ECEN 665

LNA Linearization Using Bipolar Transistors

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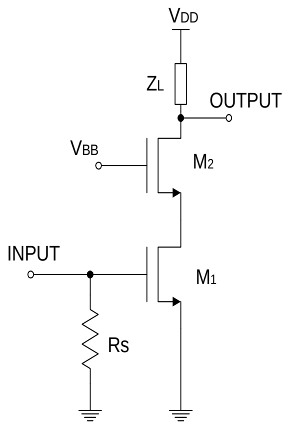
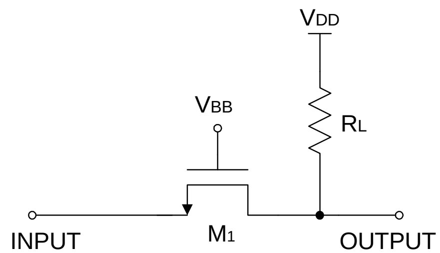
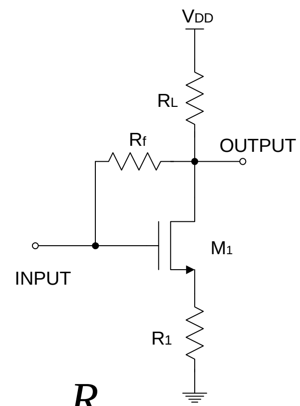
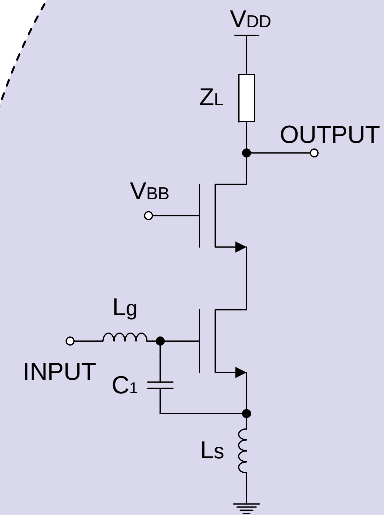
Review: Conventional CMOS LNA Topologies

Resistive Termination	Common Gate	Shunt-series Feedback	Source Degeneration









Resistive Termination

$$Z_{in} = R_s$$

$$F \geq 2 + \frac{4\gamma}{\alpha}\frac{1}{g_{m1}R_s}$$

NF: > 6dB

Common Gate

$$Z_{in} = \frac{1}{g_{m1}}$$

$$F \geq 1 + \frac{\gamma}{\alpha}$$

4.8 dB

Shunt-series Feedback

$$Z_{in} \approx \frac{R_f}{1 + \frac{R_L}{R_1}}$$

$$F = 1 + \frac{1}{1+g_m R_L} + \frac{1}{g_m R_s}\frac{\gamma}{\alpha} + \frac{1}{g_m^2 R_S R_L}$$

2~5 dB

Source Degeneration

$$Z_{in} = j\omega(L_g + L_s) + \frac{1}{j\omega C_{gs}} + \frac{g_m}{C_{gs}} L_s$$

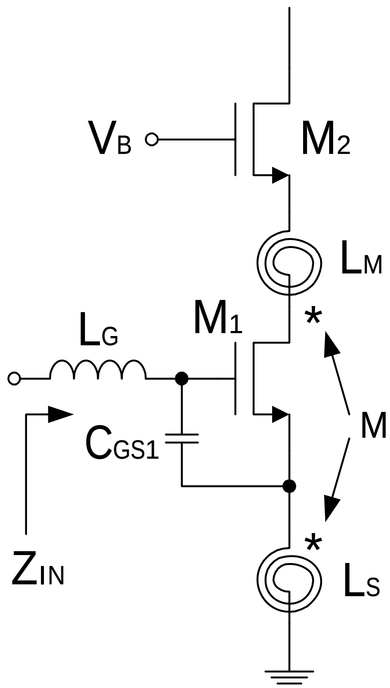
< 2dB

A Mutual Coupled Degenerated LNA

Input Impedance Matching

The goal: Input impedance match to 50 or 75 Ohm

$$Z_{in} = s(L_G + L_S) + \frac{1}{sC_{gs1}} + \frac{g_m}{C_{gs1}}(L_S \pm M)$$



At resonant Frequency:

$$\omega_o = \frac{1}{\sqrt{(L_G + L_S)C_{gs1}}} \Rightarrow R_{in} = \omega_T(L_S \pm M)$$

$$\omega_T = \frac{g_m}{C_{gs1}} \propto \frac{1}{L^2}(V_{gs1} - V_{th1})$$

$$IIP3 \propto V_{gs1} - V_{th1}$$

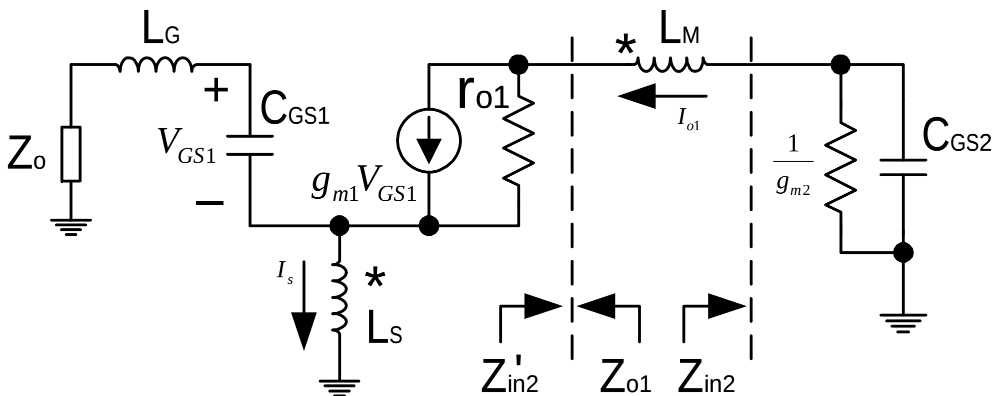
$$I_D \propto \frac{W}{L}(V_{gs1} - V_{th1})^2$$

Design trade-offs: $IIP3 \uparrow \rightarrow (V_{gs1} - V_{th1}) \uparrow \rightarrow I_D \uparrow \rightarrow (L_S - M)$

$I_D \downarrow \rightarrow (V_{gs1} - V_{th1}) \downarrow \rightarrow IIP3 \downarrow \rightarrow \omega_T \downarrow \rightarrow (L_S + M)$

Inter-stage Impedance

Verification of reduced Miller effect



$$Z_{o1} = 2r_{o1} + \frac{\omega_o^2}{\omega_T} L_s + j\frac{\omega_o}{\omega_T} Z_o \approx 2r_{o1}$$

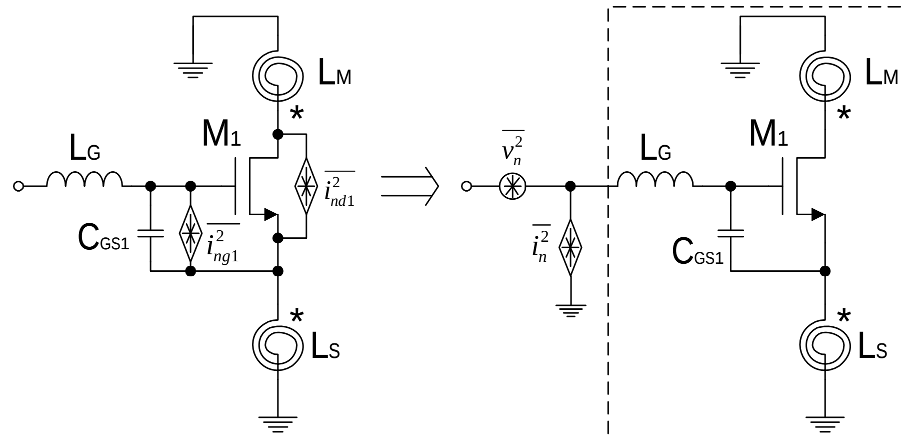
$$Z_{in2}^{'} \leq \frac{1}{g_{m2}}$$

$$Z_{in2}^{'} \approx \frac{1}{g_{m2}} \frac{1}{1+\left(\frac{\omega_o}{\omega_{T2}}\right)^2} \pm \left(\frac{\omega_o}{\omega_T}\right)^2 \omega_T M$$

- More current into cascoded stage
- Reduced inter-stage gain
- Reduced Miller effect
- Improved reverse isolation

Noise

- Noise model



$$\overline{i_n^2} = \overline{\left(i_{ng1} + i_{nd1}\, j \frac{\omega_o}{\omega_T} \right)^2}$$

$$\overline{v_n^2} = \left(-j \frac{i_{ng1}}{j\omega_o C_{gs1}} \right)^2$$

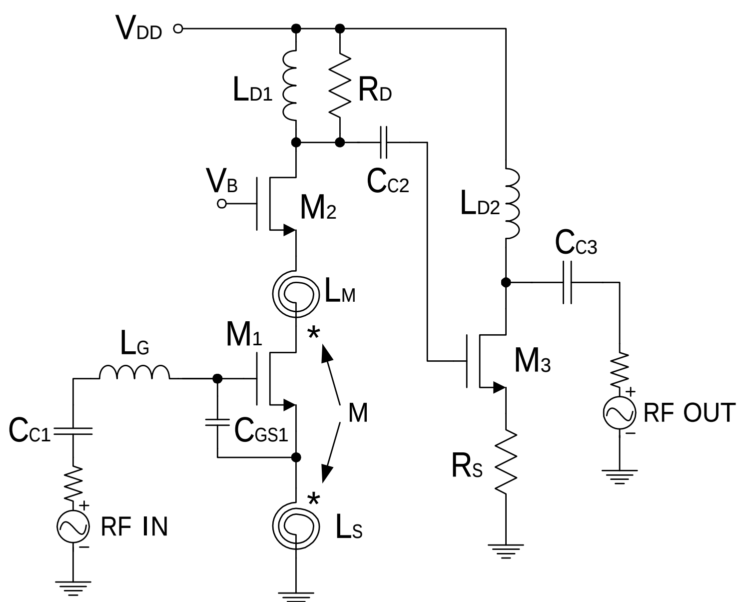
M does not appear!

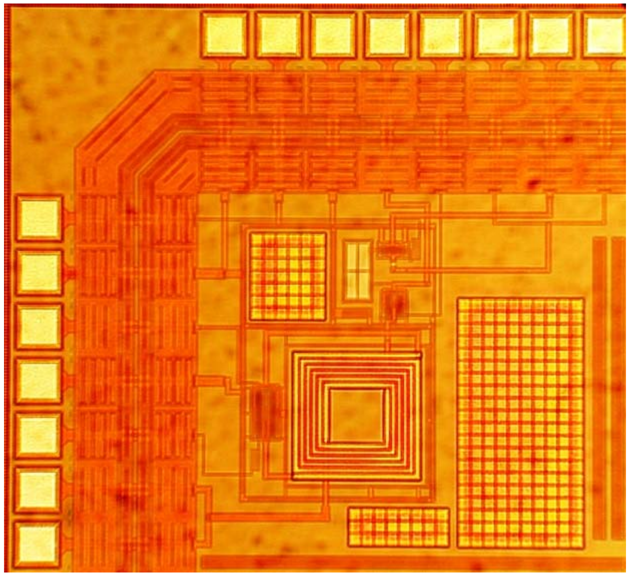
Noise figure degradation:

- Loss due to inferior inductor
- Substrate loss

A GSM LNA Using the Proposed Matching Method

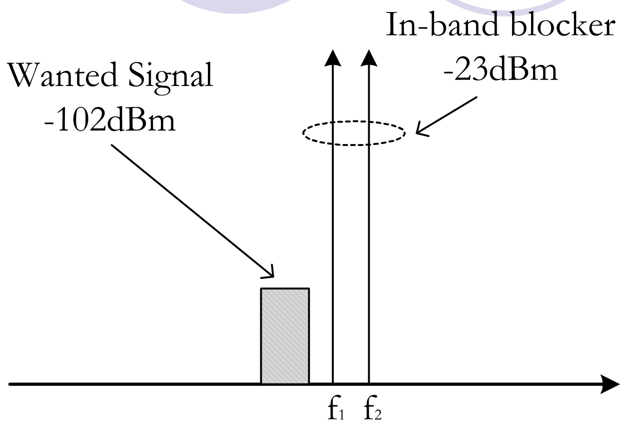
- Effective transconductance $$G_m = -j\left(\frac{\omega_T}{\omega_o}\right)\frac{1}{R_s}$$



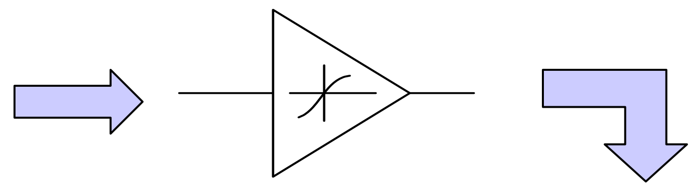


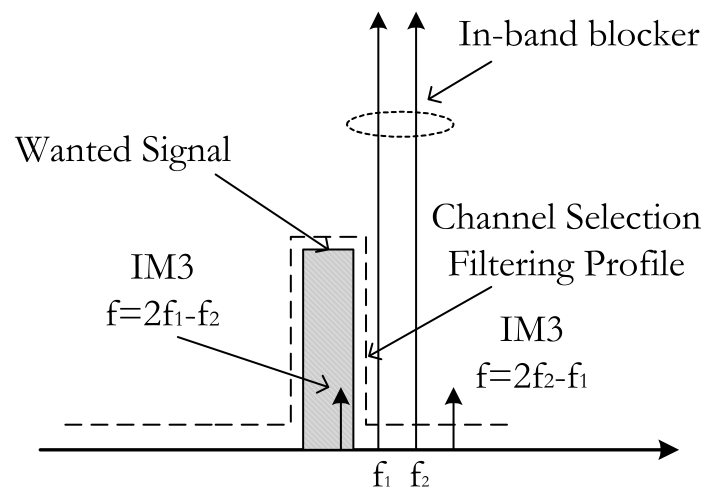
C. Xi and E. Sanchez-Sinencio, “ A GSM LNA Using Mutual-Coupled Degeneration.” *IEEE Microwave and Wireless Component Letters*, Vol. 15, No. 2, pp 68 -70, Feb. 2005

Why Linearity So Important ?



Communication system always deals with interferences.





Unwanted non-linearity will :

- Compress amplified signal
- Desensitize front-end
- Generate harmonics (filter out)
- Generate in-band interference (IMD)
- Cross-modulation

Linearity Metrics

- 1dB compression:

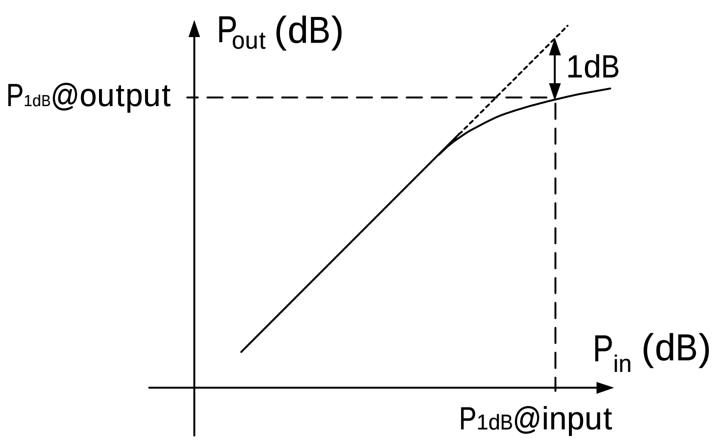
Measure gain compression for large input signal

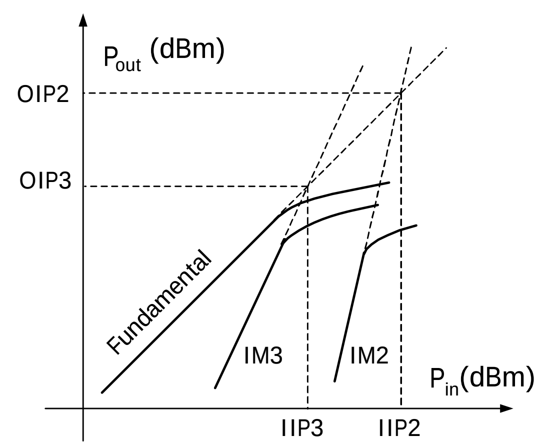
- IIP3/IIP2:

Measure inter-modulation behavior

Relationships between IIP3 and P1dB

- For one tone test: IIP3-P1dB=10dB
- For two tone test: IIP3-P1dB=15dB





Next we will consider the mobility not constant, i.e. ,

$$\mu(v_{DS}) = \frac{\mu_O}{1+\theta V_{ds}}$$

$$where\ \ V_{ds} = V_{od} = V_{GS} - V_t$$

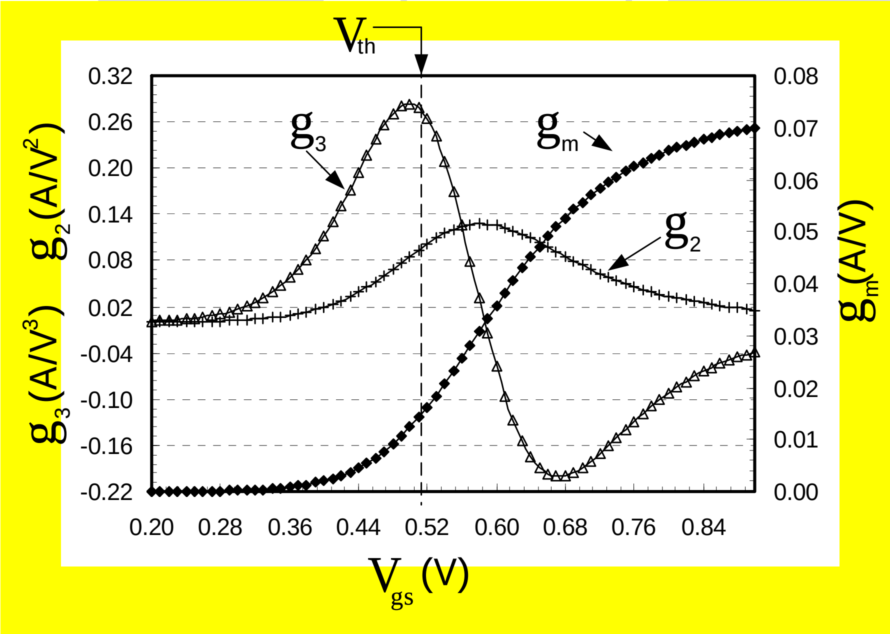
Non-linearity Terms of MOS Device

Intrinsic MOS I-V characteristic:

$$i_{ds} = K\frac{\chi^2}{1+\theta\chi}$$

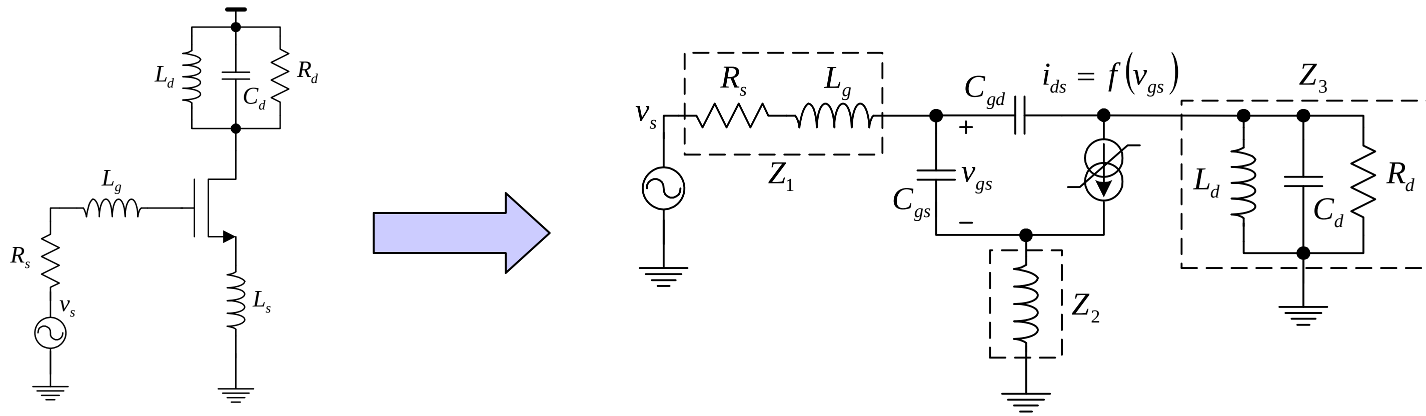
$$\chi = 2\eta\phi_t \ln\left(1+\exp\frac{V_{gs}-V_{th}}{2\eta\phi_t}\right)$$

$$i_{ds}(v_{gs}) = g_m v_{gs} + g_2 v_{gs}^{\ 2} + g_3 v_{gs}^{\ 3}$$



Inversion Level	g_m	g_2	g_3
Strong/moderate	$\frac{KV_{od}(2+\theta V_{od})}{(1+\theta V_{od})^2}$	$\frac{K}{(1+\theta V_{od})^3}$	$-\frac{\theta K}{(1+\theta V_{od})^4}$
Weak	$\frac{I_{s0}}{\eta\phi_t}$	$\frac{I_{s0}}{2(\eta\phi_t)^2}$	$\frac{I_{s0}}{6(\eta\phi_t)^3}$

Non-Linearity Analysis of Conventional Inductive Degenerated LNA



$$IIP3(2\omega_2 - \omega_1) = \frac{1}{6R_s \cdot |H(\omega)| \cdot |A_1(\omega)|^3 \cdot |\varepsilon(\Delta\omega, 2\omega)|}$$

$$\varepsilon(\Delta\omega, 2\omega) = g_3 - g_{oB}(\Delta\omega, 2\omega)$$

$$g_{oB}(\Delta\omega, 2\omega) = \frac{2}{3} g_2^2 \left[\frac{2}{g_m + g(\Delta\omega)} + \frac{1}{g_m + g(2\omega)} \right]$$

$g(\cdot)$ is a function of Z1, Z2 and Z3

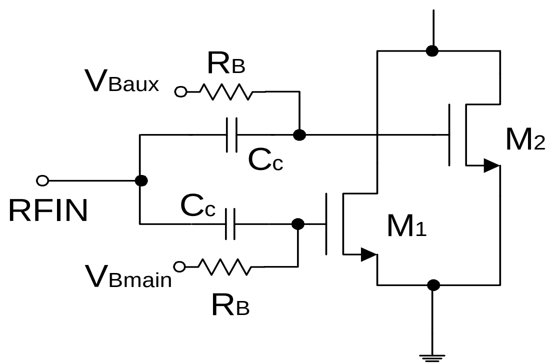
g_3	Negative
g_{oB}	Positive

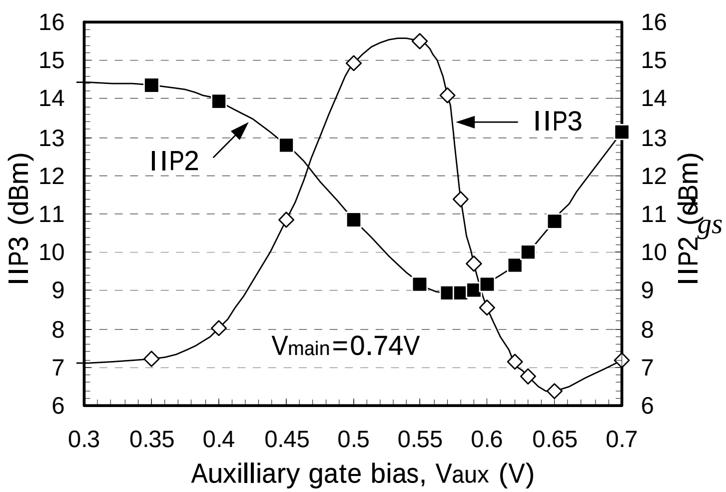
The absolute value of these two quantities should be kept small in order to achieve high linearity.

$$|g_3| \downarrow, \; |g_{oB}| \downarrow \;\Rightarrow\; IIP3 \uparrow$$

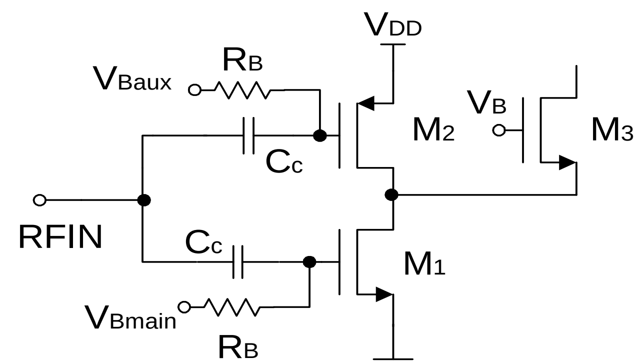
Review of Multi-Gated-Transistor Linearization

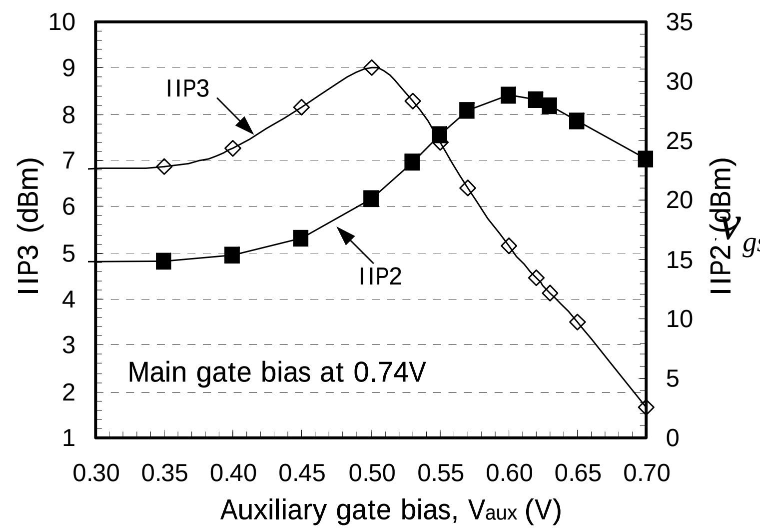
Original Configuration:



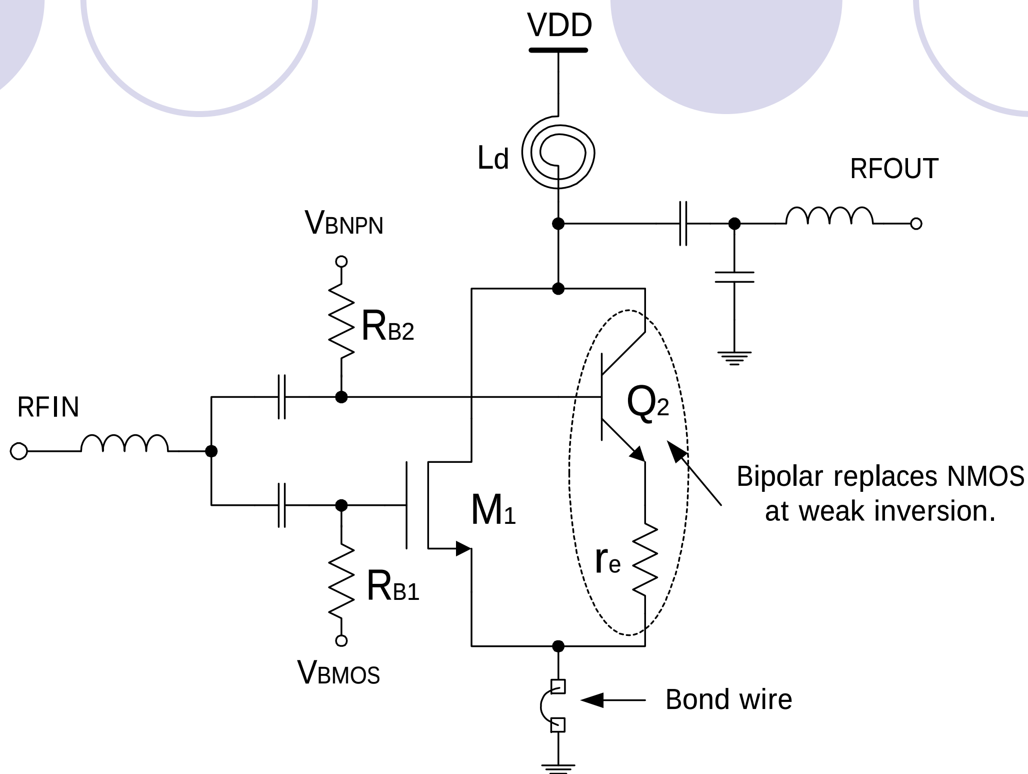


Alternate Configuration:



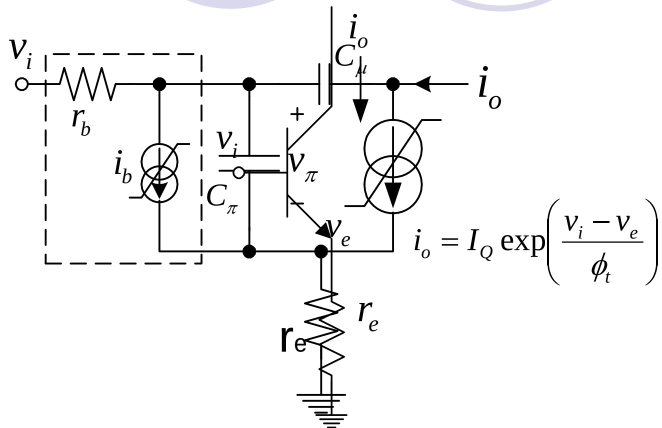


Proposed Method: Hybrid LNA

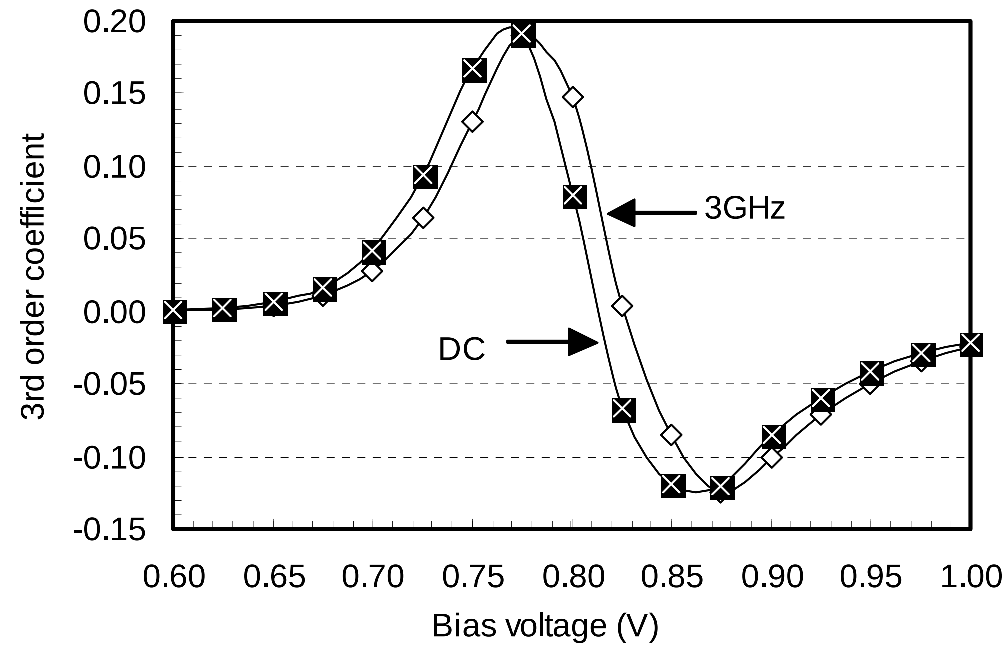


- MOS in weak inversion has speed problem
- MOS transistor in weak inversion acts like bipolar
- Bipolar available in TSMC 0.18 technology (not a parasitic BJT)
- Why not using that bipolar transistor to improve linearity ?

Linearity Analysis of the BJT



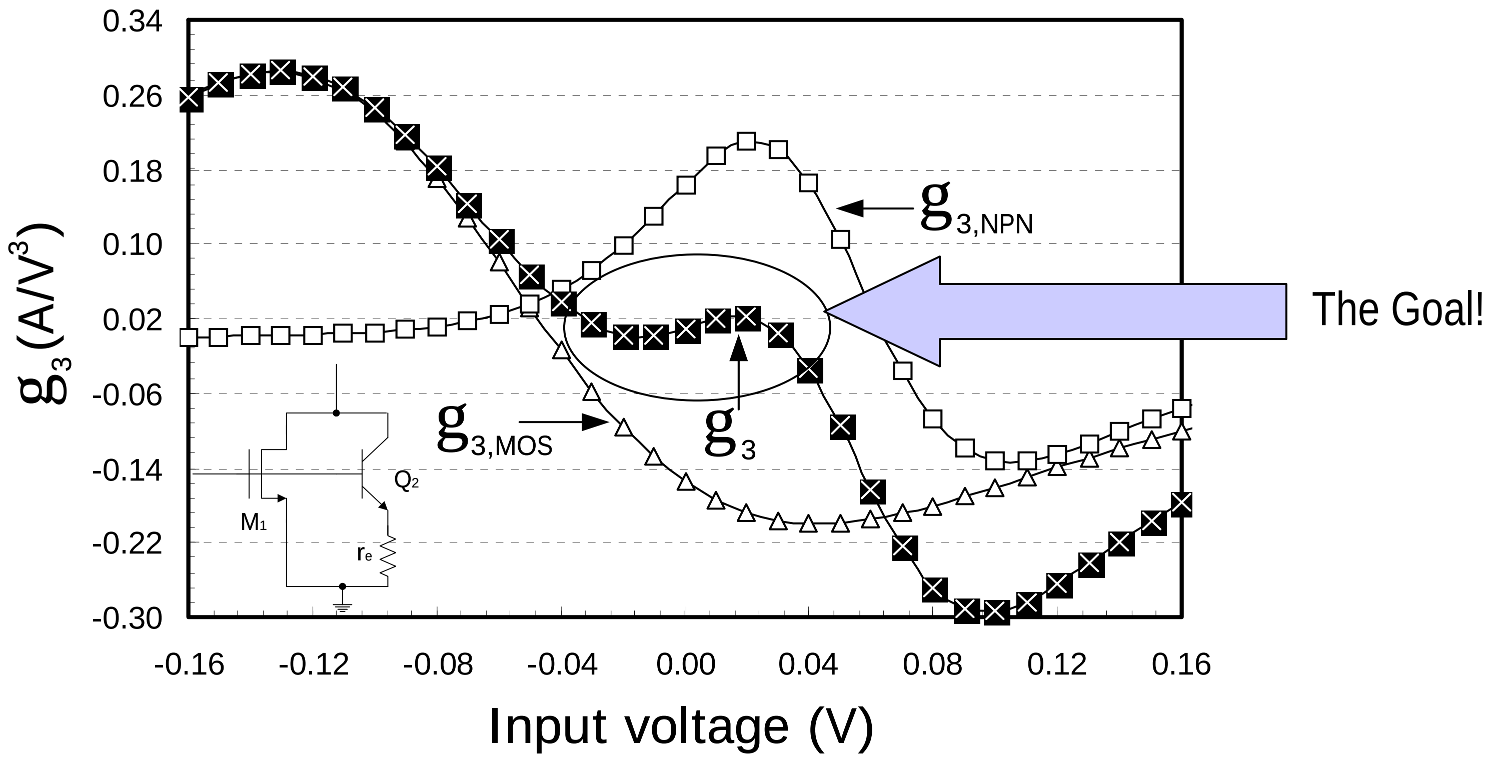
Weak memory effect



$$g_{3,bjt} = \frac{1}{6I_Q^2}\frac{g_m^3}{(1+g_m r_e)^5}(1-2g_m r_e)$$

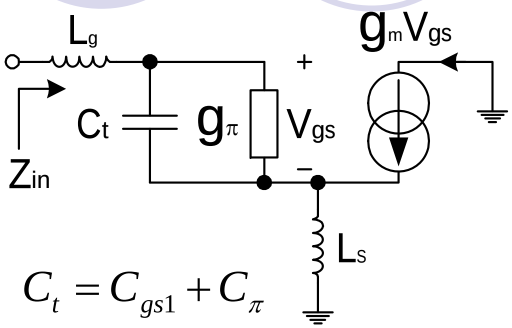
- Bipolar is more non-linear than MOS
- Degeneration used to match the 3rd order non-linear term of MOST

3rd Order Cancellation Effect

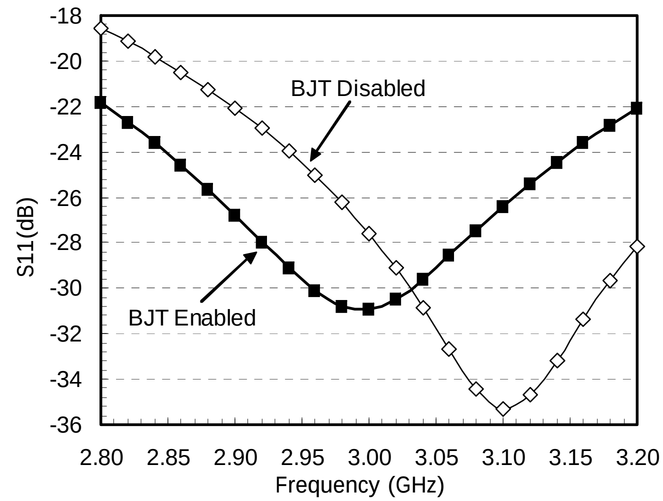


- MOS and BJT biased separately
- MOS in moderate inversion, BJT in active region

Effects on Input Impedance Matching and Noise



$$C_t = C_{gs1} + C_\pi$$



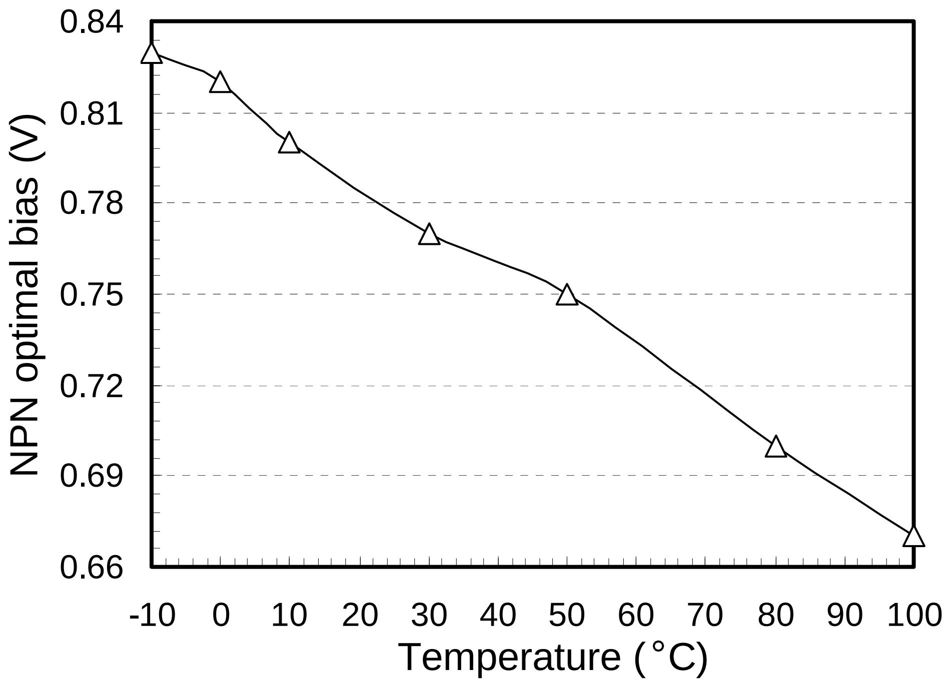
- BJT biased at low current: 320uA
- BJT noise contribution: 2.4%

Device	Noise ratio
Source Resistance	60%
MOS Transistor	14%
Bipolar Transistor	2.4%
Other	23.6%

C_π shifts the matching point to a lower frequency

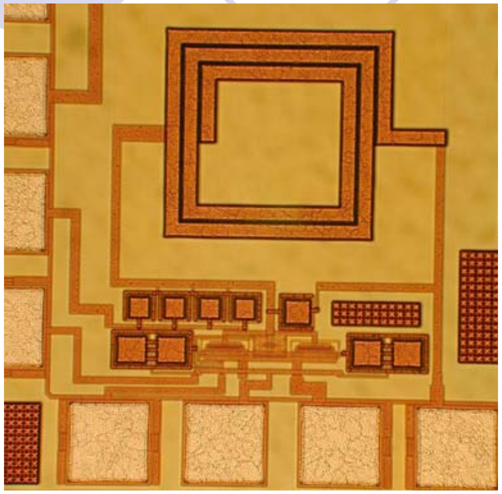
g_π moves the impedance away from the intended value

Biasing Temperature Profile

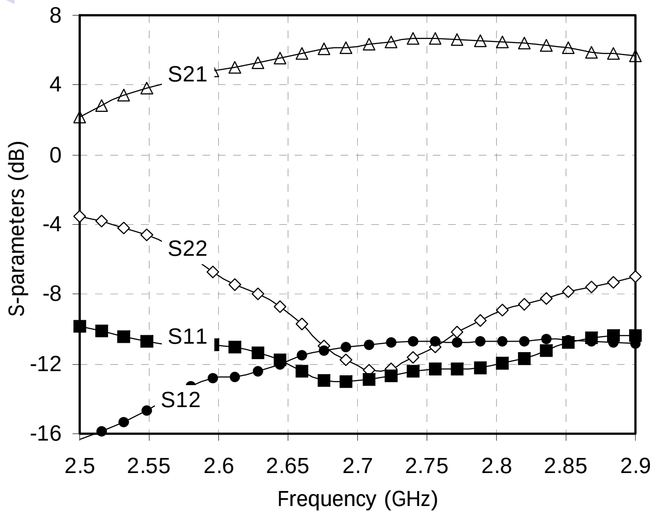


- MOS biased by constant-g_m
- BJT biased by a PTAT circuit

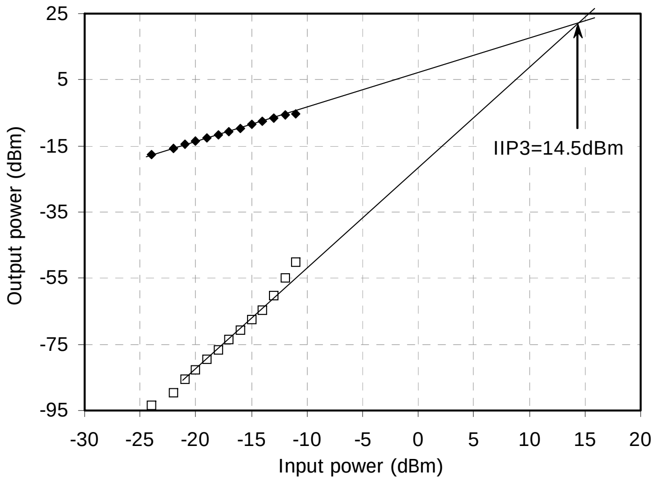
Experimental Results of the Proposed Linearized LNA



Active area: 390um x 290um



Frequency	2.7	GHz
Gain	6.4	dB
IIP3	14.5	dBm
NF	2.1	dB
Pd	8.9	mW



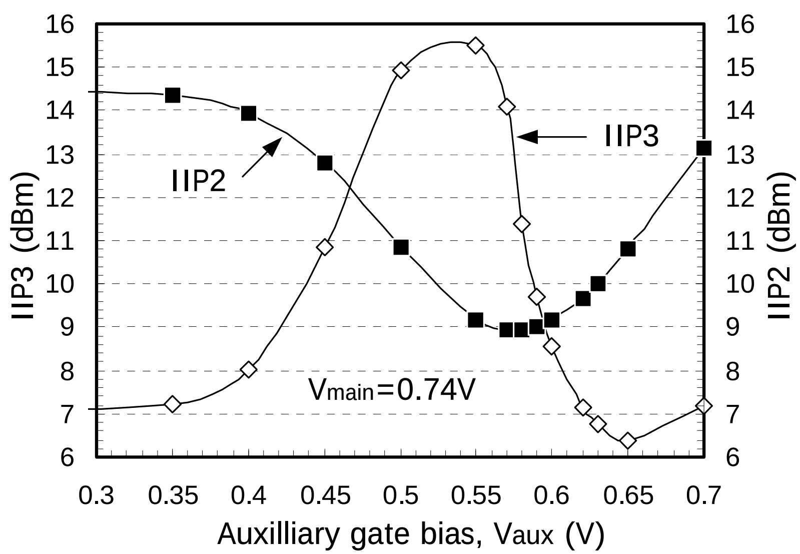
Extend to a Differential Version

- Single-ended suffers from small IIP2
- Out-of-band termination

$$IIP3(2\omega_2-\omega_1)=\frac{1}{6R_s\cdot|H(\omega)|\cdot|A_1(\omega)|^3\cdot|\varepsilon(\Delta\omega,2\omega)|}$$

$$\varepsilon(\Delta\omega,2\omega)=g_3-g_{oB}(\Delta\omega,2\omega)$$

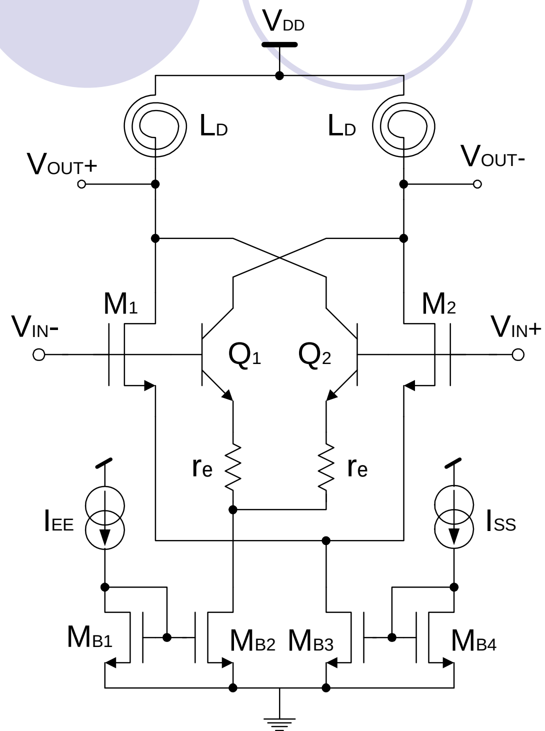
$$g_{oB}(\Delta\omega,2\omega)=\frac{2}{3}g_2^2\left[\frac{2}{g_m+g(\Delta\omega)}+\frac{1}{g_m+g(2\omega)}\right]$$

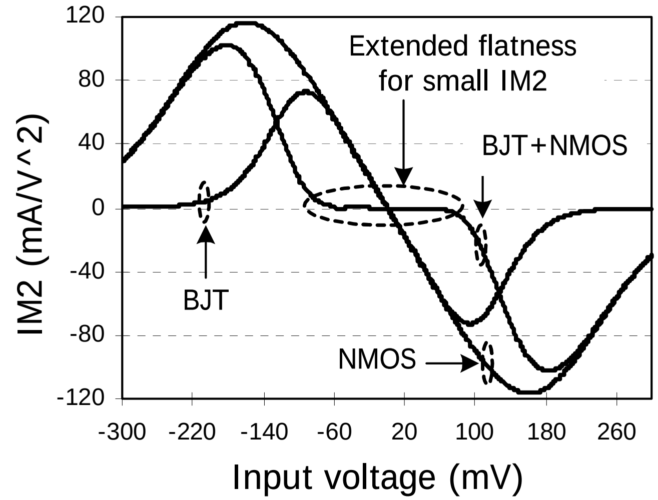


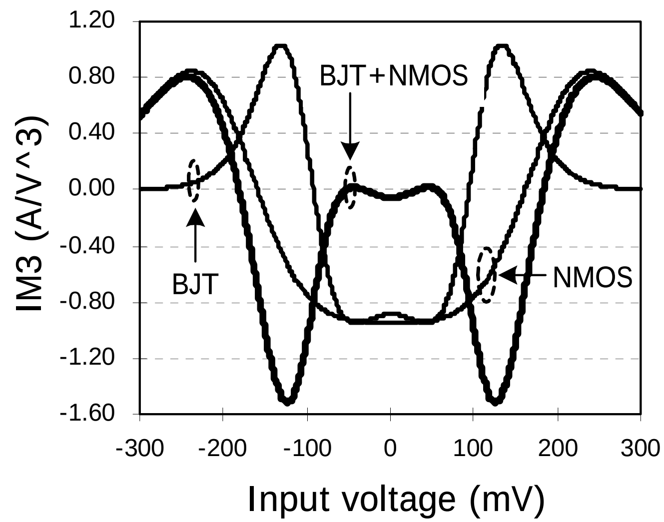
- The 3rd order term of MOS and BJT differential pair has the same sign.
- BJT is more non-linear than MOS
- Less current for BJT to present same non-linearity as MOS
- Cross-couple MOS and BJT differential pair will help

Extend to a Differential Version

(Cont'd)

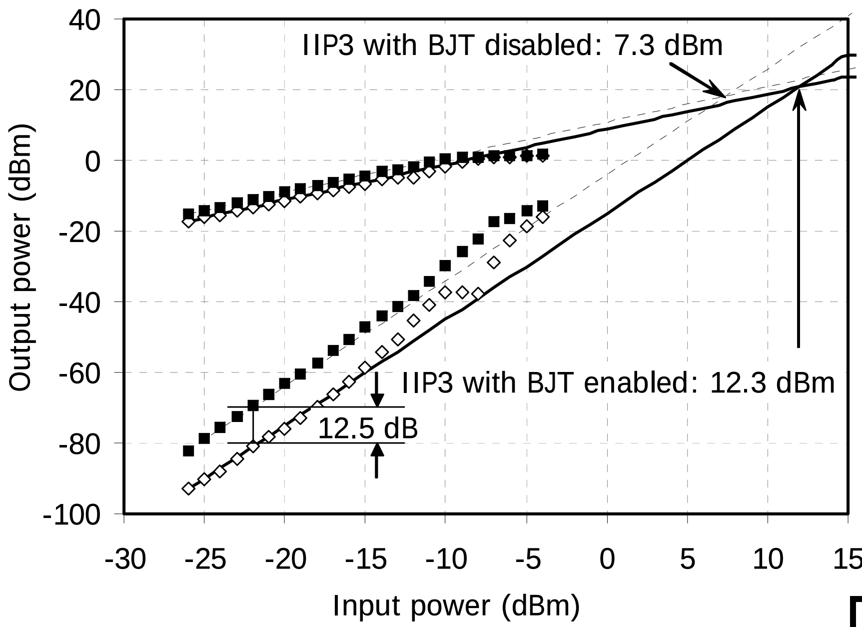


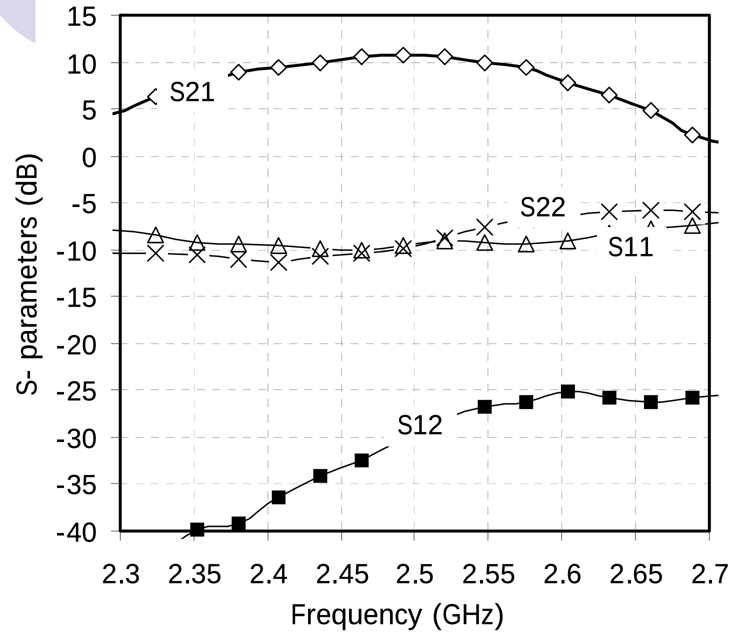


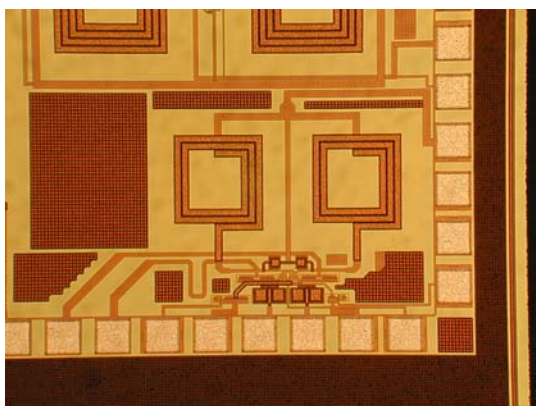


- BJT pair contributes 15% of noise
- Larger noise figure: 3.4 dB
- Larger current dissipation: 10mA
- Better reverse isolation: 25 dB
- No need out-of-band termination

Experimental Results of the Proposed Differential LNA

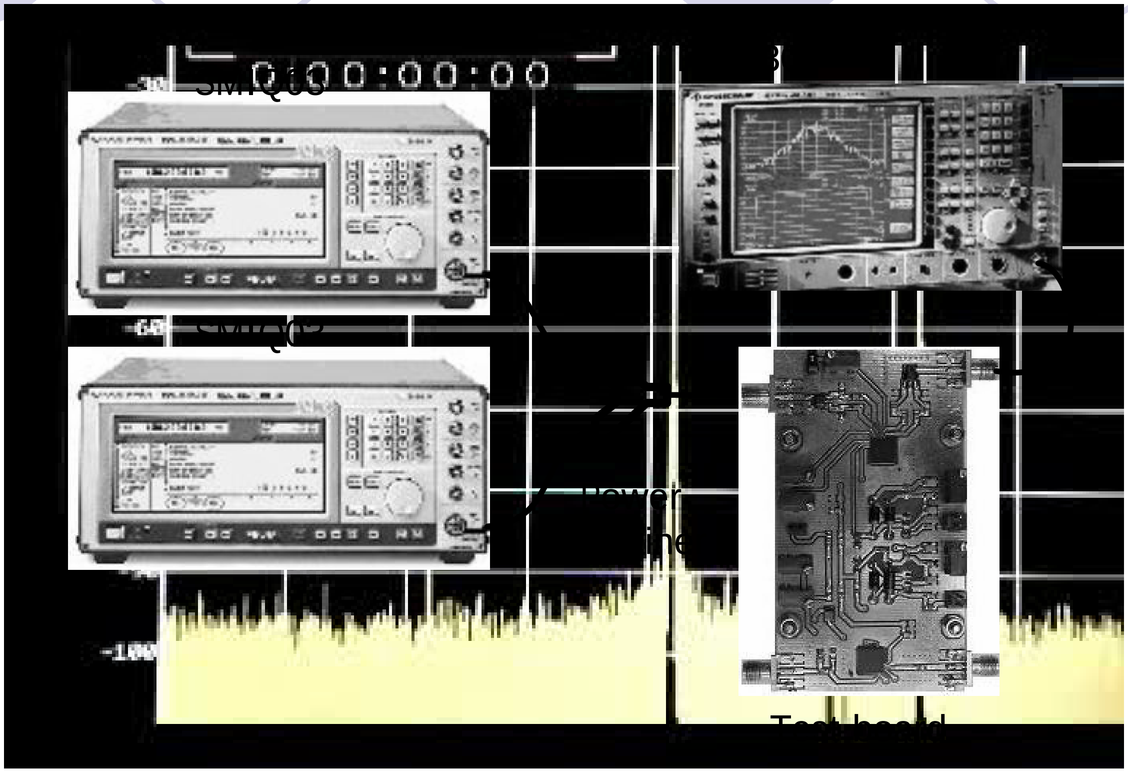






Frequency	2.5	GHz
Gain	10	dB
IIP3	12.5	dBm
NF	3.4	dB
Pd	19.8	mW

IM3 Cancellation Demo



Measurement setup

Measurement video clip shows the IM3 cancellation effect of BJT differential pair in the differential LNA.

Comparison Table

	Frequency	Gain	NF	IIP3	Pd	FOM
	GHz	dB	dB	dBm	mW	
Single-ended [1]	0.9	10	2.85	15.6	21.1	18.5
Single-ended [proposed]	2.7	6.4	2.1	14.5	8.9	22.8
Differential [2]	0.9	5	2.8	18	45	4.9
Differential [proposed]	2.5	10	3.4	12.3	19.8	7.2
BiCMOS [Simulated]	3	9.5	1.2	12	6.6	67

$$FOM = \frac{G \cdot IIP3}{(F-1)P_D}$$

C. Xin, E. Sanchez-Sinencio, "A Linearization Technique for RF Low Noise Amplifier", IEEE International Symposium on Circuits and Systems, May, 2004. ISCAS 2004 , pp. 313 -316, 2004

Conclusions

- Hybrid: Bipolar linearizes MOST
- Differential structure: no degradation on IIP2
- Better trade-offs between design parameters
- Good figure of merit