24.6 Noise Cancelling in Wideband CMOS LNAs

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Wideband low-noise amplifiers (LNAs) are used in receivers processing several signal channels such as TV cable modem, multiband mobile terminals and base stations. High sensitivity applications require sufficient gain, good isolation and a noise figure ($NF=10\cdot\log_{10}(F)$) below 3dB over a wide frequency range. Moreover, source impedance matching (i.e. $Z_{IN}=R_S$) is usually wanted to limit reflections on a cable or to avoid alterations of the characteristics of the RF filter preceding the LNA. LNAs exploiting an inductor achieve such requirements, but only in a narrow frequency band around resonance. Little is found in literature about wideband LNAs in CMOS [1,2].

Elementary wideband amplifiers such as the resistively terminated common-source (Figure 24.6.1 a), the common-gate (Figure 24.6.1 b), the common-source shunt-feedback (Figure 24.6.1 c) stages and the ones found in Reference 2 (Figure 24.6.1 d and e) show a severe trade-off between their F and the matching requirement $Z_{IN}=R_S$, failing to achieve sub-3dB NF. In these circuits, the impedance-matching device (indicated with subscript "i") is constrained by the matching requirement (i.e. $g_{m,i}=1/R_S$ or $R_i=R_S$) to generate as much equivalent input noise as that of a resistance equal to NEF·RS ($NEF=\gamma g_{d0}/g_m>1$ for a MOST and NEF=1 for a resistor). This renders $F>1+NEF\geq 2$, so NF>3dB. Global negative feedback can break the trade-off between source match and F. In principle, transformer feedback offers the best performance [3]. In practice, lossy feedback is often used due to relatively poor performance of wideband transformers in CMOS. The LNA, in Figure 24.6.1 f for instance, provides a matched input impedance $Z_{IN}=R_i/(1-A_{CL})=R_S$ ($A_{CL}=V_{OUT}/V_{IN}$) while contributing an equivalent input noise current (A_{CL}-1) times smaller than that of R_S. Sub-3dB NF is then possible provided the contribution of the MOST is small. However, sufficient gain at GHz frequencies is often not available OR multiple cascaded gain stages produce instability problems (e.g.: 3 poles in Figure 24.6.1 f). Next, frequency compensation networks can complicate design trade-offs, limiting the wideband performance of the amplifier [1].

Here, a noise-cancelling technique allows wideband sub-3dB NF while simultaneously achieving source match without needing global feedback. Thus, instability risks are relaxed. In Figure 24.6.2, the noise-cancelling principle is applied to the common source shunt-feedback stage M1-R with input impedance $Z_{IN}=1/g_{m1}$. The noise current output from the impedance matching device M1, $I_{n,1}/(1+g_{m1}R_S)$, flows through resistors R and R_S (Figure 24.6.2, dashed line), leading to noise voltages $V_{X,n,1}=R_SI_{n,1}/(1+g_{m1}R_S)$ and $V_{Y,n,1}=(R+R_S)I_{n,1}/(1+g_{m1}R_S)$ at nodes X and Y that are fully correlated. $V_{X,n,1}$ and $V_{Y,n,1}$ instantaneously have equal sign while signals at the same nodes have opposite sign (i.e. negative gain). This "difference in sign" makes it possible to cancel the output noise contribution of M1 while adding the signals. This is done, introducing a feed-forward path from node X via amplifier "A" and from node Y via an adder rendering $V_{OUT}=A\cdot V_X+V_Y$. Signal components $A\cdot V_{X,Sig}$ and $V_{Y,Sig}=(1-g_{m1}R)V_{X,Sig}$ add in-phase at the output for $g_{m1}R>1$. In contrast, the output noise due to M1, $V_{OUT,n,1}=A\cdot V_{X,n,1}+V_{Y,n,1}$, is cancelled at the negative gain $A=-V_{Y,n,1}/V_{X,n,1}=-(1+R/R_S)$ regardless g_{m1} (i.e. Z_{IN}). Cancelling is independent on any impedance from node Y to ground (i.e. g_{d1}) as it loads the two paths equally. As M1 contributes no output noise, it does not affect NF, so the trade-off with $Z_{IN}=R_S$ brakes. Now, the unmatched amplifier "A" mainly determines the LNA F. In earlier work, a limited form of noise cancellation (Figure 24.6.1 d) is observed Reference [2]. However, here this property is fully exploited by overcoming the limitation in Reference [2] and making it possible to achieve a wideband sub-3dB NF.

The transistor implementation of Figure 24.6.2 is shown in Figure 24.6.3. A CMOS inverter M1a-M1b replaces M1, providing larger g_m/I_D. To do so, capacitor C1 grounds the source of pMOST M1b. A two-inputs single-output circuit M2-M3 (dashed line) implements both the amplifier "A" and the "adder". The common source M2-M3 gain is $A=-g_{m2}/g_{m3}$, while M3 acts also as a voltage follower from node Y to the output. The gate of M3 is ac-coupled to node Y via the 2MHz high-pass filter C2-R2. Transistor M2a is cascoded to increase reverse isolation and reduce the Miller effect. The matching device M1a-M1b output noise is then zero for $|A|=g_{m2}/g_{m3}=1+R/R_S$. The noise factor is $F=1+1/(2A_{VF})+NEF(2-3A_{VF}+A_{VF}^2)/(g_{m2}R_SA_{VF}^2)$. F decreases increasing $g_{m2}\cdot R_S$ until it is limited by R to $F_{lim}=1+1/(2A_{VF})$. For $A_{VF}=-5$, NEF=1.4 and $g_{m2}\cdot R_S=4.7$ yield NF=2.04dB and NF_{lim}=0.41dB.

In practice, noise cancelling is affected by: (a) device parameter variations due to process-spread and (b) parasitic capacitances:

(a) The cancelling is robust to process-spread. The reason is twofold. First, the cancelling condition $g_{m2}/g_{m3}=1+R/R_S$ in independent on gm1 and the gd of any MOST in the signal path as well. Second, for a cancelling error $\varepsilon=|-g_{m2}/g_{m3}+1+R/R_S|<1$, the output noise power of M1 increases as $\varepsilon^2<<1$. Montecarlo simulations predict a $4\cdot\sigma_{NF}<0.18$dB at 1GHz.

(b) Due to the input capacitance CA (Figure 24.6.2), the noise current of M1 flows into a complex impedance $R_S//(1/(j\omega CA))$. The latter renders different frequency-dependent phase and amplitude variations for the noise voltage $V_{X,n,1}$ and $V_{Y,n,1}$, leading to a lower degree of cancelling as frequency increases. The noise factor can be approximated by $F(f)\approx 1+(F-1)(1+(f/f_0)^2/4)+(NEF/4)\cdot(f/f_0)^2$ with $f_0=1/(2\pi C_AR_S)$. In this design with $g_{m2}\cdot R_S=4.7$, $A_{VF}=-5$, $C_A=0.9$pF and $R_S=50\Omega$, the predicted increase of NF is only 0.11dB at 1GHz and 0.42dB at 2GHz.

The LNA in Figure 24.6.3 uses 0.25μm CMOS for 2dB wideband NF, while driving the output bond pad capacitance C_{LOAD}=0.2pF load. Noise coupled via the input/output bond pad, the MOS gate resistance, substrate resistance and the capacitance of the cascode node is minimized by layout. Figure 24.6.4 shows measured S-parameters (on-wafer). The gain $A_{VF}=V_{OUT}/V_S$ is 13.7dB with a –3dB bandwidth between 2MHz and 1.6GHz and S_{12} is <-36dB in 0.01-1.8GHz. NF and IIPs are measured with the die glued to a ceramic substrate and connected to 50Ω lines. Figure 24.6.5 shows measured $NF_{50\Omega}$, ≤ 2dB over in 0.25-1.1GHz and ≤ 2.4dB over in 0.15-2GHz matching simulations and hand calculations. Figure 24.6.6 summarizes measurements.

Acknowledgments:
Philips Research Laboratories for IC fabrication and M. Dijkstra for layout assistance.

References:
[1] J. Janssens et al., "A 2.7Volt CMOS Broad Band Low Noise Amplifier," Symp. on VLSI Circuits, pp. 87-88, Kyoto, June 1997.
[2] F. Bruccoleri, et.,"Generating All Two-MOS-Transistor Amplifiers Leads to New Wide-Band LNAs," IEEE JSSC, vol.36, No.7, pp.1032-1040, July 2001.
[3] Norton, D.E.,"High Dynamic Range Feedback Amplifiers Using Lossless Feedback," Proc. IEEE Symp. on Circuits and Systems, pp. 438-440, 1975.

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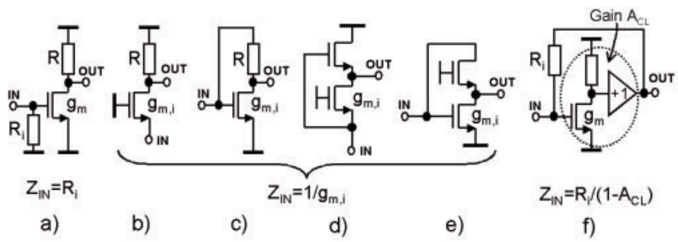


Figure 24.6.1: Common wideband CMOS LNAs (biasing not shown).

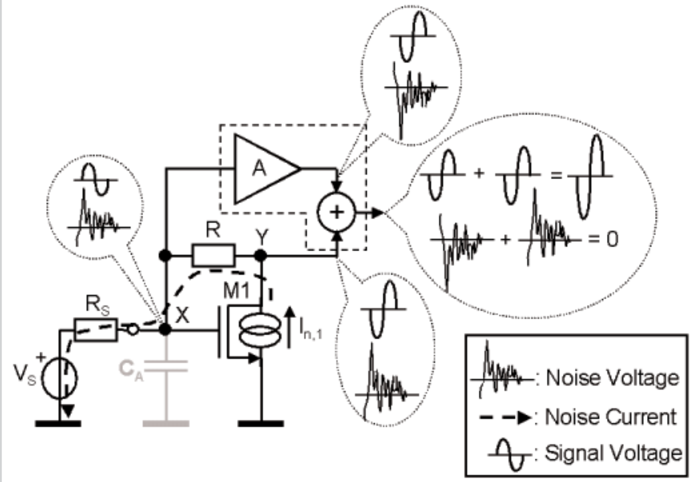


Figure 24.6.2: Noise cancelling applied to the amplifier of Figure 24.6.1c (biasing not shown).

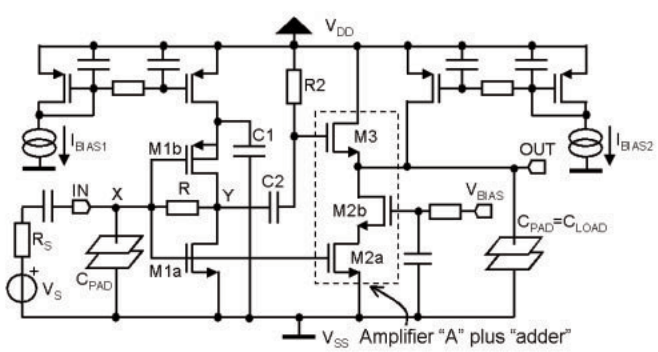


Figure 24.6.3: Schematic of the noise-cancelling wideband LNA.

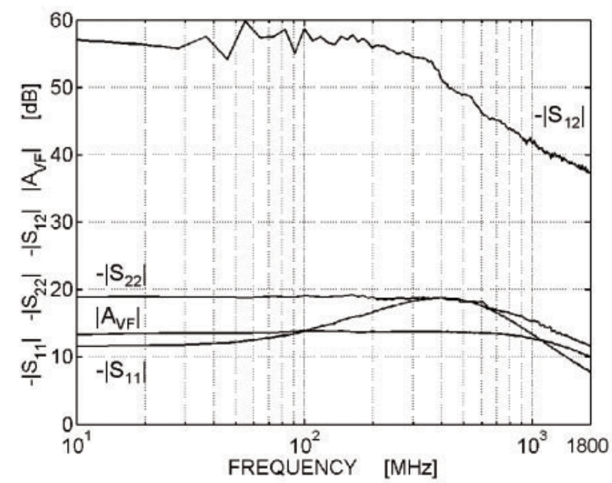


Figure 24.6.4: Measured S_{11}, S_{22}, S_{12} and $A_{VF} = V_{OUT}/V_S$ ($C_{LOAD} = C_{PAD} = 0.2pF$).

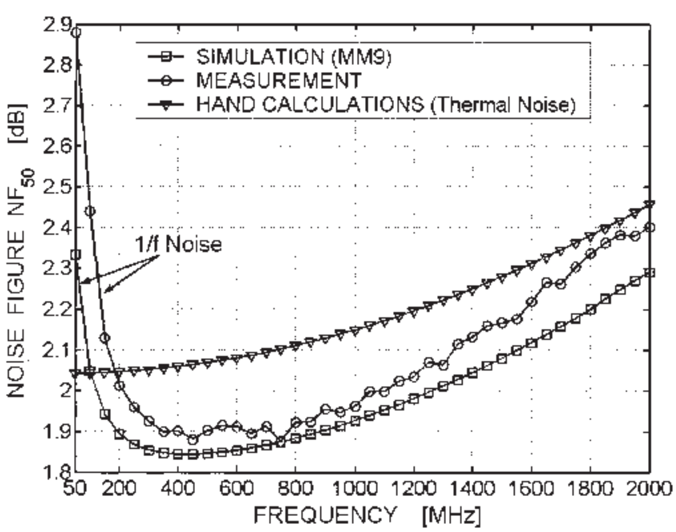
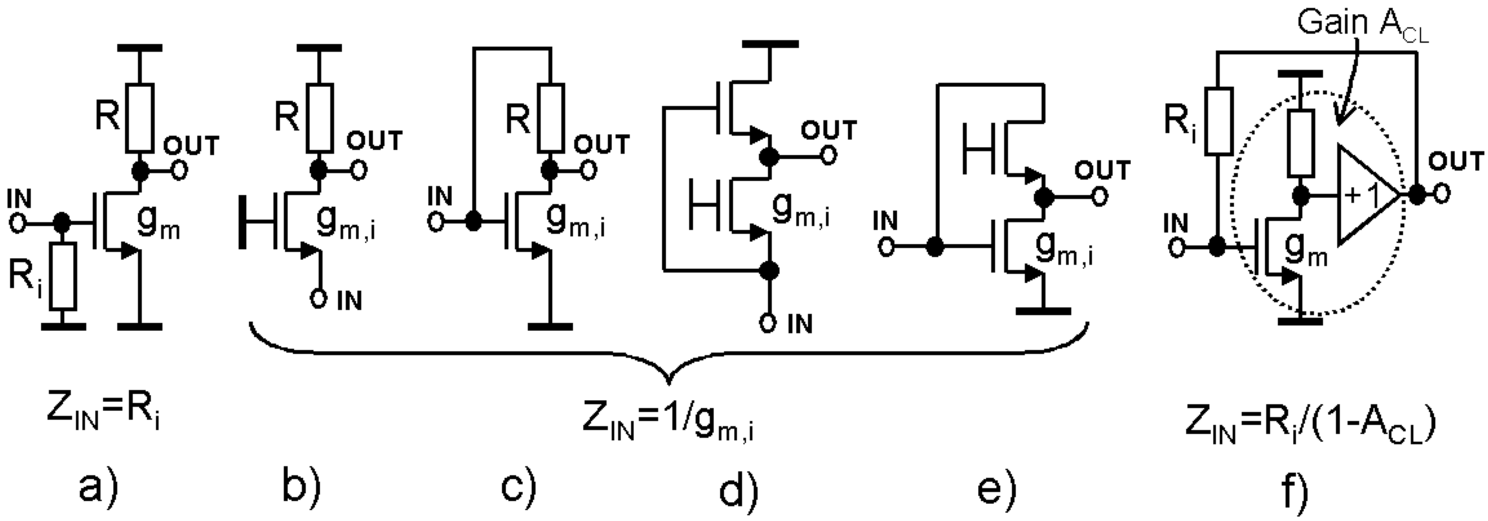


Figure 24.6.5: Measured/simulated/hand-calculated 50Ω NF.

$A_{VF}=\lvert V_{OUT}/V_S \rvert$	13.7dB
-3dB Bandwidth	2-1600MHz
$\lvert S_{12}\rvert$	<-36dB in 10-1800MHz
$\lvert S_{11}\rvert$	<-8dB in 10-1800MHz
$\lvert S_{22}\rvert$	<-12dB in 10-1800MHz
IIP3 (Input Ref.)	0dBm (f_1=900MHz & f_2=905MHz)
IIP2 (Input Ref.)	12dBm (f_1=300MHz & f_2=200MHz)
ICP1dB (Input Ref.)	-9dBm (f_1=900MHz)
$NF_{50\Omega}$	<=2dB in 250-1100MHz <+2.4dB in 150-2000MHz
I_{DD} at V_{DD}	14mA at 2.5V
Area	0.3x0.25mm^2
Technology	0.25µm CMOS

Figure 24.6.6: Summary of measurements.



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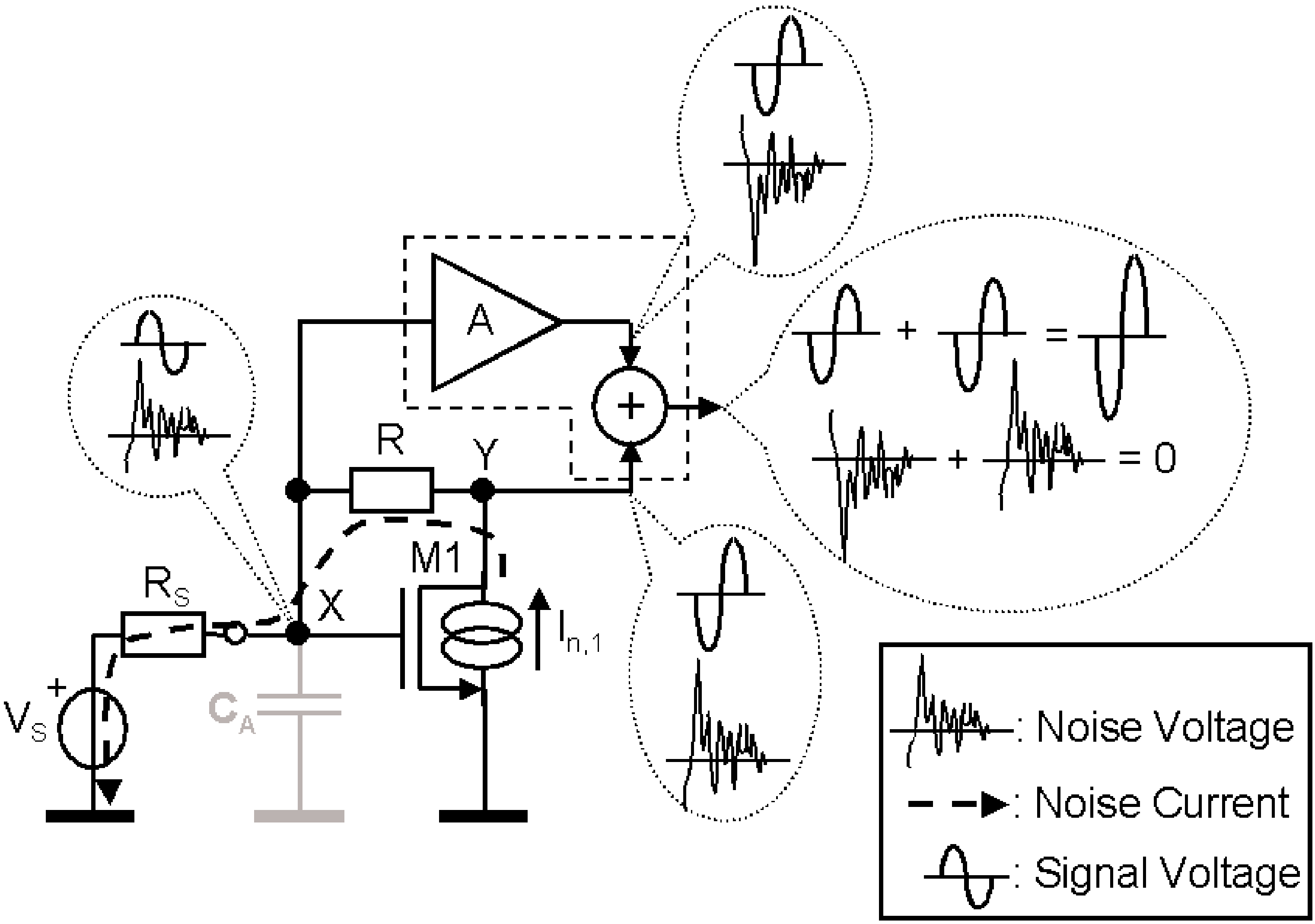

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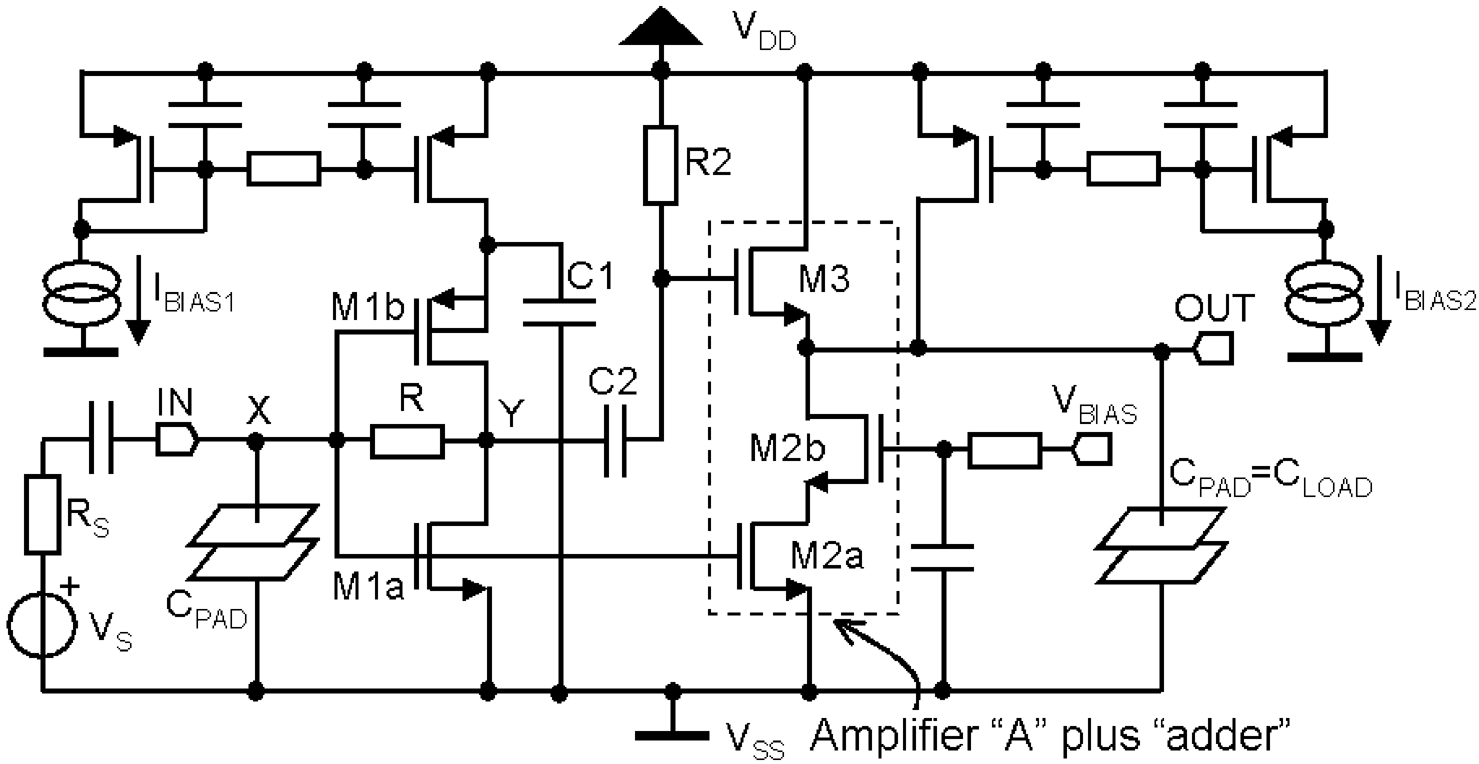

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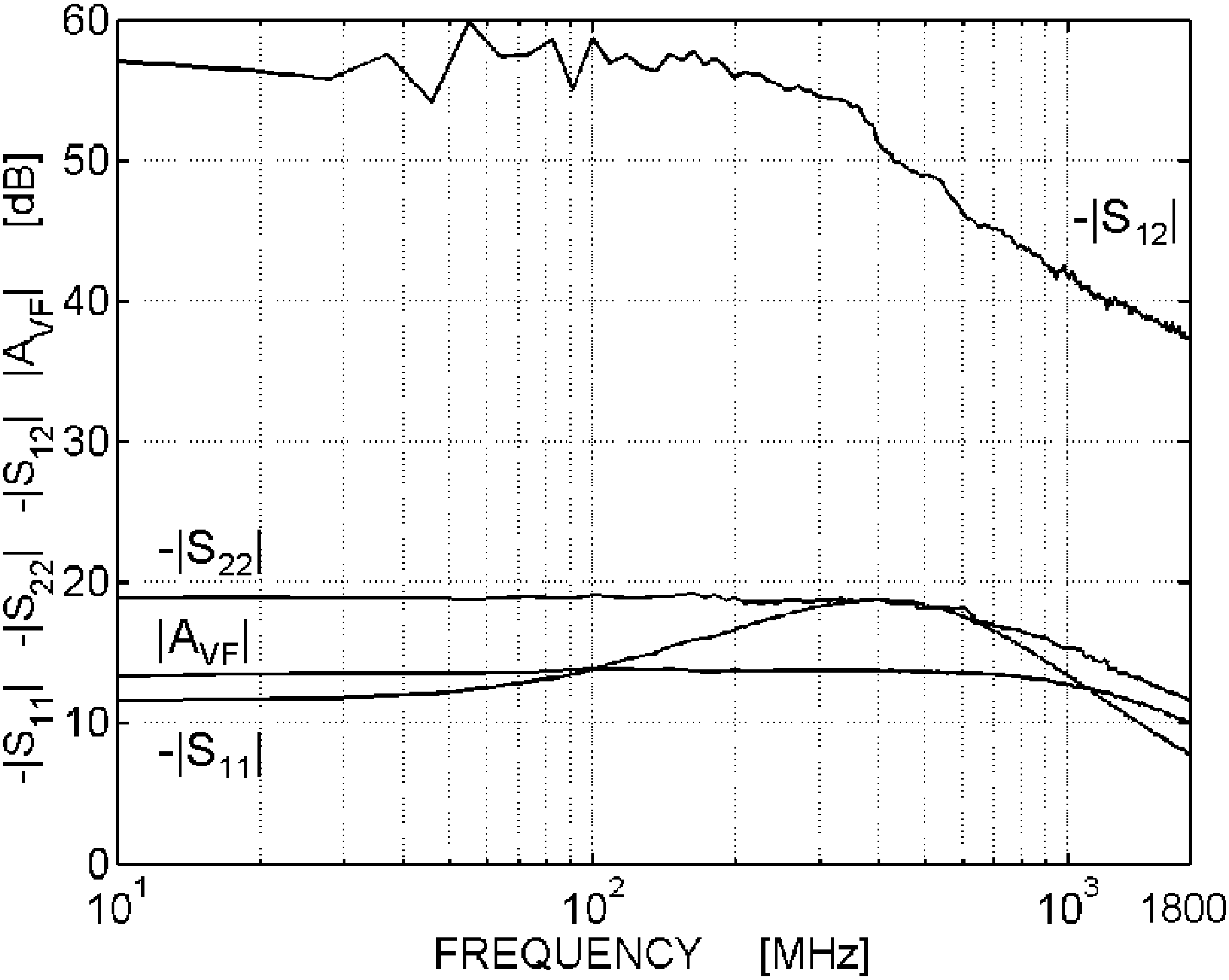

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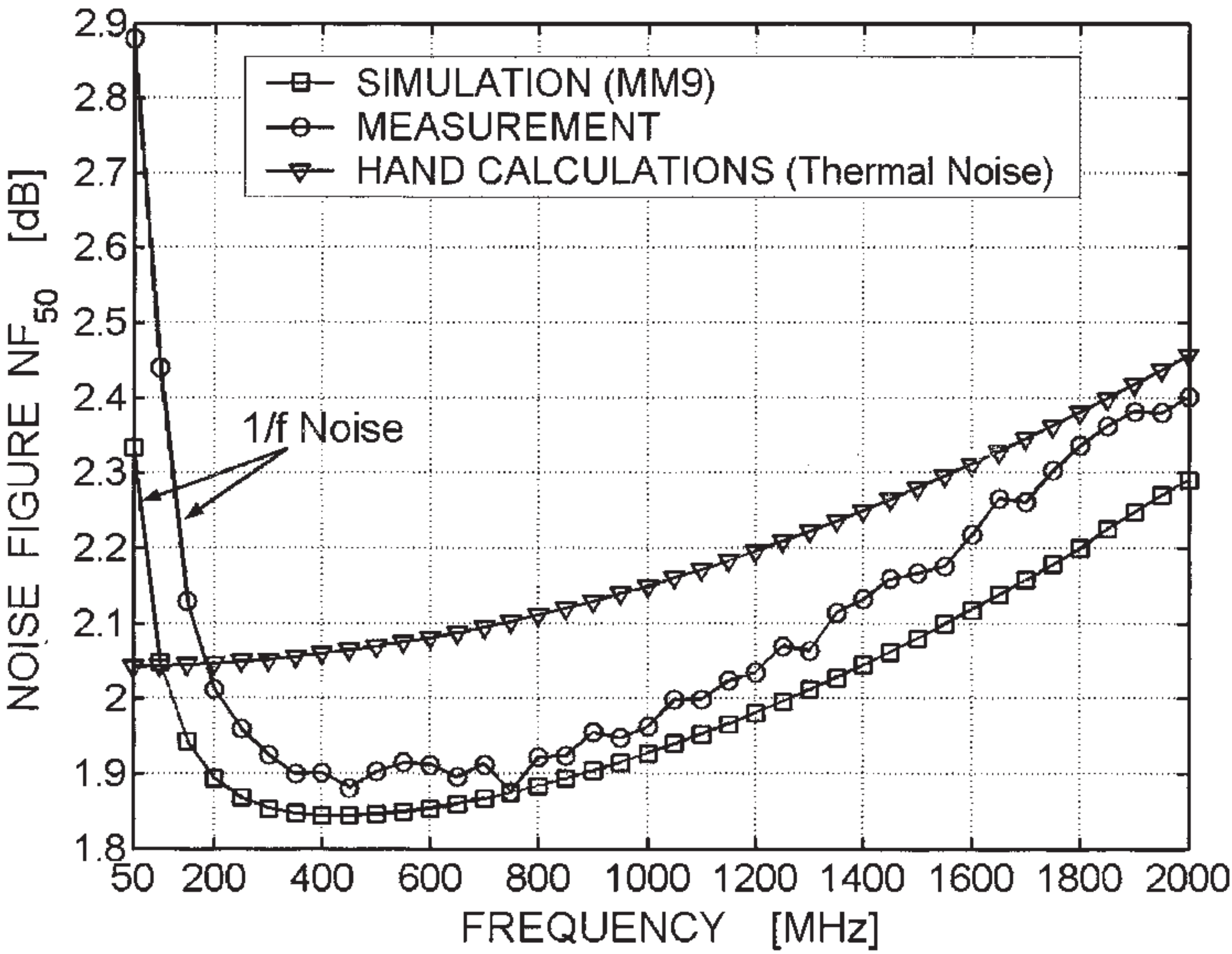

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