LO}-m.f_{RF}, m even	+	+	
-m.f_{LO}+m.f_{RF}, m even	+	+	
m.f_{LO}-m.f_{RF}, m odd	-	-	

IV. SIMULATION RESULTS

Simulations are performed using Advanced Design System software of Agilent Technologies. A computer model of the six-port circuit is used. The circuit is realized in MHMIC technology on a 125 µm substrate having a relative permittivity of 9.9.

Table II gives the values of the S parameters at 77 GHz. According to these results, the six-port exhibits good performances. The return loss at all ports is better than 24 dB. The isolation between the input ports (port 5 and 6) is 30.5 dB and the output ports (port 1 to 4) are well balanced.

TABLE II THE S PARAMETERS OF THE SIX-PORT AT 77 GHZ

S parameter	Amplitude [dB]	Phase [degré]
S11	-28.26	-
S22	-25.50	-
S33	-27.71	-
S44	-27.50	-
S55	-24.28	-
S66	-25.00	-
S16	-6.244	14.706
S26	-6.301	104.941
S36	-6.145	-72.370
S46	-6.132	-163.086
S15	-6.280	111.133
S25	-6.234	20.265
S35	-6.144	-156.045
S45	-6.169	-66.042
S56, S65	-30.5	-

Fig. 4 shows the spectrum at the IF output port. All the harmonics are suppressed. Their levels are at least 130 dB under the IF signal level. This fact is to be explained that, beside suppression of harmonics due to the balanced structure of the mixer, the six-port itself acts as a filter.

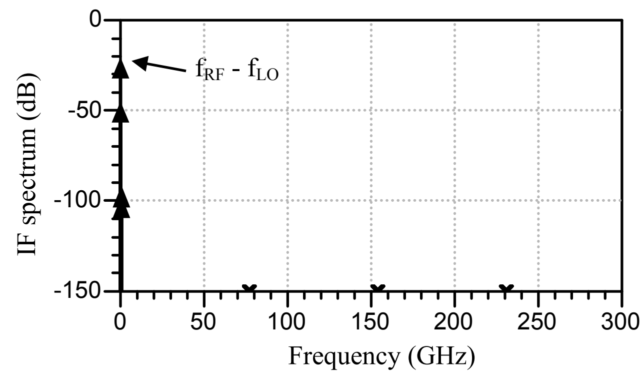

Fig. 4. Output spectrum

Fig. 5 represents the zooming of the Fig. 4 in the neighborhood of the IF frequency. The levels of all spurious products are at least 70 dB under the IF signal level and their filtering with an extra filter is no more necessary. The DC level which is -50 dB (27 dB under the level of the IF signal) is also shown in this figure. The presence of the DC current at the output is due to the fact that the six-port and the diodes are not ideal. Therefore, all the four outputs of the six-port are not ideally balanced.

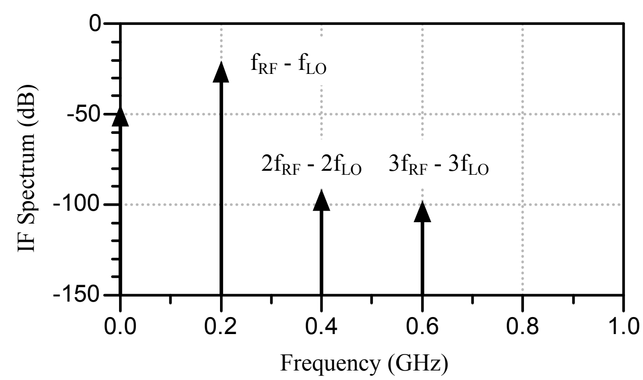

Fig. 5. The level of spurious products in the neighborhood of the IF signal

Figs. 6 and 7 show the conversion loss value versus the LO power and RF frequency, respectively. In order to improve the conversion loss, the LO power must be increased. Less than 4 dB conversion losses are obtained using a 10 dBm LO signal, as shown in Fig. 6. In addition, Fig. 7 shows a minimal conversion loss at central operating frequency and less than 6 dB conversion loss in a 1 GHz bandwidth. For comparison, simulations performed on single and double-balanced microwave diode mixers in [1] and [3]-[7] show minimal conversion loss greater or equal 6 dB.

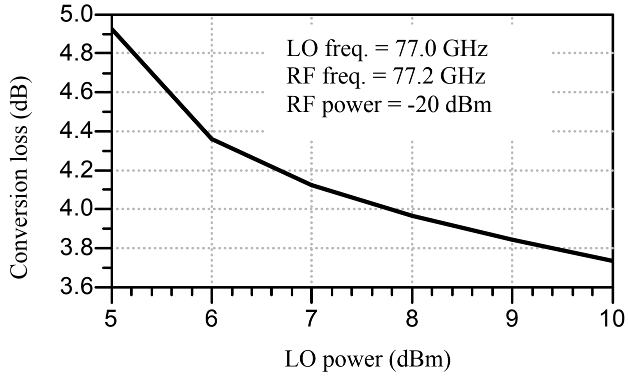

Fig. 6. The conversion loss versus LO power

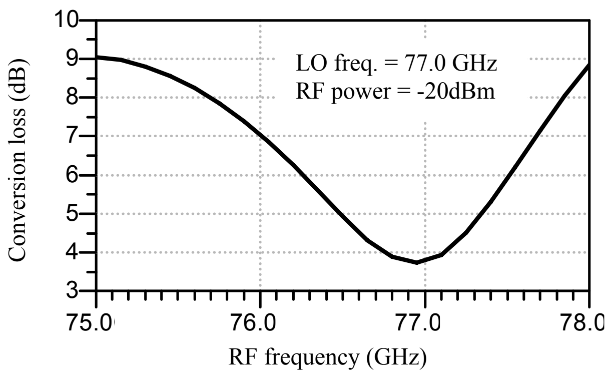

Fig. 7. The conversion loss versus RF frequency

Figs. 8 and 9 show conversion gain results versus the RF power for a 10 dBm LO signal. As seen, for a RF power less than 5 dBm, a very good linearity is obtained. If the RF input power increases above this value, the conversion loss also increases due to the increased harmonics and spurious product power. The 1-dB compression point is reached at 7.25 dBm input power (as seen in Fig. 9).

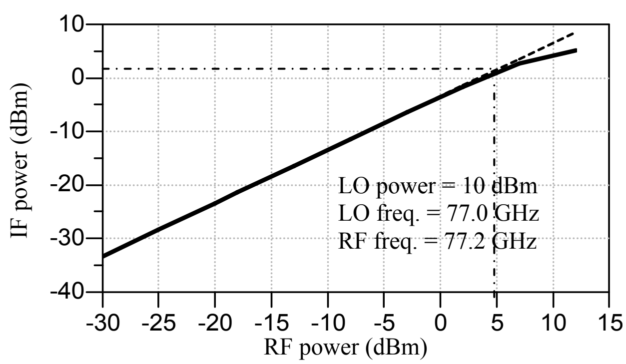

Fig. 8. The IF power versus RF power

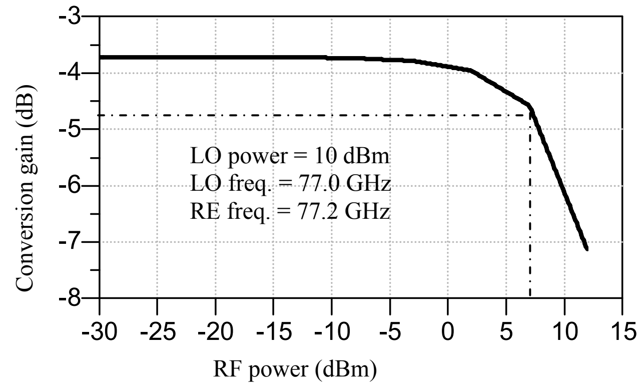

Fig. 9. The gain compression curve

V. CONCLUSION

Theoretical and simulation results show that the proposed six-port double-balanced mixer exhibits good performances at millimeter-wave frequencies. Its architecture allows to avoid air bridges at RF and LO paths, making easier the implementation of this circuit compared to the conventional one at these frequencies. The analysis of harmonics and intermodulation suppression shows that this six-port double-balanced mixer cannot be used for frequency upconversion.

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