A CMOS Even Harmonic Mixer with Current Reuse For Low Power Applications

Ming-Feng Huang, Shuenn-Yuh Lee, and Chung J. Kuo
Department of Electrical Engineering,
National Chung Cheng University,
160, San-Hsing Ming-Hsiung, Chia-Yi, Taiwan, R.O.C.
mfhuang@vlsi.ee.ccu.edu.tw, sylee@vlsi.ee.ccu.edu.tw, chung.kuo@delta.com.tw

ABSTRACT
This paper presents a novel topology for the even harmonic mixer (EHM). The proposed mixer employs current reuse and double frequency circuits in the RF input stage and LO stage, respectively, to improve its linearity and isolation. In addition, the proposed topology has the advantage of the low power consumption. In order to demonstrate the benefits of the proposed mixer, theoretical analyses of conversion gain and linearity have been described in details. The measured results reveal that the proposed mixer possesses single-end conversion gain of 8 dB and third-order input intercept point (IIP_3) of –3.8 dBm, respectively, under the supply voltage of 1.8 V and LO power of 4 dBm. The power consumption of the proposed mixer is about 1.4 mW at 900 MHz.

Categories and Subject Descriptors
B.7.0 [Integrated Circuit]: general

General Terms: Design, Verification

Keywords: Low Power, Current Reuse, Mixer

1. INTRODUCTION
A low power RF device become a tendency as applied in the portable wireless communication systems. However, the performance including linearity and conversion gain will be degraded when we reduce the power or the supply voltage of the RF mixer circuit. Hence, the implementation of the mixer with low power consumption, high linearity, and high conversion gain would be a challenge in the RF front-end circuit.

In general receiver topology, the most critical design issue is the leakage generated by the coupled LO signal. This coupling behavior is shown in Figure 1, where the one path is the LO frequency bypassed to the output; another path, where the LO leakage reflected from the antenna is amplified by the LNA. In addition, the LO signal not only directly enters the mixer but also couples into the mixer through parasitic capacitances. This amplified LO leakage and the coupled LO signal will be injected together into the input port of the mixer and down-converted to IF. Therefore, these coupling behaviors will reduce the dynamic range of the IF signal, no matter what which receiver topology is used.

The effect caused by coupling mechanism can be improved by using the even harmonic mixer (EHM) as shown in Figure 2. At the node *A*, the LO signal is cancelled by itself to improve the isolation, and produces a double frequency of the LO signal. Therefore, the RF signal is mixed with the second harmonic of LO signal and modulated as the desired output frequency ($f_{IF} = |f_{RF} - 2f_{LO}|$), where f_{IF}, f_{RF}, and f_{LO} are the IF, RF, and LO frequencies, respectively. Furthermore, no leakage component will be generated by LO signal, achieving the LO leakage-free result. In addition, the LO frequency provided by the local oscillators can be lower than the general mechanism and relax the local oscillator design.

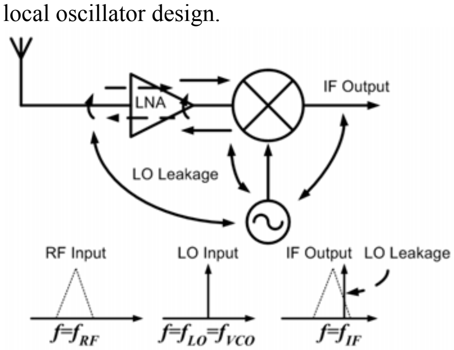

Figure 1 The general receiver topology

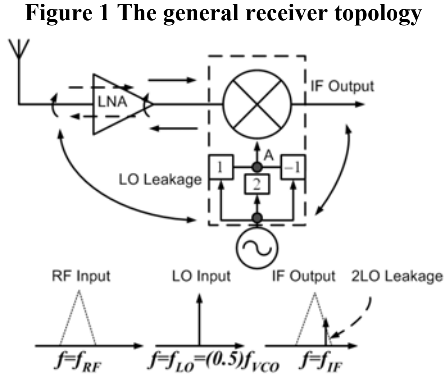

Figure 2 The leakage-improved mechanisms built by even harmonic mixer

Active EHMs [1], [2], [3], [4] suffer from low linearity and high power consumption. In this paper, the proposed EHM with current reuse technique can improve the linearity, provide high conversion gain, and possess lower power consumption than other active EHMs. Moreover, theoretical studies of linearity and conversion gain are presented in this paper to facilitate the optimal design. In addition, the benefits of this proposed EHM have been demonstrated by way of measured results.

The rest parts of this paper are organized as follows. Section 2 presents the mathematical analysis of mixer architecture, and some aspects are emphasized to assist in designing the proposed EHM. The measured results are described in Section 3 where manifests the theoretical analysis. Finally, Section 4 briefly concludes this paper.

2. MIXER ARCHITECTURE

2.1 Circuit Principle

In recent CMOS designs [3], [4], [5], [6], [7], the current reuse and charge injection techniques are addressed to increase the circuit linearity and/or conversion gain. However, the charge injection technique can only enhance the conversion gain, but can't be suitable for low power application, since the injected current must be large enough to better the conversion gain or noise figure [3], [4]. As a result, the low power EHM with current reuse technique [5], illustrated in Figure 3, is presented to improve its linearity and conversion gain. This circuit mixes an RF input signal with even harmonics of the LO signal and employs RF-dependent current source to suppress the nonlinear terms. However, the dynamic range is still limited in the RF input stage due to the nature of class-A structure.

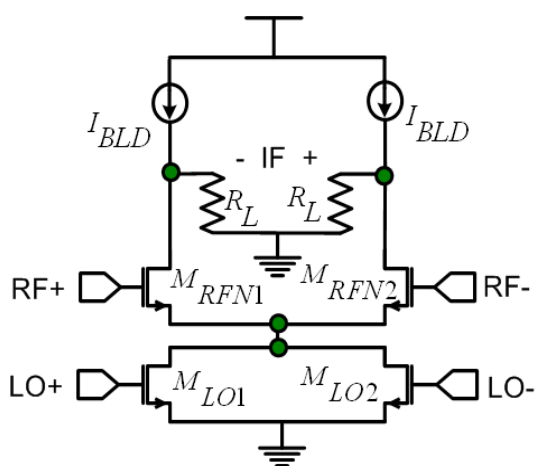

Figure 3 Topology of the CMOS even harmonic mixer with RF-dependent current source technique

The proposed EHM topology is shown in Figure 4. This topology simply provides the mixing of the RF input signal with the even harmonics of the LO signal. Since the balanced structure is used, it restrains the RF input signal from mixing with fundamental and all odd harmonics of the LO signal. Therefore, the LO leakage can be cancelled and the isolation can be promoted.

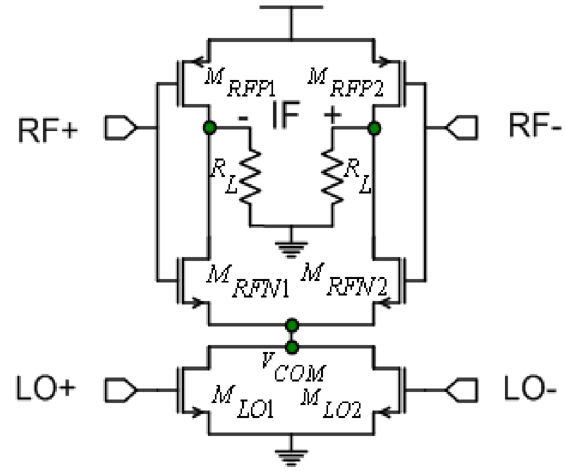

Figure 4 Proposed CMOS even harmonic mixer with current reuse technique

At the LO stage, it employs the parallel-connected transistor pair (M_{LO1} and M_{LO2}) in a double-frequency circuit as shown in Figure 5. When the differential LO signals are separately injected into the gate of transistor, the transistor pair converts the differential LO voltages into a time-varying current with the double frequency of the LO signal, at the same time, the fundamental frequency of LO signal is virtual ground at node V_{COM}. Therefore, the double frequency circuit not only provides a double frequency, but also reduces the leakage of LO frequency to improve the isolation.

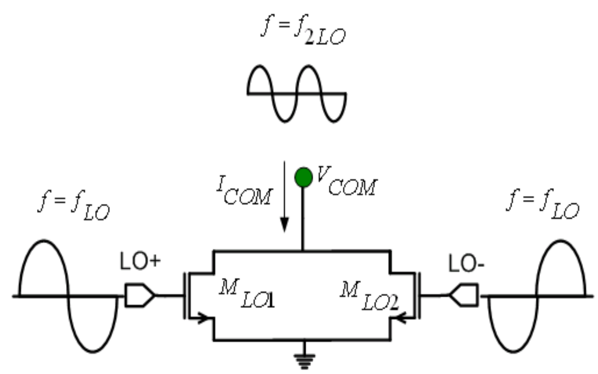

Figure 5 Operation of double frequency circuit

On the other hand, the RF stage employs the current reuse pair (M_{RFP1}, M_{RFP2}, M_{RFN1}, and M_{RFN2}) to achieve a square-law mixing between the RF signal and even harmonics of the LO signal. Simultaneously, the current reuse pair can also increase the conversion gain and improve the linearity [5]. In addition, the proposed current reuse topology is just like a push-pull circuit having an ability of gain enhancement to increase its dynamic range. The phenomenon can be interpreted by way of Figure 6, when the input signal is small, both transistors (M_{RFN} and M_{RFP}) operate in the saturation region and the conversion gain is almost constant. However, when the positive input signal is large and M_{RFN} has not yet dropped into the triode region, M_{RFP} moves toward the cut-off region, providing a greatly equivalent load, R_{RFP}, which is connected with R_L in parallel and utilized as a load in the *n*-channel common-source amplifier. On the contrary, when the negative input signal is large, the action of the transistors (M_{RFN} and M_{RFP}) is interchanged to achieve another *p*-channel common-source amplifier. Although both gain levels of the *n*- and *p*-channel common-source amplifiers may be not the same, they can be mutually compensated through adjusting the R_L value to obtain a maximum enhanced gain, in the meanwhile, the circuit linearity is improved. Therefore, even though the linearity is sensitive to the process variation, and it can be overcome by way of the tuning R_L.

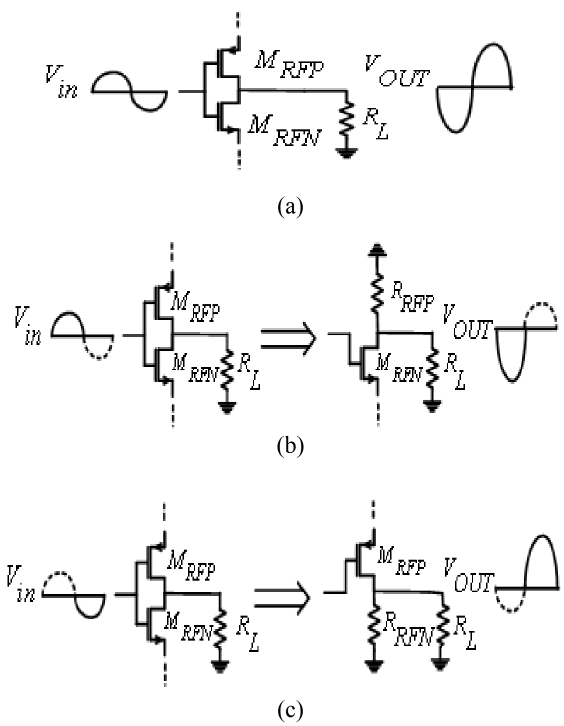

Figure 6 The operation of current reuse circuit (a) when V_{in} is small signal, (b) when V_{in} is positive large signal, and (c) when V_{in} is negative large signal

As a result, the proposed EHM has the advantages including high conversion gain, good linearity and isolation, moreover, it possesses the ability of confronting the process variation.

2.2 Mathematical Analysis

The proposed EHM is presented in Figure 4. To analyze the mixing act simply and aim for the biasing behavior, the inherent channel square-law current is used to assist the analyses of conversion gain and linearity. The drain current can be described as

$$\begin{aligned} I &= \frac{1}{2}\mu_n C_{OX}\left(\frac{W}{L}\right)\left(V_{GS}+v_{in}-V_{TH}\right)^2(1+\lambda_n V_{DS}) \\ &= \beta_n\left(\Delta V + v_{in}\right)^2 \end{aligned} \tag{1}$$

where β_n is the transconductance parameter, v_{in} and $\Delta V = V_{GS} - V_{TH}$ are defined as the input signal and overdrive voltage, respectively.

Based on the Kirchhoff's current law, the sum of the drain currents in the RF stage and the LO stage are equal, that is

$$I_{RF}^{+} + I_{RF}^{-} = I_{LO}^{+} + I_{LO}^{-}. \tag{2}$$

From (2), the relationship between ΔV_{RF} and ΔV_{LO} can be derived as

$$\Delta V_{RF} = \sqrt{\frac{2\beta_{LON}\Delta V_{LO}^2 + \frac{\beta_{LON}}{2}v_{LO}^2 - \frac{\beta_{RFN}}{2}v_{RF}^2}{2\beta_{RFN}}}. \tag{3}$$

At the same time, according to the Kirchhoff's current law again, the sum of currents in the transistors (M_{RFN1} and M_{RFN2}) and the IF stage (R_L) are equal to the drain currents in the transistors (M_{RFP1} and M_{RFP2}), that is

$$I_{RFP}^{+} + I_{RFP}^{-} = I_{RF}^{+} + I_{RF}^{-} + 2\left(\frac{V_{DD}-V_{SDP}}{R_L}\right). \tag{4}$$

From (4), the expression of ΔV_{RFP} can be derived as

$$\Delta V_{RFP} = \sqrt{\frac{2\beta_{LON}\Delta V_{LO}^2 + \frac{\beta_{LON}}{2}v_{LO}^2 - \frac{\beta_{RFP}}{2}v_{RF}^2 + 2\left(\frac{V_{DD}-V_{SDP}}{R_L}\right)}{2\beta_{RFP}}} \tag{5}$$

where V_{DD} and V_{SDP} are, respectively, the supply voltage and the source-drain voltage of transistors (M_{RFP1} and M_{RFP2}). Hence, the time-varying transconductance ($G_{m,CR}(t)$) is obtained.

$$\begin{aligned} G_{m,CR}(t) &= G_{mP}(t) + G_{mN}(t) \\ &= 2\beta_{RFP}\Delta V_{RFP} + 2\beta_{RFN}\Delta V_{RF} \end{aligned} \tag{6}$$

where $G_{mN}(t)$ and $G_{mP}(t)$ are the transconductances of the *n* and *p* channel transistors of the current reuse pair respectively, and the subscript CR denotes the overall transconductance of using current reuse technique. Furthermore, assuming $v_{LO} = a_{LO}\cos\omega_{LO}t$ and using the Taylor's series expansion, the resulting differential output voltage (v_{IF}) of the proposed EHM can be expressed as

$$
\begin{aligned}
v_{IF} &= G_{m,CR}(t)\, v_{RF} R_L \\
&\cong \left(F_1 v_{RF} + F_3 v_{RF}^3 + \cdots\right) R_L \\
&+ \left(G_1 v_{RF} \cos 2\omega_{LO} t + G_3 v_{RF}^3 \cos 2\omega_{LO} t + \cdots\right) R_L + \cdots
\end{aligned} \tag{7}
$$

According to (7), the down conversion mixing operation is completed via the second harmonic of the LO signal. As a result, the desired conversion gain, the input 1-dB compression point ($IP_{1dB,CR}$), and third-order input intercept point ($IIP_{3,CR}$) defined in [8], where express its signal level in dBm and assume the input impedance, R_{in} , of 50 ohm, can be described in (8)-(10), respectively.

$$
\begin{aligned}
G_{f_{RF}-2f_{LO},CR} &= \frac{G_1 R_L}{2} \\
&\cong \frac{a_{LO}^2}{4}\sqrt{\frac{\beta_{LON}\beta_{RFN}}{2}}\,\frac{R_L}{\sqrt{a_{LO}^2 + 8\Delta V_{LO}^2}} \\
&+ \frac{a_{LO}^2}{4}\sqrt{\frac{\beta_{LON}\beta_{RFP}}{2}}\,\frac{R_L}{\sqrt{a_{LO}^2 + 8\Delta V_{LO}^2 + 8\left(\frac{V_{DD}-V_{SDP}}{\beta_{LON} R_L}\right)}}
\end{aligned} \tag{8}
$$

$$
\begin{aligned}
IP_{1dB,CR}(dBm) &= 10\log \frac{\left(\sqrt{0.145\frac{G_1}{|G_3|}}\right)^2}{2R_{in}} + 30 \\
&\cong 10\log\left(\left(\frac{\beta_{LON} a_{LO}^2 + 8\beta_{LON}\Delta V_{LO}^2}{\beta_{RFN} + \beta_{RFP} - \sqrt{\beta_{RFN}\beta_{RFP}}}\right)\left(\frac{1}{R_{in}}\right)\right. \\
&\left. + \left(\frac{\sqrt{\beta_{RFP}}}{\sqrt{\beta_{RFN}^3} + \sqrt{\beta_{RFP}^3}}\right)\left(8\left(\frac{V_{DD}-V_{SDP}}{R_L R_{in}}\right)\right)\right) + 18.6
\end{aligned} \tag{9}
$$

$$
\begin{aligned}
IIP_{3,CR}(dBm) &\cong 10\log\left(\left(\frac{\beta_{LON} a_{LO}^2 + 8\beta_{LON}\Delta V_{LO}^2}{\beta_{RFN} + \beta_{RFP} - \sqrt{\beta_{RFN}\beta_{RFP}}}\right)\left(\frac{1}{R_{in}}\right)\right. \\
&\left. + \left(\frac{\sqrt{\beta_{RFP}}}{\sqrt{\beta_{RFN}^3} + \sqrt{\beta_{RFP}^3}}\right)\left(8\left(\frac{V_{DD}-V_{SDP}}{R_L R_{in}}\right)\right)\right) + 28.24
\end{aligned} \tag{10}
$$

2.3 Design Consideration

The proposed EHM is shown in Figure 4. According to (3), (5), (8), (9), and (10), we have several rules of thumb in the circuit design. From the conversion gain consideration, ΔV_{LO} must be small, and β_{RFN} and β_{RFP} must be large. When β_{RFN} and β_{RFP} are large enough, it is clear that ΔV_{RF} and ΔV_{RFP} move toward the weak inversion region. However, if both ΔV_{RF} and ΔV_{RFP} are located in weak inversion region, the conversion gain will increase significantly, and the unacceptable linearity and the speed degradation occur. Fortunately, the linearity and the conversion gain can be improved simultaneously by increasing the LO power. The results are different from that of the Gilbert mixer, which increasing the LO power enhances the conversion gain, but degrades the linearity. As a result, the embraceable LO power and ΔV_{LO} are necessary, and the transistor sizes and biases of the current reuse pair are trade-offs. Furthermore, we can obtain the low power consumption from biasing the ΔV_{LO} at the weak inversion region, acquire the compensated linearity from current reuse pair, and get the acceptable conversion gain from the appropriate LO power.

3. IMPLEMENTATION

The even harmonic mixer with current reuse topology is implemented by TSMC 0.18 μ m 1P6M process. The implementation of the proposed full monolithic EHM is shown in Figure 7. The source-follower is adopted as the output buffer and the mixer's characteristics are measured at an RF frequency of 900.05 MHz with an IF frequency of 50 KHz, and the LO frequency of 450 MHz.

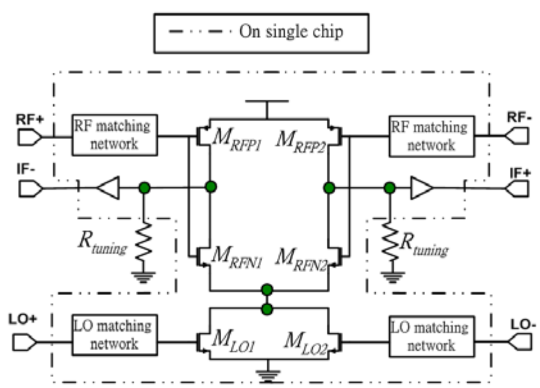

Figure 7 Schematic of the proposed full monolithic even harmonic mixer

In order to make sure that the RF and LO signals can inject into the proposed mixer and reduce the signal distortion, the on-chip matching network is designed to achieve the input return loss less than 10 dB. The measured result of input return loss is shown in Figure 8.

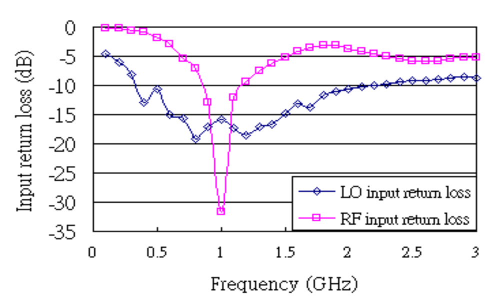

Figure 8 Performance of input return loss of RF and LO stage

By sweeping the LO power to measure the RF power versus single-end conversion gain is shown in Figure 9, the tuning resistor (R_{tuning}) is infinite in this measured condition. Apparently, it is consistent with our theoretical studies, that is, increasing LO power can enhance the conversion gain and improve the linearity. We obtain the single-end conversion gain of 9 dB and IP_{1dB} of –14.7 dBm as LO power of 4 dBm is applied.

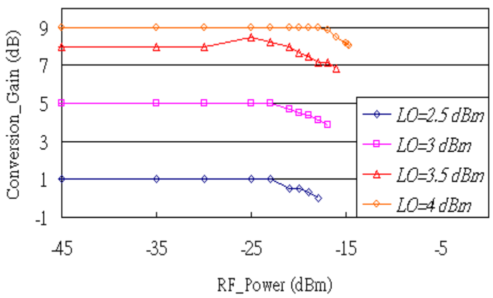

Figure 9 Measured RF power versus single-end conversion gain as different LO Power

The measured result of the optimum linear performance of EHM by adjusting the tuning resistor, R_{tuning} , to obtain the *n*- and *p*-transistors source-drain voltage drop of current reuse pair are illustrated in Figure 10 and IIP_3 is obtained by two-tone testing as plotted in Figure 11. Summarily, the measured results reveal that the proposed EHM possesses single-end conversion gain of 8 dB, IP_{1dB} of –12.5 dBm and the worse-case IIP_3 of –3.8 dBm under the LO power of 4 dBm.

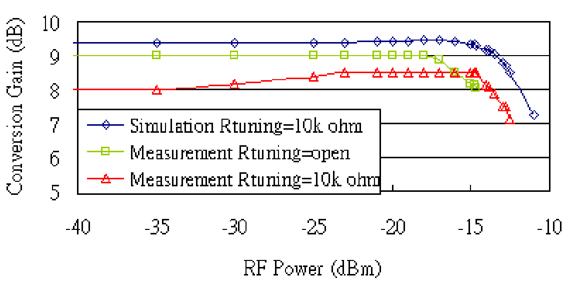

Figure 10 Measured results of conversion gain and IP_{1dB} using R_{Tuning} with 10 kohm and LO power of 4 dBm

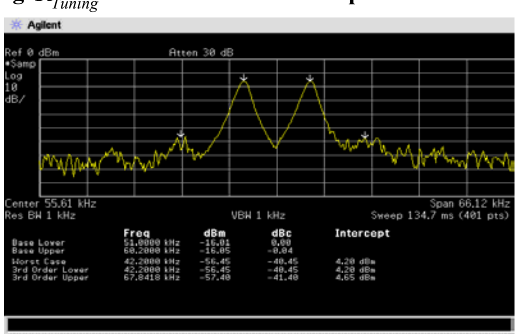

Figure 11 Measured result of the IIP_3 performance

Summary of the measured results is depicted in Table I, and the chip's photomicrograph is shown in Figure 12. Note that the LO power level is different from prelayout simulation as shown in Table I, because the circuit model in foundry, parasitic effect in bond wire, and transmission line in test key will result in circuit imbalance and make the LO power level apart from its optimal value. Nevertheless, the measured result is agreed with simulation, furthermore, this chip not only achieves high conversion gain and isolation but also provides good linearity; even, the power consumption of the individual mixer circuit is only about 1.4 mW. Finally, the comparison is shown in Table II, it is shown that the proposed mixer provides high linearity, lower power consumption and acceptable conversion gain.

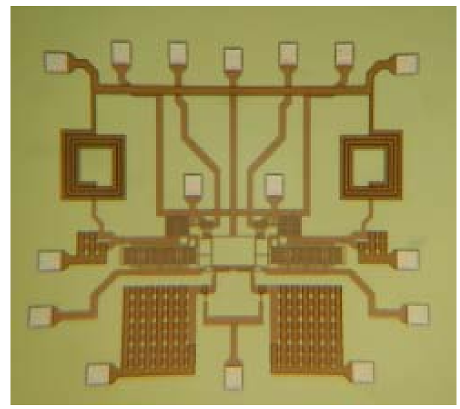

Figure 12 Photomicrograph of the proposed EHM chip

Table I Summary of the proposed even harmonic mixer

	Simulation	Measurement	
Output	Single-end		
R_{tuning}	10 kohm		Open
RF Frequency	900.05 MHz		
LO Frequency	450 MHz		
LO Power (dBm)	5	4	
Conversion gain (dB)	9.36	8	9
IP_{1dB} (dBm)	-12.5	-12.5	-14.7
IIP_3 (dBm)	-3	-3.8	<-3.8
LO-IF Isolation (dB)	---	34	32
2LO-IF Isolation (dB)	---	43	42.6
Mixer DC Current	0.79 mA	0.78 mA	<0.77 mA
Buffer DC Current	14.8 mA	10.55 mA	10.55 mA
Supply Voltage	1.8 V		
Process	TSMC 0.18 μ m 1P6M process		

Table II Summary of the comparison

	Proposed EHM	[3]	[1]	[2]	[6]
Output	Single-end	Differential-end			
RF Frequency	900.05 MHz		2 GHz		900 MHz
LO Frequency	450 MHz		1 GHz		
LO Power (dBm)	4	-15.4	-6	---	0
Conversion gain (dB)	8	13	5.6	11.61	8.8
IP_{1dB} (dBm)	-12.5	-19.9	---	---	-16.1
IIP_3 (dBm)	-3.8	-10.6	-1	-13.5	-4.1
2LO-IF Isolation (dB)	43	---	---	---	---
LO-IF Isolation (dB)	34	---	---	---	34.4
Mixer DC Current	0.78 mA	1.72 mA	5 mA	1.71 mA	2.6 mA
Supply Voltage	1.8 V	3 V	2.7 V	3 V	2.7 V

4. CONCLUSION

In this paper, we proposed an even harmonic mixer with current reuse technique. The mathematical expressions for linearity and conversion gain using Taylor's series expansion are derived. Measured results show that the even harmonic mixer with current reuse topology has extremely low power consumption, high linearity, high isolation, and large conversion gain. Hence, the proposed even harmonic mixer with current reuse topology is suitable for highly linear and low power applications such as in portable wireless communication systems.

5. ACKNOWLEDGMENTS

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