A Si/SiGe BiCMOS Mixer with 3rd-Order Nonlinearity Cancellation for WCDMA Applications (Student Paper)

Liwei Sheng, and Lawrence E. Larson

Center for Wireless Communications , University of California – San Diego
9500 Gilman Drive, La Jolla, CA 92037 USA

Abstract – **This paper presents a general analysis of the 3rd-order nonlinearity of a differential common-emitter (CE) radio frequency (RF) amplifier and an improved way to cancel the 3rd-order nonlinearity. A SiGe BiCMOS mixer is designed based on the 3rd-order cancellation scheme. The mixer achieves +6dBm IIP3, 15dB gain and 7.7dB DSB NF with only 2.2mA current at $2.1GHz$. This performance exceeds that of previously reported active mixers in this frequency range.**

1. INTRODUCTION

Low-power, high-performance, and low-cost integrated RF circuits are aiding the rapid growth of mobile wireless communications. The bipolar common-emitter and differential-pair stages are commonly used in many RF building blocks such as low-noise amplifiers(LNAs) and mixers. For a direct-conversion WCDMA receiver, the linearity requirements of the mixer are approximately $0dBm\ IIP_3$ and $35dBm\ IIP_2$ if the LNA before the mixer has a gain of approximate $16dB$ and a SAW filter in between the LNA and the mixer [1]. The inherent linearity of a common-emitter circuit does not satisfy the linearity requirements of most high-performance RF systems. Inductive or resistive degeneration is usually applied to improve the linearity of these circuits, though it sacrifices the gain or raises dc current [2]. Another way to improve the linearity is to utilize the second-order nonlinearity to cancel the third-order nonlinearity [3]. This method achieves high linearity at lower bias current, but requires a complicated nonlinear analysis. Recently several authors [3–5] analyzed the problem and showed that up to $14dB$ linearity improvement can be achieved with proper choice of source harmonic termination.

In this paper, we directly compute the nonlinear response of a differential common-emitter circuit. The direct nonlinear response was solved, then a relatively straightforward solution of the 3rd-order nonlinearity cancellation was given. A fully balanced active mixer was designed based on the 3rd-order nonlinearity distortion cancellation, and achieved outstanding results.

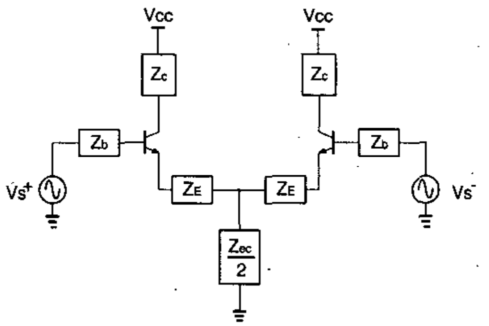

Figure 1: Large-signal model of a common-emitter differential pair.

2. NONLINEAR ANALYSIS OF A DIFFERENTIAL COMMON-EMITTER CIRCUIT

Fig. 1 shows the model used for analysis on the nonlinearity of the differential CE circuit. To simplify the analysis, the following assumptions were made, similar to that in [3]. The collector current is only a function of the base-emitter voltage. The Early effect is ignored, because the output resistance is much larger than the output load for RF applications. The base-emitter junction capacitance C_{je} is considered as a linear component, because its nonlinearity is small compared to the base-emitter diffusion capacitance C_{DE}. The base resistance r_b, extrinsic emitter resistance r_e, base-collector capacitance C_μ, collector-substrate capacitance C_{CS}, forward transit time τ and the low-frequency current gain β are all constant, because their nonlinearities are much smaller compared to the nonlinearity of the g_m.

The nonlinear components are collector current i_c, base current i_b and base-emitter capacitance current $i_{C_{DE}}$. They are are all functions of base-emitter voltage v_{be}.

$$i_c = g_m\, v_{be} + g_{m2}\, v_{be}^2 + g_{m3}\, v_{be}^3 \qquad (1a)$$

$$i_b = \frac{i_c}{\beta} \qquad (1b)$$

0-7803-7694-3/03/$17.00 © 2003 IEEE

$$i_{C_{DE}} = \tau \frac{d}{dt} i_c \quad (1c)$$

where $g_m = I_{c0}/V_t$, $g_{m2} = I_{c0}/2V_t^2$, $g_{m3} = I_{c0}/6V_t^3$, I_{c0} is the dc collector current and $V_t = kT/q$.

The first-order response is given by

$$V_{out1} = H_d(s)\, V_s \quad (2)$$

where

$$\begin{aligned} V_s &= V_s^+ - V_s^- , \\ H_d(s) &= \frac{Z_c\, \{C_\mu s\, [1 + (g_m + C_\pi s)\, Z_{ed}] - g_m\}}{L_d(s)} , \\ C_\pi &= C_{je} + \tau g_m , \\ Z_{ed} &= r_e + Z_E , \\ L_d(s) &\cong 1 + g_m(Z_{ed} + \frac{Z_b}{\beta}) + C_\pi s(Z_b + Z_{ed}) \\ &\quad + C_\mu s\, (Z_b + Z_c + (g_m + C_\pi s)\Delta_d) \;\; and \\ \Delta_d &= Z_{ed} Z_b + Z_b Z_c + Z_{ed} Z_c \end{aligned}$$

The third-order currents are generated in two ways: through the 3rd-order transistor transconductance g_{m3} and the interaction of first-order response and the second-order response through 2nd-order transconductance g_{m2}. Solving the third-order solution, we have

$$V_{out3} = G_d(s)_3 \left\{ K_d(s) V_s \cdot \left[F_d(s)_2 \left(K_d(s) V_s \right)^2 \right] \right\} \quad (3)$$

where

$$\begin{aligned} F_d(s) &= \frac{1}{L_c(s)} \{1 - 2g_m Z_e' - 2g_m \frac{Z_b}{\beta} \\ &\quad + (C_{je} - 2\tau g_m) s (Z_b + Z_e') \\ &\quad + C_\mu s[(Z_b + Z_c)(1 - 2g_m Z_e') \\ &\quad + \Delta_c (C_{je} - 2g_m \tau) s - 2g_m Z_b Z_c]\} , \\ G_d(s) &= \frac{-Z_c}{L_d(s)} [1 + C_\pi s(Z_b + Z_{ed}) + C_\mu s Z_b] , \\ K_d(s) &= \frac{1 + C_\mu s Z_c}{L_d(s)} , \\ Z_e' &= Z_{ed} + Z_{ec} , \\ L_c(s) &\cong 1 + g_m(Z_e' + \frac{Z_b}{\beta}) + C_\pi s(Z_b + Z_e') \\ &\quad + C_\mu s\, (Z_b + Z_c + (g_m + C_\pi s)\Delta_c) \;\; and \\ \Delta_c &= Z_e' Z_b + Z_b Z_c + Z_e' Z_c . \end{aligned}$$

$K_d(s) = v_{be}/v_s'$, is the transfer function from the input voltage v_s' to base emitter voltage v_{be}. $F_d(s) = 6\, V_t^3\, i_{c3}/(I_c\, v_{be}^3)$, is the transfer function from v_{be} to the 3rd-order collector current i_{c3}, the subscript of $F_d(s)_2$ in (3) implies that the operations are done on the second-order current. $G(s) = v_{c3}/i_{c3}$, is the transfer function from the collector current to the output voltage; the subscript of $G_d(s)_3$ in (3) implies that the operations are done on the third-order current. Collecting all the intermodulation terms at frequency $2\omega_2 - \omega_1$, we have

$$\begin{aligned} IM_{3d} &= \frac{I_{c0}}{96 V_t^3} A_2^2 \frac{|G_d(j(2\omega_2 - \omega_1)) K_d(j\omega_2)^2 K_d(j\omega_1)|}{|H_d(j\omega_1)|} \\ &\quad \cdot |2F_d(j(\omega_2 - \omega_1)) + F_d(j\omega_2)| \end{aligned} \quad (4)$$

where A_2 is the amplitude of the signal in a two-tone-test. A lower third-order intermodulation(IM3) is achieved when (4) is minimized, while the first-order is kept the same. By careful selection of Z_b, Z_e', Z_c, it is possible to make the last term in (4), $|2F_d(j(\omega_2 - \omega_1)) + F_d(j2\omega_2)|$, close to zero. But the last terms are functions of $(\omega_2 - \omega_1)$ and ω_2, and it is difficult to find a *general* solution for termination impedance to cancel the third-order nonlinearity. Another approach is to find the termination impedance such that $|F_d(j(\omega_2 - \omega_1))|$ and $|F_d(2j\omega_2)|$ are separately close to zero. Such termination impedances are selected as

$$\begin{aligned} Z_b(j2\omega 2) &\cong 0 \qquad & Z_b(j\Delta\omega) &\cong 0 \\ Z_e'(j2\omega 2) &\cong \frac{1}{2g_m} \qquad & Z_e'(j\Delta\omega) &\cong \frac{1}{2g_m} \end{aligned} \quad (5)$$

where $\Delta\omega = |\omega_2 - \omega_1|$. Substituting these value into $F_d(s)$, we obtain

$$\begin{aligned} |F_d(j\Delta\omega)| &\cong \frac{\Delta\omega |C_{je} - 2\tau g_m|}{3g_m} \\ |F_d(2j\omega_2)| &\cong \frac{\frac{1}{g_m}\omega_2 |C_{je} - 2\tau g_m|}{\frac{3}{2} + \frac{C_\pi \omega_2}{g_m}} \end{aligned} \quad (6)$$

For typical RF applications, $\tau\omega << 1$. As a result, $|F_d(j\Delta\omega)|$ and $|F_d(2j\omega_2)|$ are dramatically lower than the non-optimized value, and the intermodulation can be substantially improved. Since the terminations are only changed at $\Delta\omega$ and 2ω, the noise performance is virtually unaffected.

3. LOW-DISTORTION MIXER DESIGN

The termination condition (5) suggests that only the second-order currents need to be terminated at the input and Z_e' needs to be real at the second-harmonic of the input signals. By connecting two emitters of the differential pair and letting $R_e = 1/g_m - 2r_e$, the emitter impedance requirement for 3rd-order cancellation can be easily satisfied. The resistor is only added for common-mode operation, so the noise is not increased for the differential circuit [5]. Figure 2 shows a simplified down-conversion mixer. The 2nd-order base termination at frequency 2ω is through series resonance components, L_c and C_c. The base termination at frequency $\Delta\omega$ was achieved with the feedback circuit of

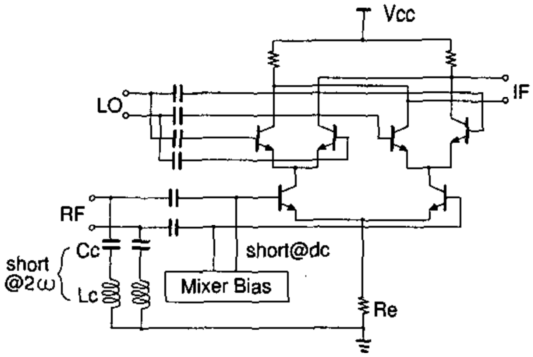

Figure 2: Down-conversion mixer with 3rd-order cancellation.

Figure 3. The impedance of the bias circuit at dc and RF frequency are

$$Z_{in}(dc) \cong \frac{R_3}{\beta_2} + \frac{R_2}{gm_1 r_{o1}} \quad (7a)$$
$$Z_{in}(\omega) \cong R_2//R_1 \quad (7b)$$

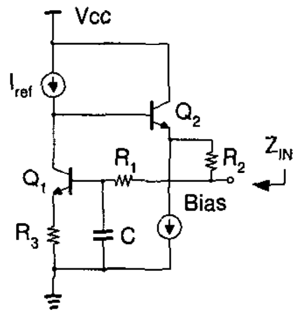

Figure 3: Mixer bias circuit to provide low impedance at dc.

4. MEASUREMENT RESULTS

The mixer was fabricated in IBM's SiGe5AM process with transistor peak $f_T = 45GHz$. The microphotograph of the mixer is shown in Figure 4.

Using the measurement setup shown in Figure 5, the mixer has been characterized at $2.1GHz$.

The output power as well as the 3rd-order intermodulation are plotted in Figure 6. Due to the nonlinearity of the bias circuit, the dc bias changes with RF input signal power, which affects the nonlinearity performance of the circuit.

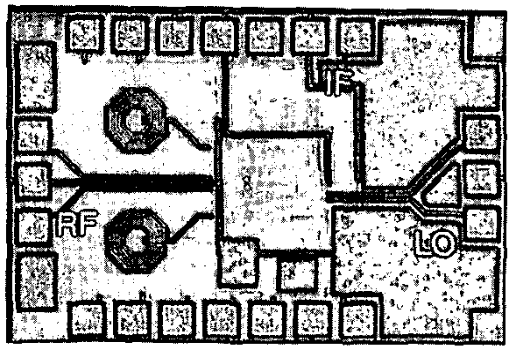

Figure 4: Microphotograph of the mixer.

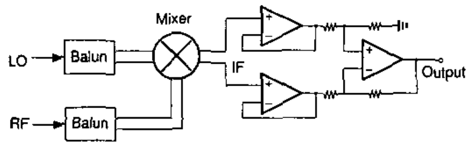

Figure 5: Measurement setup.

The bias current was optimized for approximate $-20dBm$ input power. When the input power is lower, the dc bias current drops and the 3rd-order cancellation becomes less effective. The nonlinearity of the bias circuit causes the 3rd-order distortion to remain constant at an input power range from $-22dBm$ to $-32dBm$. The changing of the dc bias also causes the 3rd-order intermodulation to rise more quickly when the input power exceeds $-18dBm$. Figure 7 is a plot of nonlinearity characteristic versus dc bias current of the mixer at an input power of $-22dBm$. Table 1 is a summary of the mixer as well as a comparison with other recent mixers. The figure of merit is defined in [7] as

$$FOM = 10log\left(\frac{IIP3(mW)}{(F-1)\cdot V_{dd}\cdot I_{dc}}\right) \quad (8)$$

The result exceeds the performance of the other previously reported results. The mixer in [8] has similar performance but is operated at $880MHz$.

The mixer also exhibits excellent 2nd-order linearity, with an IIP_2 of +45 dBm.

5. CONCLUSION

The general nonlinear responses of the CE differential-pair circuit have been developed to determine the conditions for

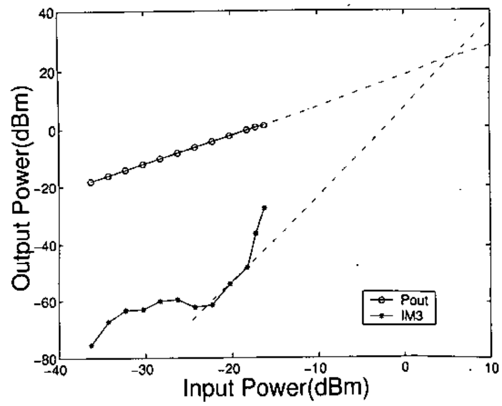

Figure 6: 3rd-order intermodulation characteristic.

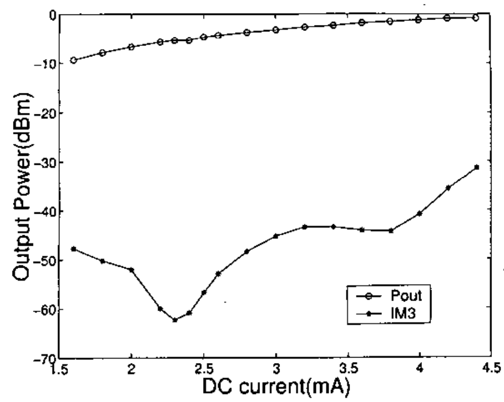

Figure 7: 3rd-order intermodulation vs. dc bias current. $P_{in} = -22dBm$.

cancellation of third-order nonlinearities. A WCDMA down conversion mixer has been designed using these techniques. The designed mixer exhibits state-of-the-art linearity at very low dc power without excessive penalty on Noise Figure.

Table 1: Comparison with other recent mixers.

Ref.	F_{RF} (GHz)	Gain (dB)	NF (dB)	IIP3 (dBm)	P_{diss} (mW)	Process	FOM (dB)
This work	2.15	15.0	7.7	6.0	5.9	0.5μSiGe BiCMOS	21.4
[8]	0.88	8.4	7.6 SSB	8.0	12.0	0.5μSiGe BiCMOS	23.4
[9]	2.0	15.0	8.5	-1.5	9.1	0.35μm BiCMOS	11.6
[10]	1.9	6.1	10.9 SSB	2.3	4.75	0.8μm SiBJT	18.0
[7]	2.0	24.2	3.2	-1.5	21.6	0.35um CMOS	14.8

6. REFERENCES

[1] O. K. Jensen, T. E. Kolding, C. R. Iversen, S. Laursen, R. V. Reynisson, J. H. Mikkelsen, E. Pedersen, M. B. Jenner and T. Larsen "RF Receiver Requirements for 3G W-CDMA Mobile Equipment" *Microwave Journal*, pp. 22-46, Feberary 2000.

[2] L. E. Larson *RF and microwave circuit design for wireless communications*, Boston : Artech House, c1996.

[3] V. Aparin "Effect of Out-of-Band Terminations on Intermodulation Distortion in Common-Emitter Circuits" *IEEE MTT-S 1999 Digest*, pp. 977-980, 1999.

[4] K. Fong and R. G. Meyer "High-Frequency Nonlinearity Analysis of Common-Emitter and Differential-Pair Transconductance Stages" *IEEE Journal of Solid-State Circuits*, pp. 548-555, Vol. 33, No. 4, April, 1999.

[5] M. Heijden, H. Graaff and L. Vreede "A Novel Frequency-Independent Third-Order Intermodulation Distortion Cancellation Technique for BJT Amplifiers" *IEEE Journal of Solid-State Circuits*, pp 1176-1183, vol.37, No. 9, September 2002.

[6] P. Wambacq and W. Sansen. *Distortion Analysis of Analog Integrated Circuits*, Kluwer Academic Publishers,1993

[7] A. Karimi-Sanjaani, H. Sjoland, A. A. Abidi "A 2GHz Merged CMOS LNA and Mixer for WCDMA" *Symposium on VLSI Circuirs Digest of Technical Papers*, pp 19-22, 2001.

[8] V. Aparin, E. Zeisel and P. Gazzerro "Highly Linear SiGe BiCOMS LNA and Mixer for Cellular CDMA/AMPS Applications" *IEEE Radio Frequency Integrated Circuits Symposium*, pp. 129-132, 2002.

[9] K. Kivekäs, A. Pärssinen, J. Jussila, J. Ryynänen, K. Halonen "Design of Low-Voltage Active Mixer for Direct Conversion Receivers" *IEEE International Symposium on Circuits and Systems*, pp. 382-385, vol. 4, 2001.

[10] J. R. Long and M. A. Copeland "A 1.9GHz Low-Voltage Silicon Bipolar Reciver Front-End for Wireless Personal Communications System" *IEEE Journal of Solid-State Circuits*, pp 1438-1448, vol.30, No. 12, December 1995.