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Ultra-Wideband Distributed Active Mixer in 0.18µm CMOS Technology

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Abstract— In this paper, a distributed topology is used for down conversion active mixers, so that by creating transmission lines in Gilbert-cell multiplier ports, bandwidth and conversion gain of the proposed mixer will increase in ultra-wideband (UWB) operations. In this process, distributed mixer in the frequency band of 3.1 to 10.6 GHz (UWB operations) using the technology of 0.18µm CMOS is designed and simulated, which the simulated conversion gain is equal to 8.7±0.5 dB, and the average of DSB noise is 8.2 dB and isolation between ports is more than 130 dB on the whole of the UWB operations and The DC power consumption of the mixer is 15.5 mw.

Keywords— Distributed Active Mixer, Gilbert-Cell Multiplier, Conversion Gain, Transmission Lines, Ultra-Wideband Operations.

I. INTRODUCTION

Rapid growth technology and the development of successful business of wireless now have a considerable impact on our daily lives. The transition from analog to digital telecommunications, progress to the third and fourth generation of radio systems and replacement of Wi-Fi and Bluetooth for wired systems enable customers to access a vast range of information from any place and at any time. With the development of new applications for wireless communication, new circuits should be designed to have the ability to operate at high frequencies [1]. In 2002, FCC provided an unlicensed band in entire of 3.1 to 10.6 GHz frequency range for the application used in ultra-wideband (UWB) systems that has caused UWB operations to catch most attention for RFIC design. UWB operations with bandwidth of 7.5 GHz is divided into 14 channels with a bandwidth of 528 MHz, comparing narrowband wireless communication with UWB operations, it is observed that this technology has made the high data rate and wideband operations in short distance and with a low cost possible [2]. Down conversion mixer in receivers, converts the radio frequency (RF) signal to an intermediate frequency (IF) signal and CMOS technology is used primarily to implement it. CMOS technology is a suitable for wireless communication applications, such as UWB operations, WLAN and GSM due to its low cost, less power consumption and high capability of integration, but nevertheless CMOS technology causes decrease and increase in the value of f_t and parasitic capacitances respectively, that produces nonlinear harmonics in circuit. These harmonics reduce the circuit conversion gain and linearity, as a result, designing the CMOS circuit at high frequencies (UWB operations) will be more difficult [3], [4]. To remove this problem, a CMOS distributed mixer is used which also improves the mixer characteristics.

II. PROPOSED DISTRIBUTED ACTIVE MIXER

In the UWB operations, parasitic capacitances of transistors produce nonlinear harmonics in the circuit and to remove them, CMOS distributed mixer is used. So that by creating the transmission lines at the gate and drain line of MOSFETs, parasitic capacitances are absorbed to transmission lines and until the cutoff frequency of transmission lines, input impedance and bandwidth of the mixer will improve [5], [6].

A. Gilbert-Cell Mixer

Fig. 1 shows a simple example of a Gilbert-cell mixer that is a combination of transconductance stage ($M_{1,2}$) and commutating stage ($M_{3\text{-}6}$). In the transconductance stage, $M_{1,2}$ are biased saturation region which convert the input RF voltage to the amplified RF current (I_{RF}) and in commutating stage, $M_{3\text{-}6}$ are biased in the triode region that perform switching by LO signal. switching transistors direct the I_{RF} with time period of LO signal to load resistance (R_L) and finally by multiplying the I_{RF} in to load resistance, the IF output voltage is generated [7],[8]. Assuming ideal behavior for commutating stage, conversion gain is expressed according to the following equation:

$$CG = \frac{V_{IF}(\omega)}{V_{RF}(\omega - \omega_0)} \approx \frac{2}{\pi} R_L g_m = \frac{2}{\pi} R_L \sqrt{2\mu_n C_{OX} \left(\frac{W}{L}\right) I_{bias}} \qquad (1)$$

In this relation, g_m is the transconductance of $M_{1,2}$ MOSFETs and I_{bias} is bias current in mixer and (W/L) is the aspect ratio of $M_{1,2}$ MOSFETs [9].

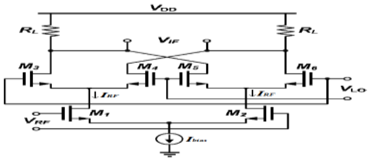


Fig. 1 Conventional Gilbert-cell mixer

According to (1), it is known to increase conversion gain of the Mixer, one can increase the I_{bias} and the aspect ratio of $M_{1,2}$ MOSFETs and the output load, but by increasing these parameters, the power consumption of circuit will also increase. The remarkable thing is that the conversion gain of Gilbert-cell mixer, is lower than what we calculated above and this is because of the parasitic capacitances of transistors which are generated in the UWB operations range. RC time constant is produced by parasitic capacitances between the mixer ports and then, great nonlinear effects are applied on the mixer. For this reason, this circuit exhibit a good conversion gain in a low frequency and this mixer has a weak conversion gain in the UWB operations.

B. Proposed Distributed Mixer Topology

In order to increase the mixer bandwidth, LC ladder networks are recently being used (including the parasitic capacitances of transistors and inductors of transmission lines). Transmission lines are formed by inductors and attached to the transistors so that they attract the parasitic capacitances to increase circuit bandwidth and decrease ports return losses. Fig. 2 shows a simplified circuit for transmission lines, in this figure, series inductors and parallel parasitic capacitances in tandem are used. As seen in Fig. 2, the transistors are distributed between the transmission lines and therefore this type of model is called the distributed structure [6], [10].

In Fig. 2, the cutoff frequency is calculated as:

$$f_c = \frac{1}{\pi\sqrt{LC}} \tag{2}$$

In this circuit, inductors (L) are fabricated on a chip and capacitances (C) are the parasitic capacitances C_{gs} of the MOSFETs. By increasing W/L (for increasing conversion gain) the value of C will have increased and finally according to (2), the f_c will have decreased. In other words to increase conversion gain of the circuit, the bandwidth is reduced. But despite L (inductors of transmission lines), and according to (2), L can be set so that the f_c increase occurs as the conversion gain amount is also desirable.

For a distributed structure with N identical sections, the total transconductance is calculated through the following equation [6], [11].

$$G_m \approx \sum_{i=1}^{N} g_{mi} \frac{\exp\left[-(i-1/2)\gamma - j\tan^{-1}(\omega/\omega_g)\right]}{\sqrt{1+(\omega/\omega_g)^2}\left[1-(\omega/\omega_g)^2\right]} \tag{3}$$

In this equation, g_{mi} is the transconductance of the i^{th} transistor and ω_g is cutoff frequency of gate circuit which is equal to $\omega_g=1/R_gC$ (losses of transmission lines are shown by R_g gate resistance, series with C capacitance). And also, the propagation constant of transmission line (γ) is defined by the following equation [6], [11]:

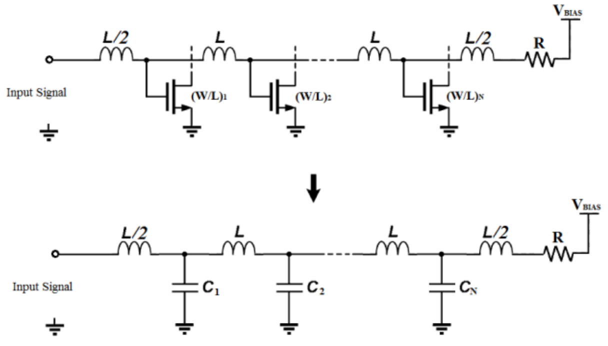


Fig. 2 Lumped model of transmission line

$$\gamma \approx \sqrt{j\omega L\left(\frac{1}{R_g}\frac{\omega^2}{{\omega_g}^2} + j\omega C\right)} = \frac{2\omega}{\omega_c}\sqrt[4]{\left(1+\frac{\omega^2}{\omega_g^2}\right)}\sin\left(\frac{1}{2}\tan^{-1}\frac{\omega}{\omega_g}\right) + j\frac{2\omega}{\omega_c}\sqrt[4]{\left(1+\frac{\omega^2}{\omega_g^2}\right)}\cos\left(\frac{1}{2}\tan^{-1}\frac{\omega}{\omega_g}\right) \tag{4}$$

ω_c is the cutoff frequency of the transmission lines.

According to (3), it is clear that there is a direct relationship between the total transconductance with g_m of each transistor, so that transistor existence with distributed structure will increase G_m, resulting in increased conversion gain of the circuit. And according to (4), transmission lines losses can be reduced by increasing the f_c, as a result, it is seen that by using a distributed structure, the bandwidth and conversion gain of the circuit can be optimally adjusted in a desirable range. But be careful that, by the excessive increase of sections between the transmission lines, the noise figure of mixer will have increased, so because of this, there are limitations in number of sections between the transmission lines [3].

C. Circuit Design

A distributed architecture is used in the design of UWB mixer, so that one can achieve flat and high conversion gain with desirable linearity in a broadband. Complete scheme of distributed mixer is shown in Fig. 3. In this work, as shown in Fig. 3, the two Gilbert-cells are used between the transmission lines, which the Gilbert-cells are the same and both have the same bias. In this figure, transmission lines of the IF, LO and RF respectively with the L_{IF}, L_{LO} and L_{RF} are specified. As mentioned, cutoff frequency of this model can be set by changing inductors of transmission lines which will increase the bandwidth of the circuit and also due to the existence of two Gilbert-cells in Fig. 3, conversion gain will increase, too. This mixer, because of the mentioned characteristics, has better bandwidth and conversion gain in the UWB operations in comparison with conventional Gilbert-cell mixer [12].

The conversion gain in the Gilbert-cell is calculated when, one of the switching pairs (e.g M_{13} and M_{16}) is in the triode region and the other pair (e.g M_{14} and M_{15}) is in the cutoff region. The following results are indicated by analyzing the small-signal model for proposed mixer:

$$A_v = \left[\frac{2}{\pi} G_{mRF} Z_L\left(f_{IF}, f_{RF}\right)\right]\left(A_{vbuffer}\right) \tag{5}$$

$$G_{mRF} = g_{m11} + g_{m21} = g_{m12} + g_{m22} \tag{6}$$

$$A_{bufferv} = \frac{g_{mIF} R_S}{1 + g_{mIF} R_S} \tag{7}$$

G_{mRF} is the total transconductance in these relationships which is composed of the sum of two transconductance of each Gilbert cell in an ideal condition (that the propagation constant and R_g is equal to zero) and just one g_m is active according to RF frequency signal in each Gilbert cell. And, $A_{vbuffer}$ is the amount of buffer voltage gain which is lesser than 1. Moreover, $Z_L(f_{IF}, f_{RF})$ is output impedance of proposed mixer which can be seen before buffer and the Fig. 4 small-signal model is used in order to calculate $Z_L(f_{IF}, f_{RF})$.

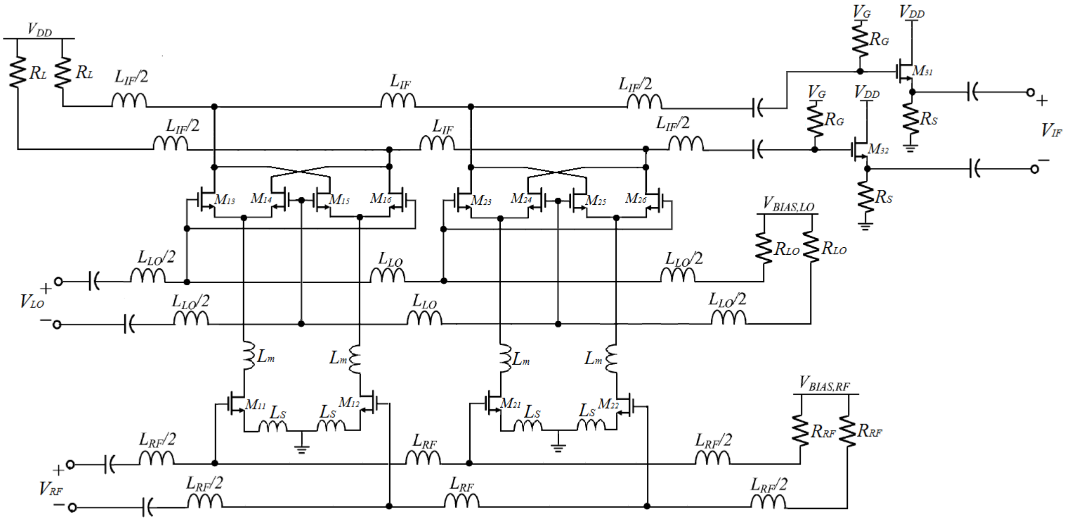


Fig. 3 Complete circuit schematic of the distributed active mixer

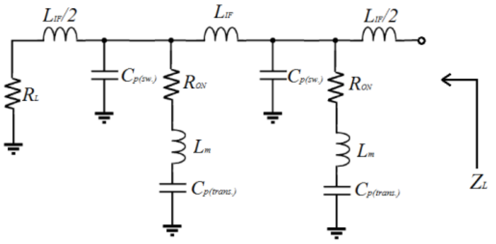


Fig. 4 The output impedance of the proposed mixer before buffer

$C_{P(sw.)}$ is parasitic capacitance in drain of the switching transistors and $C_{P(trans.)}$ is the parasitic capacitance in drain of the transconductance transistors in Fig. 4 (it should be considered that capacitive impedance of $C_{P(sw.)}$ calculates in IF fixed frequency, but capacitive impedance of $C_{P(trans.)}$ calculates in RF frequency) and R_{ON} is the on-resistance of the switching transistors. It should be noted that we considered parasitic capacitances of transistors gates as absorbed by inductors of transmission lines in this model. According to above relationships and calculating $Z_L(f_{IF}, f_{RF})$ impedance, it is clear that conversion gain depends on IF and RF frequencies.

As seen in Fig. 3, in addition to using the distributed topology, L_S (Degeneration Inductor) is used in the source of the transconductance transistors. This inductor will increase stability of the circuit by creating a negative feedback. If the small-signal model has drawn for transconductance stage despite of L_S and we will calculate the total transconductance, it will be:

$$G_m = \frac{I_{RF}}{V_{RF}} = \frac{g_m}{1 + Sg_m L_S + S^2 L_S C_{gs}} \tag{8}$$

According to (8), it is seen that the presence of L_S reduces the total transconductance and thus L_s reduces the conversion gain. However, by presence of L_S, two poles can be formed in G_m and adjusted in such a way that the mixer has good stability and thus, IIP3 increases.

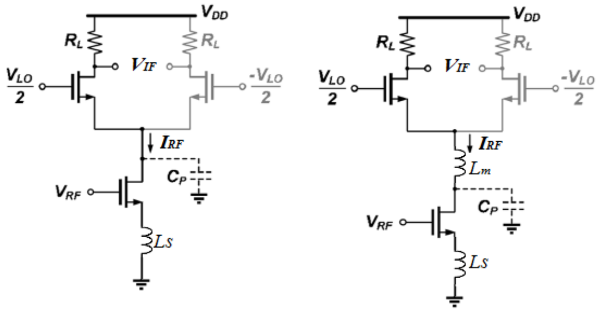
Another reason for the limited bandwidth of the mixer is the existence of parasitic capacitances at the output node of transconductance stage (C_p in Fig. 5-a). C_p creates an unwanted pole in high frequency and reduces conversion gain in the end of the band. We can solve this problem by adding additional poles in the proper place, the effect of C_p, which is the unwanted pole of parasitic capacitance will be neutralized at high frequencies. According to Fig. 5-b, series inductor (L_m) is used between the commutating and transconductance stages to compensate for the effects of the parasitic capacitances. This method in mixers is called, inductive peaking technique.

by drawing a small-signal model for the Fig. 4-a, the transfer function of the total transconductance is calculated as follows:

$$G_m = \frac{I_{IF}}{V_{RF}} = \frac{-g_m}{1 - SR_L\left(C_{p1} + C_{p2}\right)} \tag{9}$$

And small-signal transfer function for Fig. 4-b is calculated as follows:

$$G_m = \frac{I_{IF}}{V_{RF}} = \frac{-g_m}{1 - SR_L\left(C_{p1} + C_{p2}\right) - S^2 L_m C_{p1} + S^3 L_m C_{p1} C_{p2} R_L} \tag{10}$$



a) b)

Fig. 5 Simplified circuit models of the: a) conventional and b) proposed inductive peaking technique

In the (9) and (10) C_{P1} is a parasitic capacitance for the transconductance transistor and C_{p2} is a parasitic capacitance for switching transistors. According to (9), it is observed that the existence of parasitic capacitances produces an unwanted pole at high frequencies, which reduces the G_m in the circuit and a peaking inductor (L_m) is used to solve this problem. In (10) it is seen that despite the existence of a peaking inductor, three poles which are dependent to L_m will be generated, these kind of poles can be used to control the unwanted effects of C_p parasitic capacitance and improve the G_m at high frequencies.

III. SIMULATION RESULTS

Distributed mixer has been designed by Advanced Design System (ADS) in TSMC RF CMOS 0.18µm technology and frequency range of 3.1 to 10.6 GHz (UWB operations). The IF frequency is set at 500 MHz and also the voltage supply is 2-V which leads to the power consumption equal to 15.5 mW.

The proposed mixer has a flat conversion gain over the entire frequency band and as in Fig. 6 the conversion gain is drawn. As shown in Fig. 6, the average amount of simulated conversion gain in the UWB operations is 8.7 dB and a ripple of conversion gain is just ±0.5 dB.

To investigate the nonlinear behavior of the proposed mixer, the third-order intercept point (IP3) of the third-order intermodulation with two tone tests (as interferers are apart for 20 KHz) was used. The RF power and the LO power are respectively -30 dBm and 0 dBm. In the proposed mixer, third-order input intercept point (IIP3) is equal to -25 dBm as in the Fig. 7. And also, in Fig. 7, the 1dB-compression point is plotted, which is equal to -30 dBm and the equal voltage for this power by 50 ohms as an input impedance, is equal 7mV$_{pp}$. The IIP3 and OIP3 versus RF frequency are plotted in Fig. 8. As it has seen in addition to flat conversion gain, the proposed mixer has a flat IIP3 in the UWB operations.

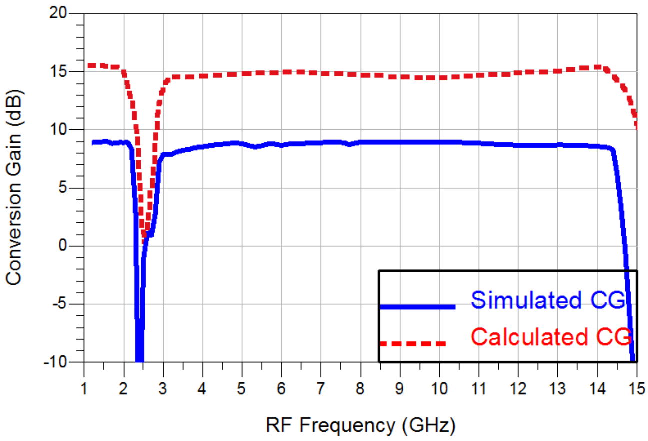


Fig. 6 Simulated and calculated conversion gain of the proposed mixer

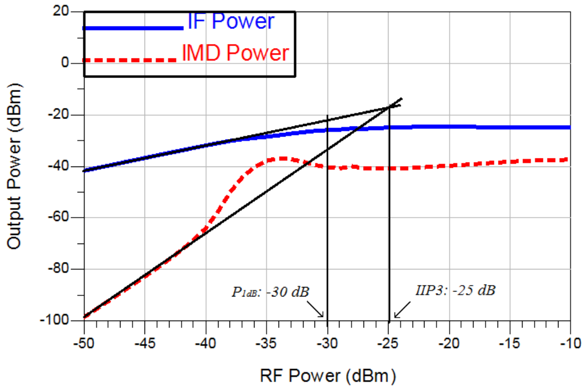


Fig. 7 IP3 with the components of the first and third harmonics in the proposed mixer (f_{RF} = 10.6 GHz)

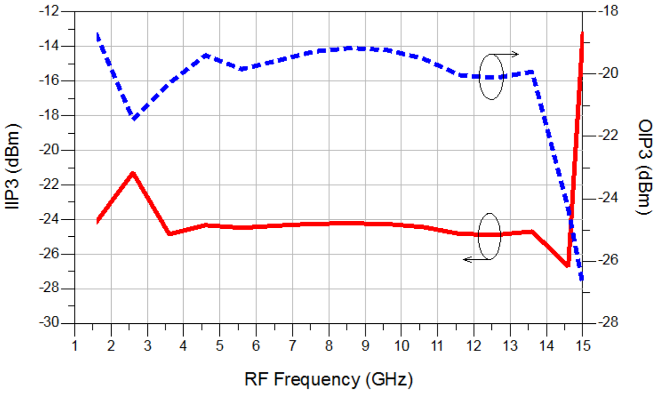


Fig. 8 IIP3 and OIP3 versus RF frequency

Fig. 9 shows the DSB noise figure against the RF frequency for the proposed mixer which the average value of the noise figure in the UWB operations is 8.2 dB. In Fig. 10, the input return loss (S_{11}) is shown in the UWB operations, which is less than -15 dB. In Fig. 11, Port to Port isolation is shown which the value of isolation is achieved more than 130 dB.

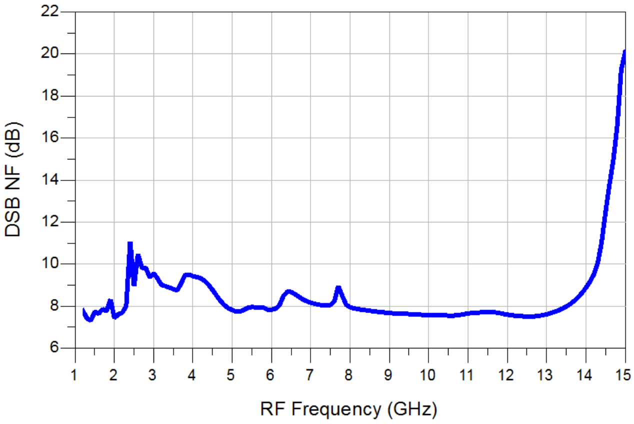


Fig. 9 DSB noise figure of the proposed mixer.

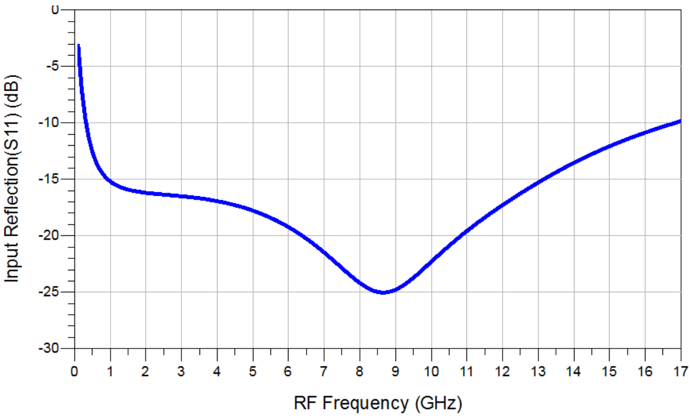


Fig. 10 Input reflection coefficient (S_{11}) of the proposed mixer

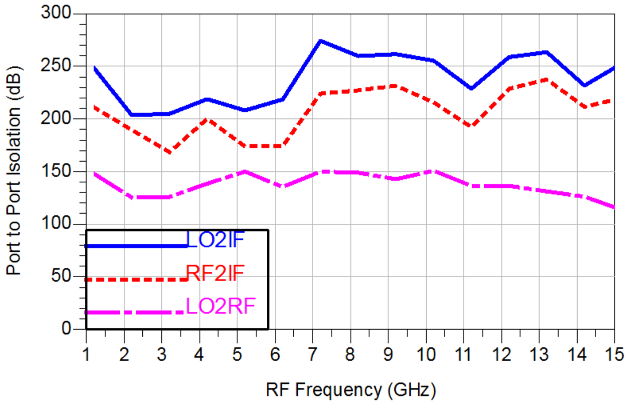


Fig. 11 Port to port isolation of the proposed mixer

The impact of L_S and L_m on stability and conversion gain has been described, in this section, as seen in Fig. 12, conversion gain decreases with increasing L_S, as mentioned earlier the reason for that is the production of poles in the transfer function of the mixer, but the IIP3 is improved as shown in Fig. 13. It should be noted that the optimum value of the L_S is 0.1 nH in this design. and as shown in Fig. 14, it is observed that by increasing the L_m to 2 nH, the conversion gain of the mixer will improve at high frequencies, and it should be noted that excess increase of L_m (more than 2 nH) reduces the conversion gain and as seen in Fig. 15, for the L_m variation, IIP3 is fixed. And it should be said that the optimum L_m is selected as 2 nH in this design.

To evaluate the performance of the proposed mixer with other available mixers, the results are summarized in Table I.

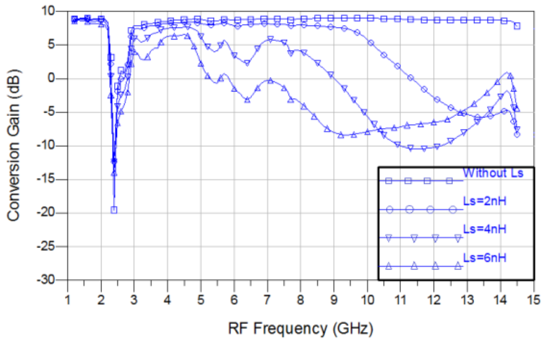


Fig. 12 Conversion gain of the proposed mixer with Ls (degeneration inductor) various

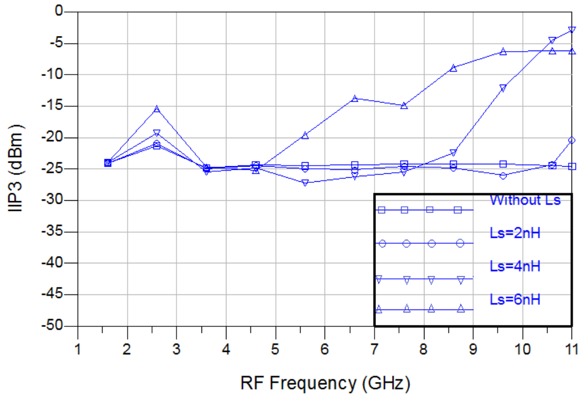


Fig.13 IIP3 of the proposed mixer with L_S (degeneration inductor) various

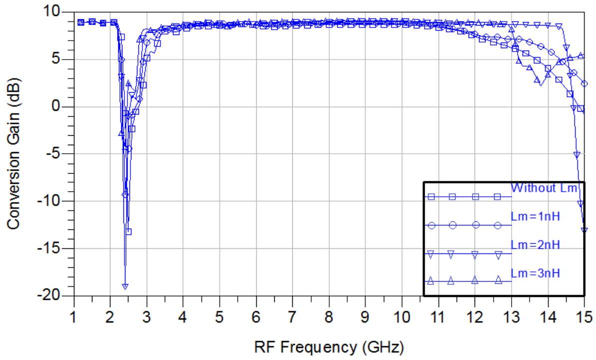


Fig.14 Conversion gain of the proposed mixer with L_m (peaking inductor) various

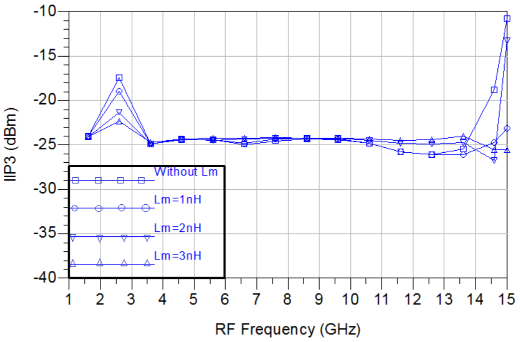


Fig.15 IIP3 of the proposed mixer with L_m (peaking inductor) various

TABLE I: PERFORMANCE SUMMARY OF THE ULTRA-WIDEBAND DOWNCONVERSION MIXER

	Technology	**IF Freq. (MHz)**	**RF Freq. (GHz)**	**Conversion Gain (dB)**	**Gain Flatness (dB)**	**IIP3 (dBm)**	**NF (dB)**	**P_{DC} (mW)**	**Isolation (dB)**
[2]	CMOS 0.18µm	528	0.2-16	6.5	±1	-10	-	15	-
[3]	CMOS 0.18µm	528	3.1-8.7	3.5	±1	5	8 (DSB)	10.4	-
[12]	-	500	3-22	4	±0.5	-	9 (DSB)	129.36	-
[13]	CMOS 0.18µm	500	2.3-6	-10	-	13.6	15 (DSB)	-	-
This work	CMOS 0.18µm	500	3.1-10.6	8.7	±0.5	-24	8.2 (DSB)	15.5	>130

IV. CONCLUSION

In the present paper, a distributed mixer with TSMC RF CMOS 0.18µm technology is used for UWB operations. So that, by applying transmission lines with inductors in the gate and drain line of the transistors in the Gilbert-cell, parasitic capacitances are absorbed to transmission lines and nonlinear harmonics are also neutralized, therefore the proposed mixer has a flat and high conversion gain with broad bandwidth in the UWB operations and other features of the mixer are also at desired levels.

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