CONCEPT OF TRANSFORMER-FEEDBACK DEGENERATION OF LOW-NOISE AMPLIFIERS

Aleksandar Tasic, Wouter A. Serdijn and John R. Long

Delft University of Technology, Department of Microelectronics
Electronics Research Laboratory, UbiCom Research Project
Mekelweg 4, 2628 CD Delft, The Netherlands
Phone: +31 (0)15 278 9423 Fax: +31 (0)15 278 5922
E-mail: a.tasic|w.a.serdijn@its.tudelft.nl

ABSTRACT

Owing to its well-known properties of possible power-noise match as well as superior noise performance, the inductively-degenerated low-noise amplifier is nowadays the most-widely used RF pre-amplifier topology. However, there is not much flexibility when setting the amplifier's input impedance to match the source impedance, as for a given biasing condition, matching is achievable only for a particular value of the degenerative inductance. Moreover, to cancel the imaginary part of the input impedance, a particular excess inductance is necessary at the input of the amplifier, directly degrading the over-all amplifier performance. On the other hand, the transformer-feedback degeneration, introduced in this paper, offers the possibility, for low-noise amplifiers, of simultaneous matching to the source impedance of both the imaginary and the real part of the input impedance, in an orthogonal way. What is more, by controlling the amount of feedback, the amplifier is relieved from the burden of having the lossy input inductor. Detailed analysis shows a novel representation of the input impedance as well as a novel power-matching scheme of the transformer-feedback degenerated low-noise amplifier.

1. INTRODUCTION

Being multi-objective in nature, the design of low-noise amplifiers (LNA) imposes strict trade-offs among the performance parameters such as gain, noise-figure, linearity, and power consumption. Some of the challenging goals that designers of LNAs encounter are: (1) providing sufficient gain to minimize the noise contribution of the following stages in the front-end receive chain while at the same time not degrading over-all system linearity; (2) optimizing the noise-figure of the LNA itself with simultaneous noise and power match to the receive antenna; (3) operating at lowest possible power levels so as to ensure long battery life of portable systems.

Facilitating any of the fore-mentioned requirements will subsequently lead to a less complex design procedure. Setting the performance parameters in an orthogonal way would be the ultimate goal, where a simultaneous match of the real and the imaginary part of the input impedance to a source impedance can be considered as a first step towards.

Even though the inductively-degenerated LNA [1] is the most widely used amplifier scheme, it doesn't offer much flexibility for the power match at the input of the LNA. Namely, for a given biasing, resulting from the optimum noise condition, the 50Ω input impedance can be set for only one inductance value in the emitter of the amplifier's input transistor. What is more, for the full matching, a certain excess inductance must be placed at the input of the amplifier. Since integrated passive components are rather lossy, putting them in the direct signal path can seriously degrade the overall LNA performance [2].

The technique of transformer-feedback degeneration, proposed in this paper, offers the possibility for the low-noise amplifiers to achieve *matching* of both the real and the imaginary part of the input impedance in fully *orthogonal* way. Controlling the amount of feedback, the LNA is relieved from the burden of having the very-input lossy inductor. What is more, the power match is, unlike inductive degeneration, in this case, rather independent of the transistor's transition frequency (f_T), being possible even for the highest values of f_T.

The paper is divided into four sections. A novel input-impedance model is presented in Section 2. The concept of transformer-feedback degeneration is the subject of Section 3, while the conclusions are summarized in Section 4.

2. NOVEL INPUT-IMPEDANCE MODEL

For the last few years, there has been shown hardly any effort to depart from the traditional RF amplifier topologies, the inductive-degenerated (ID) LNA [1][2][3], relying on series feedback via the inductor in the emitter of the amplifier's input transistor, being one of them. Especially, the ID topology is effective in realizing power-noise match, due to the fact that optimum noise resistance and device input resistance can be adjusted independently, by changing the size and biasing conditions of the LNA's input transistor, respectively. However, once the optimum-noise biasing point is determined, there is no freedom in determining the emitter inductance, being responsible for the input power-match. What is more, for the full input match, a few times larger inductance is necessary at the input of the amplifier, additionally degrading the performances of the LNA.

Therefore, the transformer-feedback degenerated (TFD) LNA is introduced, offering much more design flexibility than

0-7803-7762-1/03/$17.00 ©2003 IEEE

conventional approach [1]. Functionally, TFD fully encompasses ID, as ID appears to be just the simplest form of TFD.

A schematic of transformer-feedback degenerated low-noise amplifier is shown in Fig. 1, without a complete biasing.

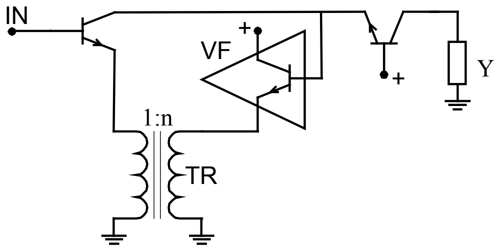

Fig. 1 Transformer-feedback degenerated LNA.

This amplifier topology is a traditional cascode configuration, with the addition of the feedback around the input transistor, that is realized by means of a voltage-follower (VF), in its simplest form a single transistor in a common-collector configuration and a transformer TR.

To appreciate the novel power-matching scheme, we will introduce an equivalent feedback function and an input impedance model. However, to make the forthcoming analysis tractable and insightful, the model of the equivalent input circuit to be dealt with will be presented first.

2.1. Equivalent input circuit

The circuit, shown in Fig. 1, can easily be simplified to the one of Fig. 2,

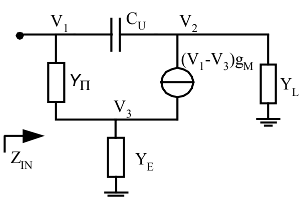

Fig. 2 Simplified input circuit of the TFD LNA.

where Y_Π is the base-emitter admittance, dominated by capacitance C_Π for high frequencies, C_U is the Miller capacitance, g_m the transconductance of the bipolar transistor, Y_L the input admittance of the following stage, in this case the input admittance of the common base transistor, and Y_E the equivalent admittance seen at the emitter of the first transistor. In the above configuration the transconductances of both transistors are assumed to be equal.

Applying Kirchof's current law to the schematic of Fig. 2, the input admittance is calculated to be:

$$Y_{IN} = Y_\Pi (1 - V_3/V_1) + sC_U (1 - V_2/V_1) \tag{1}$$

With the assumption that at the frequency of interest $sC_U << g_m$ and $sC_U << Y_L = g_m = g_m$, the equivalent input admittance becomes:

$$Y_{IN} = Y_\Pi \cdot f(Y_E) + sC_U [1 + g_m f(Y_E)/Y_L] \tag{2}$$

where $f(Y_E)$ is the *feedback function* equal to:

$$f(Y_E) = 1 - V_3/V_1 \tag{3}$$

Now, the input impedance can be estimated, simply, by using Eq. (2) that accounts for the feedback over capacitance C_U, and the function f, that can however be calculated without taking into account the Miller effect.

2.2. Feedback function

Not only is the feedback function f used for the calculation of the input impedance, but also it can be used for the estimation of the other LNA performance parameters, power gain being one of them. At this point, note as well that the properties of the function f depend to a large extent on the transformer parameters, that directly determine the amount of feedback in the amplifier.

As already indicated in the previous sub-section, the characterization of the function f is now rather facilitated, as the capacitance C_U does not appear in f. The equivalent circuit of the transformer-feedback degenerated LNA, to be used for the calculation of the function f, with a first order transformer model [4], is shown in Fig. 3,

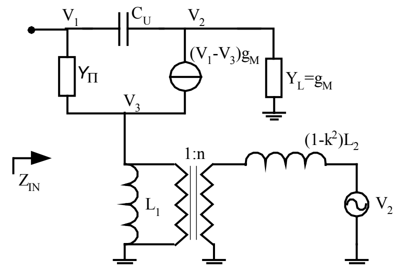

Fig. 3 Detailed schematic of the TFD LNA.

where L_1 and L_2 are the transformer primary and secondary inductors, n is the transformer turn ratio, and k is the coupling factor between the transformer inductors.

As the primary and the secondary inductors of the transformer TR are, by definition, related as $L_2/L_1 = n^2/k^2$, it is rather straightforward to calculate the voltage transfer from node V_1 to node V_3, and subsequently derive the function $f(Y_1, Y_2)$ as:

$$f(Y_1, Y_2) = \frac{Y_1 + n^2 Y_2}{Y_\Pi + Y_1 + n^2 Y_2 + g_m (1 \pm n Y_2 / Y_L)} \tag{4}$$

with $Y_1 = 1/sL_1$ and $Y_2 = 1/s(1-k^2)L_2$. Depending on the orientation of the transformer, the feedback can be either negative or positive, which is the origin of the ± sign in Eq. (4).

2.3. Input impedance model

Rearranging Eq. (1), the equivalent input impedance can be given a form:

$$Y_{IN} = Y_{IN}(C_U = 0) + sC_U + C_U / C_\Pi \cdot Y_{IN}(C_U = 0) \tag{5}$$

where $Z_{IN}(C_U=0) = 1/Y_{IN}(C_U=0)$ is calculated with the aid of the function f as:

$$Z_{IN}(C_U = 0) = Z_\Pi \left[1 + \frac{Y_\Pi + g_m (1 \pm n Y_2 / Y_L)}{Y_1 + n^2 Y_2} \right] \tag{6}$$

or in the final form as:

$$Z_{IN}(C_U = 0) = R + j \left[\frac{\omega_0}{\omega_T} R \pm \frac{\omega_T}{\omega_0} \frac{k^2}{n} \frac{1}{g_m} - \frac{\omega_T}{\omega_0} \frac{1}{g_m} \right] \tag{7}$$

with $R = \omega_T (1-k^2) L_1$ being the real part of the impedance and $\omega_T = 2\pi f_T$ and $\omega_0 = 2\pi f_0$ being the corresponding angular frequencies.

The equivalent of Eq. (7) in the circuitry is shown in Figs. 4a and 4b for both positive and negative feedback. One can notice, that the effect of TFD, as compared to ID, is the additional reactance seen at the input of the LNA, a capacitance in case of negative feedback and an inductance in case of positive feedback. How this extra reactance can be used for the input power-match is to be discussed in the next section.

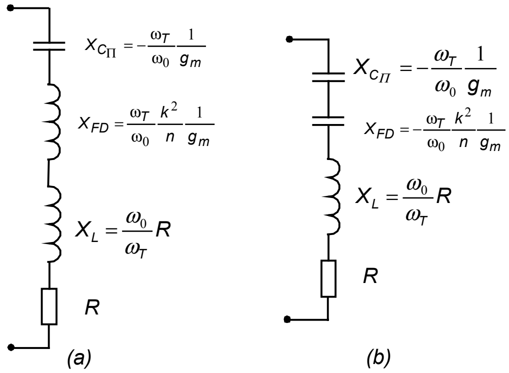

Fig. 4 $Z_{IN}(C_U=0)$ for (a) positive and (b) negative feedback.

Now the over-all input impedance model, following Eqs. (5) and (7) and Figs. 4a and 4b, can be shown as depicted in Fig. 5.

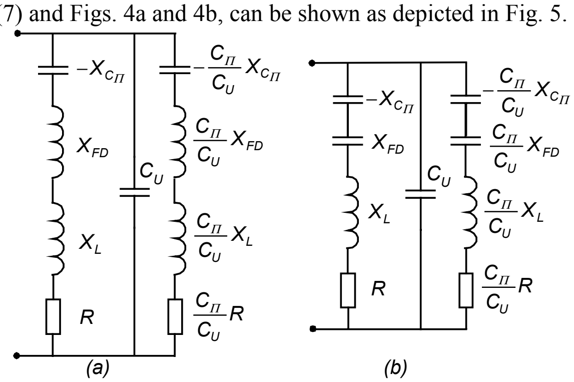

Fig. 5 Input impedance for (a) positive, (b) negative feedback.

Obviously, following the same procedure for the ID LNA, as has been done for the TFD case, the input impedance model is the same as the one shown in Fig. 5a or 5b, with the omission of the feedback reactance X_{FD}. This is, as expected, in line with [1].

Finally, apart from the development of the input-impedance model for the TFD LNA, the foregoing analysis has also shown that it is justifiable to omit the Miller capacitance and its effect when qualifying the performances of cascode amplifiers. Namely, as $C_\Pi/C_U>>1$ and $1/\omega_0 C_U>>R$, the models of Figs. 5a and 5b, reduce to the ones, already shown, in Figs. 4a and 4b ("C_U *free*" models), to be used in the coming analysis.

3. NOVEL POWER-MATCHING SCHEME

If meant for narrowband applications, the matching condition must be satisfied for the operating frequency of the front-end system. For that purpose, as well as due to their noiseless nature, reactive components have found their use in the feedback configurations, the inductively-degenerated, being one of them. In this case, a two step procedure is followed: setting the real part of the input impedance to a 50Ω by means of the inductor in series with the emitter of the input transistor and setting the imaginary part to zero by means of the inductor, usually rather large and lossy, in the base of the input transistor.

However, not only does the transformer-degeneration make the matching easier and more flexible, but also it enables the matching to be performed completely out of the direct signal path, i.e., in the feedback. For all these reason, the new topology appears to be a very promising candidate for RF applications.

Before we proceed evaluating the newly introduced scheme, we will first get acquainted with the constraints that this topology imposes. Namely, as there are two regions of operation, one for the negative and the other for the positive feedback, estimation of the loop-gain can give us full information regarding the applicability of the TFD LNA.

3.1. Loop-gain constraint

To evaluate the loop-gain of the TFD LNA, shown in Fig. 1, we will refer to the simplified schematic of Fig. 3, with a difference that node V_1 is, via the source resistance R_S, connected to the ground and voltage controlled current source $g_m V_{BE}$ is replaced by a uncontrolled one I. Now, the square of the module of the loop-gain, for the "*critical*" positive loop-gain, is calculated to be:

$$\left|LG\right|^2 = \left|\frac{g_m V_{BE}}{I}\right|^2 = \frac{(k^2/n)^2 + (g_m R\omega_0/\omega_T)^2}{1+(g_m R\omega_0/\omega_T)^2} \tag{8}$$

As for safe operation of the amplifier, its loop-gain must be below one, the condition to be satisfied reduces to:

$$k^2/n < 1 \tag{9}$$

which indicates that for the unquestionable stability the transformer turn ratio should be larger than the square of the coupling coefficient k.

3.2. Power-matching condition

With the aid of the input-impedance model for a slightly positive feedback (Fig. 4b), the equivalent input circuit can be shown as depicted in Fig. 7,

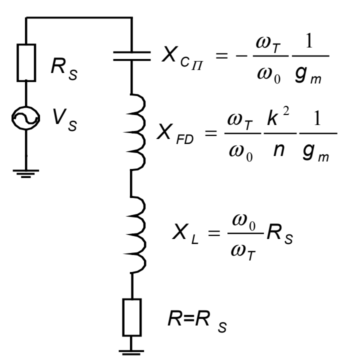

Fig. 7 Antenna and input of the TFD LNA.

where R_S is the source impedance, usually given a value of 50Ω.

Now, the condition for the match of the real part of the input impedance to a 50Ω source impedance is derived from Eq. (6) to be:

$$\omega_T(1-k^2)L_1 = R_S \tag{10}$$

which is the same as in the case of the ID.

However, setting the imaginary part of the input impedance to zero is facilitated, simply because the feedback-resulting inductance $(\omega_T/\omega_0)(1/g_m)(k^2/n)$, given in Fig. 7 and Eq. (7), enables the immediate *cancellation* of the capacitance C_Π. The condition to be satisfied is again derived from Eq. (7), by setting the imaginary part to zero, and has a form:

$$\frac{k^2}{n} = g_m \left[\frac{1}{g_m} - (\frac{\omega_0}{\omega_T})^2 R_S \right] \tag{11}$$

This condition implies that the stability criterion is not violated, even more favorizing the proposed scheme in comparison with the ID. What is more, small bond-wire inductance, if used, relaxes the loop-gain constraint to the extent that stability is not an issue any more.

Another property of the proposed topology is the *orthogonal match* of the real and the imaginary part of the input impedance. As indicated in Eq. (10), by choosing a certain value for the coupling factor k and the primary inductance of the transformer L_1, we are adapting the real part of the impedance. On the other hand, choosing the right value for the transformer turn ratio n, according to Eq. (11), the imaginary part is set to zero.

This becomes obvious when the matching conditions, i.e., Eqs. (10) and (11), are translated into the transformer parameters, as shown in Fig. 9,

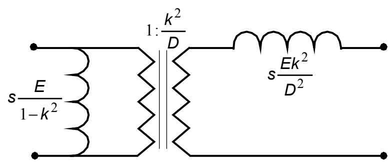

Fig. 9 Transformer power-matching model ($s=j\omega$).

where $E=R_S/\omega_T$ and $D=1-g_mR_S(\omega_0/\omega_T)^2$. This transformer model is suitable for simulation purposes, where the real part of the input impedance depends on parameter E, while the imaginary part depends on parameter D. The striking property of the transformer-feedback degeneration is its full independence of the parameter k, once the values for E and D are properly chosen. Namely, any change in the coupling coefficient k, for the model shown in Fig. 9, has no influence on input matching, as long as it has the value different from zero. Only in that case, the transformer-feedback degeneration reduces to the inductive-degeneration, where, as already indicated, the property of the simultaneous cancellation of the imaginary part, by means of controlling the feedback, is lost.

3.3. Example

To prove the validity of the introduced concept, a fully realistic example, also tested with the SpectreRF simulation tool, is presented.

Referring to the TFD LNA of Figs. 1 and 2, let us assume the following operating conditions of the amplifier: transition frequency f_T=24GHz, frequency of operation f_0=2.4GHz and collector current I_C=7mA. If the operating conditions are the same for the ID LNA, being the amplifier of Fig. 1 with the omission of the feedback and the addition of the inductors in the emitter and the base of the input transistor L_E and L_B, respectively, the following 50Ω matching parameters are obtained:

LNA\par.	L_E [nH]	L_B [nH]
ID	0.33	2.12

Table 1. Parameters of the power-matched ID LNA.

LNA\par.	k	E[e-9]	D	L_1[nH]	L_2[nH]	n
TFD1	0.9	0.33	0.86	1.74	1.9	0.95
TFD2	0.7	0.33	0.86	0.66	0.47	0.6
TFD3	0.5	0.33	0.86	0.44	0.16	0.3

Table 2. Parameters of the power-matched TFD LNA.

The results show that, unlike ID, in case of TFD the power-matching is possible not only for one, but for a number of different transformer's parameters values. Also, it is seen that parameters E and D of the, so-called, transformer power-matching model of Fig. 9, are indeed constant and the final choice of the primary and the secondary inductance of the transformer depends only on the factor k.

Finally, note that the other performance parameters of the transformer-feedback degenerated LNA, such as noise-figure, power-gain and linearity, are not lagging their inductively-degenerated counterparts. However, this aspect of performance characterization is left for the forthcoming analyses.

4. CONCLUSIONS

The emerging complexity of the nowadays RF front-end systems urges for immediate structurization, systematization and orthogonalization of the design procedure. This is the only way to tackle the multitude of the interwoven objectives that designers are faced with. Accordingly, the proposed concept of the transformer-feedback degeneration of low-noise amplifiers, is just a first step towards these goals, enabling simultaneous and orthogonal input power match.

In this paper, the input-impedance model of the amplifier has first been introduced, proving to be very suitable for the interpretation of the effect of the applied feedback on the input matching.

Also, a new power-matching scheme is proposed, where owing to the applied feedback, it is possible to match simultaneously both the imaginary and the real part of the input impedance to the source impedance, in an orthogonal way.

Finally, the matching components are removed from the direct signal path to the feedback, thus not additionally degrading the performances of the low-noise amplifier.

5. REFERENCES

[1] S.P.Voinigescu et al. "A Scalable High-Frequency Noise Model for Bipolar Transistors with Application to Optimal Transistor Sizing for Low-Noise Amplifier Design", *IEEE Journal SSC*, vol. 32, no. 9, pp. 1430-1439, Sep. 1997.

[2] H.Hashemi and A. Hajimiri, "Concurrent Multiband Low-Noise Amplifiers – Theory, Design and Applications" *IEEE Journal MTT*, vol. 50, no.1, pp. 288-301, Jan. 2002.

[3] D.K.Shaeffer and T.H.Lee, "A 1.5V 1.5GHz CMOS Low Noise Amplifier", *IEEE Journal SSC*, vol. 32, no. 5, pp. 745-759, May 1997.

[4] J.R.Long, "Monolithic Transformers for Si RF IC Design", *IEEE Journal SSC,* vol. 35, no. 9, pp. 1368-1382, Sept. 2000.