A MICROPOWER CONTINUOUS-TIME CMOS OTA FILTER OPERATING IN SUBTHRESHOLD REGION

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Abstract

A CMOS OTA designed to operate in the subthreshold (weak inversion) region for continuous–time filtering application is presented. Power consumption of a second–order filter designed using less than 1 uW at 2 kHz. A large tunning capability in the audio–frequency range was obtained. Advantages of weak inversion operation are described, and experimental results are presented.

Introduction

In recent years much effort has been made to reduce the power consumption of active elements, such as the operational amplifier (OA) or the operational transconductance amplifier (OTA).[1,2,3] The need for micropower systems has been extended to many battery powered medical system such as hearing aids, pace-makers and various types of small size portable instruments. Most of those applications require a low voltage power supply and low current consumption, which historically lead to serious performance limitations of the active devices (noise and dynamic range). A CMOS OTA operating in the weak inversion region intended for micropower applications is introduced and applied in a second–order filter.

Figure 1 shows the I–V characteristic of the MOSFET, which can be devided into three regions; subthreshold (region I), transition (region II) and the strong inversion region (region III). It is well known that the drain current is exponentially proportional to the gate voltage when the gate to source voltage (V_{gs}) is below threshold voltage(V_{th}), while drain current has the usual parabolic transfer characteristic in strong inversion region (region III)[4]. In the transition region, no accurate simulation model is available at the present time[5,6].

The Subthreshold Region

The characteristics of each region can be explored by considering the voltage gain and 3–dB bandwidth of the basic inverter as a function of drain current. Figure 2 shows the basic inverter with an active load. Characteristics of the maximum gain, required power supply(V_{dd}) and power consumption per unit gain at operating points are presented in Figure 3. In these plots, V_{dd} was adjusted to achieve maximum gain at each drain current. Device sizes remained fixed and are indicated in the Figure 2. The 3–dB frequency of the inverter and the power per unit gain are shown in Figure 4 with V_{dd} set for maximum gain as before. Trade-offs between frequency range and power consumption must be made. For low frequency and low power applications 10nA–1uA range are chose in where the input transistor M1 is in the weak inversion and load M2 is in the strong inversion region(see figure 1).

Consider now the transconductance gain of the MOSFET. In the weak inversion region, the drain current is given by the expression

$$I_{drain} = I_{do}exp((V_{gs} - V_{th})/nV_T)(1 - exp(-V_{ds}/V_T) \quad (1)$$

It follows tahat the transconductance gain is proportional to bias current as indicated in (2).

$$g_m = \frac{\partial I_{drain}}{\partial V_{gs}} = 1/nV_T I_d \quad (2)$$

In the strong inversion saturation region

$$I_d = K'(V_{gs} - V_{th})^2 \quad (3)$$

which results in a quadratic relationship between I_d and g_m as indicated in (4)

$$g_m = 2\sqrt{K' I_d} \quad (4)$$

In the MOSFET–based filter structures appearing in the literature, all devices are generally biased to operate in the strong inversion saturation region. Because of the quadratic dependence of I_d in this region, the practical adjustment range of g_m is considerably less than is attainable when operating in weak inversion which is characterized by a linear dependence on g_m. This paper describes a weak inversion OTA and its application in a a second–order OTA–based.

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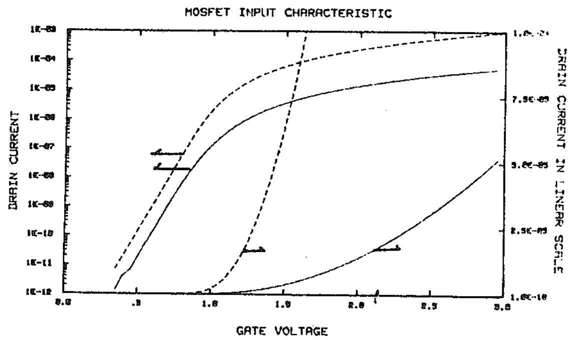


Fig. 1. I-V characteristic of n-channel MOSFET.

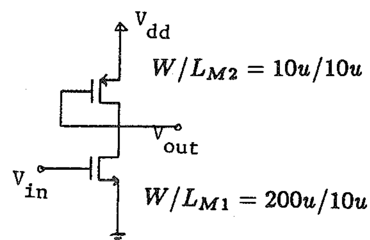


Fig. 2. Inverter with active load.

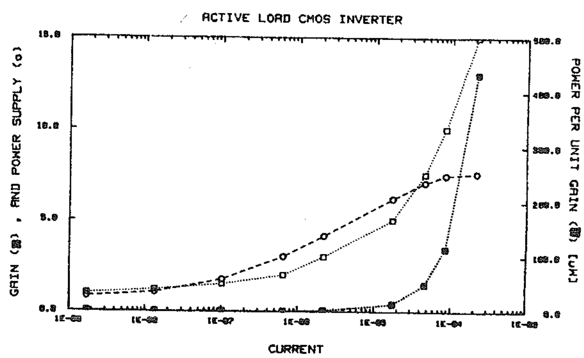


Fig. 3. Characteristics of Inverter.

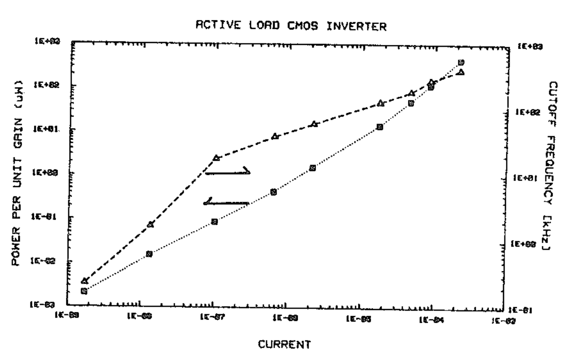


Fig. 4. Frequency characteristics of Inverter.

Operational Trance. Amplifier

The transconductance element is very useful in voltage controlled filters, amplifiers, and nonlinear circuits [7]. Very simple filter structures containing only OTA's and capacitor are practical to implement. High frequency operation and convenient post–fabrication tuning capability are readily abtainable. A symbol of an ideal OTA is shown in Fig. 5. An ideal OTA is characterized by the expression

$$I_{out} = G_m(V_1 - V_2) \tag{5}$$

where G_m is the transconductance gain. It is generally a goal of the OTA designer to make G_m dependent upon an external dc voltage or current to provide for conveniently tuning OTA. A schematic of a basic OTA is shown in Fig. 6 [2]. It can be shown that the output impedance of this circuit is very high and that the input/output relation ships are characterized by (5), for either weak inversion or strong inversion saturation operation of M1 and M2. This CMOS OTA has been considered as a voltage amplifier[2]. Milkovic has investigated a similar structure [9]. Krammanecker[2] used this structure to achieve a high voltage gain in which input pairs are biased to operate in weak inversion region for maximum voltage gain. Hostica [8] applied dynamic biasing scheme to this simple CMOS OTA structure to reduce power consumption in switched–capacitor filters. Milkovic [9] exploited the high units gain bandwidth capabilities. The applicability of this structure as a voltage amplifier is limited due to the high output impedance characteristics. Emphasis here is on applications which exploite the high output impedance of the basic OTA structure.

The Weak–Inversion OTA

It can be shown from (1) and (2) that the differential drain currents of the source coupled pair of Fig. 6 are related to the differential input voltage, V_i=V_1-V_2, by the expression

$$I_{d1} - I_{d2} = K' V_i \sqrt{\frac{I_{ss}}{K'} - V_i^2} \tag{6}$$

for M1 and M2 in strong inversion saturation region and by the expression

$$I_{d1} - I_{d2} = I_{ss} tanh \frac{V_i}{2V_T} \tag{7}$$

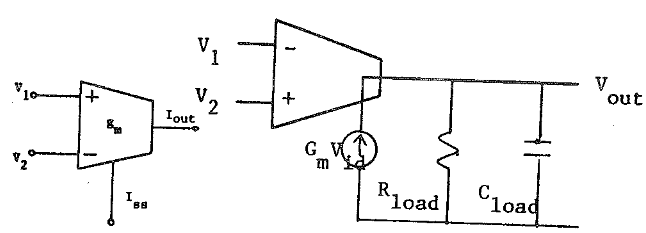


Fig. 5. (1) Symbol of OTA and (2) macromodel

for M1 and M2 operating in weak inversion. For small inputs, (6) and (7) can be approximated by

$$I_{d1} - I_{d2} = \sqrt{K' I_{ss} V_i} \tag{8}$$

for $V_i < 2V_T$ and $V_i < \sqrt{\frac{I_{ss}}{K'}}$.

$$I_{d1} - I_{d2} = \frac{I_{ss} V_i}{2V_T} \tag{9}$$

for $V_i < 2V_T$. The additional constraint that $V_i < 2V_T$ which is necessary to obtain the linear relationship of (8) places a practical lower bound on I_{ss} and correspondingly upon power dissipation. This I_{ss} dependence is not required for the weak inversion operation region. This is an important consequence because one practical way to reduce the power consumptions is to decrease I_{ss}. The low power dissipation attainable in subthreshold is not, however, obtained without penalty. The matching characteristics of 24 equally sized MOSFETs are compared in Fig. 7. It can be seen that the standard deviation of the drain currents increase as the gate–to–source voltage V_{gs} decreases. Technically this deviation characteristic can be reduced by careful layout, but it is an important factor in the overall characteristics of low power OTAs.

Frequency responce is one of the important factors in designing the active circuit. As seen in Figure 6, the output node is the only high impedance node in the whole OTA structure. In designing a filter using OTA, the load capacitor should be sufficiently large to have dominant pole at the output node. Figure 5.2 shows a macromodel of the cascode OTA of Figure 6. A_{IM} denotes the gain of the current mirror, $A_{IM} = I_o/(I_{d2} - I_{d1})$, and g_m is the transconductance gain of the differential input pair as given by (2) and (4). The dominant pole p_1 is located at the frequency

$$\omega_1 = \frac{1}{R_{out} C_{load}} \tag{10}$$

and the unity gain bandwidth GBW is given by

$$GBW = \frac{g_m A_{IM}}{C_{load}} \tag{11}$$

From equation(11), the unit gain bandwidth is directly proportional to the g_m of the input devices, when the load capacitance is large enough to neglect the parasitic capacitances. In audio frequency application, small g_m is required since capacitor is bounded by the silicon area. By sizing the widths of the input transistors 20 times larger than that of the diode connected transistors, the input transistors are forced to operate in the subthreshold region where the g_m/I ratio reaches its maximum value. When input devices in the circuit of Figure 6 are biased in this region, large g_m/I and wide g_m range are available with a small change of biasing current. In figure 8, the output current characteristics are shown at three different biasing currents.

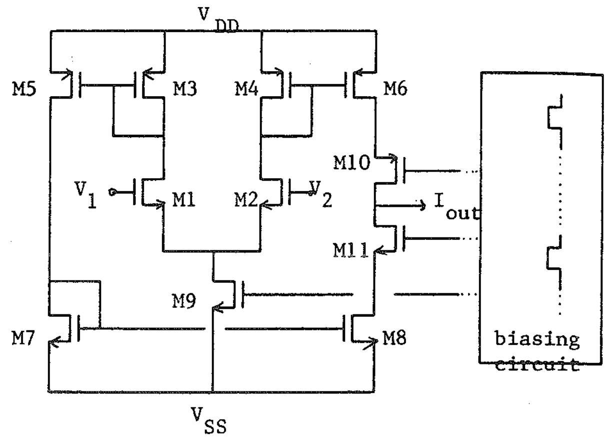


Fig. 6. CMOS OTA circuit schematic diagram.

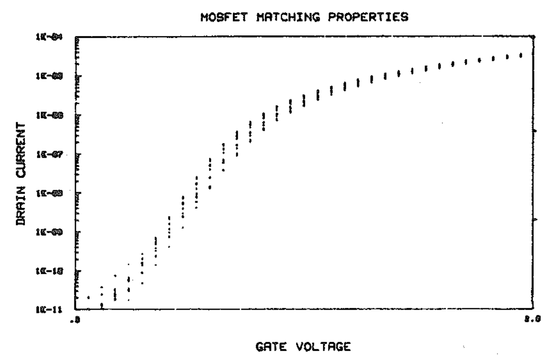


Fig. 7. Standard deviation of drain current.

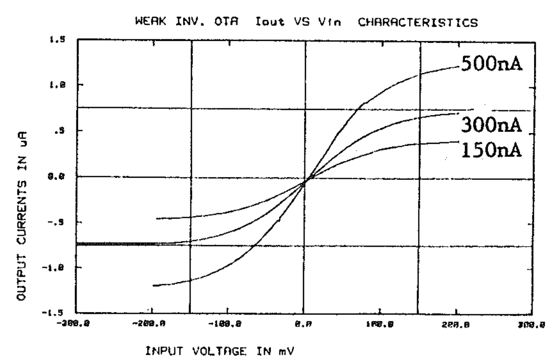
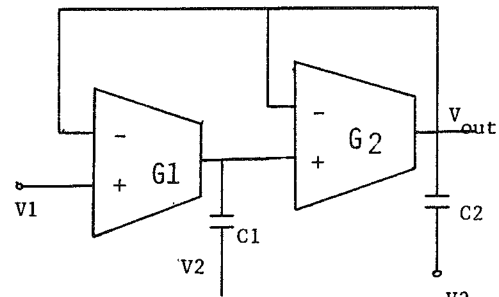


Fig. 8. Output currents of CMOS OTA.

It becomes obvious that the weak inversion OTA has its advantage in low frequency applications since g_m is small and its tunnable G_m range is much wider than that of the conventional strong inversion OTA. Offset voltages are also dependent bias currents, but these variations are only a few mV. Since its output stage is cascode connected, the output conductance of this stage is reduced and its voltage gain can be enhanced up to 70dB. A filter of figure 9 was designed to operate from 300 Hz to 20 kHz and fabricated where M1,M2 and M12 are in seperate wells. Figures 10 shows the band



$$V_{out} = \frac{C1\ C2\ V3\ s^2 + C1\ G2\ V2\ s + G1\ G2\ V1}{C1\ C2\ s^2 + C1\ G2\ s + G1\ G2}$$

$$w_o = G1/\ C1\ C2 \qquad Q = C1/C2$$

Fig. 9. Second–order filter.

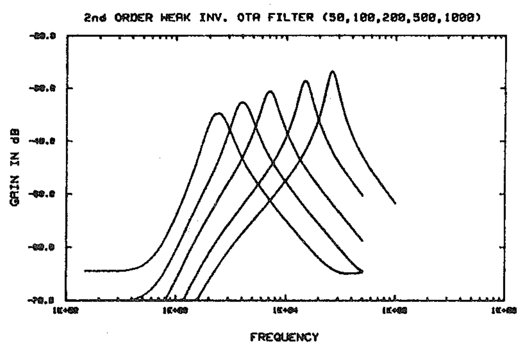


Fig. 10. Band–pass response of Fig. (9).

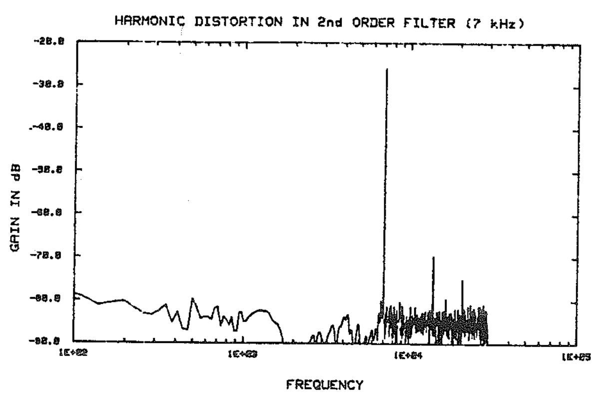


Fig. 11. Distortion spectrum for 750 mv at 7 kHz.

pass responses at different biasing currents. The distortion characteristics of the filter were tested. Spectrum analyzer output measured at 7 kHz is shown in Figure 11. In this case, maximum input to maintain total harmonic distortion within 1% was 75 mV.

Conclusion

The design and characteristics of a weak inversion operational transconductance amplifier has been presented. A fabricated continuous– time filter shows good agreement with the theory over the 200–20 kHz range, and is capable of handling the 75 mv signal with 1% THD using $\pm$ 1.5 V power supllies. The linear range and GBW are large enough for low frequency applications with only 5uw power consumption which can be reduced down to the several hundreds nw.

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