

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

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Abstract – This paper presents a general analysis of the 3rd-order nonlinearity of a differential common-emitter (CE) radio frequency (RF) amplifier and a novel 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 NF with only 2.2mA current at 2.1GHz. This performance substantially 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. The inherent linearity of such a 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 [1]. Another way to improve the linearity is to utilize the second-order nonlinearity to cancel the third-order nonlinearity [2]. This method achieves high linearity at lower bias current but requires a complicated nonlinear analysis. Recently several authors [2–4] 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. A direct nonlinear response of an input voltage was solved, then a relatively easy 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.

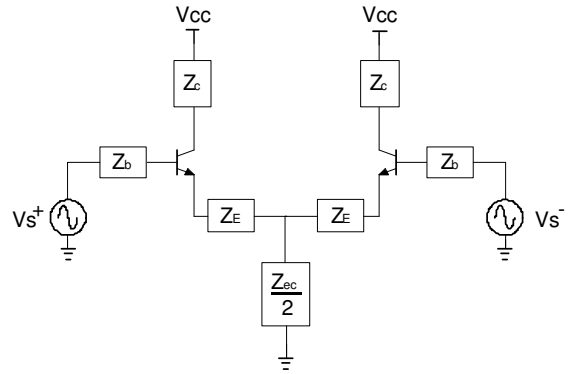


Figure 1: Large-signal model of a common-emitter differential pair.

2. NONLINEAR ANALYSIS OF A DIFFERENTIAL COMMON-EMITTER CIRCUIT

Figure 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 [2]. 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 g_m .

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

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

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

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

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

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

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

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

where

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

$K_d(s) = v_{be}/v'_s$, is the transfer function from input voltage v'_s to base emitter voltage v_{be} . $F_d(s) = 6V_t^3 i_{c3}/(I_c v_{be}^3)$, 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. $G(s) = v_{c3}/i_{c3}$, is the transfer function from collector current to output voltage,

the subscript of $G_d(s)_3$ in (3) implies that the operations are done on the third order. Collecting all the intermodulation terms at frequency $2\omega_2 - \omega_1$, we have

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

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

$$\begin{aligned} Z_b(j2\omega_2) &\cong 0 & Z_b(j\Delta\omega) &\cong 0 \\ Z'_e(j2\omega_2) &\cong \frac{1}{2g_m} & Z'_e(j\Delta\omega) &\cong \frac{1}{2g_m} \quad (5) \end{aligned}$$

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

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

For typical RF applications, $\tau\omega \ll 1$. 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 unchanged.

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 let $R_e = 1/g_m - 2r_e$, 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. Figure 2 shows a simplified down-conversion mixer. The 2nd-order base termination at frequency 2ω is through resonance components, L_c and C_c . The base termination at frequency $\Delta\omega$ was achieved with the feedback circuit of Figure 3.

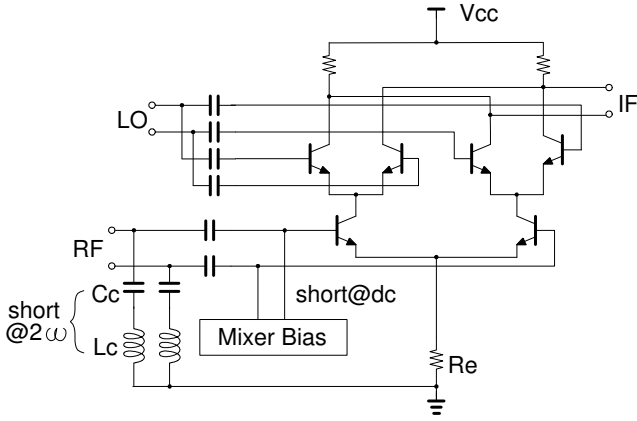


Figure 2: Down-conversion mixer with 3rd-order cancellation

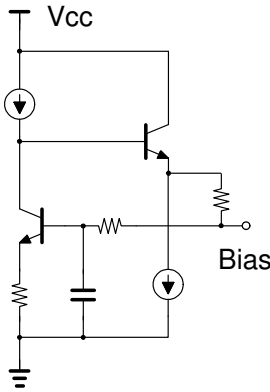


Figure 3: Mixer bias circuit to provide low impedance at frequency $\Delta\omega$

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. 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

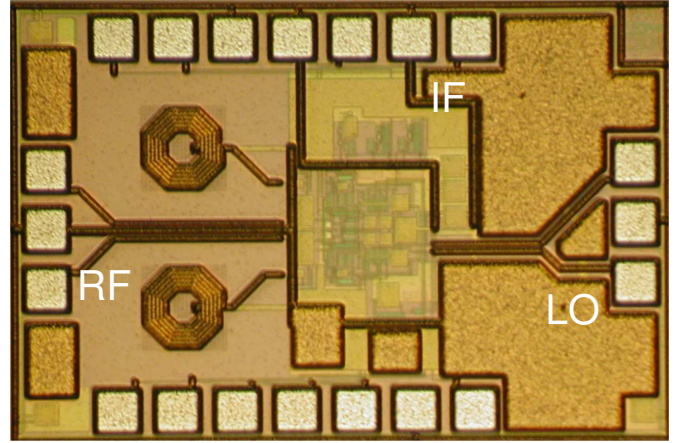


Figure 4: Microphotograph of the mixer

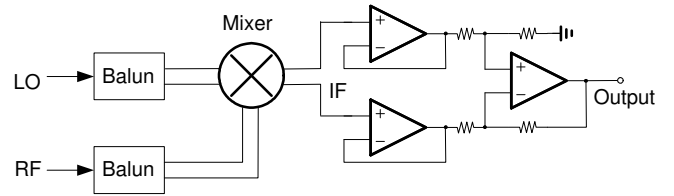


Figure 5: Measurement setup

quickly when the input power exceeds $-18dBm$. Figure 7 is a plot of nonlinearity characteristic versus dc bias current of the mixer. Table 1 is a summary of the mixer as well as a comparison with other recent mixers. The figure of merit is defined in [6] as

$$FOM = 10 \log \left(\frac{IIP3(mW)}{(F-1) \cdot Vdd \cdot Idc} \right) \quad (7)$$

The result exceeds the performance of the other previously reported results. The mixer in [7] has similar performance but is operated at $880MHz$.

5. CONCLUSION

The general nonlinear responses of the CE differential-pair circuit have been derived. A down-conversion mixer has been designed upon the 3rd-order nonlinearity cancellation. The designed mixer exhibits state-of-the-art linearity at very low dc power without excessive penalty on Noise Figure.

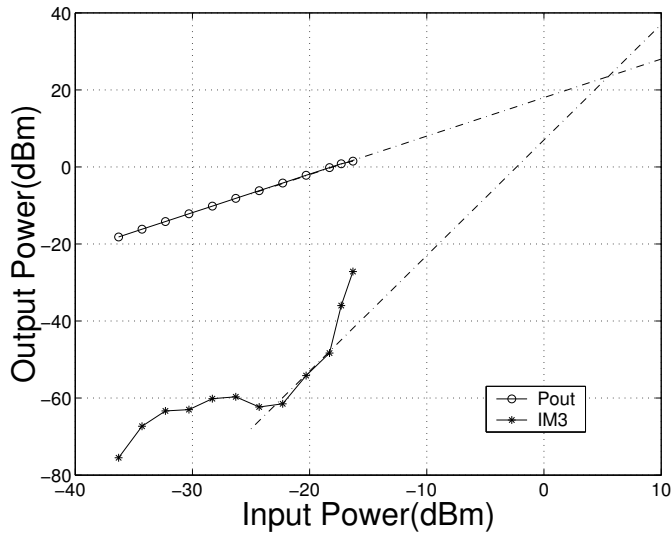


Figure 6: 3rd-order intermodulation characteristic

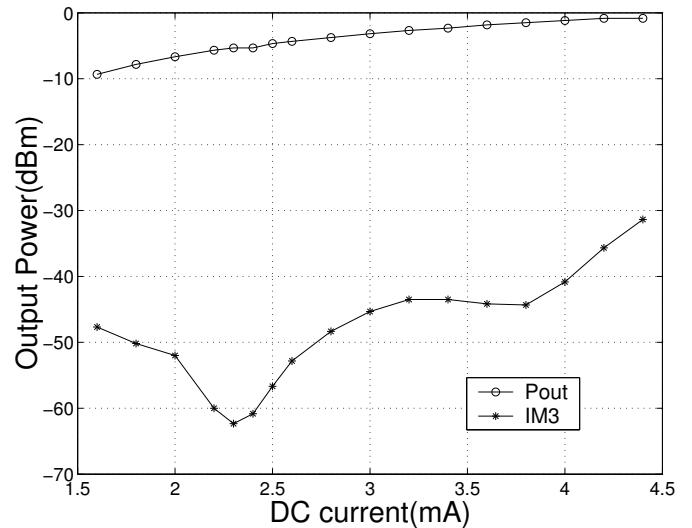


Figure 7: 3rd-order intermodulation vs. dc bias current

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	0.5 μ SiGe BiCMOS	21.4
[7]	0.88	8.4	7.6 SSB	8.0	12.0	0.5 μ SiGe BiCMOS	23.4
[8]	2.0	15.0	8.5	-1.5	9.1	0.35 μ m BiCMOS	11.6
[9]	1.9	6.1	10.9 SSB	2.3	4.75	0.8 μ m SiBJT	18.0
[6]	2.0	24.2	3.2	-1.5	21.6	0.35 μ m CMOS	14.8

6. REFERENCES

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