Thermal Noise Canceling in LNAs: A Review

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Abstract — **Most wide-band amplifiers suffer from a fundamental trade-off between noise figure NF and source impedance matching, which limits NF to values typically above 3dB. Recently, a feed-forward noise canceling technique has been proposed to break this trade-off. This paper reviews the principle of the technique and its key properties. Although the technique has been applied to wideband CMOS LNAs, it can just as well be implemented exploiting transconductance elements realized with other types of transistors.**

I. INTRODUCTION

Wide-band Low-Noise Amplifiers (LNAs) are used in receiving systems where the ratio between the bandwidth and its center frequency can be as large as two, for instance in analog cable TV (50-850 MHz), and satellite (950-2150 MHz) or terrestrial digital (450-850 MHz) video broadcasting. Moreover, a wide-band low-noise amplifier can replace several LC-tuned LNAs in multi-band or multi-mode narrow-band receivers. A wide-band solution saves chip-area and fits better to the trend towards flexible radios with as much signal processing (e.g.: channel selection, image rejection, etc.) as possible in the digital domain (towards "software radio").

High-sensitivity integrated receivers require LNAs with sufficiently large gain, noise figure NF well below 3dB, adequate linearity and source impedance matching $Z_{IN}=R_S$. The latter is to avoid signal reflections on a cable or alterations of the characteristics of the RF filter preceding the LNA, such as pass-band ripple and stop-band attenuation. Fig.1a-d shows well-known wide-band amplifiers capable of matching a real source impedance R_S. These amplifiers suffer from a fundamental trade-off between their noise factor F ($NF=10\log_{10}(F)$) and impedance matching, $Z_{IN}=R_S$. Assuming large gain, low F requires a large g_{mi} or R_i. However, impedance matching demands a fixed $g_{mi}=1/R_S$ or $R_i=R_S$. Modeling transistors as a transconductance g_m, and assuming current noise spectral density $4kT\cdot NEF\cdot g_m$, analysis renders [4]:

$$F_{LNA} \geq 1 + NEF \tag{1}$$

NEF for a long channel MOSFET is theoretically 2/3, but for a practical deep-submicron MOSFET between 1 and 2, whereas resistive degeneration of a transconductor results in NEF=1. Thus, practical MOSFET LNAs are limited to F>2 or NF>3dB, even for high gain.

To be best of our knowledge, only amplifiers exploiting global negative feedback (shunt-feedback) can break this trade-off between NF and impedance matching, but they are prone to instability [1]. In contrast, we propose a feed-forward thermal-noise canceling technique enabling low NF and source impedance matching, without instability problems [2,3,4]. In earlier work [5,6], LNA circuits with partial noise cancellation have been found, via systematic circuit topology generation [7]. However, those circuits still have constraints on NF upon $Z_{IN}=R_S$. In contrast, the technique presented in this paper can, at least in principle, achieve arbitrarily low NF, at the cost of power consumption. This paper reviews the basis of the technique and discusses its key properties. Furthermore several examples of circuits exploiting full or partial noise cancellation will be given.

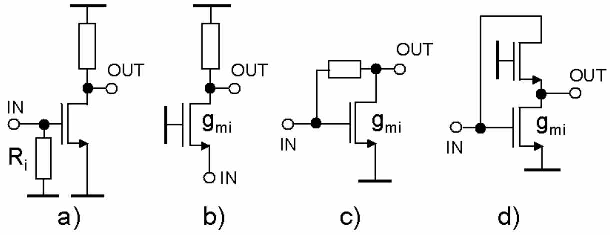

Fig. 1. Well-known wide-band LNAs capable of impedance matching for $R_i=R_s$ or $g_{mi}=1/R_s$ (biasing not shown).

The paper is organized as follows. Section II reviews the principle of the noise canceling technique, while section III discusses its key properties and limitations. Section IV shows some practical examples of amplifiers exploiting noise cancellation. Finally, section V draws conclusions.

II. NOISE CANCELING PRINCIPLE

To understand the principle of noise canceling, consider the amplifier stage of fig. 1c redrawn in fig. 2. Its input impedance is $Z_{IN}=1/g_{mi}$ and the voltage gain is $A_{VF,MS}=V_Y/V_X=1-g_{mi}R$ where the index "MS" refers to the matching stage in fig.1c. For $Z_{IN}=R_S$ its F is larger than 1+NEF, as discussed in the previous section. Let's now analyze the signal and the noise voltages at the input node X and output node Y, both with respect to ground, due to the noise current $I_{n,i}$ of the impedance matching MOSFET. Depending on the relation between $Z_{IN}=1/g_{mi}$ and R_S, a noise current $\alpha(R_S,g_{mi})\cdot I_{n,i}$, flows out of the matching MOSFET through R and R_S (fig. 2a), with $0<\alpha<1$. This current causes two instantaneous noise voltages at nodes X and Y, which have equal sign. On the other hand, the signal voltages at nodes X and Y have opposite sign (fig. 2b), because the gain $A_{VF,MS}$ is negative, assuming $g_{mi}R>1$. This *difference in sign for noise and signal* makes it possible to *cancel the noise* of the matching device, while simultaneously *adding the signal* contributions constructively. This is done by creating a new output, where the voltage at node Y is added to a scaled negative replica of the voltage at node X. A proper value for this scaling factor renders noise canceling at the output node, for the thermal noise originating from the matching device.

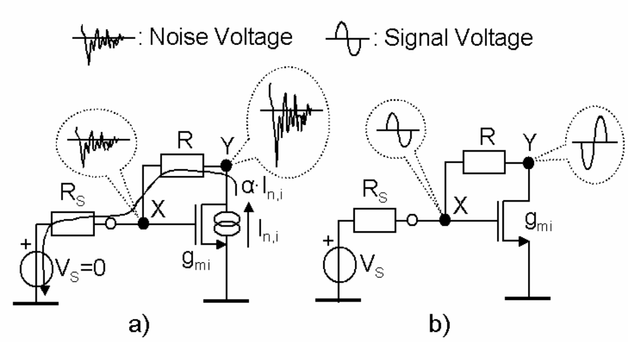

Fig. 2 Matching MOSFET noise (a) and signal (b) voltage at nodes X and Y for the amplifier in fig. 1c (biasing not shown).

Fig. 3a shows a straightforward implementation using an ideal feed-forward voltage amplifier "A" with a gain $-A_v$ (with $A_v>0$). By circuit inspection, the matching device noise voltages at node X and Y are:

$$V_{X,n,i}=\alpha(R_S,g_{mi})\cdot I_{n,i}R_S=\alpha\cdot I_{n,i}R_S \qquad (2)$$
$$V_{Y,n,i}=\alpha\cdot I_{n,i}(R_S+R)$$

The output noise voltage due to the noise of the matching device, $V_{OUT,n,i}$ is then equal to:

$$V_{OUT,n,i}=V_{Y,n,i}-V_{X,n,i}\cdot A_v=\alpha\cdot I_{n,i}(R+R_S-A_vR_S) \qquad (3)$$

Output noise cancellation, $V_{OUT,n,i}=0$, is achieved for a gain A_v equal to:

$$A_{v,c}=\frac{V_{Y,n,i}}{V_{X,n,i}}=1+\frac{R}{R_S} \qquad (4)$$

where the index "c" denotes the cancellation. On the other hand, signal components along the two paths add constructively, leading to an overall gain (assuming $Z_{IN}=1/g_{mi}=R_S$ and $A_v=A_{v,c}$):

$$A_{VF,c}=\frac{V_{OUT}}{V_X}=1-g_{mi}R-Av,c=-g_{mi}R-\frac{R}{R_S}=-2\frac{R}{R_S} \qquad (5)$$

From equation (4), two characteristics of noise canceling are evident:

1. Noise canceling depends on the absolute value of the *real* impedance of the source, R_S (e.g.: the impedance seen "looking into" a terminated coax cable).
2. The cancellation is independent on $\alpha(R_S,g_{mi})$ and on the quality of the source impedance match. This is because any change of g_{mi} equally affects the noise voltages $V_{X,n,i}$ and $V_{Y,n,i}$.

Fig. 3b shows a simple implementation of the noise-canceling LNA in fig. 3a. Amplifier "A" and the adder are replaced with the common-source stage M2-M3, rendering an output voltage equal to the voltage at node X times the gain $A_v=g_{m2}/g_{m3}$. Transistor M3 also acts as a source follower, copying the voltage at node Y to the output. The superposition principle renders the addition of voltages with an overall gain $A_{VF}=1-g_{mi}R_S-g_{m2}/g_{m3}$.

Note that *any* small signal that can be modeled by a current source between the drain and source of the matching device is cancelled too (e.g.: 1/f noise, thermal noise of the distributed gate resistance and the bias noise-current injected into node Y). However the noise of R is not cancelled. This can be seen splitting its noise current $I_{n,Ri}$ in two correlated sources to ground, at the output node Y and the input node X. The former is cancelled for $A_v=A_{v,c}$, the latter is not.

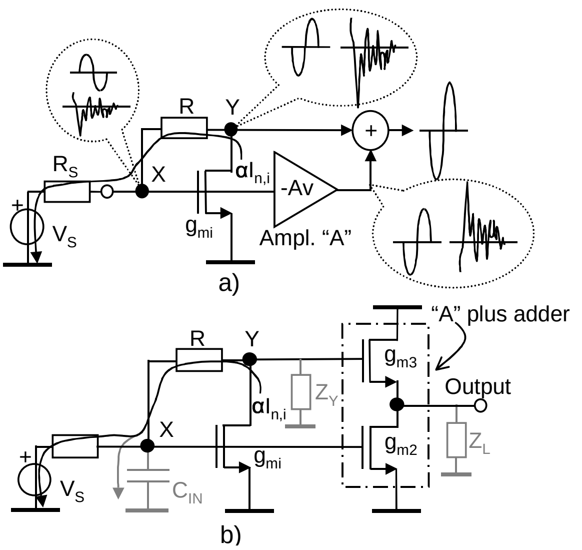

Fig. 3. (a) Wide-band LNA exploiting noise-canceling, (b) Elementary implementation of amplifier "A" plus adder (biasing not shown).

III. PROPERTIES AND LIMITATIONS

We will now discuss some key properties of noise cancellation, starting with the achievable noise factor.

A. Noise factor

The noise factor F of the LNA in fig. 3a can be written as:

$$F=1+EF_{MD}+EF_R+EF_A \qquad (6)$$

where the "excess noise factor" EF is used to quantify the contribution of different devices to F, and index MD refers to the matching device, R to the resistor R, and A to amplifier "A". For the implementation in fig. 3b, expressions for EF for $Z_{IN}=R_S$, assuming equal NEF, and cancellation for $A_v=A_{v,c}$ become:

$$EF_{MD,c}=0 \qquad (7)$$
$$EF_{R,c}=-\frac{2}{A_{VF,c}}=\frac{R_S}{R}$$
$$EF_A=\frac{NEF}{g_{m2}}\left(\frac{1}{R_S}+\frac{3}{R}+\frac{2R_S}{R^2}\right)$$

The noise factor at cancellation, F_c, is thus only determined by $EF_{A,c}$ and $EF_{R,c}$, which are both not constrained by the matching requirement. $EF_{A,c}$ can be made arbitrarily smaller than 1 by increasing g_{m2} of its input stage, at the price of power dissipation. The minimum achievable F_c is now determined by $EF_{R,c}$. The latter can also be significantly smaller than 1 when the gain $|A_{VF,c}|$ is large, which is desired anyhow for an LNA. In practical design, F_c can be lowered below 2 (i.e. 3dB) by increasing $g_{m2}R_S$ until it saturates to $F_{c,min}=1+EF_{R,c}=1+R_S/R$.

B. Robustness for component spread

The noise canceling technique is relatively robust to device parameter variations. The cancellation depends only on a *reduced* set of device parameters. For instance, the impedance from node Y to ground Z_Y (e.g.: g_d of the matching device), the load Z_L (e.g.: g_d and g_{mb} of M3) and g_{mi} of the matching device in fig. 3b do *not* affect the cancellation because they "load" the two feed-forward paths in the same fashion. On the other hand, any deviation of the source resistance R_S and the gain A_v from their nominal values $R_{S,NOM}$ and $A_{v,c}$ affect the cancellation, as shown by equation (3). For a typical practical case with NEF=1.5 and $A_{v,c}=7$, $\delta R_S/R_{S,NOM}$ and $\delta A_v/A_{v,c}$ as large as ±20% are needed in order to rise EF_{MD} to only 0.1 [4], one tenth of the contribution of the input source. Thus, the sensitivity to variations of R_S and the gain A_v is low.

C. Advantages compared to negative feedback

As shown in the previous section, the noise canceling technique is capable of NF well below 3dB upon $Z_{IN}=R_S$. Similar noise performance could also be achieved exploiting shunt-feedback. However, noise cancellation offers several advantages:

- It is a feed-forward technique *free* of global feedback, so instability risks are greatly relaxed.
- To first order, Z_{IN} depends only on g_{mi}. Thus, Z_{IN} is less sensitive to process spread.
- Implementing variable-gain at $Z_{IN}=R_S$ is more straightforward due to the orthogonality between the gain A_{VF} and Z_{IN} (changing the value of R and Av changes the gain, but not Z_{IN}).

Furthermore, it can be shown [3] that simultaneous noise and power matching is achieved.

D. Frequency dependence of noise cancellation

Parasitic capacitors not only limit the signal bandwidth but also degrade noise cancellation at high frequencies. The simplified case of fig. 3b with $C_Y=C_L=0$ appears to be adequate to model the main trend. Here, C_{IN} accounts for the parasitic capacitance contributed to the input node mainly by the matching device and amplifier "A". This simple model is realistic because: (a) C_Y and the load C_L in fig. 3b do not affect the cancellation and (b) C_L does not affect the F of the LNA standalone. The noise current $\alpha\cdot I_{n,i}$ flowing out from the matching device "sees" a complex source impedance $Z_S(f)= R_S/(1+j2\pi fC_{IN})$ as shown in fig. 3b. In this case, the output noise due to the matching device, $V_{OUT,n,i}(f)$, is obtained replacing R_S with $Z_S(f)$, resulting in a frequency dependent noise factor, $F_c(f)$, which can be written as:

$$F_c(f) = F_c + \left(F_c - 1 + NEF\right)\cdot\left(f / f_0\right)^2 \quad (8)$$

where F_c is the low-frequency noise factor as given in (6) and $f_0=1/(\pi\cdot R_S\cdot C_{IN})$ is the input pole. For F_c smaller than 1+NEF, $F_c(f)$-F_c increases with f/f_0 mainly because the cancellation degrades. However, this effect and the increase of $F_c(f)$ with the frequency can be modest up to relatively high frequencies because of the low input-node resistance $R_S/2$. Equation (8) shows the importance of maximizing f_0 (i.e. minimizing C_{IN}) in order to mitigate the degradation of noise factor. This can be done by increasing V_{GS}-V_{T0} of Mi and M2, cascoding to reduce the Miller effect, by frequency compensation, e.g. so-called shunt-peaking technique or using amore advanced deep sub-micron CMOS process with higher f_T.

IV. PRACTICAL CIRCUIT EXAMPLES

A. Wide-band Noise Cancellation LNA

A wide-band LNA according to the concept of fig. 3b was designed in a 0.25μm standard CMOS process. Table 1 summarizes the achieved performance and fig.4 shows the simulated and the calculated noise figure using formula (8). The measured NF is below 2.4dB over more than one decade (150-2000 MHz) and below 2dB over more than 2 octaves (250-1100 MHz). At low frequency, NF rises due to a AC-coupling high-pass filter.

$A_{VF,TOT}=\lvert V_{OUT}/V_S\rvert$	13.7dB
-3dB Bandwidth	2 MHz -1600 MHz
$\lvert S_{12}\rvert$	<-36dB in 10-1800 MHz
$\lvert S_{11}\rvert$	<-8dB in 10-1800 MHz
$\lvert S_{22}\rvert$	<-12dB in 10-1800 MHz
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 250-1100 MHz <= 2.4dB in 150-2000 MHz
I_{DD}@V_{DD}	14mA@2.5Volt
Area	0.3x0.25 mm^2
Technology	0.25μm CMOS

Table. 1. Performance summary of the CMOS LNA [2,4].

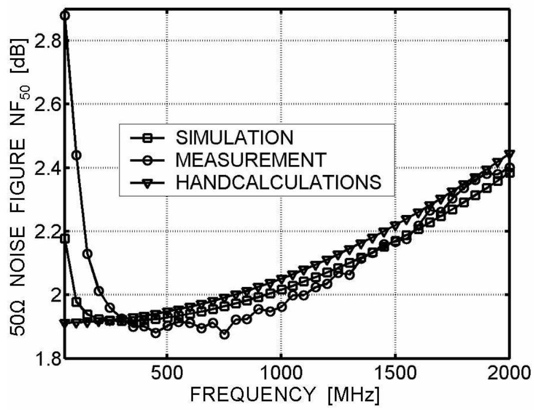

Fig. 4. Noise Figure [dB] as a function of frequency for the CMOS LNA [2,4].

B. Wide-band Variable Gain LNA

Even partial noise cancellation can result in advantages, as exemplified by the "Amp1" topology shown in the low part of fig.5. Compared to three other wide-band LNAs, which are systematically generated [5,6,7], its noise performance is remarkable. Even though the minimum noise figure is not below 3dB, it is relatively small compared to the other LNAs operating at the same gain. Detailed analysis [5,7] shows that this is due to (partial) noise cancellation of the noise of the common gate device (note that there are again two paths to the output, one via the common gate device Ma, one via the common drain device Mb). Moreover, the noise figure is almost independent of the gain, which is useful in LNAs requiring variable gain. Measurements on an LNA realized in a 0.35μm standard CMOS process show NF<4.4dB for 6-11dB gain, very good linearity (IIP3=15dBm) at only 1.5mA current consumption [6].

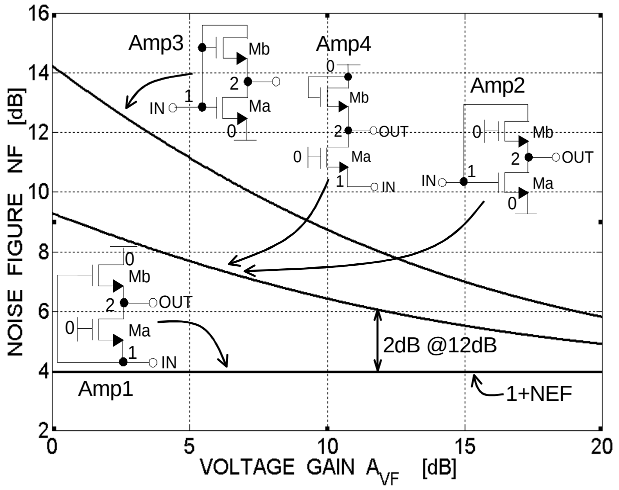

Fig. 5. Generalized noise cancellation concept and alternative circuit implementation [6].

C. Other Noise Cancellation Configurations

The concept of noise canceling can be generalized to other circuit topologies according to the model shown in fig. 6a. It consists of the following functional blocks: (a) An amplifier stage providing the source impedance matching, $Z_{IN}=R_S$. (b) An auxiliary amplifier *sensing* the voltage (signal and noise) across the real input source. (c) A network combining the output of the two amplifiers, such that noise from the matching device cancels while signal contributions add.

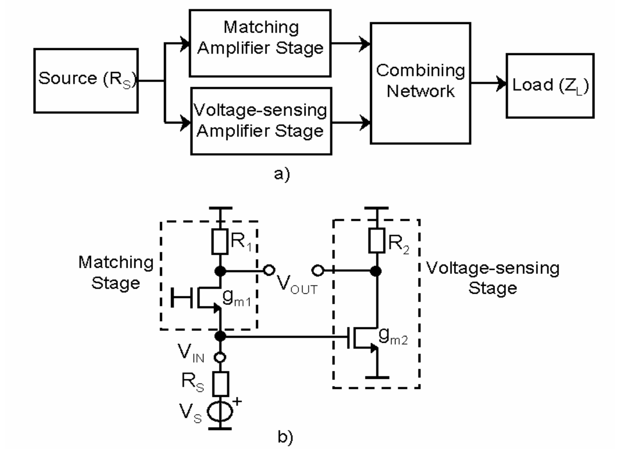

Fig. 6. Generalized noise cancellation concept and one possible alternative circuit implementation.

Fig. 6b shows an implementation example (biasing not shown) among several alternatives [3]. Noise cancellation occurs for $R_1=g_{m2}R_SR_2$, while low F requires high g_{m2}. The 2-MOSFETs configuration in fig. 6b is a well-known transconductor [8], also used in active mixers [9]. However, in both cases, noise canceling was apparently not recognized. Recently, a differential version of the same basic topology has been proposed, but now with degenerated bipolar transistors [10]. A noise figure close to 3dB was achieved, most probably partly due to noise cancellation.

V. CONCLUSION

In this paper, noise canceling was reviewed as a circuit technique, which is able to break the trade-off between noise factor F and source impedance matching. This is done placing an auxiliary voltage-sensing amplifier in feed-forward to the matching stage such that the noise from the matching device cancels at the output, while adding signal contributions. In this way, one can minimize the LNA noise figure, at the price of power dissipation in the auxiliary amplifier. By using this technique in an LNA, low noise figures over a wide range of frequencies can be achieved, without the instability issues that are typically associated with wide-band negative feedback amplifiers.

Other attractive assets of the technique are:

- Simultaneous cancellation of noise and distortion terms due to the matching device.
- Robustness to variations in device parameters and the external source resistance R_S.
- Simultaneous noise and power matching for frequencies where the effect of parasitic capacitors can be neglected.
- Orthogonality of design parameters for input impedance and gain, allowing easier implementation of variable gain at constant input match.
- Applicability in other IC technologies and amplifier topologies.

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