V-band Six-Port Down-conversion Techniques

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Abstract-**This paper presents comparative down-conversion results using two different topologies. The conventional approach using anti-parallel diodes acting as LO-driven switches is compared with the six-port one. The proposed down-converter which provides an intermediate frequency of 900 MHz uses a six-port circuit and two pairs of parallel Schottky diodes acting as power detectors connected to two differentials amplifiers. Simulation results of down-converter passive circuits in terms of I/Q IF power and phase are presented and discussed. In addition, comparative analyses are performed in terms of local oscillator and RF power required for the frequency conversion and phase stability. These results show the advantages of the proposed frequency conversion six-port technique.**

***Index Terms*—**Millimeter wave mixers, Multiport circuits, Schottky diode.

I. Introduction

The "Wireless Revolution" has created a large number of new opportunities; the design of compact and low cost wireless millimeter-wave communication front-end, which can offer convenient terminal mobility and high capacity channels. In order to meet the increasing demand for high data rates, the IEEE 802.15.3c industrial standard based on millimeter–wave technology has been recently introduced for WPLAN. In the last decade some initial research has been made, especially in terms of designing new millimeter wave components operating over the V-band [1]-[5]. In order to improve performances of the communications receivers, many architectures of front end based on multi-port mixer was explored and designed [3]-[6].

The six-port technology, originally used to perform complex reflection coefficient measurement of a microwave device, is an inexpensive alternative to the network analyzer. It is an excellent candidate for the future millimeter-wave wireless communication systems.

Recently, the six-port technology is performing the millimeter-wave frequency conversion, avoiding the conventional millimeter-wave active costly mixers [7].

This paper addresses the two topologies of down converter based on the six-port circuit, constructed with four hybrid microstrip ring couplers interconnected with transmission lines. A 125 microns ceramic substrate having a relative permittivity of 9.9 is used. This passive structure is optimized to operate at 60 GHz, using ADS (Advanced Design System of Agilent Technologies). Excellent return loss, isolation between RF/LO inputs and transmission S parameters were obtained [8].

The main purpose of this paper is to show that the down-conversion architecture using two pairs of parallel diodes and differential amplifiers can improve mixing signals in terms of conversion loss and I/Q phase stability regarding the quadratic reference of 90^0, versus the swept of the power of the local oscillator and the RF signal. In these simulations the gain of amplifiers for both mixers is fixed at the same value of 10 dB.

II. Six-Port Architecture

It is well known that the mixer based on six-port passive circuit is performing down conversion instead of the conventional active one, for its well known advantages in terms of quality and low-cost [8].

The six-port block diagram is shown in Figure 1. The circuit is composed of four 90 hybrid couplers and a 90^0 phase shifter. Two hybrids couplers acting as power divider of LO (a_5) and RF (a_6) signals. The two others provide a plurality of different linear combinations of the signals that derives from the power dividers.

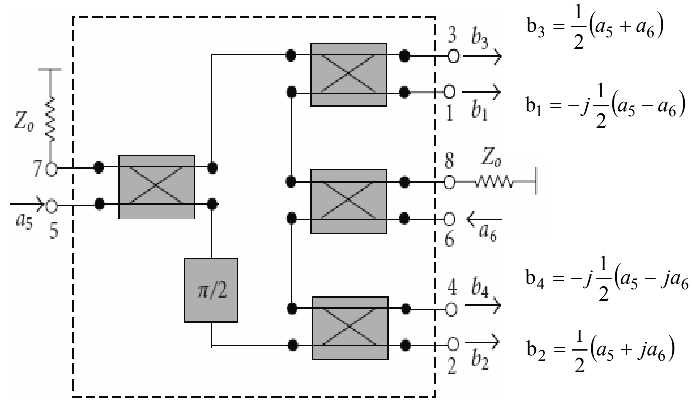

Fig. 1. Six-port block diagram

Let us assume that there are two normalized wave inputs a_5 (LO) and a_6 (RF) with different amplitudes and phases,

$$a_5 = a \cdot \exp[j(\omega_0 \cdot t + \varphi_5)] \tag{1}$$

$$a_6 = \alpha(t) \cdot a \cdot \exp[j(\omega \cdot t + \varphi_6)] \tag{2}$$

Therefore, we can calculate the normalized wave outputs using six-port S-parameters [8].

$$b_i = a_5 \cdot S_{5i} + a_6 \cdot S_{6i} \quad \text{for i =1, 2, 3, 4} \tag{3}$$

The low IF outputs signals are obtained using for power detectors connected to the six-port outputs. As known, the output voltage of an ideal power detector is proportional to the square magnitude of the RF input signal.

$$V_i = K_i |b_i|^2 \qquad \text{for } i = 1, 2, 3, 4 \qquad (4)$$

where K_i are constants and measured in V/W. Considering four identical detectors ($K_i = K$, for i = 1 to 4), the output detected voltages can be obtained as follows:

$$V_{1,3} = K\frac{a^2}{4}\cdot\left\{1+\alpha(t)^2 -/+ 2\cdot\alpha(t)\cdot\cos[-\Delta\omega\cdot t+\Delta\varphi(t)]\right\} \quad (5)$$

$$V_{2,4} = K\frac{a^2}{4}\cdot\left\{1+\alpha(t)^2 -/+ 2\cdot\alpha(t)\cdot\sin[-\Delta\omega\cdot t+\Delta\varphi(t)]\right\} \quad (6)$$

In the previous equation, $\Delta\omega = \omega_0 - \omega$ represents the frequency difference between the six-port inputs (LO and RF), and $\Delta\varphi = \varphi_6 - \varphi_5$ is the phase difference between the same signals. Using equations (5) and (6), and considering the differential approach, the output I/Q signals are:

$$i(t) = V_3(t) - V_1(t) = K\cdot\alpha(t)\cdot|a|^2\cdot\cos[-\Delta\omega\cdot t+\Delta\varphi(t)] \quad (7)$$

$$q(t) = V_4(t) - V_2(t) = K\cdot\alpha(t)\cdot|a|^2\cdot\sin[-\Delta\omega\cdot t+\Delta\varphi(t)] \quad (8)$$

The previous equations show that the proposed mixer based on the six-port circuit with four power detectors and two differential amplifiers can successfully generate the I/Q signals.

Down-conversion results using differential approach will be compared to a quasi-conventional one. As known, the block diagram of a conventional I/Q mixer is composed of two mixers and related coupler and power divider (see Fig.2). The 3 dB 90 ^{0} hybrid coupler and the in phase power divider consists on two hybrid ring couplers and a 90 ^{0} phase shifter. We assume that mixer 1 and 2 are single-balanced diode mixer, composed with the 90 ^{0} hybrid couplers and two anti-parallel diodes. The input RF signal is fed to the in phase divider and split before feeding the RF ports of the two double balanced mixers. The LO signal is fed to the hybrid and split with a 90 degree phase shift between the two outputs before feeding the LO ports of the two double balanced mixers. The two double balanced mixers provide the IF outputs (I and Q) that are equal in amplitude, but in quadrature.

Effectively since the quadrature hybrid introduces a 90 ^{0} phase difference, mixer1 multiplies (cos ω_0t)(cos ωt), while mixer2 multiplies (sinω_0t)(cos ωt). The trigonometry results in sum and difference frequencies, of which the difference frequency is chosen. Therefore four hybrid ring couplers and a 90 ^{0} phase shifter are included in this architecture composing a six-port circuit. This quasi-conventional mixer based on the six-port circuit in a double balanced structure with two pairs of anti-parallel diodes can also generate the I/Q signals.

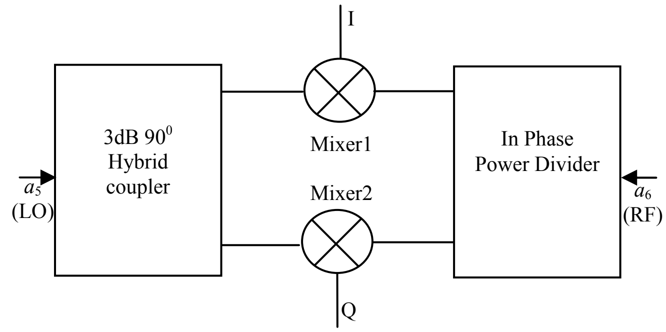

Fig. 2. Block diagram of I/Q conventional mixer

The phase and the magnitude of the transmission S parameters are of main interest to obtain the requested four "qi points" of the six-port circuit (see the block diagram of Fig. 1). The difference of phase between the required transmission S parameters for RF and LO inputs is found to be equal to 90^0 multiples over the frequency band as suggested in Figure 1.

In order to evaluate the losses of the six-port circuit, and then the conversion loss of the circuit a supplementary loss of around 0.3 dB appears at the central frequency compared to the ideal multi-port model. Similar results related to the magnitude of transmission S parameters between the LO input port and the four outputs are also obtained [8].

In order to avoid leakage of LO signal detected in the antenna, excellent isolation between RF inputs and return losses were obtained in a 2 GHz frequency band centered at the 60 GHz operating frequency, return loss is better than 20 dB, and isolation better than 30 dB [8].

III. MIXER TOPOLOGIES

Using the results of previous section we can draw the block diagrams of both I/Q mixers.

The proposed mixer consists of the six-port circuit, two pairs of parallel Schottky diodes acting as power detectors connected to two differential amplifiers, as showed in Fig.3.

The quasi-conventional mixer use couplers and anti-parallel pairs of Schottky diodes acting as LO-driven switches. As seen in section II, in order to provide comparative results with the proposed differential architecture, the same six-port circuit is used instead of conventional mixer couplers. In addition the amplifiers have the same gain as the differential amplifiers of the first mixer. The block diagram of this architecture is illustrated in Fig. 4.

As seen, the main difference between these topologies is the use of differential amplifiers operating at intermediary frequency. The down-conversion in the case of the proposed mixer is implemented without the need of switching diodes reducing the requested power of the LO, as will be demonstrated in the next section.

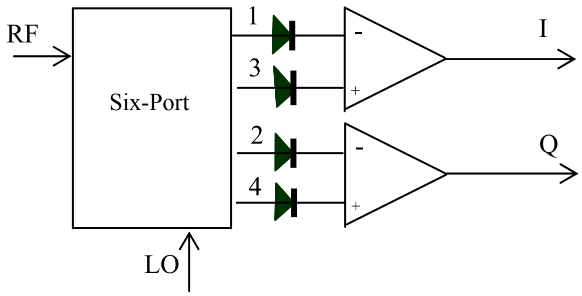

Fig. 3. Mixer using parallel diodes and differential amplifiers

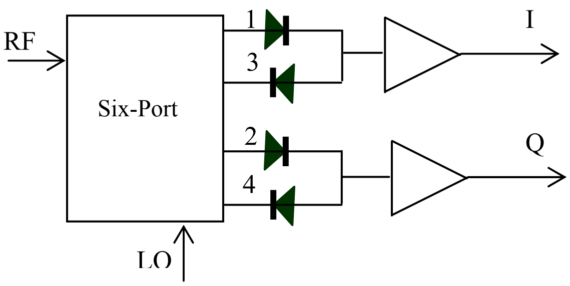

Fig. 4. Mixer using anti-parallel diodes

IV. SIMULATIONS RESULTS

Most modern diode mixer designs use Schottky diodes. The main reason for this is that the Schottky diode is a majority carrier device which means, it has a higher switching speed than p-n junction diodes. The model of the electrical is shown in Figure 5.

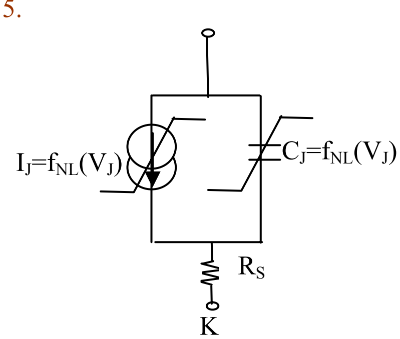

Fig. 5. Schottky diode Non-Linar scheme

The model shows the two non linearity's, the courant source I_j and the junction capacity C_j. R_S series is the parasitic resistance due to the ohmic contact. The I_J courant circulating from the metal to the semi-conductor and driven by V_J is expressed as follows :

$$I_J = I_S\left(\exp\left(\frac{qV}{nKT}\right)-1\right) \tag{9}$$

Where I_0 is the saturation current and n the ideality factor, q is the electron charge, K , the Boltzmann's constant, and T is the absolute temperature.

The junction capacitance C_J wich depends on V_J and the threshold tension V_b is expressed as below:

$$C_J = \frac{C_{J0}}{\sqrt{1-\frac{V_J}{V_b}}} \tag{10}$$

The MS8150-P2613 used here in simulations as nonlinear element, is a GaAs Schottky diode, its high cut-off frequency insures good performance at frequencies to 100 GHz. The spice model parameters are regrouping in the table below.

Table
SPICE MODEL PARAMETERS OF MS8150-P2613

I_S	R_S	N	TT	C_{J0}	M	EG	V_J	B_V	IB_V
A	Ω		s	pF		eV	V	V	A
2.10^{-13}	3	1.2	0	0.045	0.5	1.42	0.85	4	1.10^{-5}

In order to estimate the conversion loss for both mixers, and the stability of I/Q phase, Harmonic Balance simulations are performed using ADS.

In the first step, we analyze the behavior of the two topologies versus the LO drive in a range from 0 dBm to 15 dBm. The RF power is fixed at –15 dBm and the amplifier gain at 10 dB for both mixers.

Through the graphic in Fig. 6, we notice that for all the LO power range, the variation of the output I/Q power of the proposed mixer is only 8 dBm and about 50 dBm for the conventional one. In addition the LO driving power for the quasi-conventional mixer should be at least 10 dBm to be able to move into saturation area.

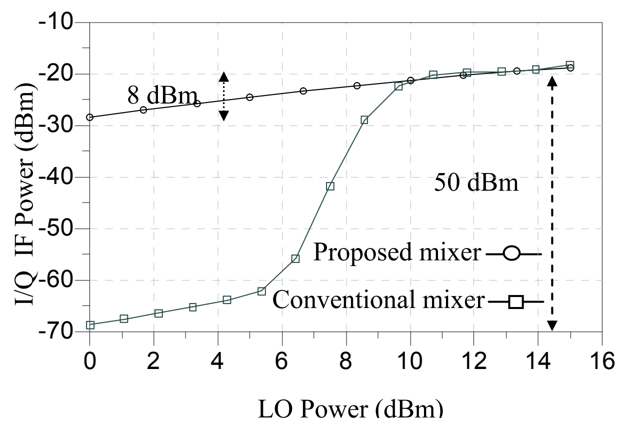

Fig. 6. Simulation Results of Conversion Loss versus LO

In the graphic below Fig.7, we observe that the I/Q phase in the quasi-conventional mixer is about 90^0 until the power of LO of 5 dBm, over that, the I/Q phase is about 60^0, that mean 30^0 of disparity regarding the quadratic reference of 90^0.

The proposed mixer has a disparity of only 0.5^0 in the entire range of LO drive, then I/Q phase is closed to the quadratic reference of 90^0.

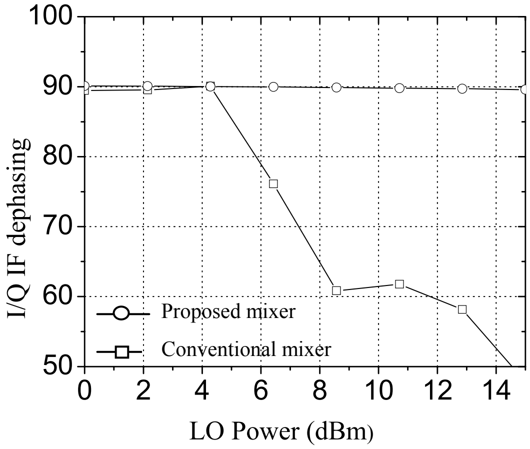

Fig. 7. Simulation Results of I/Q dephasing versus LO

As known, low-cost receivers require reduced LO power, and high I/Q phase stability, so the proposed mixer is an excellent candidate for these applications.

The second step consists of analyzing the two topologies versus the RF drive in the range of -50 to 0 dBm., LO power is fixed at 0 dBm for the proposed mixer and 5 dBm for the quasi-conventional one, the amplifier gain is set at 10 dB for both mixers.

Fig. 8 shows the results of HB analysis in terms of I/Q IF power versus RF drive. Excellent results of the proposed mixer are obtained. For RF power value of – 50 dBm, the (I/Q) power has reached approximately -60 dBm, –90 dBm for the quasi-conventional one. For RF power of 0 dBm, we got -14 dBm and -34 dBm of (I/Q) power, for the proposed mixer and the quasi-conventional one respectively.

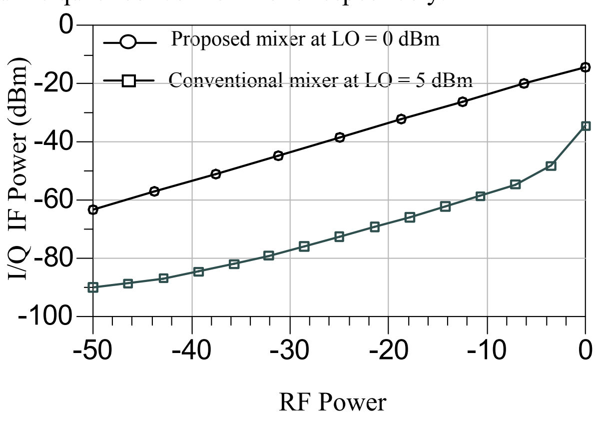

Fig. 8. Simulation Results of Conversion Loss versus RF

By keeping the same parameters for the mixers, we analyze the (I/Q) phase as shown in Figure 9. The (I/Q) phase is practically equal to the quadratic reference of 90^0 in the whole range of -50dBm to 0 dBm. Otherwise, for the quasi-conventional mixer, let us consider two regions depending of RF values. For the swept of RF power from – 38 dBm to – 5 dBm, The I/Q phase presents a deviation of 2^0 from the quadratic reference. For RF power less than – 38 dBm, the I/Q phase decrease rapidly, and for a RF power of – 45 dBm we obtained an I/Q phase around 50^0.

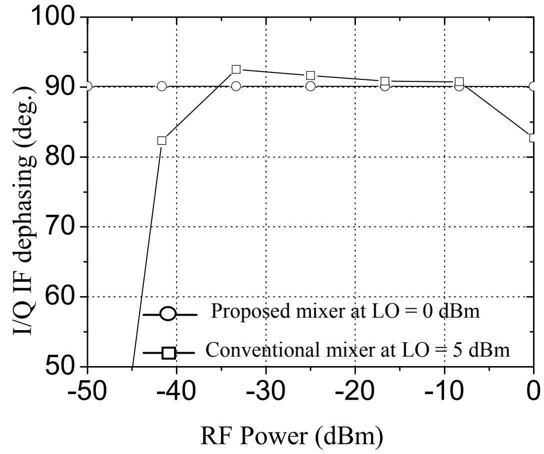

Fig. 9. Simulation Results of I/Q dephasing versus RF

Therefore an improved conversion loss and an I/Q phase closed to the quadratic reference of 90^0 were obtained with low LO power with the proposed mixer.

V. CONCLUSION

Two different down-conversion architectures are presented and compared in terms of conversion loss, required LO/RF power levels and phase stability over the LO and RF range. Comparative results show advantages of the use of power detectors and differential amplifiers instead of anti-parallel diodes operating as LO-driven switches. Therefore the proposed down-converter is suitable for V-band wireless communication and can successfully replace a conventional mixer in a low-cost millimeter-wave heterodyne receiver.

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