

A Wideband CMOS Low Noise Amplifier Employing Noise and IM2 Distortion Cancellation for a Digital TV Tuner

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Abstract—A wideband CMOS low noise amplifier (LNA) with single-ended input and output employing noise and IM2 distortion cancellation for a digital terrestrial and cable TV tuner is presented. By adopting a noise canceling structure combining a common source amplifier and a common gate amplifier by current amplification, the LNA obtains a low noise figure and high IIP3. IIP2 as well as IIP3 of the LNA is important in broadband systems, especially digital terrestrial and cable TV applications. Accordingly, in order to overcome the poor IIP2 performance of conventional LNAs with single-ended input and output and avoid the use of external and bulky passive transformers along with high sensitivity, an IM2 distortion cancellation technique exploiting the complementary RF performance of NMOS and PMOS while retaining thermal noise canceling is adopted in the LNA. The proposed LNA is implemented in a 0.18 μm CMOS process and achieves a power gain of 14 dB, an average noise figure of 3 dB, an IIP3 of 3 dBm, an IIP2 of 44 dBm at maximum gain, and S11 of under -9 dB in a frequency range from 50 MHz to 880 MHz. The power consumption is 34.8 mW at 2.2 V and the chip area is 0.16 mm^2 .

Index Terms—Cable, CMOS, complementary device, current mirror, digital terrestrial, IIP2, IM2 distortion canceling, low noise amplifier, noise canceling, tuner.

I. INTRODUCTION

WITH the current global trend toward digitalization, many countries are starting to replace analog TV broadcasting with digital TV broadcasting. There are presently several digital TV standards such as advanced television systems committee (ATSC) in the USA and Korea, digital video broadcasting-terrestrial (DVB-T) in Europe, and integrated services digital broadcasting-terrestrial (ISDB-T) in Japan. Commercial TV receivers and set-top boxes supporting various digital TV standards require universal tuners capable of processing analog terrestrial, digital terrestrial, and cable broadcasts with low cost and small size [1]. Considering that most digital TV receivers adopt a CMOS digital demodulator, the

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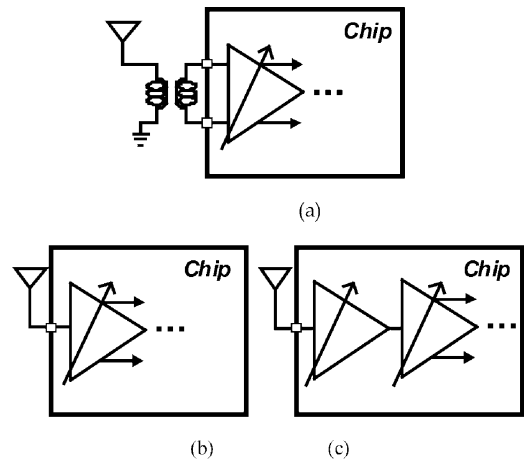


Fig. 1. Three methods to convert a single-ended signal from an antenna to differential signals in a wideband RF receiver by (a) a differential LNA using external bulky passive transformer, (b) a single to differential LNA, (c) a single-ended LNA and single to differential LNA.

RF tuners should be integrated with the CMOS digital demodulator into a single chip in order to support multi-standards and multi-functions with low cost and small size. Therefore, CMOS technology is believed to be the most promising technology for digital TV tuners.

The RF tuner must be capable of handling signals with broadband input frequencies ranging from 54 MHz to 880 MHz and thus requires a wideband LNA with sufficiently high gain and a low noise figure (NF) well below 3 dB in order to obtain high sensitivity. In the case of the cable TV standard, since hundreds of broadcast channels can enter the LNA without any filtering, the LNA should have higher linearity compared to the terrestrial TV standard. Furthermore, unlike narrow-band receivers such as CDMA, GSM, Bluetooth, and terrestrial digital multimedia broadcasting (T-DMB), which cover channels where the ratio of maximum frequency to minimum frequency is less than two, the second-order input-referred intercept point (IIP2) as well as the third-order input-referred intercept point (IIP3) of the LNA is important in the digital TV tuner. Therefore, since a LNA with single-ended input and output has poor IIP2 performance, most tuners adopt a fully differential LNA or a single to differential amplifier as the first stage of the receiver [2]–[6].

Although the fully differential LNA shown in Fig. 1(a) can provide excellent IIP2 performance over a wideband, one of its main drawbacks is that passive transformers are required to convert a single-ended signal to a differential signal. For wideband

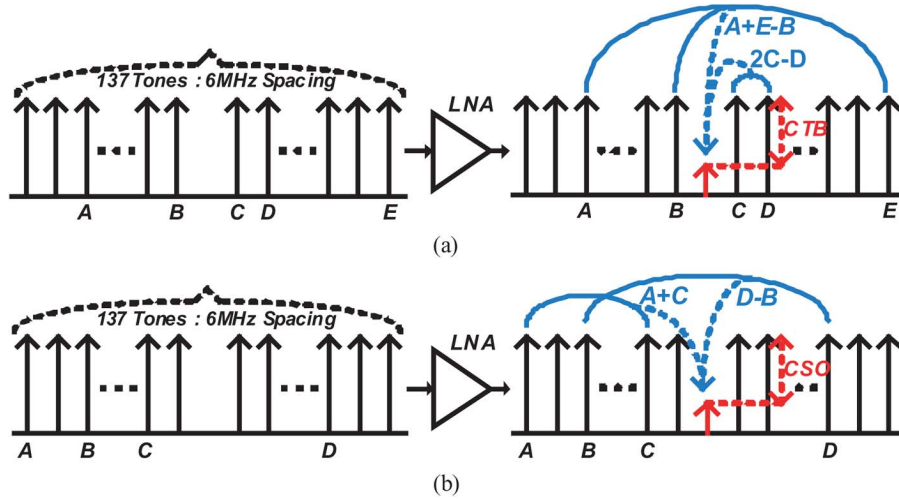


Fig. 2. Distortions generated by multiple interferers in LNA: (a) composite triple beat (CTB), (b) composite second-order distortion (CSO).

applications of a few hundred megahertz, however, the on-chip transformers are too bulky to be integrated on a chip and their insertion loss degrades the receiver NF and the sensitivity. For this reason, RF tuners adopting fully differential LNA still use low-loss and expensive external passive transformers. Placing a single to differential LNA as the first stage of the receivers, as shown in Fig. 1(b), is a good configuration for integration, and it can provide moderate IIP2 performance. However, due to the relatively high NF in this case, it is difficult to place the LNA on the first stage in a wideband tuner owing to sensitivity degradation, as shown in [7].

Thus, as shown in Fig. 1(c), in a wideband digital TV tuner, it is desirable to adopt a LNA with single-ended input and output in front of a single to differential amplifier in order to achieve excellent NF and also allow for the removal of external transformers. However, since the IIP2 burden lies in the LNA with single-ended input and output, improving the poor IIP2 performance of the conventional single-ended LNA presents a considerable challenge in the context of designing a fully integrated high sensitivity digital TV tuner without external transformers.

In this paper, we propose a wideband CMOS low noise amplifier with single-ended input and output employing noise and IM2 distortion cancellation in order to achieve an excellent noise figure and high linearity represented by the IIP2 and IIP3. Linearity issues such as composite triple beat (CTB) and composite second-order distortion (CSO) of a digital TV tuner are described in detail in Section II. The use of conventional wideband LNAs as the first stage of a digital TV tuner is described in Section III. The noise cancellation concept based on current amplification and a noise canceling wideband LNA, which were introduced in our previously published paper [8], are explained in Section IV. In order to improve the IIP2 performance without using external transformers, the complementary CMOS parallel push-pull (CCPP) concept is applied to the noise canceling wideband LNA described in Section IV and a new wideband LNA is proposed and designed in Section V. Section VI presents experimental results of the

proposed wideband LNA employing noise and IM2 distortion cancellation, followed by conclusions in Section VII.

II. LINEARITY ISSUES OF DIGITAL TV TUNERS

In a digital terrestrial TV tuner, the linearity of the RF front-end block is mainly related to desensitization by the one-tone blocker or third-order inter-modulation distortion (IMD3) by two-tone blockers. Since the IIP3 is directly related to desensitization and IMD3, it is the most important linearity factor. However, since two in-band blockers can generate second-order inter-modulation distortion (IMD2) at the desired signal band, the IIP2 is as important as the IIP3 in a broadband system.

On the other hand, in case of a cable TV tuner, hundreds of broadcast channels enter the RF front-end of a tuner without any filtering and the third-order or second-order inter-modulation distortions caused by multiple interferers can fall into the desired band. CTB is generally known as composite third-order distortion and these distortion products are caused by one, two, three or more carriers experiencing third-order nonlinearity, as shown in Fig. 2(a). CSO is a result of one or two carriers experiencing second-order nonlinearity, as shown in Fig. 2(b)[9]. The CTB and CSO are the ratio, expressed in dB, of the power of the desired RF carrier signal to the average power level of distortion components centered on the carrier. The simple relationship between CTB and third-order input-referred intercept point (IIP3), and that between CSO and second-order input-referred intercept point (IIP2), can be derived as follows:

$$CTB[dB] = -2(IIP3 - P_{sig}) + 6 \text{ dB} + 10 \log \left(\frac{3N^2}{8} \right) \quad (1)$$

and

$$CSO[dB] = -(IIP2 - P_{sig}) + 10 \log(N) \quad (2)$$

where N is the total number of entered tones and P_{sig} is the power of each entered tone [9].

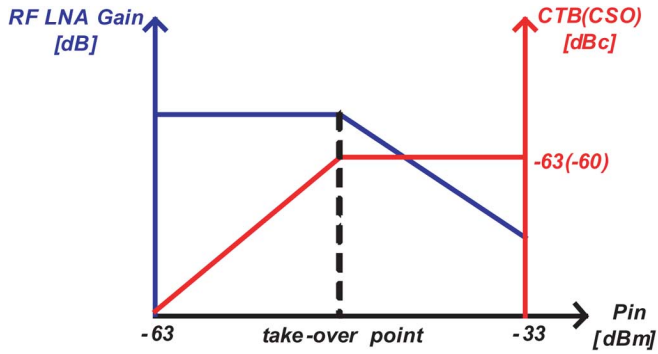


Fig. 3. CTB and CSO according to RF LNA gain control in the digital terrestrial and cable TV tuner.

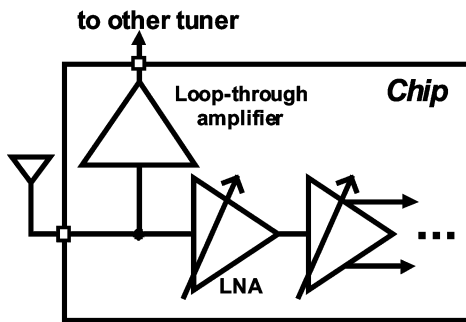


Fig. 4. Block diagram of RF front-end in a digital terrestrial and cable TV tuner IC.

In order to handle multiple distortions, the RF front-end of a tuner should have variable gain according to the power of the received signal. Since the CTB and CSO performance is degraded under target specifications if the RF front-end receives a signal with average power over some critical point (called the take-over point) at high gain mode, the gain of the LNA should be lowered at the take-over point, maintaining the performances of the third-order output-referred intercept point (OIP3) and the second-order output-referred intercept point (OIP2), as shown in Fig. 3. Considering the LNA should satisfy CTB and CSO performance under -63 dBc and -60 dBc, respectively, while receiving 137 multi-tones, each of which has strong power of -33 dBm, the required IIP3 and IIP2 of the LNA are approximately 20 dBm and 50 dBm by (1) and (2). In that situation, the RF LNA operates with approximately 0 dB gain considering the required signal to noise ratio (SNR) in the demodulator. Therefore, since the OIP3 (or OIP2) is the sum of IIP3 (or IIP2) and gain, in order to maintain the CTB and CSO under -63 dBc and -60 dBc respectively, the required OIP3 and OIP2 of the LNA are approximately 20 dBm and 50 dBm throughout the overall gain range. In addition, as shown in Fig. 4, in the case of a loop-through amplifier with a gain of 3 dB that passes the received signals to other tuners and generally employs a single-ended structure, since the loop-through amplifier should have linearity performance similar to that of the LNA, it is important to enhance the IIP2 performance of the single-ended amplifier.

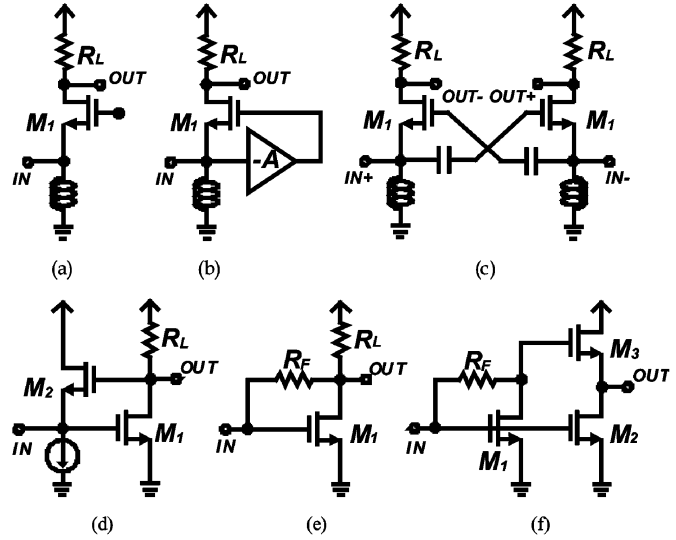


Fig. 5. Conventional low noise amplifiers: (a) common gate amplifier, (b) g_m -boosting common gate amplifier, (c) capacitor cross-coupled differential common gate amplifier [12], (d) common drain feedback amplifier, (e) resistive shunt feedback amplifier, (f) thermal noise canceling amplifier [11].

III. REVIEW OF CONVENTIONAL WIDEBAND LNAs

The common gate (CG) amplifier, common drain feedback amplifier, and resistive shunt feedback amplifier are widely used amplifiers that can achieve wideband input matching as well as flat and moderate gain over a wide operating frequency. While the CG amplifier in Fig. 5(a) can achieve 50 ohm input impedance (Z_{IN}) in wideband, its NF when the input impedance is well matched to the source impedance (R_S) is

$$F \approx 1 + \gamma \frac{g_{d01}}{g_{m1}} \quad (3)$$

where γ represents the ratio of the value of thermal noise at given drain bias to the value of thermal noise at $V_{DS} = 0$ V [10], g_{m1} is the trans-conductance of transistor M_1 , and g_{d01} is the channel conductance for $V_{DS} = 0$ V. Although the noise generated by R_L is ignored, (3) easily goes beyond 4 dB due to the tradeoff between the noise factor (F) and the impedance matching requirement ($Z_{IN} = R_S$) in a short channel device [11]. For the Gm-boosted CG LNA in Fig. 5(b), its noise factor is reduced by $1/(1+A)$ compared to the conventional CG LNA. However, since an amplifier with gain A may introduce additional noise, the Gm-boosting technique is only useful in a capacitor cross-coupled differential CG LNA [12], where $A = 1$ is easily achieved by naturally available inverting gain, as shown in Fig. 5(c).

The severe tradeoff between the noise factor and source impedance matching can be relaxed by negative feedback. Typically, there are two types of negative feedback topologies in designing a broadband LNA: the common drain feedback type and resistive feedback type. In the case of the common drain feedback LNA shown in Fig. 5(d), its input impedance is

$$Z_{IN} \approx \frac{1}{g_{m2}} \frac{1}{1 + g_{m1}R_L} \quad (4)$$

and the noise factor (ignoring the noise contributed by R_L) is

$$F \approx 1 + \frac{2}{3}g_{m2}R_S + \left(\frac{1}{R_L} + \frac{2}{3}g_{m1} \right) \frac{(1 + g_{m2}R_S)^2}{g_{m1}^2 R_S} \quad (5)$$

where g_{m2} is the transconductance of transistor M_2 , R_L is the load impedance, and R_S is the input source impedance. In order to achieve input matching and low noise figure, g_{m2} should be decreased as much as possible. However, this degrades the linearity considerably at high gain mode due to the nonlinearity of the transistor M_2 , as shown in [13].

The input impedance and noise factor (ignoring the noise contributed by R_F and R_L) of the resistive feedback amplifier in Fig. 5(e) are

$$Z_{IN} \approx \frac{R_F + R_L}{1 + g_{m1}R_L} \quad (6)$$

and

$$F \approx 1 + \frac{R_S}{R_F} \left(1 + \frac{1}{g_{m1}R_S} \right)^2 + \frac{g_{d01}\gamma}{g_{m1}^2 R_S} \quad (7)$$

where R_F is the impedance of the feedback resistor. In order to achieve a minimum noise figure, R_F and g_{m1} should be increased as much as possible under the condition that the sum of R_F and R_L is a constant value of $Z_{IN}(1 + g_{m1}R_L) \approx Z_{IN}A_V$, where A_V is the given gain of the LNA. This can degrade the gain flatness over a wide frequency range and ultimately it negatively affects the linearity.

In an effort to circumvent the trade-off between the noise factor and the requirement on input matching in the resistive feedback amplifier, a noise canceling technique was proposed in [11]. Excellent noise performance can be achieved in the LNA of Fig. 5(f) by canceling the thermal noise generated by input main transistor M_1 through the combining network composed of transistor M_2 and M_3 . However, relatively poor linearity performance is obtained despite heavy power consumption. In particular, since the IIP2 performance is poor, this LNA is unsuitable for the first stage of digital terrestrial and cable TV tuners.

IV. NOISE CANCELING AND LINEAR WIDEBAND LNA BASED ON CURRENT AMPLIFICATION

Fig. 6(a) shows our previously proposed wideband LNA adopting noise canceling technique based on current amplification. This technique combines a CG amplifier (M_1) and a common source (CS) amplifier (M_2) based on the current mirror (M_3 and M_4). The noise current generated by input transistor M_1 flows from the diode connection load consisting of PMOS M_3 to the choke inductor. Since the polarity of noise at the source of input transistor M_1 is opposed to that of noise at diode connection load M_3 , a noise canceling effect can be achieved at the output by adding the two paths, i.e., the CG path and the CS path, through the current mirror.

The input impedance of the proposed LNA can be approximately described as

$$Z_{IN} \approx \frac{1}{g_{m1}} \quad (8)$$

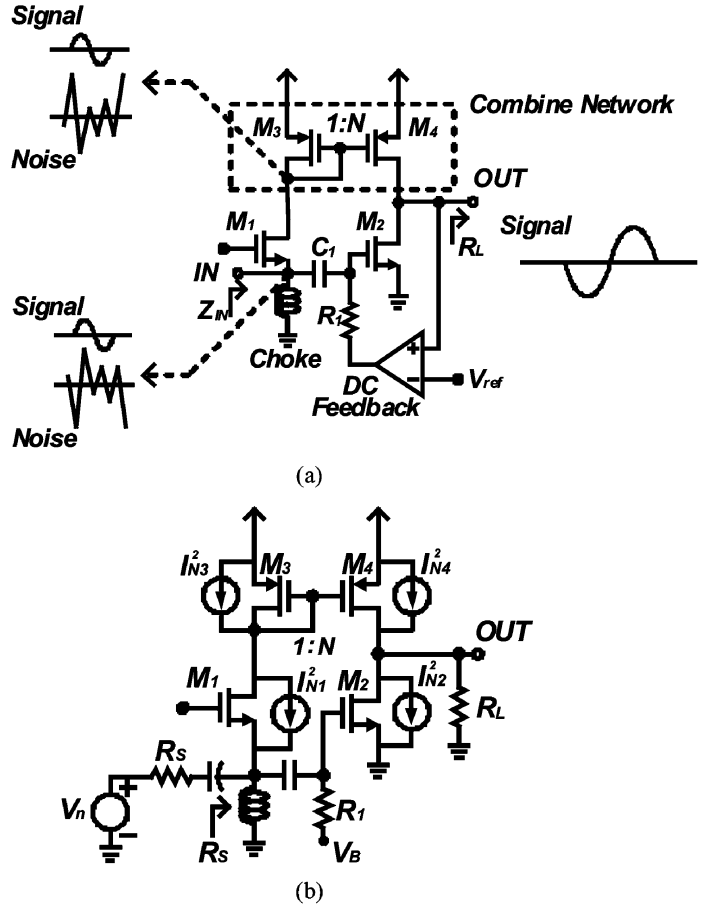


Fig. 6. (a) Our previously proposed noise canceling LNA with a structure combining a common source amplifier with a common gate amplifier by current amplification. (b) Schematic for noise analysis.

and wideband input matching can be easily achieved. The gain of the LNA is given by the sum of the CG amplifier gain term and the current mirroring ratio N , and the CS amplifier gain term. When the input impedance of the LNA is well matched to the source impedance, the gain can be expressed as

$$G|_{Z_{IN}=R_S} \approx -\frac{1}{2}(g_{m1}N + g_{m2})R_L \quad (9)$$

where N is the scaling ratio of the two current mirrors, M_3 and M_4 . Considering noise only by M_1 , M_2 , M_3 , and M_4 , the total noise voltage at the output node under an input matched condition, as shown in Fig. 6(b), can be calculated as

$$V_{n,out}^2 \approx 4kTR_S G^2 + \left[\frac{1}{4}I_{n1}^2 R_S^2 g_{m2}^2 - N^2 I_{n1}^2 \right] R_L^2 + (N^2 I_{n3}^2 + I_{n4}^2 + I_{n2}^2) R_L^2 \quad (10)$$

where G is the gain of the LNA at the input matched condition and I_n is the transistor's drain current noise. For canceling the second term, which is the dominant noise source of the common

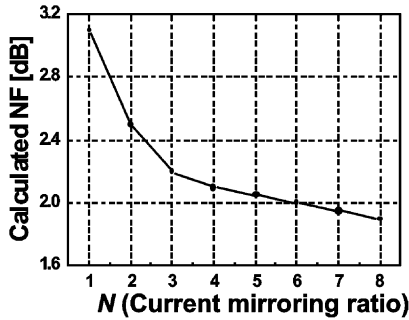
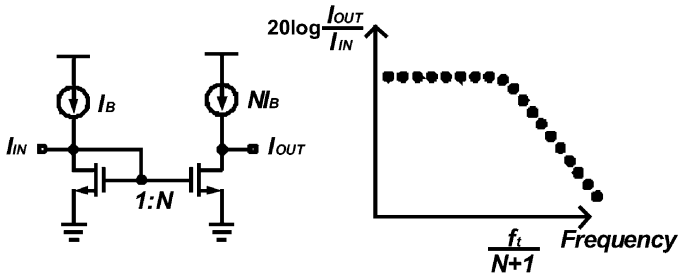
Fig. 7. Calculated NF by varying current mirroring ratio N .

Fig. 8. Current gain-bandwidth of a current mirror.

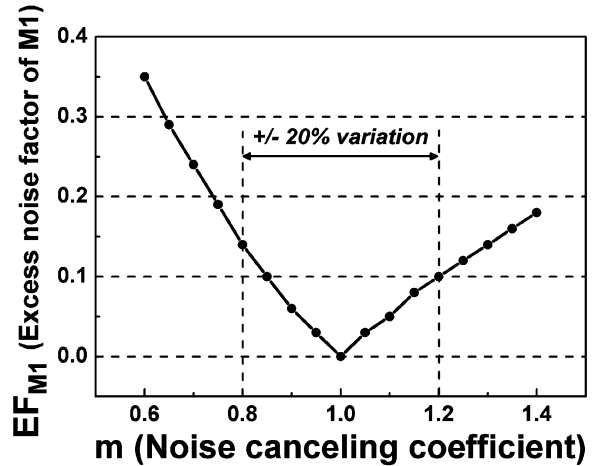
gate transistor M_1 , if g_{m2} is equal to $2Ng_{m1}$, the noise factor of the proposed LNA can be given as

$$\begin{aligned}
 F &= \frac{V_{n,out}^2}{4kTR_S G^2} \\
 &\approx 1 + \frac{(2N^2 I_{n3}^2 + I_{n2}^2) R_L^2}{4kTR_S G^2} \\
 &= 1 + \frac{\frac{8}{3} (2N^2 g_{m3} + g_{m2})}{R_S (g_{m1} N + g_{m2})^2} = 1 + \frac{8 (2N^2 g_{m3} + g_{m2})}{27N^2 g_{m1}} \\
 &= 1 + \frac{16}{27} \left(\frac{g_{m3}}{g_{m1}} + \frac{1}{N} \right). \quad (11)
 \end{aligned}$$

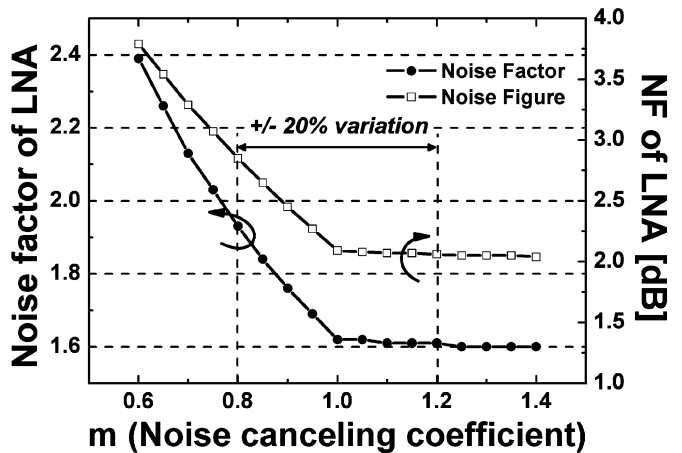
Fig. 7 shows the calculated noise figure by varying the current mirroring ratio N at the input matched condition and $g_{m3}/g_{m1} = 0.8$. Adopting the current mirror as a network that combines the CG path and the CS path in order to cancel thermal noise, we can obtain high gain and low NF well below 3 dB without considering the input matched condition by increasing N . Since the limited bandwidth of the simple current mirror is given by $f_t/(N+1)$, as shown in Fig. 8, where f_t is the unit gain frequency of a current mirror transistor, the proper scaling ratio value of N should be carefully selected [14]. Furthermore, a long-channel transistor should be used in the current mirror if possible under the allowed gain bandwidth and supply voltage, because harmonic distortions of the current mirror are generated by the mismatch of threshold voltage [15].

Noise canceling based on current amplification is relatively robust to the applied bias point. If the ratio $g_{m2}/2Ng_{m1}$ is defined as m , which is called the noise canceling coefficient (NCC) in this paper, the noise factor of the LNA is given as

$$\begin{aligned}
 F &\approx 1 + EF_{M1} + EF_{ETC} \\
 &= 1 + \frac{8}{3} \frac{|1 - m^2|}{(1 + 2m)^2} + \frac{16}{3(1 + 2m)^2} \left[\frac{g_{m3}}{g_{m1}} + \frac{m}{N} \right] \quad (12)
 \end{aligned}$$



(a)



(b)

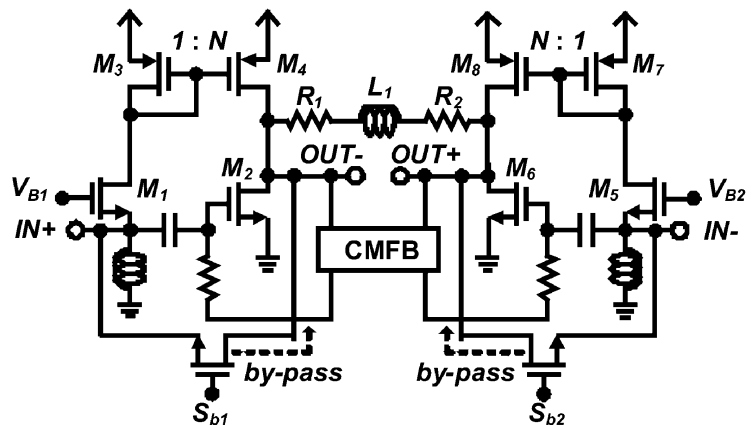
Fig. 9. Robustness of noise canceling by the deviation of m (noise canceling coefficient) in the LNA of Fig. 6: (a) EF_{M1} (excess noise factor of $M1$), (b) noise factor and NF.

Fig. 10. Our previously proposed noise canceling differential LNA.

where the excess noise factor EF_{M1} is the dominant noise contribution of the common gate transistor $M1$ in Fig. 6 and EF_{ETC} is the noise contribution of the remaining transistors $M2$, $M3$, and $M4$ in Fig. 6. It can be seen that (12) is precisely equal to (11) when NCC is 1. Fig. 9 shows the EF_{M1} and noise factor of the LNA versus the variation of NCC under

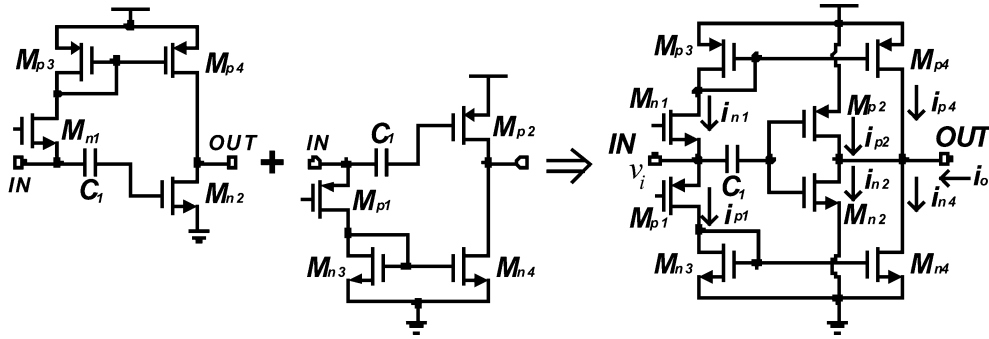
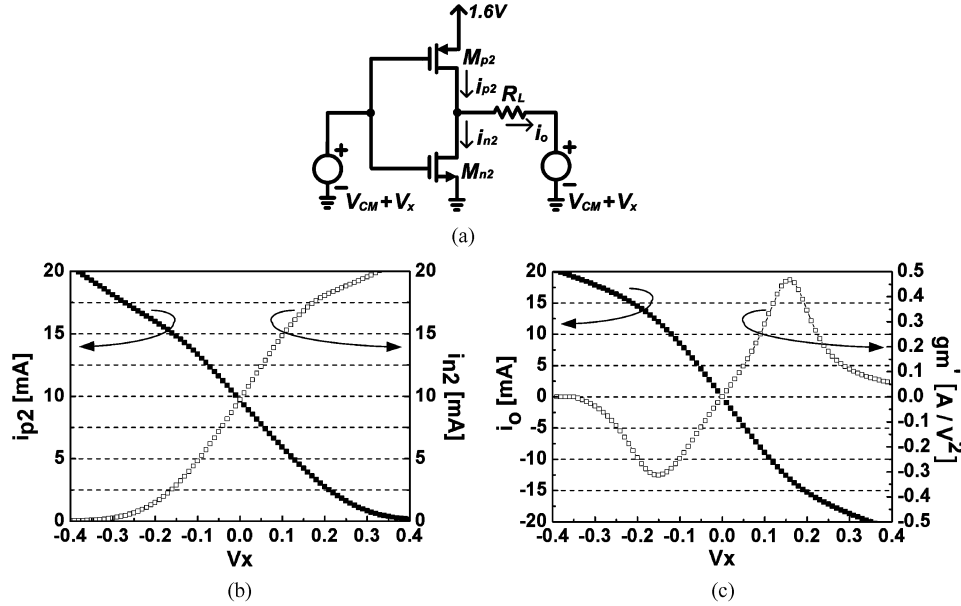


Fig. 11. Concept of newly proposed LNA employing IM2 distortion cancellation.


 Fig. 12. Inverter type amplifier using NMOS M_{n2} and PMOS M_{p2} and its current behavior: (a) inverter amplifier, (b) drain current of M_{n2} and M_{p2} over input voltage V_x , (c) difference between M_{n2} and M_{p2} drain current and its second derivative over input voltage V_x .

the condition of $g_{m3}/g_{m1} = 0.8$ and $N = 4$. As shown in Fig. 9(a), change as large as $\pm 20\%$ in the NCC increases EF_{M1} by only 0.15, which is one-tenth of the contribution of the noise factor at the optimum bias point for noise canceling. As NCC goes beyond one, the value of EF_{ETC} decreases, because the LNA has higher power consumption. Therefore, as shown in Fig. 9(b), the total noise factor of the LNA becomes flat in the region where NCC is larger than one, since the degree to which EF_{ETC} is ameliorated compensates the degree to which EF_{M1} is deteriorated. Considering the total noise factor and power consumption, power efficient NF optimum performance in the LNA can be obtained by selecting NCC as one. As seen in Fig. 9, the sensitivity to variation of NCC is low, and thus the total noise factor of LNA is insensitive to device parameter variation.

Although relatively high IIP3 over a wideband can be obtained through current amplification, which can eliminate $V-I$ converters that result in nonlinearity, the IIP2 performance is still poor owing to its inherent structure with single-ended input and output. Therefore, in [8], a differential structure, which degrades the NF due to the loss of an external bulky transformer and the added noise of the dc common mode circuit, was un-

avoidably adopted. The schematic of the proposed differential LNA is shown in Fig. 10. This LNA shows an average NF of 3.6 dB at a high gain mode. However, although this degradation can be acceptable in a cable tuner, which has more flexible NF specifications compared to a terrestrial tuner, this degraded performance must be addressed in the case of a commercial digital terrestrial tuner that requires high sensitivity. Furthermore, an expensive external transformer should be eliminated. Therefore, it is essential to improve the IIP2 performance of the proposed noise canceling single-ended LNA.

V. NEW WIDEBAND LNA ADOPTING IM2 DISTORTION CANCELLATION

The importance of second-order nonlinearity in both a direct conversion receiver and a general broadband system was presented in [16]. This paper described a technique called complementary CMOS parallel push-pull (CCPP) to enhance the IIP2 of the amplifier with single-ended input and output. In spite of the structure with single-ended input and output, the remarkable IIP2 performance could be acquired by utilizing the complementary characteristic of NMOS and PMOS operated as a differential mode. Therefore, we apply this concept, combining

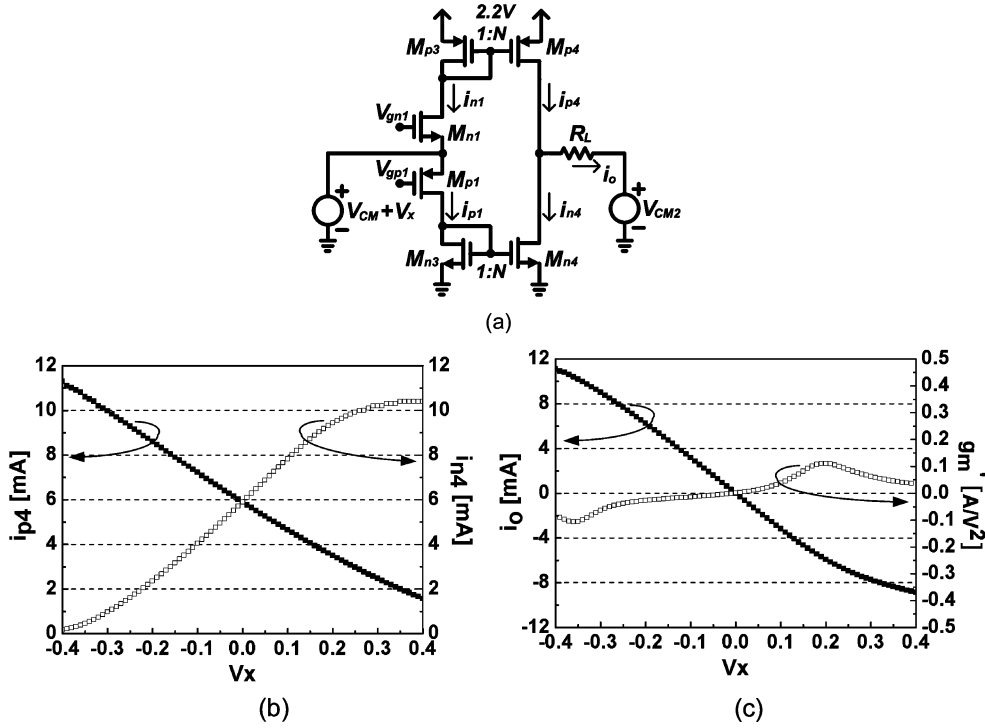


Fig. 13. Complementary CG amplifier with current mirrors and its current behavior: (a) complementary CG amplifier with current mirrors, (b) drain current of M_{n4} and M_{p4} over input voltage V_x , (c) difference between M_{n4} and M_{p4} drain current and its second derivative over input voltage V_x .

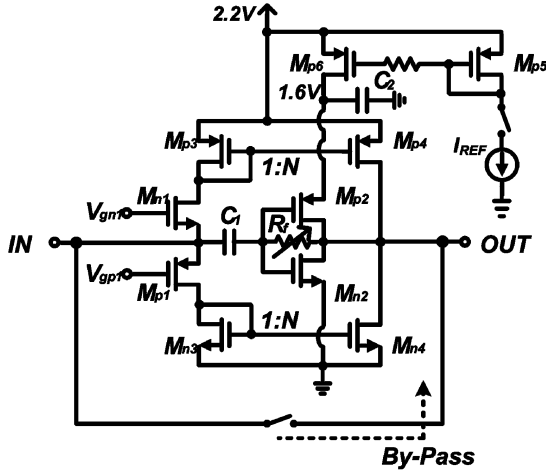


Fig. 14. Complete circuit schematic of the newly proposed LNA.

the in-phase nodes of the NMOS and PMOS as the outputs, i.e., the drain nodes of the NMOS and PMOS, to the proposed noise canceling wideband LNA described in Section IV in order to enhance the IIP2 performance. To fully utilize the RF performance of NMOS and PMOS in a complementary fashion while retaining the thermal noise canceling, complementary components consisting of transistors M_{p1} , M_{p2} , M_{n3} , and M_{n4} are added to the existing configuration made up of transistors M_{n1} , M_{n2} , M_{p3} , and M_{p4} , as shown in Fig. 11. The input matching, gain, and noise property of the newly proposed LNA can be easily calculated by $g_{m,n} + g_{m,p}$ for $g_{m,n}$ or $g_{m,p}$ of the equations given in Section IV. The current consumption of the proposed LNA is reduced for the same input impedance and NF as a result of reusing current. In addition, because the proposed

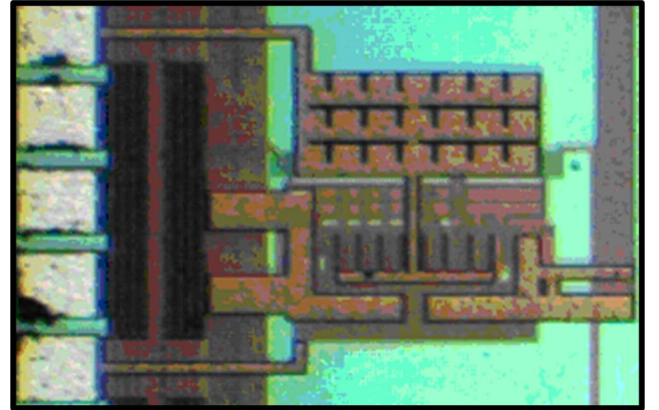


Fig. 15. Chip photograph of the proposed LNA.

LNA is composed of the noise canceling single-ended LNA described in Section IV and a complementary version based on current amplification, the IIP3 of the LNA is compatible to that of the noise canceling single-ended LNA.

If a small signal input voltage Δv_i is applied to the proposed amplifier and the additional distortions generated by the current mirror are ignored, the small signal drain currents of NMOS (M_{n2} and M_{n4}) and PMOS (M_{p2} and M_{p4}) can be expressed as

$$\begin{aligned}
 i_{n2} &= g_{m,n2}\Delta v_i + g'_{m,n2}(\Delta v_i)^2 + \dots, \\
 i_{p2} &= g_{m,p2}(-\Delta v_i) + g'_{m,p2}(-\Delta v_i)^2 + \dots, \\
 i_{n4} &= N i_{p1} = N [g_{m,p1}(\Delta v_i) + g'_{m,p1}(\Delta v_i)^2 + \dots], \\
 i_{p4} &= N i_{n1} \\
 &= N [g_{m,n1}(-\Delta v_i) + g'_{m,n1}(-\Delta v_i)^2 + \dots] \quad (13)
 \end{aligned}$$

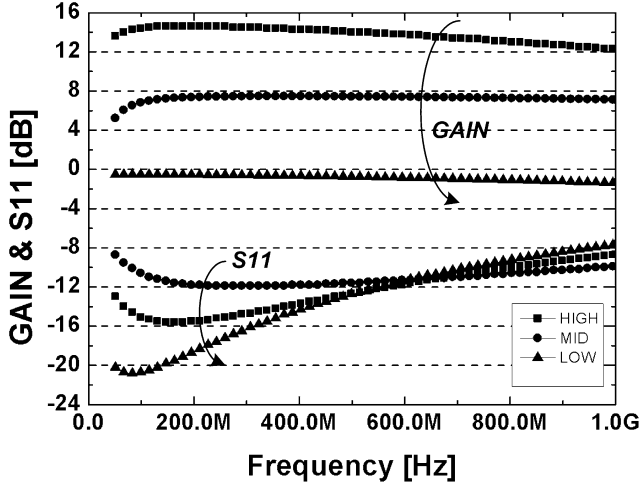
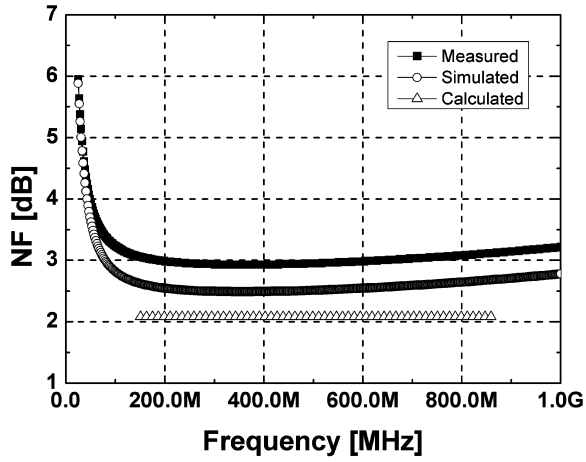
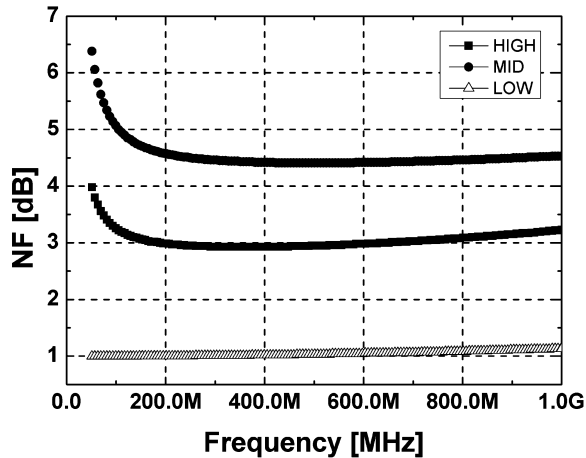


Fig. 16. Measured power gain and S11 of the proposed LNA.



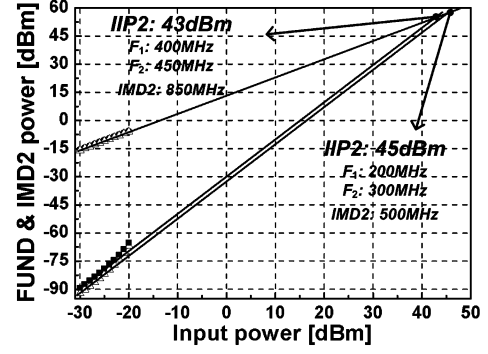
(a)



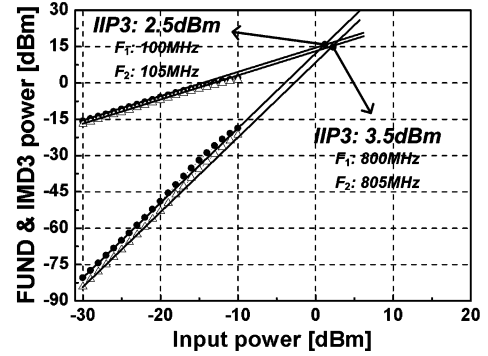
(b)

Fig. 17. Noise figure: (a) calculated NF, simulated NF, and measured NF at high gain mode, (b) measured NF of the proposed LNA according to gain control.

where g'_m is the first derivative of the transconductance g_m with respect to the gate-to-source voltage v_{gs} . The second-order non-linearity of CMOS RF circuit is dominated by the g'_m non-linearity. The process of calculating $g'_{m,n}$ and $g'_{m,p}$ is described in



(a)



(b)

Fig. 18. (a) IP2 measurement plot and (b) IP3 measurement plot at high gain mode.

detail in [16]. Thus, the small signal output current (i_o) is given as follows:

$$\begin{aligned} i_o &= (i_{n4} - i_{p4}) + (i_{n2} - i_{p2}) \\ &= N \left[(g_{m,n1} + g_{m,p1}) \Delta v_i \right. \\ &\quad \left. + (g'_{m,p1} - g'_{m,n1}) (\Delta v_i)^2 + \dots \right] \\ &\quad + \left[(g_{m,n2} + g_{m,p2}) \Delta v_i \right. \\ &\quad \left. + (g'_{m,n2} - g'_{m,p2}) (\Delta v_i)^2 + \dots \right]. \end{aligned} \quad (14)$$

From (14), the input-referred voltage intercept point for second-order inter-modulation of the proposed amplifier can be expressed as

$$V_{IIP2} = \left| \frac{N(g_{m,n1} + g_{m,p1}) + g_{m,n2} + g_{m,p2}}{N(g'_{m,p1} - g'_{m,n1}) + (g'_{m,n2} - g'_{m,p2})} \right|. \quad (15)$$

Equation (15) states that V_{IIP2} of the proposed amplifier can be infinite if the following holds:

$$\begin{aligned} g'_{m,n1} &= g'_{m,p1} \text{ and} \\ g'_{m,n2} &= g'_{m,p2}. \end{aligned} \quad (16)$$

Therefore, the IIP2 of the proposed amplifier can be improved by choosing the proper NMOS (M_{n1} and M_{n2}) and PMOS (M_{p1} and M_{p2}) transistor size.

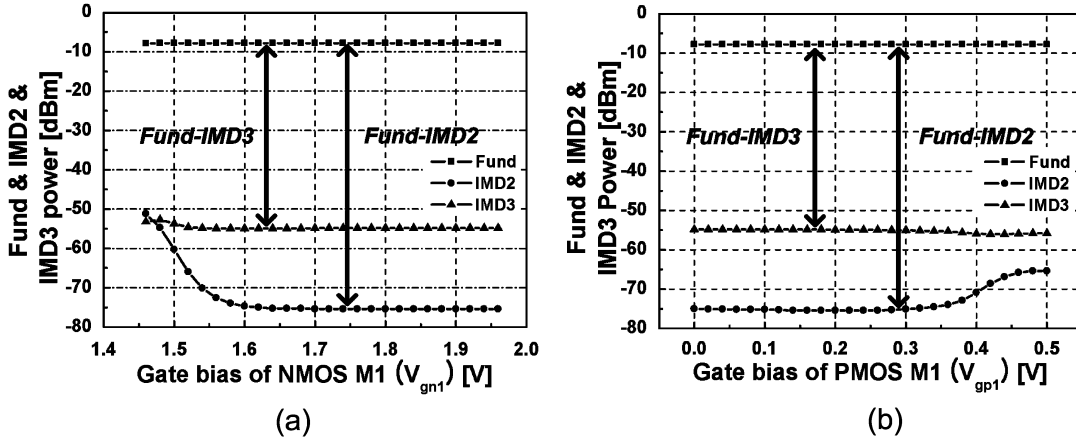


Fig. 19. (a) Measured IMD2 and IMD3 power versus the gate bias variation of M_{n1} and (b) gate bias variation of M_{p1} at high gain mode.

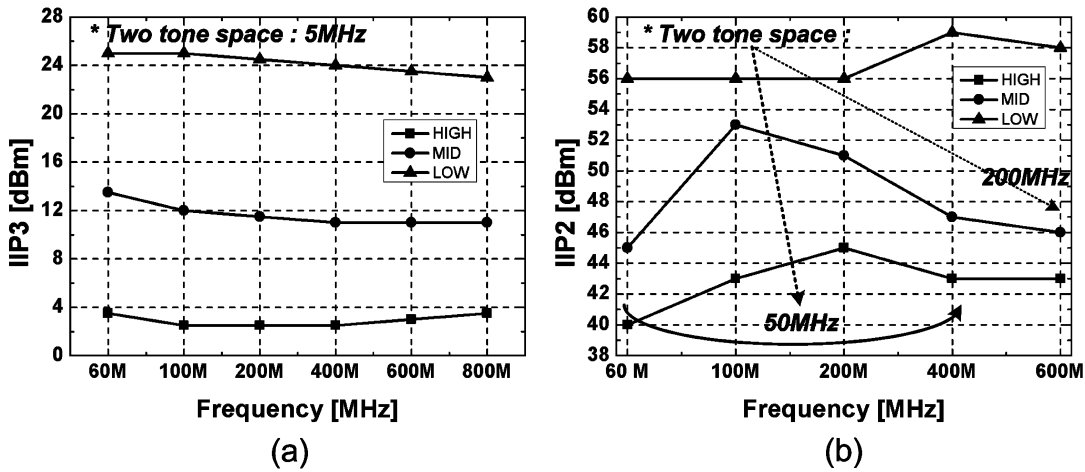


Fig. 20. (a) IIP3 measurement and (b) IIP2 measurement over frequency according to gain control.

On the other hand, in order to achieve noise and IM2 distortion canceling properties simultaneously, we should satisfy the following two conditions:

$$(g_{m,n2} = 2Ng_{m,n1}, g_{m,p2} = 2Ng_{m,p1})$$

$$\text{and } (g'_{m,n1} = g'_{m,p1}, g'_{m,n2} = g'_{m,p2}). \quad (17)$$

Unfortunately, it is impossible to meet these two conditions simultaneously, because the optimum condition for noise canceling is different from the optimum condition for IM2 distortion canceling. However, in Section IV, it was shown that the NF degradation is relatively insignificant when the applied bias point deviates from the optimum bias points for noise canceling. Since the optimum bias points for IM2 distortion cancellation can be found within $\pm 20\%$ variation of NCC, it is possible to enhance the OIP2 performance remarkably while satisfying the allowed target NF performance by adjusting the bias condition.

The size of the transistors in the proposed LNA is chosen considering both RF performances as noted above and power consumption. To obtain symmetrical characteristics, the size of the PMOS counterparts is determined to be double that of the NMOS counterparts in the proposed LNA. If proper gate biases of M_{n1} , M_{p1} , M_{n2} , and M_{p2} , satisfying (16), are chosen in the

proposed LNA circuit, an I-V curve similar to that of a differential circuit can be acquired. Figs. 12(a) and 13(a) show the simulation environment for observing the drain currents of M_{n2} and M_{p2} and the drain currents of M_{n4} and M_{p4} over the input voltage V_x , respectively. Figs. 12(b) and 13(b) illustrate the simulated I-V curve that is similar to that of fully differential circuits. As shown in Figs. 12(c) and 13(c), the second derivative of the output current (i_o) with respect to the input voltage V_x has approximately odd symmetry. Therefore, the excellent IIP2 performance can be expected in the proposed LNA.

The complete proposed LNA employing noise and IM2 distortion cancellation simultaneously is shown in Fig. 14. The feedback resistor R_F is used to provide stable dc output voltage and gain flatness, maintaining the stability, NF, and linearity. To reduce the sensitivity of the IM2 distortion cancellation, the inverter type amplifier using NMOS and PMOS simultaneously is biased via a current mirror while a large MOS capacitor grounds the source of M_{p2} over a wideband covering VHF low frequency.

In order to handle multiple distortions generated by strong adjacent channels, the LNA has three gain modes. By changing the current mirror ratio N and feedback resistor R_F by the switch control, we can make high and mid gain mode. In addition, the LNA operates in low gain mode through by-pass transistor.

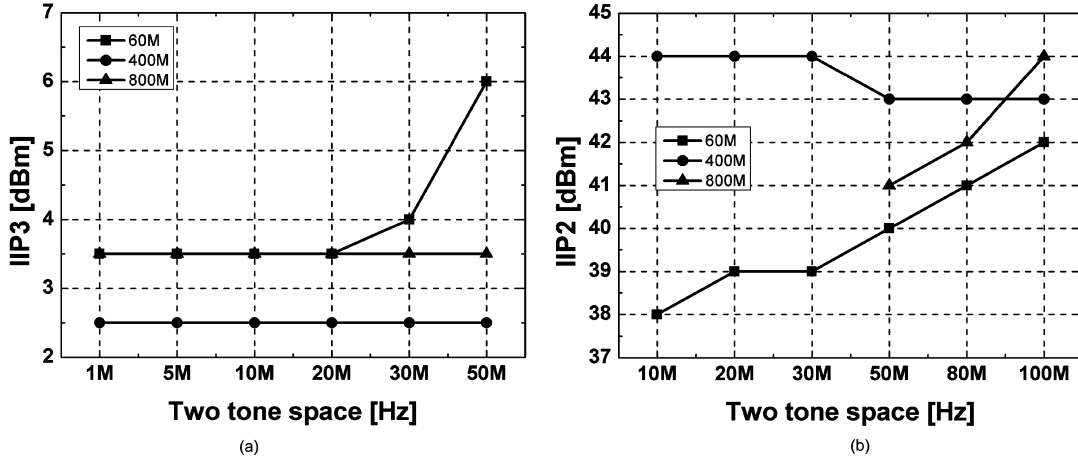


Fig. 21. (a) IIP3 measurement and (b) IIP2 measurement over the spacing of applied tones.

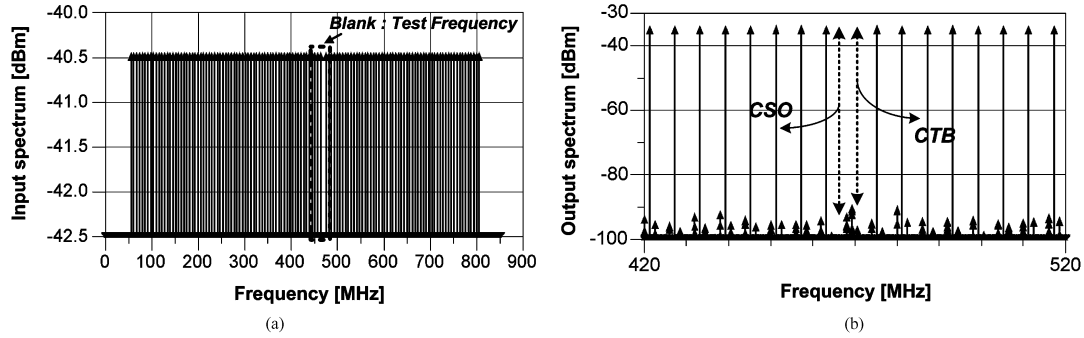


Fig. 22. Distortion simulation by multiple interferes: (a) input spectrum, (b) output spectrum.

VI. EXPERIMENTAL RESULTS

The proposed LNA was implemented in a $0.18 \mu\text{m}$ CMOS technology as part of a digital terrestrial and cable tuner receiver. Because the TV tuner receiver uses an internal 2.2 V regulator driven by external 3.3 V supply voltage, the LNA operates in regulated 2.2 V supply voltage. Fig. 15 shows the chip photograph of the proposed LNA. The characteristic impedance of the TV tuner is 75 ohms. However, all measurements are executed on single-ended 50 ohms termination for fair comparison with other published papers.

The power gain and input matching results of the proposed LNA are shown in Fig. 16. The input S11 parameter is less than -8 dB from 48 MHz to 880 MHz and the 3 dB gain bandwidth is above 1 GHz irrespective of the gain control. Fig. 17(a) shows calculated NF, simulated NF, and measured NF at high gain mode. At low frequency, the NF rises due to flicker noise and a high pass-filter composed of $C_1 - R_F$. The measured NF shows good agreement with the simulation result obtained after de-embedding the interconnection loss. It can be seen that the simulated NF is slightly larger than the calculated NF when N is 4 at 400 MHz by 0.4 dB, and this discrepancy results from the feedback resistor R_F that is not included in the calculated NF for a simple calculation. The measured NF according to gain control is shown in Fig. 17(b). The average NF from 48 MHz to 880 MHz is 3 dB at the high gain mode, 4.6 dB at the mid gain mode, and 1 dB at the low gain mode.

Two tone measurement for intermodulation distortion is carried out for the proposed LNA over the operating frequency. As shown in Fig. 18, 43–45 dBm of the IIP2 and 2.5–3.5 dBm of the IIP3 is obtained at high gain mode, respectively. The power measurement results of IMD2 and IMD3 versus the gate bias variation of M_{n1} and M_{p1} are presented in Fig. 19(a) and (b). Two-tone interferers at 200 MHz and 250 MHz with power of -22 dBm are used and the power of the IMD2 at 450 MHz and the IMD3 at 300 MHz are measured respectively. The power of IMD2 and IMD3 remains unchanged over sufficiently large gate bias windows while the gain and dc current are also almost unchanged. These results show that the windows are large enough for possible mass production. Fig. 20 shows the measured IIP3 and IIP2 according to gain control. Considering the gain flatness, the IIP3 is almost unchanged over the wide operating frequency. Although IIP2 is slightly degraded at low-end frequency band (60 MHz) where the source of M_{p2} is not sufficiently ac-grounded by MOS capacitor C_2 , the measured IIP2 still satisfies the target specification. The linearity performance over the spacing of applied tones at high gain mode is shown in Fig. 21. Since the sum of two tone is out of band, the difference of two tone is measured as the IMD2 component at 800 MHz. Both measured IIP3 and IIP2 do not show strong dependency on two tone space.

Fig. 22(a) shows the input spectrum of 137 multi-tones, each of which has a power of -40 dBm that is the allowed maximum input power level at mid gain mode in our system. 57 dBc of the

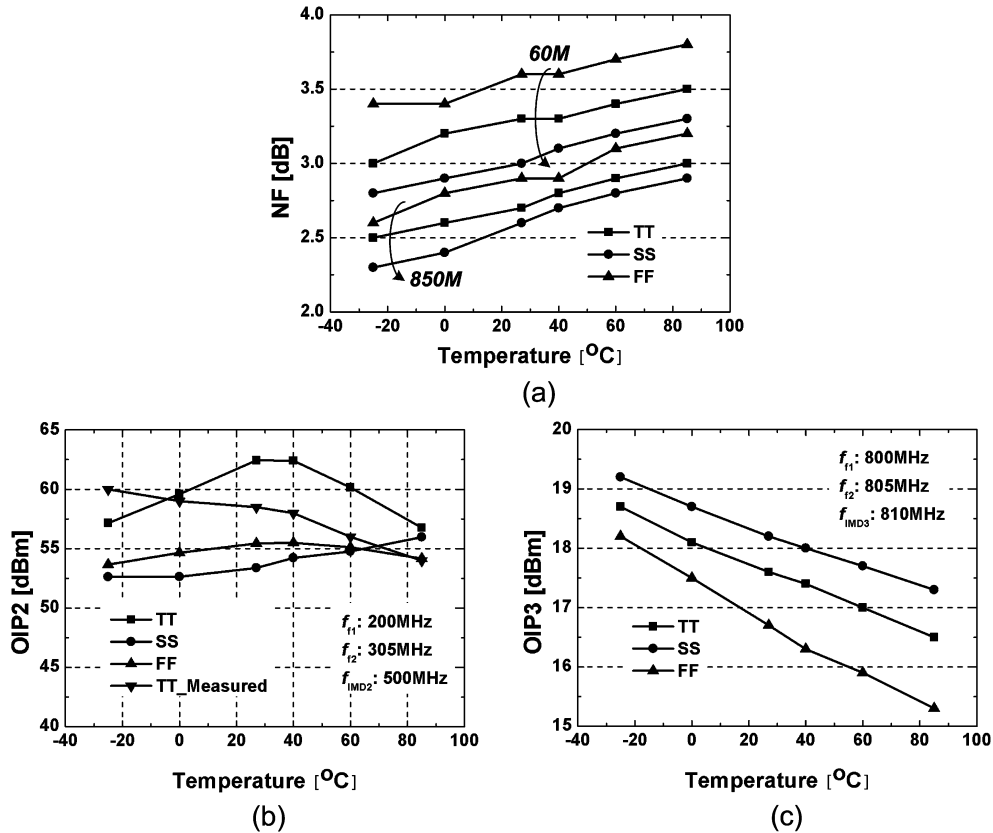


Fig. 23. Simulated performance of the proposed LNA over process and temperature variation: (a) NF, (b) OIP2, (c) OIP3. TT, SS, and FF stand for typical, slow, and fast model, respectively.

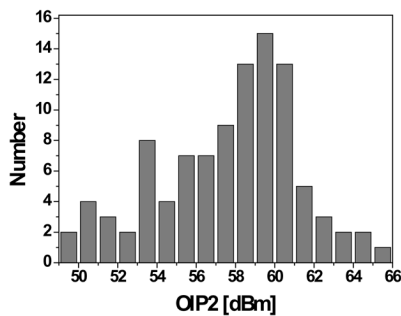


Fig. 24. Histogram of OIP2 performance according to process variation and mismatch effect through Monte Carlo simulation with 100 runs.

CTB and 59 dBc of the CSO are obtained in simulation as shown in Fig. 22(b). These show quite good agreement with the hand calculation by (1) and (2).

Fig. 23(a) shows the simulated NF performance of the proposed LNA for process and temperature variation at 60 MHz and 850 MHz. The degradation is below 0.5 dB compared to normal condition (TT and 27°C) over all variations. In case of OIP2, more than 50 dBm of OIP2 can be obtained in the LNA irrespective of process and temperature variation as shown in Fig. 23(b). The OIP3 is more than 15 dBm in all cases as shown in Fig. 23(c).

Fig. 24 shows the histogram of OIP2 performance according to process variation and mismatch effect through Monte Carlo simulation with 100 runs. Even in the presence of both process variation and mismatch effect, more than 50 dBm of OIP2 can be obtained.

Table I summarizes and compares the performances of the proposed LNA against those of other reported LNAs. Because the IIP2 performance of wideband single-ended LNAs is generally not reported, comparison of IIP2 performance is made with some previous works [8], [11], [17]. The OIP2 of the proposed LNA is at least 20 dB–30 dB higher than that of other conventional single-ended LNAs, thus showing IIP2 performance comparable to our previously published differential LNA. Therefore, in spite of the inherent single-ended structure, the proposed LNA can improve the IIP2 performance remarkably by IM2 distortion cancellation with low NF by noise canceling and high IIP3 based on current amplification.

VII. CONCLUSION

A wideband CMOS low noise amplifier with single-ended input and output employing noise and IM2 distortion cancellation for application to a digital terrestrial and cable TV tuner is proposed. By adopting thermal noise canceling based on current amplification, the proposed LNA can acquire low NF and high IIP3 performance without degradation of gain and wideband input matching. Furthermore, by employing an IMD2 cancellation technique to utilize the complementary characteristics of NMOS and PMOS, the proposed LNA greatly improves IIP2 performance while retaining the thermal noise canceling. In spite of its single-ended structure, the IIP2 of the LNA is compatible to that of a differential LNA. Therefore, the proposed LNA is suitable for a first stage of broadband RF receivers that

TABLE I
MEASURED PERFORMANCE SUMMARIES OF PROPOSED WIDEBAND LNA AND COMPARISON TO PREVIOUSLY REPORTED WIDEBAND LNAs

| | [11] | [17] | [18] | [19] | Our previous work [8] | This Work |
|-------------------------------------|-------------------|-------------|------------------------------------|-------------------|-----------------------|---------------------------------|
| Power Gain [dB] | 13.7 | 18 | 12 | 20.3 | 15.5 | 14 |
| 3-dB Bw [MHz] | 2~1600 | 48~1000 | 100~1600 | 40~900 | 54~880 | 48~1200 |
| NF @ Max Gain [dB] | 2 | 1.2 | 2.6 | 4 | 3.6 | 3 |
| IIP3 @ Max Gain [dBm] | 0 | 8 | 16 (cf. two-tone space: 170MHz) | -11 | -1 | 3 (cf. two-tone space: 5MHz) |
| IIP2 @ Max Gain [dBm] | 12 | 17 | - | - | 38 | 44 |
| OIP3 [dBm] | 13.7 | 26 | 28 | 9.3 | 14.5 | 17 |
| OIP2 [dBm] | 25.7 | 35 | - | - | 53.5 | 58 |
| S11 [dB] | -8 | -10 | -10 | -8.5 | -10 | -9 |
| Differential | No | No | No | YES | YES | No |
| Area [mm ²] (Core Only) | 0.075 | - | 0.1 | 0.57 | 0.54 | 0.16 |
| Technology | 0.25 μ m CMOS | GaAs MMIC | 0.13 μ m CMOS | 0.18 μ m CMOS | 0.18 μ m CMOS | 0.18 μ m CMOS |
| Power consumption | 14mA @ 2.5V | 30mA @ 3.3V | 11.6mA @ 1.5V | 24mA @ 1.8V | 28mA @ 1.8V | 15.8mA @ 2.2V |

require high sensitivity and linearity without requiring external bulky passive transformers.

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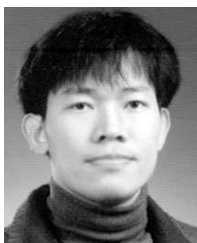


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