A K-band High Gain Linearity Mixer with Current-Bleeding and Derivative Superposition Technique

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Abstract— This paper presents the design and analysis of high linearity RF mixers in a 65-nm CMOS process for K-band down-conversion receivers. Current-bleeding is useful for high gain mixer. However, the circuit linearity is degraded by the third-order transconductance of the current-bleeding cell. The proposed circuit uses Derivative Superposition (DS) technique to both g_m-stage and current-bleeding. Circuit simulation results show conversion gain of 9.7 dB and Third Order Input Intercept Point (IIP3) of 5.0 dBm.

I. Introduction

Mixers are becoming more and more important on the RF front end in the K-band. For example, it has been demanded that high data-rate communication, millimeter-wave (MMW) integrated circuits, and systems. While used in communication circuits, RF circuit has been applied to radar systems such as automotive, medical imaging, robot vision, security safety, tank level meters, and human–device interaction systems[1]. However, as the usage of K-band increases, signals generated by other devices act as strong interference. Under strong interference, the mixer produces the third-order intermodulation (IM3), which tends to overlap the signal of interest and affect sensitivity[2].

Conventional Gilbert mixers demonstrate good conversion gain (CG) and port-to-port isolation[3] and widely used in MMW transceiver. However, the stack topology requires high supply voltage and high DC power consumption. In order to improve the linearity, a current-mirror structure in and a multiple-gate-transistor technique are proposed for low-frequency operation [4], [5]. However, the technique cannot been used for high-frequency. Modified Gilbert-cell and source-driven mixer using weak inversion biasing is proposed in [6] and high linearity mixer using Distributed Derivative Superposition (DS) technique is proposed in [2], but these mixers have low conversion gain. On the other hand, high conversion gain and linearity mixer is proposed by the source follower current-bleeding [7].

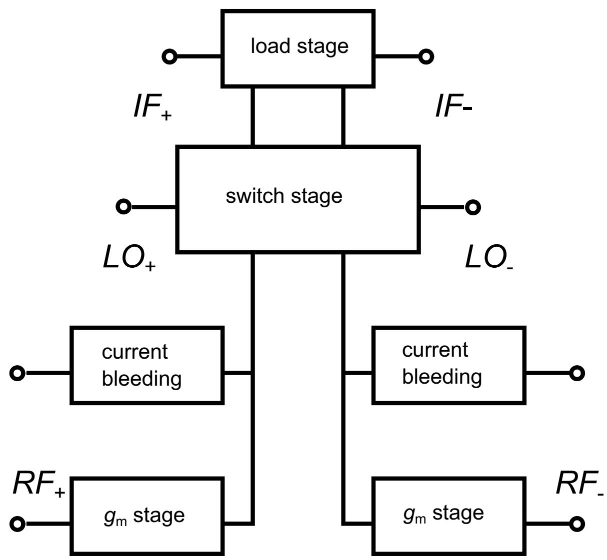

Fig. 1. Block diagram of current-bleeding Gilbert-cell mixer

In this paper, we adopt the source follower current-bleeding to neutralize the parasitic gate-to-drain capacitor (C_{gd}) and to cancel out the nonideal return path. The current-bleeding using the DS technique is proposed to achieve high-linearity

II. Conventional Current-Bleeding circuit

Fig. 1 shows the block diagram of the proposed double-balanced Gilbert-cell mixer with current-bleeding. It consists of input g_m stage, current-bleeding cell, local oscillator (LO) switch stage, and load stage. The g_m stage converts RF voltage to RF current. The differential RF current generated by the g_m stage flow to the switch stage through AC-coupling inductors. The differential RF cur-

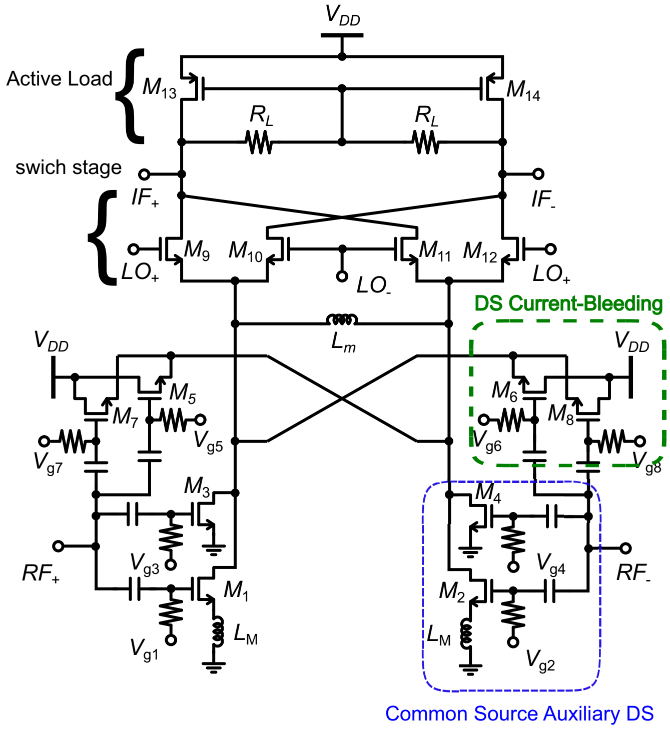

Fig. 2. Schematic of the proposed mixer

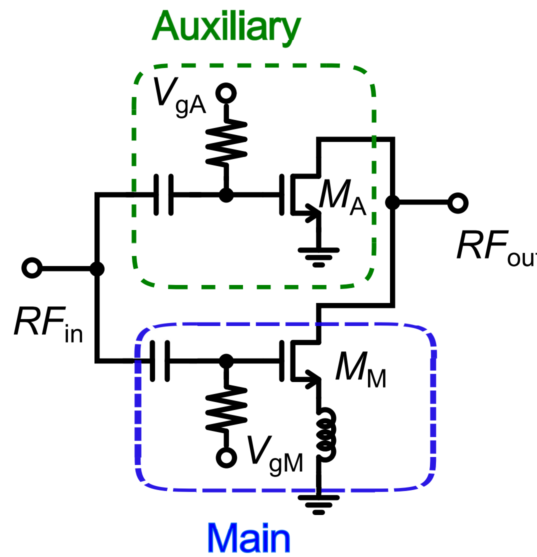

Fig. 3. Schematic of DS

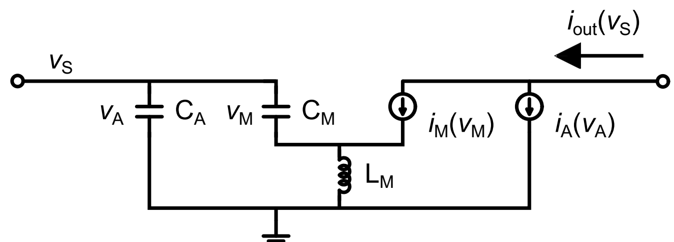

Fig. 4. Small-signal equivalent circuit of DS

rent are switched to the IF currents. The differential IF current by the differential LO drive signals and converted to IF emissions through the load stage. Current-bleeding supplies DC current to g_m stage.

The current-bleeding has the advantage of higher CG. CG is improved by increasing load stage resistance. However, load stage resistance cannot be increased because the voltage supplied to the g_m stage is decreased by a large voltage drop and current at the load and switch stage. Current at the load stage should be decreased to improve CG, whereas the g_m stage needs the current to the convert operation. The current-bleeding decreases the load stage current while maintaining the g_m stage current. However, The current-bleeding has disadvantage by the third-order transconductance of current-bleeding cell. The linearity of the mixer is decreased by the third-order transconductance.

III. Circuit Design

A. *Current-Bleeding Mixer*

The schematic of the proposed mixer based on the current-bleeding architecture is shown in Fig. 2. The g_m stages with common source auxiliary DS method consist of the complementary pairs M_1, M_3 and M_2, M_4, which provide high linearly signal for the differential input RF signals[8]. However g_m stages generate harmonic mixing feedback effect, which third-order inter modulation (IM3)[9]. Transistors M_9 $-M_{12}$ building the switch stage are driven by the differential LO signals. An inductor L_m is inserted between the two switching common nodes[10]. At the operating frequency, L_m resonates with the parasitic capacitances at the switching common nodes to suppress the leakage of the differential RF currents and also cancel the parasitic C_{gd} of M_1–M_4 to reduce the LO-RF leakage for the isolation enhancement [7]. The load stage is composed of Active Load M_{13}, M_{14} and R_L. Is has the large equivalent load impedance and high CG[11]. Transistors M_5-M_8 act as the current-bleeding. Source follower is used to the current-bleeding for blocking harmonic mixing feedback effect. C_{gs} of M_5-M_8 cancel out the Miller effect in C_{gd} of M_1-M_4[7].

B. *Common Source Auxiliary Derivative Superposition Method*

Fig. 3 shows the schematic of common source auxiliary DS method[8]. In common source auxiliary design, the third-order transconductance ($g_{m3,M}$, $g_{m3,A}$) of the main transistor M_M and auxiliary transistor M_A are cancelled

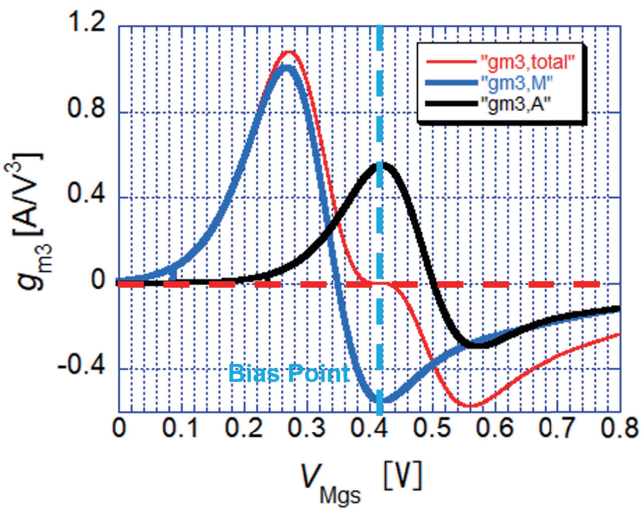

Fig. 5. The third-order transconductance

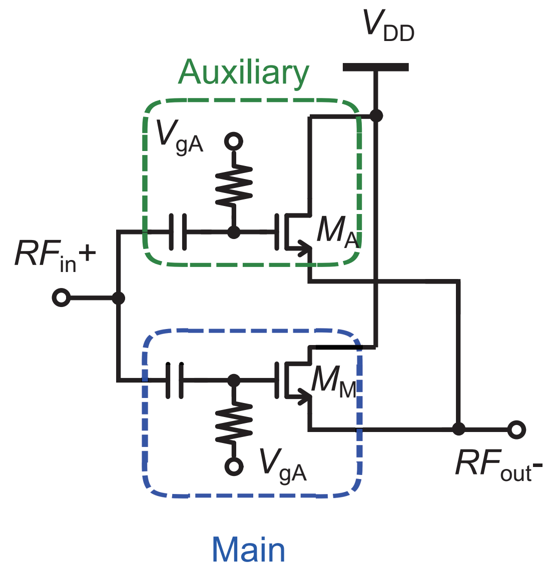

Fig. 6. Schematic of a current-bleeding with DS technique

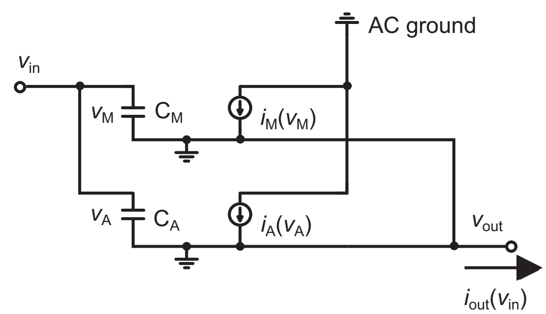

Fig. 7. Small-signal equivalent circuit model current-bleeding with DS technique

in a particular bias region [8]. g_m-stage is designed based on common source auxiliary DS method for the high IIP3. Fig.4 shows the schematic of small signal equivalent circuit of DS. According to [8], the i_M and i_A are express as

$$i_M(v_M) = g_{m1,M}v_M + g_{m2,M}v_M^2 + g_{m3,}v_M^3 + \cdots , \quad (1)$$
$$i_A(v_A) = g_{m3,A}v_A^3 \quad . \quad (2)$$

The IIP3 of a common source auxiliary DS is expressed as

$$\mathrm{IIP}_3 = \frac{4g_1^2\omega^2 L(C_M + C_A)}{3|\varepsilon|} \quad , \quad (3)$$

where

$$\varepsilon = 2g_{m3,A}\left(1 + j\omega L_M g_{m1,M}\right)\left[1 + (\omega L_M g_{m1,M})^2\right] + g_{m3,M} - \frac{2g_{m2,M}^2}{3g_{m1,M}} \frac{1}{1 + \frac{1}{j2\omega L_M g_{m1,M}}} \quad , \quad (4)$$

$g_{m1,M}$, $g_{m2,M}$ and $g_{m3,M}$ are the fundamental, second, and third-order transconductance of M_M. $g_{m3,A}$ is the third-order transconductance of M_A. L_M is a degeneration inductance. C_M and C_A are gate-source capacitances, respectively v_M and v_A are gate-source voltage of the M_M and M_A.

Fig. 5 shows $g_{m3,M}$, $g_{m3,A}$ and total g_{m3}. V_{Mgs} and V_{Ags} are the bias voltage applied to the gate in M_M and M_A, respectively. V_{Ags} is 150 mV lower than V_{Mgs}. since bias point is zero total g_{m3}, V_M is 415 mV.

C. *Proposed DS Current-Bleeding*

Figs. 6 and 7 show the proposed DS current-bleeding and small-signal equivalent circuit model a circuit respectively. In conventional mixer, IIP3 is decreased by the third-order transconductance generated from the main MOSFET. Therefore, the auxiliary MOSFET is added to the current-bleeding to cancel the third-order transconductance. In the proposed current-bleeding, i_M and i_A are same as Eqs. (3) and (4). v_M and v_A are expressed as

$$v_M = v_A = v_{in} - v_{out} \quad . \quad (5)$$

According to Eq. (5), the v_{gs} of M_M and M_A are the same signal. Therefore, the proposed circuit is considered to be DS method with the major amplifier. The input tone amplitude at the intercept point (IIP3) is given by

$$\mathrm{IIP}_3 = \sqrt{\frac{4}{3}\left|\frac{g_{m1,M}}{g_{m3,M} + g_{m3,A}}\right|} \quad , \quad (6)$$

The high linearity mixer is realized by the current-bleeding when bias point is applied to become lower g_{m3} as shown in Fig. 5. Since g_{m3} becomes smaller, IM3 generated from current-bleeding can be suppressed from Eq. (6).

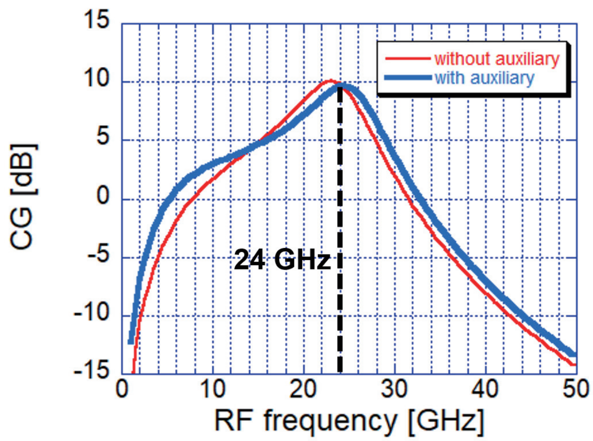

Fig. 8. Conversion gain versus RF frequency at fixed 100 MHz IF bandwidth

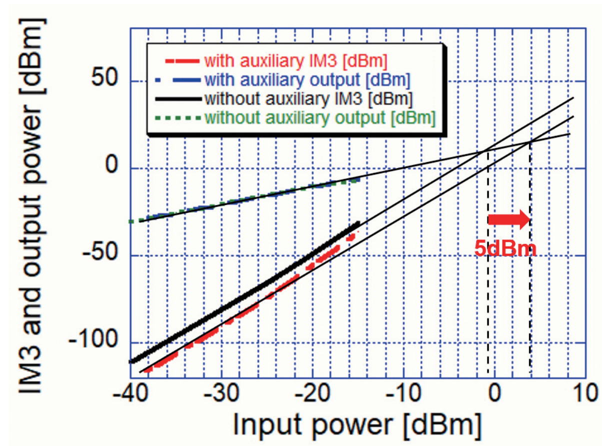

Fig. 10. Two-tone comparison of current-bleeding mixers.

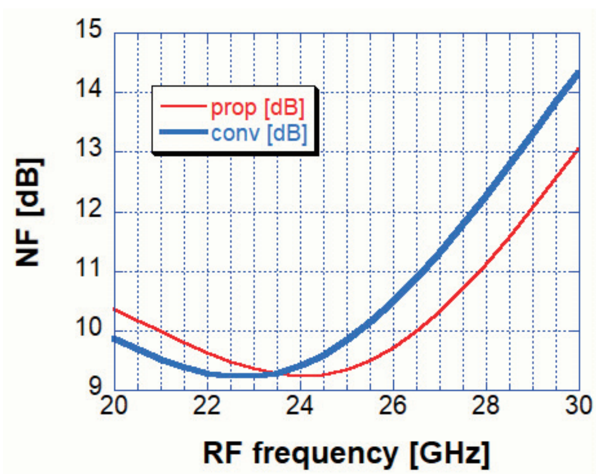

Fig. 9. NF versus RF frequency at fixed 100 MHz IF bandwidth

IV. SIMULATION

The mixer is simulated using spectre in the 65-nm CMOS process. The simulation condition is shown in Table I. Supply voltage is 2 V. Simulations are performed with and without auxiliary MOSFETs (M_7 and M_8).

Fig. 8 shows simulation results of conversion gain (CG). The CG of the proposed mixer achieves 9.7 dB at 24 GHz. The CG of the proposed mixer is 1 dB lower than that of the conventional mixer, because RF signal is lost by parasitic capacitances of Auxiliary MOSFETs.

Fig. 9 shows simulation results of noise figure (NF).The noise figure (NF) of the proposed mixer achieves 9.2 dB at 24 GHz. The simulation of twotone test with frequency of 24 GHz ± 10 MHz is shown in Fig. 10. When the current-bleeding auxiliary is off, the IIP3 of the mixer is 0 dBm. When the auxiliary is on, the IIP3 is increased by 5 dB. The conventional total DC power consumption is 12.8 mW, and the proposed total DC power consumption is 12.6 mW at 2 V power supply voltage. IM3 generated from current-bleeding is suppressed by Auxiliary MOSFETs.Table II summarizes previously reported CMOS mixers with literature.

V. CONCLUSION

This paper presents a high linearity down-conversion mixer with a 65-nm CMOS process. By using the current-bleeding with DS technique, the proposed mixer has 5.0 dBm IIP3 improvement and 9.7 dB CG and 12 mW DC power consumption. The proposed mixer using current-bleeding and DS technique has high linearity and conversion gain with same power consumption as conventional mixers.

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TABLE I
SIMULATION CONDITIONS

M_1-M_2, M_5-M_6	W/L	2 μm/60 nm
	# of finger	16
M_3-M_4	W/L	1.9 μm/60 nm
	# of finger	9
M_7-M_8	W/L	1.9 μm/60 nm
	# of finger	9
M_9-M_{12}	W/L	2 μm/60 nm
	# of finger	32
M_{13}-M_{14}	W/L	6 μm/180 nm
	# of finger	8
	R_{L}	100 Ω
Inductance w/ Auxiliary	L_{m}	900 pH
Inductance w/o Auxiliary	L_{m}	600 pH

TABLE II
PERFORMANCE COMPARISON

	This work	[2]	[12]	[13]
Process	65 nm	0.18 μm	90 nm	90 nm
RF Freq (GHz)	24	23-25	20-50	23-30
Conversion Gain (dB)	9.7	-4.5 ± 0.6	0 ± 2	-3.2 ± 0.5
LO power (dBm)	4	5	0	1
NF (dB)	9.2	N/A	16	N/A
IIP3 (dBm)	5.0	23	9.5	21
DC power (mW)	12.6	16	6	10

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