

Linearization of Monolithic LNAs Using Low-Frequency Low-Impedance Input Termination

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Abstract:

This work investigates the linearization technique based on terminating the LNA input with a low impedance at the low frequencies of the 2nd-order mixing terms. This technique is analysed using the Volterra series and is shown to be very effective in linearizing BJTs but not FETs if the latter are biased in the strong inversion region. Several methods to generate the low-frequency low-impedance input termination are reviewed. A SiGe BiCMOS cellular band CDMA LNA using this linearization technique is described. The LNA achieves +11.7dBm IIP3, 15.7dB gain and 1.4dB NF with a current consumption of only 3.9mA@3V.

1. Introduction

Many wireless communication systems operate in a hostile jammer environment and require exceptionally linear receiver front-ends. In CDMA phones, narrowband jammers degrade the receiver sensitivity due to the cross modulation distortion that is generated by the LNA nonlinearities in the presence of the transmitted signal leakage [1]. 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 able 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 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.

Traditionally, FETs (both GaAs and silicon) were perceived to be more linear devices than BJTs. At the early stages of the wireless revolution, GaAs-based HEMTs were commonly used in low-distortion amplifiers either as discrete devices or in ICs [2]. 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.

Si MOSFETs are a less expensive alternative to GaAs FETs. Their use as the main amplifying devices in CDMA front-end designs started only recently, after their f_T improved significantly thanks to the shrinking

gate length. However, despite a more linear dc transfer characteristic than that of a BJT, MOSFETs require a high dc current to achieve a high input-referred 3rd-order intercept point (IIP3) [3].

To reduce the dc current, linearization techniques must be employed. One of these techniques is the feed-forward linearization used to achieve a very high IIP3 of a differential CMOS LNA reported in [4]. The addition of an auxiliary gain stage in this LNA degraded its gain by 2.5dB and NF by 0.2dB. This technique is also very sensitive to mismatches between the main and auxiliary gain stages and the errors in the signal scaling.

The derivative superposition method [5] has also been used to linearize FET LNAs [6], [7]. This method is based on superimposing different regions of the FET dc transfer characteristics with opposite signs of the 3rd-order derivative. It requires a precise manual bias control of parallel FETs and their accurate scaling to achieve the distortion cancellation.

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 [8]. The demonstrated IIP3 improvement of a discrete Si BJT LNA was achieved without degrading the gain and NF and increasing the dc current but was confined to a narrow frequency range due to the frequency dependence of the distortion cancellation effect.

This paper describes a simpler version of the optimum out-of-band tuning technique that is based on using a low-impedance low-frequency termination of the amplifier input. The RF theory of this technique is laid out here using the Volterra series analysis. Several methods to generate a low-frequency low-impedance input termination without affecting the circuit in-band performance are reviewed. The measured data collected on a SiGe BiCMOS LNA designed for cellular CDMA applications confirm the effectiveness of the low-frequency low-impedance input termination technique.

2. Volterra Series Analysis of Common-Emitter BJT and Common-Source FET Amplifiers

Consider a common-emitter BJT and a common-source FET gain stages whose nonlinear ac equivalent circuits are shown in Fig. 1 (a) and (b) respectively where C_{jE} and C_{DE} are the base-emitter junction and diffusion capacitances respectively, r_B and r_E are the base and emitter resistances respectively, β is the dc current

gain, τ_F is the forward transit time and C_{GS} is the gate-source capacitance. The only nonlinear element in both circuits is considered to be the output current (i_c or i_d) that is expanded into a Taylor series in terms of the input voltage (v_π or v_{gs}). g_1 , g_2 and g_3 are the expansion coefficients where g_1 is the small-signal transconductance and the higher-order coefficients define the strengths of the corresponding nonlinearities of the dc transfer characteristic. In addition to the main nonlinearities of $i_c(v_\pi)$, a BJT has so-called "tracking" nonlinearities due to the dependence of the base current i_b and the current through the base-emitter diffusion capacitance $i_{C_{DE}}$ on the collector current i_c as shown in Fig. 1 (a).

A Volterra series analysis shows that IIP3 of both stages is given by the same formula [8]:

$$\text{IIP3} = \frac{1}{6\text{Re}(Z_1(\omega)) \cdot |H(\omega)| \cdot |A_1(\omega)|^3 \cdot |\epsilon(\Delta\omega, 2\omega)|} \quad (1)$$

where ω is the angular center frequency of the input two tones, $\Delta\omega$ is their frequency separation,

$$\epsilon(\Delta\omega, 2\omega) = g_3 - \frac{2g_2^2}{3} [2k(\Delta\omega) + k(2\omega)], \quad (2)$$

and $H(\omega)$, $A_1(\omega)$ and $k(\omega)$ are given in Table I. The quantity $\epsilon(\Delta\omega, 2\omega)$ shows the effect of the transconductance nonlinearities on IM3 distortion. As can be seen from (2), this distortion is generated not only by the 3rd-order nonlinearity represented by g_3 but also by the 2nd-order nonlinearity represented by g_2 . The g_2 contribution is explained by the feedback through the degeneration impedance Z_2 and through the base current i_b and base-emitter diffusion capacitance current $i_{C_{DE}}$ that are functions of the collector current i_c . These feedback paths allow the 2nd-order distortion terms of the output current to be remixed with the fundamental tones contributing to the IM3 distortion. This remixing is called 2nd-order interaction. Since the amplitude and

phase of the 2nd-order distortion terms depend on the termination impedances Z_1 and Z_2 at the corresponding frequencies ($\Delta\omega$ and 2ω in this case), these out-of-band impedances affect the IM3 distortion.

In the absence of the feedback paths causing the 2nd-order interaction, $\epsilon(\Delta\omega, 2\omega) = g_3$. It can be shown that, if the real parts of Z_1 and Z_2 are positive (which is the case in a stable amplifier design), the term in brackets of (2) has a positive real part as well. Therefore, $|\epsilon(\Delta\omega, 2\omega)|$ can be reduced below $|g_3|$ only if g_3 is positive. This is the case in BJT's. For FET's operating in the strong inversion region, g_3 is negative and the out-of-band tuning can not reduce $|\epsilon(\Delta\omega, 2\omega)|$ below $|g_3|$. In the weak and moderate inversion regions, g_3 is positive but the low transconductance and f_T in these regions make them a poor choice for RF amplification.

3. Theory of BJT Linearization by Low-Impedance Low-Frequency Input Termination

It can be shown that, if a BJT is strongly degenerated ($|Z_2(\omega)| \gg 1/g_1$) and its f_T is much larger than the 2nd-harmonic frequency of the input excitation and the driving impedance at the 2nd-harmonic frequency is fairly small ($|Z_1(2\omega)| \ll \gamma|Z_2(2\omega)|$ where γ is equal to either β or $\omega_T/2\omega$ whichever is smaller), then

$$k(2\omega) = \frac{1}{g_1} \quad (3)$$

and (2) is simplified to

$$\epsilon(\Delta\omega, 2\omega) \cong g_3 - \frac{4g_2^2}{3} k(\Delta\omega) - \frac{2g_2^2}{3g_1} \quad (4)$$

For BJT operating 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 and V_T is the thermal voltage. In this case, the first and last terms of (4) cancel

Table I. Expressions for $H(\omega)$, $A_1(\omega)$ and $k(\omega)$

BJT common-emitter stage	FET common-source stage
$H(\omega) = \frac{1 + j\omega C_{JE}[Z_1(\omega) + Z_2(\omega)]}{g_1}$	$H(\omega) = \frac{1 + j\omega C_{GS}[Z_1(\omega) + Z_2(\omega)]}{g_1}$
$A_1(\omega) = \frac{1}{1 + g_1 Z_2(\omega) + (g_1/\beta + j\omega\tau_F g_1 + j\omega C_{JE})[Z_1(\omega) + Z_2(\omega)]}$	$A_1(\omega) = \frac{1}{1 + g_1 Z_2(\omega) + j\omega C_{GS}[Z_1(\omega) + Z_2(\omega)]}$
$k(\omega) = A_1(\omega) \{ Z_2(\omega) + (1/\beta + j\omega\tau_F)[Z_1(\omega) + Z_2(\omega)] \}$	$k(\omega) = A_1(\omega) Z_2(\omega)$

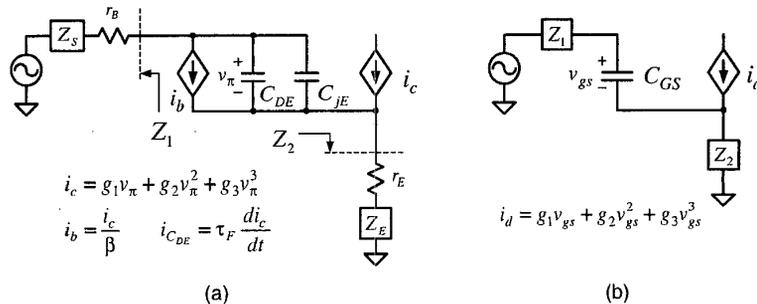


Fig. 1 Equivalent circuits of BJT common-emitter (a) and FET common-source (b) amplifiers.

out resulting in

$$\varepsilon(\Delta\omega, 2\omega) = -\frac{I_C^2}{3V_T^4} k(\Delta\omega). \quad (5)$$

In the case of an inductive emitter degeneration commonly used in BJT LNAs, with negligible reactances at $\Delta\omega$,

$$k(\Delta\omega) \approx \left\{ g_1 + \frac{1}{r_E + [r_B + Z_S(\Delta\omega)]/\beta} \right\}^{-1}. \quad (6)$$

Substituting (6) into (5) and (5) into (1), we obtain

$$IIP_3 \approx \frac{\frac{V_T^4}{I_C^2} \left| g_1 + \frac{1}{r_E + [r_B + Z_S(\Delta\omega)]/\beta} \right|}{2 \operatorname{Re}(Z_1(\omega)) \cdot |H(\omega)| \cdot |A_1(\omega)|^3}. \quad (7)$$

According to (7), the IIP3 of a BJT common-emitter amplifier depends on the source impedance at the difference frequency in such a way that reducing $Z_S(\Delta\omega)$ results in a higher IIP3. Since negative real values of $Z_S(\Delta\omega)$ result in instabilities, the minimum useful value of $Z_S(\Delta\omega)$ is 0Ω in which case the BJT IIP3 is maximum and limited by r_E and r_B . Selecting a larger emitter length results in smaller parasitic resistances and a higher IIP3.

It should be emphasized that (7) was derived here for a strongly-degenerated BJT operating at a frequency well below its f_T and terminated at the input with a relatively small impedance at the 2nd-harmonic frequency. These conditions make the IIP3 insensitive to the 2nd-harmonic termination. However, in a more general case, $Z_S(2\omega)$ can be optimally tuned together with $Z_S(\Delta\omega)$ to achieve a precise cancellation of the distortions generated by the 2nd and 3rd-order BJT nonlinearities resulting in a much higher IIP3 than the one predicted by (7) [8].

4. Low-Frequency Low-Impedance Input Termination Techniques

There are three commonly-used techniques to generate a low-frequency low-impedance input termination. They are summarized in Fig. 2. In the LC trap technique, the trap can either be a dedicated LC network [9] as shown in Fig. 2(a) or formed by the input matching components as shown in Fig. 2(b). In the first case, the purpose of L is to create a high impedance at the fundamental frequency such that the trap does not affect the in-band operation of the amplifier. In the second case, L can be a matching inductor. The purpose of C can be dual. In the dedicated LC trap shown in Fig. 2(a), C can be selected to resonate with L at a certain difference frequency of the input two tones. In this case, IIP3 is improved only for the tones with the specified frequency separation or very close to it. Since the frequency separation of the jammers received by a phone is random, the resonating LC trap is rarely used. Besides, in many cases, the cross modulation distortion is more dangerous to the receiver sensitivity than the inter-modulation distortion of multiple jammers, and the low frequency range that affects the cross modulation distortion in the desired signal band can be as wide as 285kHz to 45MHz [1].

More often, the capacitor in the LC trap is selected to have a very low impedance in a wide range of low

frequencies. However, to cover the lowest frequency affecting the cross modulation distortion (285kHz), a very large capacitor value (typically several μF) is required. Since LNAs in many wireless systems are designed to have multiple gain modes, selecting such a large value capacitor in the LC trap significantly slows down the LNA gain switching due to the charging and discharging action of C through the bias resistor. Some high-data rate applications require the LNA gain switching time to be less than 10 μs limiting the capacitor value selection well below 1nF. If the gain switching speed is not critical, the LC trap is a very attractive and commonly-used method due to its simplicity and negligible effect on the LNA gain, NF and stability.

The gain switching time can be reduced by using an active inductor bias shown in Fig. 2(c) [10]. The bias circuit uses an operational amplifier with a negative resistive feedback. The feedback resistors isolate the opamp input and output from the signal path at the fundamental frequency. The opamp dc gain is selected such that its output impedance within its 3dB bandwidth is very small. The major challenge of the active inductor design is to ensure that the opamp has enough gain at the highest frequency affecting the LNA cross modulation distortion (the separation frequency of the TX and RX channels) to provide a low-impedance termination at that frequency and has a low enough gain at the fundamental frequency to prevent its shorting of the LNA input. This condition is typically very difficult to satisfy without sacrificing the opamp phase margin. As a result, the bias circuit often shunts the LNA input leading to a lower gain and higher NF [10].

A compromise between the opamp stability and the LNA performance can be achieved if the opamp is isolated from the LNA input by an RF choke as shown in Fig. 2 (d). In this case, the opamp can be made as broadband as possible without affecting the LNA gain and NF. Since there is no need for a large value capacitor in the input matching circuit, the gain switching can be made very fast. This is the approach we have taken here.

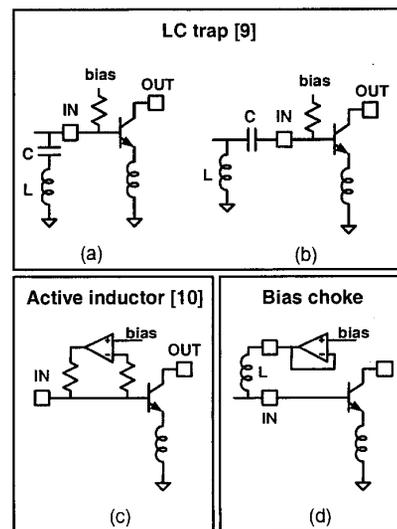


Fig. 2 Low-frequency low-impedance input termination techniques.

5. Highly-Linear SiGe BiCMOS LNA

Fig. 3 shows a simplified schematic diagram of a cellular band LNA designed in a low-cost 0.5 μ m SiGe BiCMOS technology. Q1 is the main amplifying transistor degenerated by the on-chip inductor L1. The FET bypass switch M1 is used to implement a low-gain mode. The bias circuit is connected to the LNA input through an external RF choke L2 and consists of a current mirror Q2 with a beta helper Q3. Due to the negative feedback through Q3 and a high dc voltage gain of Q2, the output impedance of the bias circuit is very low. At frequencies well below fundamental, the LNA input is loaded by the bias circuit output impedance (L2 is almost a dc short) and the output impedance of the antenna duplexer. For a broad range of ceramic and SAW cellular duplexers, their RX port impedance can be accurately modeled by a shunt 8-10pF capacitance below 500MHz. This capacitance controls the location of the non-dominant pole of the bias circuit. To make the bias circuit as broadband as possible (to cover 45MHz) without reducing its phase margin, the non-dominant pole is pushed to higher frequencies by selecting a fairly large dc current through Q3.

The measured LNA gain and NF are 15.7dB and 1.4dB respectively and the current consumption is 3.9mA without the bias circuit and 5.4mA with the bias circuit from 3V supply. The gain switching time is less than 1 μ s. The measured fundamental and IM3 output powers as functions of the input power are plotted in Fig. 4. As can be seen, the LNA achieves +11.7dBm IIP3 in a wide range of the input power. The dynamic range figures-of-merit (DRM) defined as $OIP3 / [(NoiseFactor - 1) \cdot P_{DC}]$ [11] for this and other state-of-the-art LNAs are compared in Table II. As can be seen, this LNA exhibits a better DRM than all other monolithic LNAs in any other technologies, and is inferior in DRM only to the discrete Si BJT LNA of [8].

6. Conclusion

We have investigated the linearization technique that is based on terminating an amplifier input with a low impedance at low frequencies. This technique is shown to significantly reduce the distortion in BJT amplifiers while FETs were proved to be less sensitive to the out-of-band terminations when biased in the strong inversion region. The designed LNA achieved state-of-the art IIP3.

7. References

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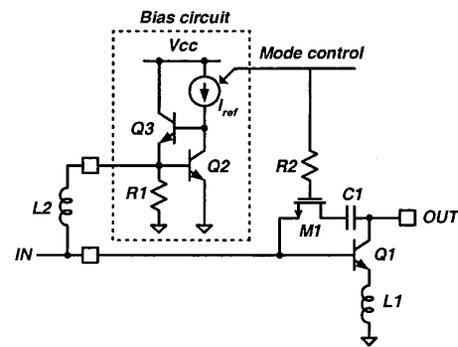


Fig. 3 Simplified schematic diagram of LNA.

Table II. Comparison of State-of-the-Art LNAs

Work	Technology	Freq GHz	Gain dB	NF dB	IIP ₃ dBm	I _{DC} mA@V	DRM
[8]	Si BJT	2	16.0	1.7	+16	5.0@2.7	245
This Work	0.5 μ m SiGe BiCMOS	0.88	15.7	1.4	+11.7	3.9@3.0	123
[6]	0.6 μ m GaAs MESFET	0.9	17	1.6	+8.5	4.7@2.7	62.8
[11]	0.5 μ m SiGe BiCMOS	0.9	15.0	1.4	+12	8.0@3.0	54.9
[7]	0.25 μ m Si CMOS	2.2	14.9	3	+16.1	9.4@2.5	53.8
[2]	0.4 μ m GaAs PHEMT	0.88	12.5	1.0	+8	5.0@3.0	28.9

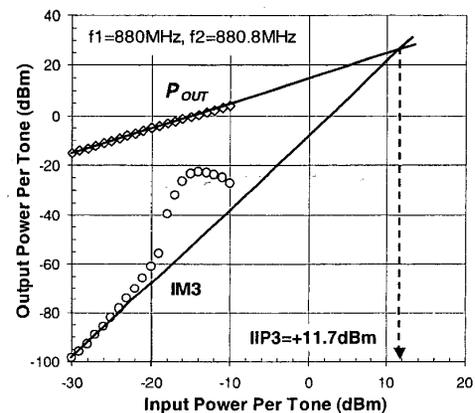


Fig. 4 Measured 2-tone transfer characteristics.