Highly Linear SiGe BiCMOS LNA and Mixer for Cellular CDMA/AMPS Applications

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Abstract — **SiGe BiCMOS LNA and mixer designed for cellular CDMA/AMPS applications are described. The circuits exhibit very high linearity thanks to low-impedance low-frequency input terminations. The LNA achieves +12.2dBm IIP3, 16.3dB gain and 1.5dB NF with 7.7mA current in the high-gain high-linearity mode. The mixer achieves +14.7dBm IIP3, 10.9dB conversion power gain and 6.7db SSB NF with 8.4mA current in the CDMA mode.**

I. INTRODUCTION

CDMA phones operate in the full duplex mode that introduces an additional interference to the phone receiver - a signal leakage from the transmitter (TX) through the antenna duplexer. This TX leakage is generally not a problem on its own. But, in the presence of a strong in-band jammer (a signal transmitted by a nearby AMPS or TDMA base station), the TX leakage modulation is transferred to the jammer by the LNA nonlinearities widening the jammer spectrum. This cross modulation (XM) contaminates the RX channel adjacent to the jammer degrading the receiver sensitivity. The maximum allowed XM sets the linearity goal for a CDMA LNA. Since the TX leakage is filtered by the RF image-reject filter following the LNA, XM is negligible in the mixer compared to the 3rd-order intermodulation (IM3) of two in-band jammers that sets the mixer linearity goal.

The single-tone desensitization requirement of the IS-98 standard demands an exceptional LNA linearity that should be achieved

1. without degrading the LNA NF and gain since the receiver should be capable to receive a weak desired signal from a distant CDMA base station while transmitting at a maximum power level and being jammed by a nearby AMPS or TDMA base station,
2. with a low power consumption to extend the phone talk time, and
3. with a low-cost technology allowing a high level of integration.

GaAs based HEMTs are known for their high linearity and have been successfully used in CDMA receiver ICs [1]. However, high material cost and relatively low yield make HEMT circuits more expensive than their Si and SiGe counterparts. A poor compatibility with analog and digital functions limits their integration level. MOSFETs are also perceived more linear than BJTs. But they still need a high dc current to achieve a high IIP3 [2].

To reduce the dc current, linearization techniques must be employed. One of these techniques is the feed-forward linearization used in [3]. Despite a very high IIP3, the reported CMOS LNA suffers from a low power gain, high NF and an unacceptably high dc current due to the use of an auxiliary gain stage. A more promising technique is based on optimizing the out-of-band input and output terminations of a circuit to improve its IIP3. The underlying theory of this technique is explained in [4]. The demonstrated IIP3 improvement of a Si BJT LNA was achieved without degradation in the gain and NF and increase in the dc current but was confined to a narrow frequency range due to the frequency dependence of the distortion cancellation effect. A simpler version of this technique is based on using just a low-impedance low-frequency termination of the LNA input [5], [6].

This paper presents the theory of the BJT linearity improvement by the low-impedance low-frequency input termination. It describes a cellular LNA and mixer with such terminations designed for a single-chip multi-band multi-mode SiGe BiCMOS RF-to-baseband receiver [7].

II. DERIVATION OF LNA IIP3 REQUIREMENT

The sources of interference in a CDMA receiver include

- N_{RX} – the receiver thermal noise,
- N_{TX} – the transmitter thermal noise in the RX band coupled through the duplexer to the LNA input,
- N_{ϕ} – the receiver LO phase noise that appears at IF after the jammer is downconverted by the RX LO (reciprocal mixing),
- P_{XM} – the jammer cross modulation power in the adjacent RX channel.

Let N_{RX}, N_{TX}, N_{ϕ} and P_{XM} be the available powers referred to the antenna connector and integrated over 1.23MHz-wide RX channel. The maximum total thermal noise power, $I_{oc,th} = N_{RX} + N_{TX}$, is set by the receiver sensitivity and dynamic range requirement of the IS-98 standard [8]. The receiver frame error rate (FER) must not exceed 0.5% when the received forward-link CDMA signal power at the antenna $\hat{I}_{or} = -104$dBm/1.23MHz including the dedicated

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traffic channel whose ratio to the total received power $Traffic\_E_c/I_{or}$= −15.6dB. As specified in Table 3.3.3.3-1 of [8], to guarantee 0.5% FER for the traffic channel in AWGN, the ratio of the average energy per data bit to the effective noise power spectral density, $Traffic\_E_b/N_t$, should be 4.5dB or more. The maximum allowed thermal noise at the antenna can be computed as

$$I_{oc,th} = \hat{I}_{or} \frac{128 \cdot Traffic\_E_c / I_{or}}{Traffic\_E_b / N_t} = -103 \frac{\text{dBm}}{1.23\text{MHz}}. \quad (1)$$

With a 3dB margin,

$$N_{RX} + N_{TX} \le 10^{\frac{-106}{10}} \text{ [mW]}.$$

The maximum total interference power, $I_{oc,tot}$ = N_{RX} + N_{TX} + N_ϕ + P_{XM}, is set by the single-tone desensitization requirement. As defined in [8], the receiver FER must not exceed 1% when $\hat{I}_{or}$ = −101dBm/1.23MHz. Assuming that the total interference has a Gaussian distribution, the minimum required E_b/N_t is 4.3dB for 1% FER. Using (1) for $I_{oc,tot}$, we find that the total interference power referred to the antenna must not exceed −99.8dBm/1.23MHz, i.e.

$$10^{\frac{-106}{10}} + N_\phi + P_{XM} \le 10^{\frac{-99.8}{10}} \text{ [mW]}. \quad (2)$$

In the single-tone desensitization test of a cellular receiver, the jammer is 900kHz away from the RX channel center. So, the RX LO phase noise that appears in the RX channel due to the reciprocal mixing is confined to the 285-1515kHz frequency-offset range. This range is above the PLL loop bandwidth and, therefore, a free-running VCO phase noise should be used in calculations. Dual-band CDMA phones with a single IF of 183.6MHz use an UHF VCO tuned to 2105.3–2173.6MHz. Its phase noise is typically 133.5dBc/Hz at 900kHz offset and is proportional to $1/f^2$ in the mentioned frequency-offset range. Integrating the phase noise over this range and taking into account 3dB noise improvement due to the LO frequency divider by 2 yield 72.9dBc. Referencing this phase noise to the jammer power P_J = −30dBm at the antenna, we get N_ϕ = −102.9dBm/1.23MHz.

Using the results from [9], it can be shown that

$$P_{XM}\text{[dBm]} = P_J + 2(P_{TX} + L_{TX} - L_{TX-RX}) - 2IIP_3 - 2.4$$

where P_{TX} is the maximum TX power at the antenna (+23dBm typ.), L_{TX} is the duplexer TX-antenna insertion loss, and $L_{TX\text{-}RX}$ is the duplexer TX-RX isolation in the TX band. Substituting P_{XM} and N_ϕ into (2) results in the following relationship between the duplexer characteristics and the LNA minimum IIP3 satisfying the single-tone desensitization requirement:

$$IIP_{3,\min}\text{[dBm]} = 59.5 - L_{TX-RX} + L_{TX}.$$

For a typical cellular SAW duplexer, $L_{TX\text{-}RX}$ = 53dB and L_{TX} = 2.7dB in the worst case. Thus, the required minimum IIP3 is +9.2dBm for a cellular LNA.

III. Theory of BJT Linearity Improvement by Low-Impedance Low-Frequency Input Termination

As shown in [4], IIP3 of a common-emitter stage with dominant 2nd and 3rd-order nonlinearities is

$$IIP_3 \propto |\varepsilon(\Delta f, 2f)|^{-1} \quad (3)$$

where

$$\varepsilon(\Delta f, 2f) = g_3 - \frac{2g_2^2}{3}\left[\frac{2}{g_1 + g(\Delta f)} + \frac{1}{g_1 + g(2f)}\right] \quad (4)$$

and g_1, g_2 and g_3 are the expansion coefficients of the collector current into a Taylor series in terms of the base-emitter voltage. The 1st-order coefficient g_1 is the transconductance of the BJT. f is the center frequency of the input two tones and Δf is their spacing. $g(f)$ is a function of the circuit impedances and BJT parameters given in [4]. $\varepsilon(\Delta f, 2f)$ shows how the collector current nonlinearities contribute to its IM3 response. The first term in (4) comes from the 3rd-order nonlinearity and the terms in brackets come from the 2nd-order nonlinearity. The latter interacts with itself by first generating 2nd-order products and then mixing them with the fundamental tones yielding 3rd-order products. This self-interaction is due to multiple feedbacks in the circuit: through the emitter-degeneration network, the base-emitter and base-collector junctions. It can be shown that $g(2f)$ is negligible compared to g_1 for a wide range of 2nd-harmonic terminations. So, (4) can be simplified to

$$\varepsilon(\Delta f, 2f) \cong g_3 - \frac{2g_2^2}{3} \cdot \frac{2}{g_1 + g(\Delta f)} - \frac{2g_2^2}{3g_1}.$$

If the BJT operates below the strong injection region,

$$g_1 = \frac{I_C}{V_T}, \quad g_2 = \frac{I_C}{2V_T^2}, \quad g_3 = \frac{I_C}{6V_T^3}$$

where I_C is the collector dc current. In this case,

$$\varepsilon(\Delta f, 2f) = -\frac{I_C^2}{3V_T^4} \cdot \frac{1}{g_1 + g(\Delta f)}. \quad (5)$$

With negligible reactances at Δf, $g(\Delta f)$ is given by:

$$g(\Delta f) = \frac{1}{r_E + [r_B + Z_S(\Delta f)]/\beta} \quad (6)$$

where r_E is the emitter resistance including the dc resistance of an emitter degeneration inductor, r_B is the base resistance, β is the forward dc current gain and $Z_S(\Delta f)$ is the source impedance presented to the base of the BJT at the difference frequency. Substituting (6) into (5) and (5) into (3), we obtain

$$IIP_3 \propto \frac{3V_T^4}{I_C^2}\left| g_1 + \frac{1}{r_E + [r_B + Z_S(\Delta f)]/\beta} \right|. \quad (7)$$

According to (7), the maximum IIP3 is achieved with zero $Z_S(\Delta f)$ (negative real values result in instabilities) and is limited by r_E and r_B/β. However, if $g(2f)$ wasn't assumed negligible in (4), the resulting formula for IIP3 would show that it is possible to achieve a theoretically infinite IIP3 by optimum tuning both the difference-frequency and 2nd-harmonic terminations [4].

IV. Highly Linear LNA and Mixer Designs

Fig. 1 shows a simplified schematic of a cellular LNA designed in 0.5um SiGe BiCMOS technology. It can operate in the high-gain/high-linearity (HG/HL), high-gain/low-linearity (HG/LL) and low-gain (LG) modes. The HG/HL mode is used during a CDMA call when the receiver is subject to the cross-modulation distortion. The HG/LL mode is used when the receiver monitors a forward-link paging channel waiting for an incoming call. In this idle mode, the phone transmitter is off and the cross modulation is no longer an issue. Half of the LNA dc current is turned off. In both modes, the amplifying transistor Q1 is biased in the forward active mode by the bias circuit with very low output impedance at low frequencies. The external RF choke L2 isolates the bias circuit from the base of Q1 at the operating frequencies but acts as a short circuit at low frequencies. The LNA is switched to the LG mode to handle a strong received signal. In this mode, Q1 is disabled and bypassed through the M1 switch resulting in a high IIP3 with zero current.

The described method of providing a low $Z_S(\Delta f)$ was chosen because it allows a fast gain switching (within the required 100us). A low-frequency LC trap similar to the one described in [5] was also considered and dismissed because, to present a low $Z_S(\Delta f)$ in a wide Δf range, it requires a capacitor of about 100nF or higher. A slow discharge of this capacitor through a bias resistor would prevent a fast gain switching. An active inductor approach described in [6] eliminates the need for the external RF choke L2 but at the expense of an unacceptably high NF, low gain and potential instability of the active inductor.

The cellular mixer uses a single-balanced topology shown in Fig. 2. Q1 is the RF transconductor whose current is switched between the CDMA and FM mixer cores, Q4/Q5 and Q6/Q7, by the current steering switch Q2/Q3. The base of Q1 is terminated by a low $Z_S(\Delta f)$ thanks to a shunt matching inductor L1 connected to the mixer input through a 100nF dc blocking capacitor C1. The single-balanced mixer topology inherently doesn't reject the LO and its noise at IF. The latter is reduced by an LC-tank loading of the LO buffer. This tank can be discretely tuned to a low or high-side LO by switching C2 and C3. The LO leakage to IF is attenuated by the on-chip shunt capacitors C4-C7. The required mixer gain in the FM mode is lower than that in the CDMA mode. The gain is reduced by turning half of the Q1 current off and by switching its emitter degeneration inductor L3 to a higher value with M1. The open-collector differential CDMA and FM IF outputs are biased to V_{CC} through external chokes.

V. Measured Results

The LNA and mixer performances measured at 3V power supply are summarized in Tables I and II respectively. Figures 3 and 4 show the 2-tone transfer characteristics of the LNA in the HG/HL mode and the mixer in the CDMA mode respectively. State-of-the-art LNAs are compared in Table III where DRM is a dynamic range merit [10] defined as OIP3/[(NoiseFactor-1)$\cdot P_{DC}$]. The reported here LNA is inferior in DRM only to the LNA of [4]. However, it doesn't rely on the distortion cancellation and, therefore, the achieved high IIP3 is fairly frequency independent and consistent from part to part.

VI. Conclusion

The designed LNA and mixer exhibit state-of-the-art linearity thanks to the low-impedance low-frequency input terminations as predicted by the theory. To our knowledge, the mixer IIP3 is the highest among published active mixers reaching that typically reported for passive mixers.

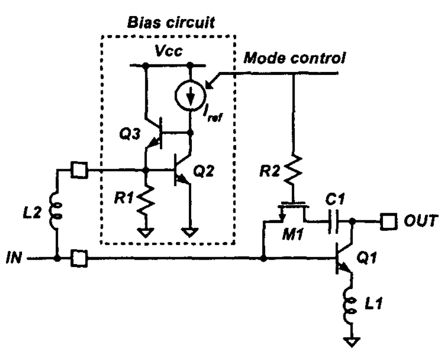

Fig. 1. Simplified schematic diagram of cellular LNA.

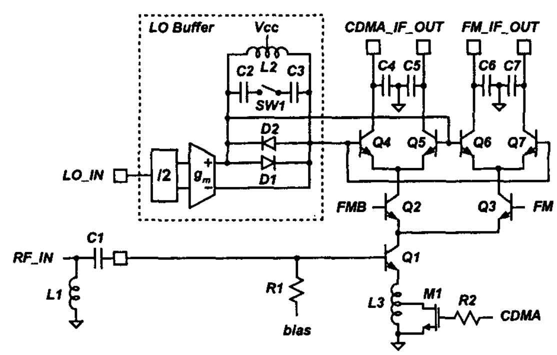

Fig. 2. Simplified schematic of cellular mixer.

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TABLE I

CELLULAR LNA MEASURED PERFORMANCE SUMMARY

Parameter	HG/HL mode	HG/LL mode	LG mode
Power Gain	16.3dB	15.7dB	-3.1dB
NF	1.5dB	1.4dB	3.2dB
IIP3	+12.2dBm	+10.3dBm	+23.0dBm
Current	Q1: 7.7mA Total: 9.4mA	Q1: 3.9mA Total: 5.4mA	Total: 0mA

TABLE II

CELLULAR MIXER MEASURED PERFORMANCE SUMMARY

Parameter	CDMA mode	FM mode
RF/IF/LO Frequency	880/183.6/1063.6 MHz	
LO Input Power	0dBm	
Conv. Power Gain	10.9dB	8.4dB
SSB NF	6.7dB	7.6dB
IIP3	+14.7dBm	+8.0dBm
LO-to-IF Leakage	-19.6dBm	-32.2dBm
Current	Q1: 8.4mA Total[1]: 17.8mA	Q1: 4.0mA Total: 12.9mA

[1]The total mixer current includes the input LO buffer current.

TABLE III

COMPARISON OF STATE-OF-THE-ART LNAS

Work	Technology	Freq GHz	Gain dB	NF dB	IIP3 dBm	I_{DC} mA@V	DRM
[4]	Si BJT	2	16.0	1.7	+16	5@2.7	245
This work HG/LL	0.5um SiGe BiCMOS	0.88	15.7	1.4	+10.3	3.9 @3.0	89.5
This work HG/HL	0.5um SiGe BiCMOS	0.88	16.3	1.5	+12.2	7.7 @3.0	74.3
D.Ahlgren 1999	0.5um SiGe BiCMOS	0.9	15.0	1.4	+12	8 @3.0	54.9
G. Wevers 2000	70GHz SiGe HBT	1.96	14.7	1.0	+10.3	8.9 @3.0	43.8
[1]	0.4um GaAs PHEMT	0.88	12.5	1.0	+8	5 @3.0	28.9
P. Shah 2000	0.5um SiGe BiCMOS	1.96	15.3	1.9	+7.6	5 @2.7	26.3
[2]	0.5um Si CMOS SOI	2	16.5	2.0	+10	17.5 @3.0	14.5
J. Lin 2001	0.25um Si BiCMOS	0.9	13.0	1.4	+9.3	12.9 @3.0	11.5
D. Wang 2001	0.5um SiGe BiCMOS	2.5	12.0	1.6	+8	7.5 @2.75	10.9

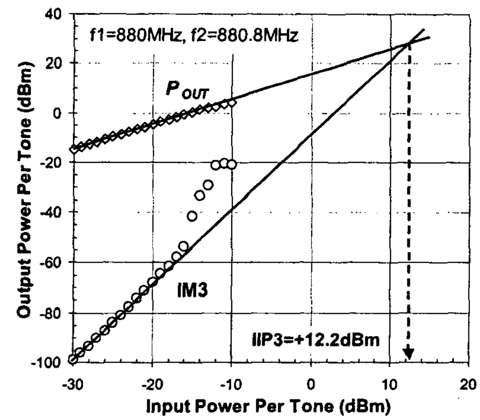

Fig. 3. Two-tone transfer characteristics of cellular LNA.

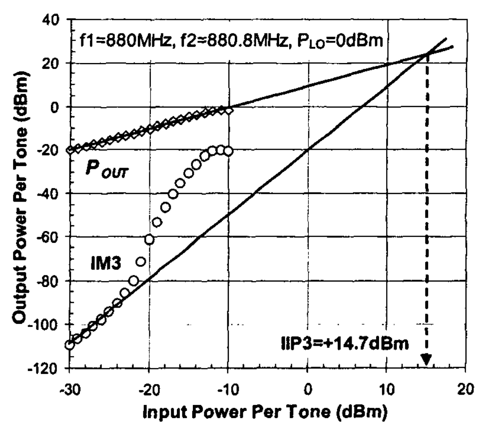

Fig. 4. Two-tone transfer characteristics of cellular mixer.