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Common Gate Transformer Feedback LNA in a High IIP3 Current Mode RF CMOS Front-End

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***Abstract*- A new topology of transformer based low noise amplifier is presented. The structure realizes a low noise input match and a current gain greater than one by a current to current positive feedback closed around a common gate stage. The amplifier is inserted in a high linearity current mode RF front-end receiver working between 4.15-4.4GHz with a NF of 4.2dB, a gain of 24.2dB and an IIP3 of -2dBm.**

I. INTRODUCTION

The extension of the frequency range and the request of wide channel bandwidth and reconfigurability have contributed to modify significantly the design of RF front-ends. High working frequencies minimize the area of passive devices allowing a more extensive use of electro-magnetic structures such as integrated transformers, furthermore wider band systems need novel design approaches avoiding the traditional narrow band resonant solutions [1],[2]. In this scenario, the inductively degenerated amplifier has proved inadequate to satisfy wide bandwidth and multi-standards requirements and recently several solutions derived from common gate stage were investigated [1-3].

The feedback and feedforward common gate topologies presented in literature, generally, need to develop sufficient voltage gain at their output node in order to close the loop with enough strength to reduce or cancel noise contributes [2],[3]. This requirement translates in the need of achieving a sufficiently high impedance at the output node, which in turn limits the amplifiers bandwidth at high frequencies where parasitic capacitances need to be resonated to synthesize high impedances. The new topology presented here overcomes these problems closing a current-current feedback around a common gate stage with an integrated transformer (Fig.1). This kind of solution, not only can provide a wide-band input matching (avoiding any resonant network), but also achieves a current gain greater than one through a positive feedback.

Although CMOS submicron technologies have reached very high cut-off frequencies, the continuous channel length reduction creates new challenges for analog design because of the finite output resistance that limits amplification especially at RF, where minimum channel devices must be used. To circumvent this fundamental problem, the transformer feedback LNA together with the two I&Q passive current mixers that follow it avoid developing any voltage amplification at RF moving it entirely in the base-band stage. As an added benefit, this current mode solution, minimizing the number of V-I conversions at RF, increases the front-end linearity.

The paper is organized as follows. In section II the transformer feedback amplifier is introduced providing analytical expressions for its input impedance, gain, noise and distortion. In section III the current mode, direct conversion front-end is described in detail while experimental results are reported in section IV. Finally some conclusions are drawn in section V.

II. TRANSFORMER FEEDBACK LNA

The positive feedback transformer amplifier is reported in Fig.1. The structure derives directly from the positive feedback common gate topology presented in [3]. In this case, the loop is closed by an integrated transformer (ideally noiseless) that senses the output current with the primary winding Lp and feeds-back the signal to the input through the secondary coil Ls.

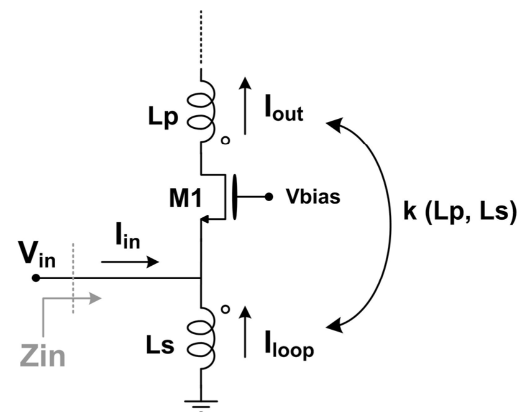

Fig. 1 Positive Transformer Feedback LNA

When a current I_{in} is injected in the input node, the loop produces a current I_{loop} proportional to I_{in} and in phase with it. As a consequence the input impedance of the amplifiers is increased and the output current becomes the sum of the current I_{in} (provided by the source) and I_{loop}, making the current gain (from the source to load) larger than one. Furthermore, the feedback reduces also the amplifier's noise figure compared to a standard common gate LNA.

A. Input Matching

The input impedance can be evaluated starting form the schematic depicted in Fig. 1. Assuming infinite output resistance for the MOS transistor (r_{ds}) and representing the transformer with its inductance matrix [4] the input impedance expression becomes:

$$Z_{in}(\omega) = \frac{j\omega Ls}{1 + j\omega(Ls - M)g_{m1}} \tag{1}$$

where Ls is the secondary winding self inductance, M the transformer mutual inductance and g_{m1} the transistor transconductance. Notice that the input impedance expression shows the typical high pass behavior given associated with the use of a transformer in shunt with its input [4].

Since this LNA avoid the use of resonant loads to synthesize the input impedance, it can operate over a wide frequency range, although the parasitic capacitances introduced by the transistor and the integrated transformer set an upper bound for the effective bandwidth. However, when the presence of strong interferers recommends a high-Q filtering action, a narrowband transfer function can be realized resonating transformer coils, improving efficiency [4].

In the amplifier band, i.e. for jω>> $1/(g_{m1}(Ls-M))$, the input impedance can be rewritten as:

$$Z_{in}(\omega)=\frac{1}{g_{m1}(1-M/Ls)}=\frac{1}{g_{m1}(1-k_m/n)} \quad (2)$$

where k_m is the coefficient of magnetic coupling and n the transformer turns ratio [4]. In (2) the dependency from jω ηασ disappeared consistently with the fact that the loop gain is only a function of k_m and n. These two parameters depend on the transformer geometry and generally have a low spread, assuring a good control of the loop gain. Moreover, if the turn ratio is chosen equal to one, the system stability is intrinsically guaranteed by the fact that k_m cannot be greater than one [4].

Setting Z_{in}=Rs in (2) it is possible to define the value of gm that ensures matching condition in the working band:

$$g_{m1}=\frac{1}{Rs(1-k_m/n)} \quad (3)$$

Again, since no voltage gain between the drain and source is required to implement a low noise matching, the Miller effect on the transistor output impedance (r_{ds}) is minimized. This feature is particularly relevant in deep-scaled technology where minimum channel transistors (with low output impedance) must be used to maximize the amplifier bandwidth.

B. Transconductance Gain

Under matching condition, the common gate LNA suffers from a poor gain, since its transconductance gain is forced to be equal to 1/Rs. Due to this common gate LNAs are generally avoided in current mode RF front-ends since their low gain increases the input referred noise contributed by the mixer and the base band circuits.

In the present solution, the additional current path, provided by the feedback (Fig. 1), produces a current gain (from the source to the output) greater than one or equivalently a transconductance G_m greater than 1/Rs. In fact, G_m is given by:

$$G_m=\frac{I_{out}}{V_{in}}=g_{m1}=\frac{1}{Rs(1-k_m/n)} \quad (4)$$

This expression points out how the gain can be made arbitrarily large making k_m/n close to unity, although this way lead to system instability. However, as explained in the previous section, the amplifier stability can be intrinsically guaranteed if a transformer 1:1 is used.

C. Noise Figure

The amplifier noise figure will be evaluated considering the noise of the input transistor and of the integrated transformer used in the feedback network. The schematic used for the noise calculation is drawn in Fig. 2. The transformer noise depends form the primary and secondary windings quality factors called Q_p and Q_s respectively.

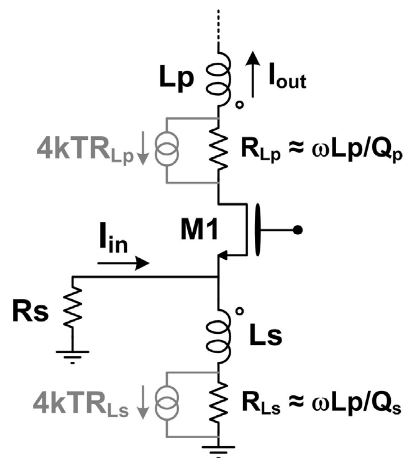

Fig. 2 Transformer noise sources

The two transformers spirals play a different role in the noise analysis: Ls injects its noise at the input of the amplifier, where a low impedance is present, while Lp is degenerated by the high impedance seen at the drain of M1 and its contribution is inversely proportional to the transistor r_{ds}. As a first approximation, assuming infinite the transistor output resistance the LNA noise factor results:

$$F=1+\frac{\gamma}{g_{m1}Rs}+\frac{Rs}{\omega LsQ_s} \quad (5)$$

where γ is the transistor thermal noise factor. Using the matching condition given by (3), equation (5) can be rewritten as:

$$F_{match}=1+\gamma(1-k_m/n)+\frac{Rs}{\omega LsQ_s} \quad (6)$$

This new expression points out the noise factor reduction given by the use of the transformer loop.

D. Linearity

If the integrated transformer is considered linear, the distortion in the amplifier output current is only due to the transistor characteristic (which depend on its bias point) and the source-gate signal-swing. As discussed with reference to the shunt feedback topology reported in [3], the linearity is independent of the loop gain and the expression for the amplifier IIP3 results

$$IIP3=\left|\frac{16V_{ov}{}^2(2+\theta V_{ov})^2}{3Rs}\right| \quad (7)$$

where V_{OV} is the voltage overdrive, θ a technology dependent fitting parameter and Rs the source resistance. This expression is equal to that of the IIP3 for a common gate LNA and indicates that is possible to improve the linearity of the stage acting on the dc-voltage overdrive, independently from the loop gain used. In actuality, a larger loop gain forces to have a larger g_{m1} (equation (3)) a therefore a larger current consumption to achieve the required V_{OV}.

III. CURRENT MODE FRONT-END

The transformer feedback amplifier was inserted in a fully-differential current mode RF front-end prototype for direct conversion architecture receiver (Fig. 3) working in the frequency range between 4.15GHz and 4.4GHz.

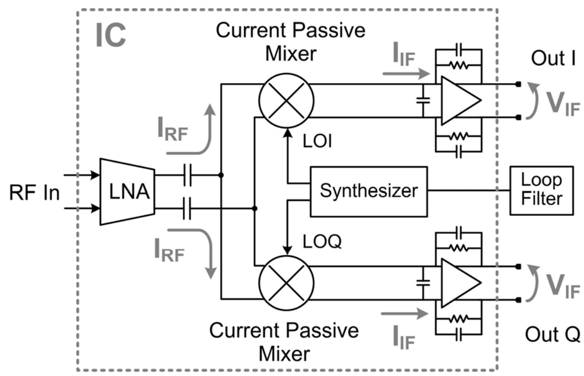

Fig. 3 Current Mode RF Front-End

The RF output current provided by the LNA is down-converted by a couple of current mode passive mixers [5] and finally amplified at low frequencies. The base band stage is realized by an RC load in feedback around an op-amp synthesizing the virtual ground for the current mode mixer. The prototype includes also a frequency synthesizer tunable from 4.15GHz to 4.5GHz [5].

A. Differential transformer feedback LNA

The pseudo-differential LNA structure was realized mirroring the single ended topology of Fig. 1 as shown in Fig.4.

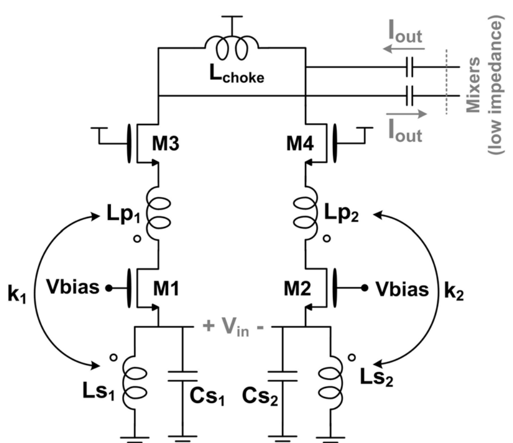

Fig. 4 Pseudo-differential transformer feedback LNA

In this case, to improve the transformer efficiency, a resonant winding approach was used although this strategy reduced the input matching bandwidth. In particular, the capacitances Cs_1 (Cs_2,) present at the input of the stage, resonate with the secondary coils Ls_1 (Ls_2). The transformer turn ratio used is 1:1 to guarantee an intrinsic system stability, while the k_m is close to 0.8 providing a transconductance gain G_m around 100 mS.

A cascode stage, formed by M3 (M4), was inserted between the LNA and the mixer to reduce the LO leakage at the input of the receivers. Finally an L_{choke} inductor was inserted at the LNA output to provide a dc-path for the amplifier bias current, without increasing the value of required voltage supply.

B. Passive Current Mixer and Base Band

The current mode RF front-end is completed by a couple of passive mixer topology and two virtual ground stages similar to those presented in [5]. The main features of this structure are the 1/f noise reduction compared to the classical Gilbert cell solution and the high linearity. In particular, the operation in current domain decreases the distortion associated with the large voltage swing at the mixer output nodes in passive voltage mode, while the absence of dc-current in the switching pairs reduces their flicker noise contribution.

Furthermore, the use of transistor in the triode region instead of saturation takes advantage from the technology scaling. In fact, while the short channel length reduces switch's on-resistance improving passive mixer solutions, the smaller r_{ds} (in saturated region) reduce the isolation between the output load in Gilbert mixers.

IV. EXPERIMENTAL RESULTS

The front-end fabricated standard 0.13μm CMOS technology from STMicroelectronics is shown in Fig. 5. The chip has an active area of 1.8mm^2 and all the pads are electrostatic discharge (ESD) protected.

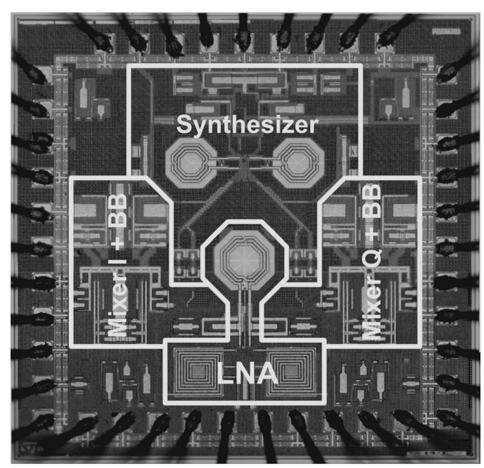

Fig. 5 Transformer LNA and Current Mode Front-end

The chip was bonded on a dedicated double-sided RF board, realized in a ROGERS4003 0.020-inch-thick substrate (εr = 3.38, tanδ = 0.027). An external SMD balun was used at the RF input to implement the single-ended-to-differential conversion. 50 Ω strip lines, optimized by means of EM simulations, carry the signal from the SMA connectors to the package inputs.

All the integrated coils are of the spiral type with ground shields to increase their quality factor. To minimize common-mode signals, induced by parasitic bond-wire inductances, multiple pads with multiple bonding are used for ground and voltage supply connections. Moreover, large on-chip bypass capacitors filter the noise on the supply voltage with respect to ground.

The input reflection coefficient has been tested by means of an Anritsu 37347C vector network analyzer. Fig. 6 reports the module representation of the S11 in the range of the frequency synthesizer.

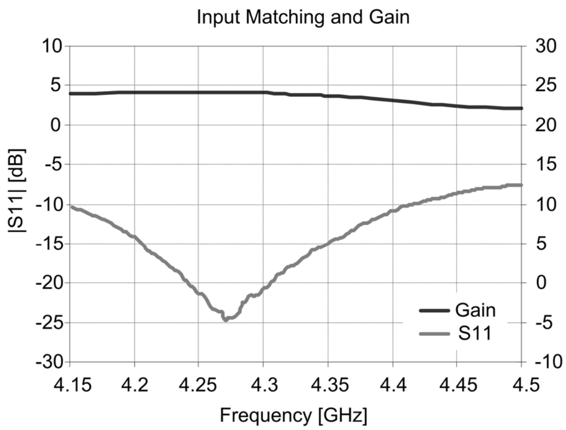

Fig. 6 S11 and RF Gain Measurement

The S11 magnitude is below -10dB from 4.15GHz to 4.4GHz reaching a minimum of -25dB. Fig. 6 shows also the gain as a function of the RF input frequencies, measured while keeping a relative LO frequency offset of 1.5MHz. The RF gain shows good flatness in the frequency range cover by the synthesizer with a maximum value of 24.2dB. This behaviour is guaranteed by the current mode approach that avoid high-Q resonant load. The base-band stage provides an output pole at 11.5MHz.

The noise figure was evaluated by means of an HP346B noise source. At the receiver output, a high-speed low-distortion differential amplifier was used to convert the differential mixer output signal to a single-ended format, drive the measurement equipment's 50 input ports, and raise the front-end output noise above the sensitivity level of the HP8564E spectrum analyzer that is used to capture its frequency content. The front-end noise figure is 4.2dB with a 1/f corner at 450kHz.

As explained in [5], in such kind of solution, the receiver IIP3 may depend on the frequency distance between the two tones used for the test if there is an insufficient isolation between synthesizer and mixer.

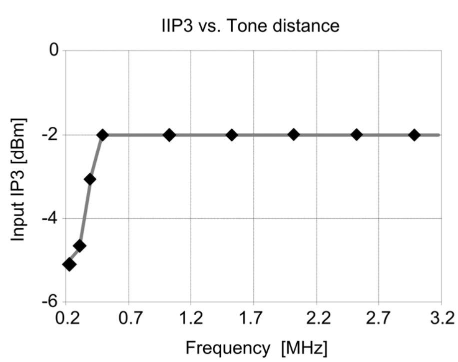

Fig. 7 IIP3 versus tone distance

For this reason, the IIP3 has been plotted versus the tones spacing and reported in Fig. 7. In this case, the IIP3 for two interferers spaced by 500kHz or more is -2dBm while it is around -5dBm if they are spaced 200kHz or less (in the PLL bandwidth). The improvement compared to the previous solution is due to the buffers introduced between local oscillators and switching pair's as predicted in [5]. The buffer introduction (not deeply optimized) explains the significant increment of power consumption compared to the previous work. All the measurement results are summarized in Table I.

TABLE I.
FRONT-END MEASUREMENT RESULTS

Gain [dB]	24.2
Noise Figure [dB]	4.2
1/f Corner [kHz]	450
IIP3 [dBm]	-2
1dB Compressio Point [dBm]	-12
Supply Voltage [V]	1.2 / 2.5
Total Power Dissipation [mW]	100
LNA Power Dissipation [mW]	12
Die Active Area [mm^2]	1.8
Technology	0.13 µm CMOS

V. CONCLUSION

A new transformer based positive feedback LNA was introduced. The use of integrated transformer guarantees an intrinsic amplifier's stability and a wide-band current gain without developing any voltage amplification at RF.

The amplifier was successfully merged with a couple of passive current mixers obtaining a high linearity current mode font-end for direct conversion architecture receivers.

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