Wideband Microwave OTA with Tunable Transconductance using Feedforward Regulation and an Active Inductor Load

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***Abstract*—A microwave operational transconductance amplifier (OTA) using a feedforward-regulated cascode stage and an active inductor load is proposed in this paper. The use of the feedforward regulation mechanism provides the OTA with a very wide bandwidth of over 10 GHz while the active inductor load contributes to the OTA's high linearity performance. Furthermore, the transconductance of the OTA can be varied from 5 mS to 20 mS. The OTA is designed in a 0.13-μm CMOS technology and it draws a maximum of 5.5 mA of current from a 1.2 V supply when the transconductance is set to 20 mS. The chip occupies an area of 110 μm $\times$ 66 μm.**

I. Introduction

The operational transconductance amplifier (OTA) is an important building block in anolog circuit design and it has been extensively used in low frequency applications [1]–[6]. However, with the rapid development of wireless communications, microwave OTAs that can operate in the gigahertz range are of high interest.

High-frequency OTAs have recently been used in RF filters [7], oscillators [8] and phase shifters [9], for example. While CMOS technology continues to move toward smaller feature sizes and power supply voltages are reduced, this can compromise the linearity performance of circuits using those deep-submicron devices. Yet, circuits used in RF applications still have to meet their specifications, some of which demand high linearity performance.

One approach to addressing linearity requirements in RF and microwave circuits is to process signals in the current domain. Since the inner nodes in current-mode circuits usually have low impedance, the voltage gain is small and the poles associated with these nodes are located at high frequency, which result in high linearity and high operating frequency, respectively.

In this paper, a high linearity, wideband OTA with tunable transconductance is presented. The circuit uses the principle of feedforward regulation [10] which leads to its wideband performance. An active inductor load is used to enhance the linearity of the OTA. The operating principles and circuit implementation of the OTA are discussed in Section II. The performance results are presented in Section III and Section IV concludes the paper.

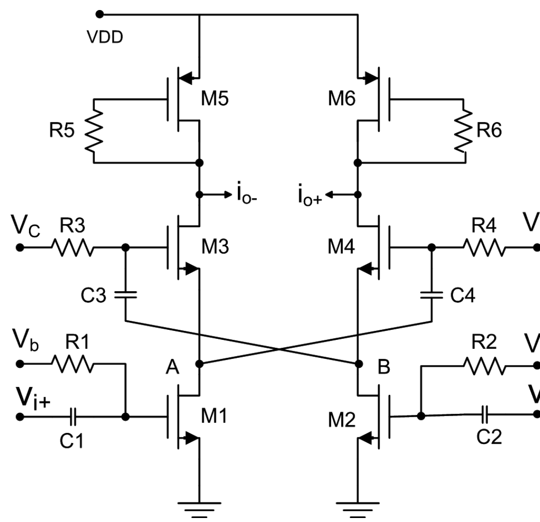

Fig. 1. The proposed OTA

II. Circuit Implementation

The schematic of the proposed OTA is shown in Fig. 1 and it is a variant of the amplifier designs presented in [10] [11]. The OTA in Fig.(1) is a low-power, high-speed circuit that uses an active inductor load. The feedforward-regulated cascode topology makes it more appropriate for the low supply voltage while retaining the characteristics of high linearity and high speed performance [1]–[6]. A folded active inductor topology is employed here to improve the high frequency performance [12]. The input transconductor stage is formed by the transistors M1 and M2 operating in the triode region. M3 and M4 are utilized as the regulated cascode stage to enhance the linearity and to boost the output impedance. The gates of transistors M3 and M4 are biased with the same voltage, V_c. The transconductance, G_m, of the OTA can be changed by varying the value of V_c. The folded active inductor, consisting of M5, M6, R5 and R6, is used as the active load to allow the OTA to work at high frequencies.

A. Input Transconductor

Using a short channel model to characterize the drain current to the gate-to-source voltage of a MOS transistor in

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triode region, its transconductance can be expressed as follows:

$$G_m = \mu_n C_{ox} \frac{W}{L} V_{DS} \cdot \frac{1}{1 + (V_{DS}/E_{sat}L)} \tag{1}$$

where W and L is the width and length of the transistor, respectively; μ is the channel mobility; C_{ox} is the oxide capacitance; E_{sat} is the saturated electrical field. As we can see in Eq:(1), the transconductance of the triode transistor is only related to V_{DS} assuming all the physics-related parameters are constant. So if we can keep the V_{DS} constant as the amplitude of the input small signal varies, then the high linearity will be obtained. However, the channel mobility μ is actually not constant. It is related to gate voltage due to mobility reduction effect. The mobility compensation technique to solve this problem has been specified in detail in [4], [5].

B. Feedforward-Regulated Cascode Stage

The feedforward regulation is implemented by cross coupling between the differential cascode pairs M1/M3 and M2/M4. When the differential input signal increases, the drain voltage of M1 decreases while the drain voltage of M2 increases. The signal at the drain of M2 is then cross coupled to the gate of the cascode transistor M3. As V_{GS3} increases, M3 inversely injects more current into M1, which then elevates V_A and counteracts any change of V_A. Therefore, by creating an inter-locking regulation mechanism between V_A and V_B, we get a stable DC voltage at the drains of M1 and M2. Besides, The cascoding transistor M3 also increases the total output impedance, which is beneficial for the OTA.

In [13], this function is performed by a feedback amplifier with gain of -A directly from node A to the gate of M3. Higher gain means higher accuracy, but feedback and high-gain amplifier both result in low operating frequency. In this paper, feedforward instead of feedback is employed by using just two capacitors to cross-couple the signal from nodes B and A. The variation at node B will be directly fed forward to the gate of M1 and then compensate the variation at node A instantly, which is simpler and faster. At the same time, the RF signals at nodes A and B are further amplified by the capacitative cross coupled cascode transistors M3 and M4. They perform small signal amplification with the same sign as the RF input signals, and together with M1 and M2 they are added to increase the total transconductance.

C. Active Inductor Load

The active load consisting of M5 and R5 is actually the PMOS version of the folded active inductor proposed in [12]. To derive the parameters of the RLC equivalent circuit of the folded active inductors, its small-signal equivalent circuit is shown in Fig. 2. To simplify analysis, we neglect C_{gd}, g_o along with other parasitic capacitances of the transistor. It can be shown that the input impedance is given by :

$$Z_{in} = \frac{sRC_{gs} + 1}{sC_{gs} + g_m} \tag{2}$$

It becomes evident that Z_{in} has a zero at the frequency $\omega_z = \frac{1}{RC_{gs}}$ and a pole at $\omega_p = \frac{g_m}{C_{gs}}$. The active load is resistive at

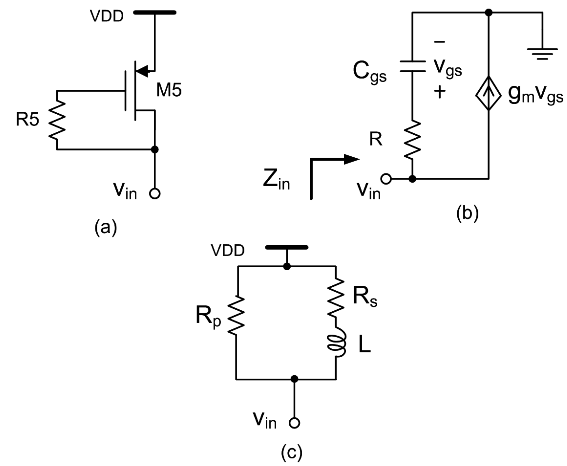

Fig. 2. The Active Inductor Load

low frequencies $\omega < \omega_z$ with resistance $R \approx \frac{1}{g_m}$ and inductive when $\omega_z < \omega < \omega_p$. To derive the RLC equivalent circuit for the network in Fig. 2b, we examine the input admittance of the circuit,

$$Y_{in} = \frac{1}{R} + \frac{1}{s\frac{RC_{gs}}{g_m - \frac{1}{R}} + \frac{1}{g_m - \frac{1}{R}}} \tag{3}$$

Eqn. (3) can be represented by a series RL network in parallel with a resistor R_p as shown in Fig. 2c, and the component values are given by,

$$\begin{aligned} R_p &= R \\ L &= \frac{RC_{gs}}{g_m - \frac{1}{R}} \\ R_s &= \frac{1}{g_m - \frac{1}{R}} \end{aligned} \tag{4}$$

It becomes evident that $g_m > \frac{1}{R}$ is required in order to have $L > 0$ and $R_s > 0$.

From a different perspective, the resistive self-biasing active load also has the function of common mode feedback. The operating mechanism is as follows. Assuming a reduction in the DC voltage appears at the output node, this will result in an equal increase in $|V_{GS5}|$. The PMOS device M5 will inject more DC current into the NMOS devices M3 and M1. To accommodate the variation of the current, M1 and M3 have to increase their drain voltage to compensate the reduction of DC voltage at the output node. This is another benefit of the resistive self-biasing active load, which removes the high-gain amplifier used in common-mode feedback circuits, for instance.

III. Simulation Results

The small-signal transconductance of the OTA as a function of frequency and control voltage, V_c, is shown in Fig. 3. The proposed OTA works well up to and beyond 10 GHz. To further investigate the behavior of the OTA's transconductance versus V_c, the process corner simulation including three sigma variations was carried out at a fixed frequency of 5.4 GHz and the results are shown in Fig. 4. TT,FF,FS,SF and SS represent TypicalNMOS/TypicalPMOS, FastN/FastP, FastN/SlowP,

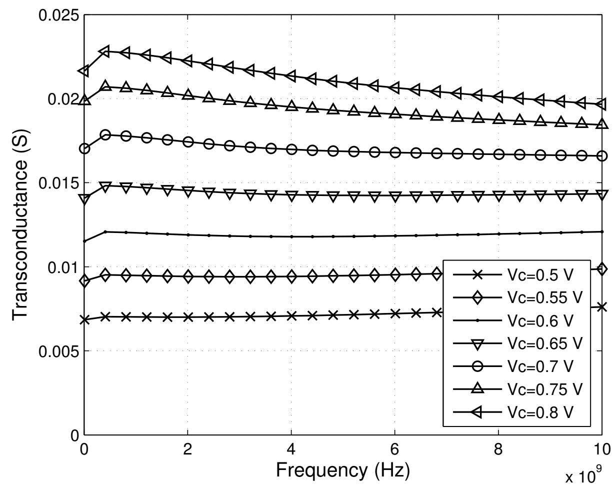

Fig. 3. Simulated OTA transconductance versus frequency and control voltage V_c

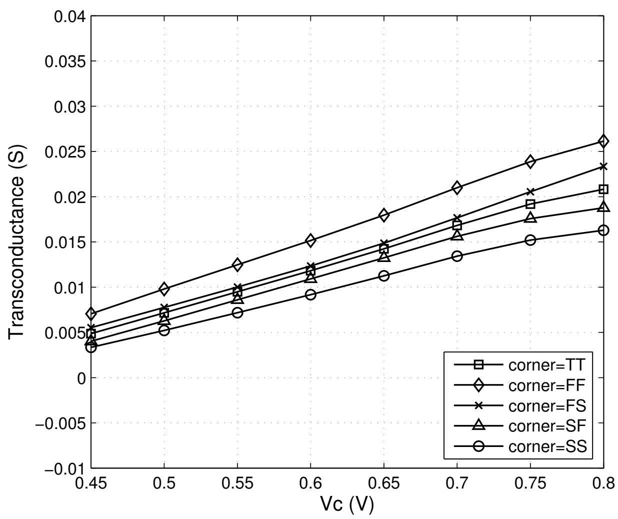

Fig. 4. Simulated OTA transconductance versus the control voltage V_c at 5.4 GHz

SlowN/FastP and SlowN/SlowP, respectively. We observe that the transconductance has a very linear response versus bias control voltage and it can be tuned from 5 mS to 20 mS as V_c varies from 0.45 V to 0.8 V for TT corner situation.

As a transconductance amplifier that translates the input voltage signal into output current , the OTA is usually loaded with low impedance such as the switches of the passive mixer or transimpedance amplifier (TIA) stage. In the following linearity simulation, the OTA is loaded with 50 Ω resistor, which will lower the G_m somewhat compared with the ideal case of a short-circuit load.

As the transconductance is the most important parameter of the OTA, we use the G_m varying along with the magnitude of the input voltage signal V_{in} to demonstrate the linearity performance. A differential voltage signal with magnitude of V_{in} is applied to the input port and the transconductance is calculated from the output current. G_m compresses as V_{in}

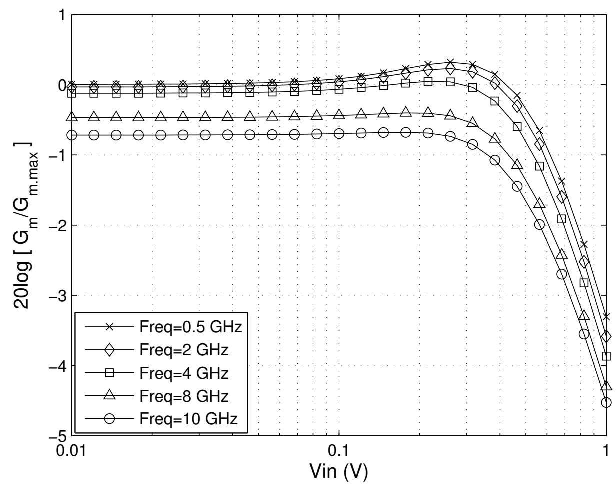

Fig. 5. Simulated OTA transconductance versus input voltage at V_c=0.7 V

increases. Fig.(5) shows the normalized transconductance of the OTA versus the magnitude of the RF input signal at $V_c = 0.7V$. $G_{m.max}$ is the transconductance obtained under condition of small input signal and low frequency, such as V_{in}=10 mV and Freq= 500 MHz.All the simulated data are normalized with respect to $G_{m.max}$ and re-represented in $20log(\frac{G_m}{G_{m.max}})$. From 500 MHz to 10 GHz, G_m decreases by 0.73 dB, which equals to a reduction of 8.8%. The 1dB compression point for G_m is V_{in} =0.5 V at 10 GHz and 0.6 V at 500 MHz. In a 50 Ω system, they correspond to +4 dBm and +5.6 dBm, respectively.

When the OTA is tuned to get different transconductance, the linearity performance changes due to the variation of the DC operating points. Fig.(6) shows the input 1 dB compression point for G_m at different tuning voltage. This figure is simulated at the frequency of 2.4 GHz. The highest input 1dB compression point for G_m is $V_{in} = 0.628V$ when the tuning voltage is set to $V_c = 0.65V$. The minimum point is $V_{in} = 0.362V$ at $V_c = 0.8V$. At the lower end of the tuning range, the linearity degrades due to the decrease of the overdrive voltage of the regulated cascoding transistor, while at the higher end of the tuning range, the linearity degrades due to transformation of the operation region of the input transistors gradually from triode to saturation .

The layout of the proposed OTA is shown in Fig. 7. The circuit occupies an area of 110 μm $\times$ 66 μm and consumes 5.5 mA current from a 1.2 V supply, or 6.6 mW, when it is biased to produce 22 mS of transconductance. The power consumption drops to 1.56 mW when the OTA is biased to produce 5 mS of transconductance.

IV. CONCLUSIONS

In this paper, a new high-speed high-linearity low-power microwave OTA has been presented that uses a feedforward regulated cascode topology with active inductor load. It can operate properly up to 10 GHz with only 0.73 dB Gm reduction. The input 1dB compression point for G_m reaches

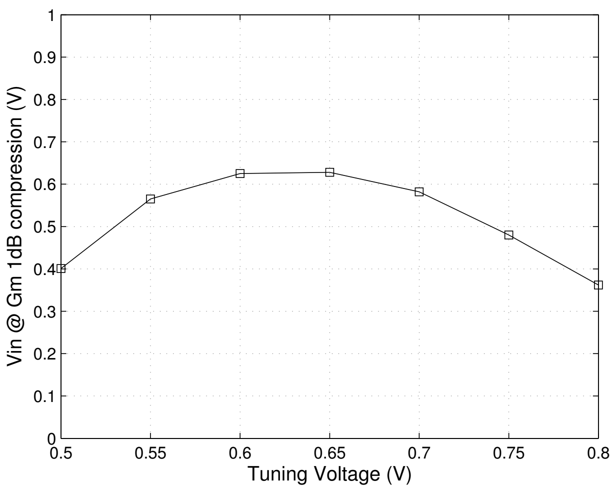

Fig. 6. Simulated OTA linearity versus the control voltage V_c at 2.4 GHz

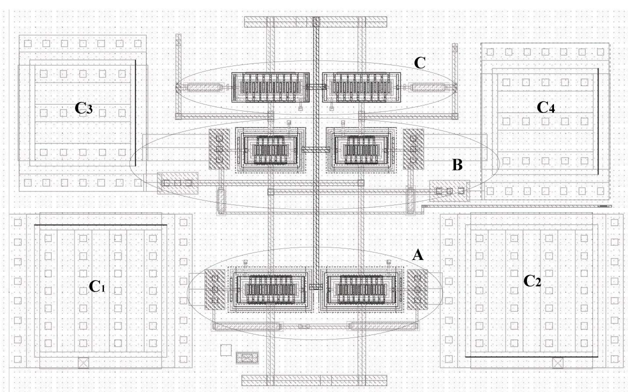

Fig. 7. Layout of the OTA:Input Transconductor (A);Feedforword-Regulated Cascode stage (B);Active Inductor Load (C)

$V_{in} = 0.628V$ at $V_c = 0.65V$. It draws a maximum DC current of 5.5 mA from a single 1.2 V supply and occupies an area of only 110 μm $\times$ 66 μm. Furthermore, the transconductance of the OTA can be tuned in a very linear fashion versus an applied control voltage. This OTA is suitable for various RF and microwave applications up to and above 10 GHz.

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