A CMOS rail-to-rail linear VI-converter

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Abstract: A linear CMOS VI-converter operating in strong inversion with a common-mode input range from the negative to the positive supply rail is presented. The circuit consists of three linear VI-converters based on the difference of squares principle [1-3]. Two of these perform the actual V to I conversion, while the third changes the bias currents of the first two in response to changes in the input common-mode level.

The resulting circuit has a large signal transconductance which is constant to within 3% over the entire common-mode input range. It can operate from a single supply voltage of 2.2Volts.

I. Introduction

The common-mode input range of a single (N-type) transconductor reaches from a certain voltage level above the negative supply rail up to a certain level above the positive supply rail. In order to cover the entire rail-to-rail common-mode input range, an N-type VI-converter and a complementary P-type have to be driven in parallel. A special circuit is needed to combine the output currents of the individual VI-converters. The combined large signal

transconductance can be made independent of the common-mode input level by biasing the N-type and P-type VI-converters with the individual output currents of a third VI-converter after current limiting. This third VI-converter is single-ended and driven by the common-mode input voltage. Figure 1 shows the structure of the proposed VI-converter circuit.

II. Principle

The individual output currents of a linear VI-converter based on the difference of squares principle can be described as:

$$I_{d1} = \frac{K}{2}(V_{dc} + V_{in})^2 \text{ and } I_{d2} = \frac{K}{2}(V_{dc} - V_{in})^2 \quad (1)$$

$$\text{with } K = \mu C_{ox} \frac{W}{L}$$

and therefore:

$$I_{d1} - I_{d2} = g_m V_{in} \text{ with } g_m = 2K V_{dc} \quad (2)$$

If the DC-voltage source V_{dc} is formed by the V_{gs} of a MOSFET with a constant drain current I_{dc} the large signal transconductance value of the VI-converter is proportional to $\sqrt{KI_{dc}}$. If the

Fig. 1 Structure of the VI-converter circuit.

total g_m of the circuit of figure 1 has to be constant, then the condition:

$$g_{mN} + g_{mP} = \text{constant} \quad (3)$$

has to be satisfied. Since g_m is proportional to $\sqrt{KI_{dc}}$ and assuming $K_N = K_P$ equation 3 can be rewritten as:

$$\sqrt{I_{dcN}} + \sqrt{I_{dcP}} = \text{constant} \quad (4)$$

It now happens that the currents of equation 1 satisfy this condition:

$$\sqrt{I_{d1}} + \sqrt{I_{d2}} = 2\sqrt{\frac{K}{2}} V_{dc} \quad (5)$$

if V_{dc} is constant. This means that these currents can be provided by a similar VI-converter.

If one of both currents becomes zero, the other one has to be limited to a constant value in order to keep the total transconductance constant, this explains the need for the current limiters in figure 1.

III. Circuit description

A possible realisation of a transconductor based on the difference of squares principle is the circuit shown in figure 2. Here the DC-voltage sources are formed by MOSFET's M3 and M4. Their drain currents are kept constant by the current sources M5 and M6 and the feedback loops formed by M7, M9 and M8, M10. The transistors M7 and M8 disconnect the drains of M3 and M4 from the gates of M9 and M10 and so enlarge the common-mode input range. This transconductor has been selected for its low minimal supply voltage and its efficient use of current.

Fig. 2 Schematic of a square-law transconductor

After some modification the transconductor of figure 2 can also be used as the biasing circuit. In

figure 3, one half of the circuit of the transconductor of figure 2 is shown together with a circuit that limits its output current to a value of $4I_{dc}$. The current limiting is based on the principle that in a series connection of a current sink and a current source the lowest current always dominates [4]. The transistor that wants to carry the larger current is forced into the triode region in order to carry the lower current. In figure 3 the output of the current mirror formed by M11-M14 act as the current source and is connected in series with the drain of transistor M1 which acts as the current sink. A mirrored copy of the drain currents of M1, M13 and M14 is available at the drain of transistor M15.

Fig. 3 Part of the transconductor with current limiter.

A combination of the circuit of figure 3 and its complementary version can be used to bias the two linear transconductors as is shown in figure 4. The complementary version of the circuit of figure 3 is used in order to eliminate the need for one of the current mirrors in figure 1. The input voltages for the biasing circuit will be a reference voltage (normally $V_{dd}/2$) and the common-mode input voltage.

The transconductor of figure 2 and its complementary version are used for the transconductors of figure 1. The transconductor with the N-type input transistors (like the one in figure 2) is biased as a function of the common-mode input level by using the circuit of figure 3. The complementary version of the transconductor (with P-type input transistors) is biased by a complementary version of the circuit of figure 3. This results in the complete linear transconductor with a rail-to-rail common-mode input range, shown in figure 4.

Fig. 4 Complete circuit of the linear VI-converter.

IV. Simulations:

Several PSPICE simulations have been performed with the circuit of figure 4, using the parameters of a 2.5 μ m CMOS process and a supply-voltage of 3 Volts. The value of I_{dc} was 8 μ A. The length of the transistors used for the actual VI-converters equals 10 μ m. The width of the PMOS transistors was taken 3.2 times the width of the NMOS transistors in an attempt to make K_P equal K_N . Under these conditions a

small-signal bandwidth of 4.8 MHz can be achieved. The adaptive biasing circuit reduces the variation of the transconductance from 33% to less than 3%. Using an I_{dc} of 2 μ A the minimal supply voltage is 2.2 Volts.

Table 1 gives an overview of several performance characteristics of the presented circuit. The figures are given for a nominal value for V_{common} which is equal to $V_{dd}/2$ and the value over the entire common-mode input range (V_{common} varies between 0 to V_{dd}).

	nominal value	over entire common-mode range	
DC-transconductance	116	110 - 116	μ A/V
Small signal bandwidth (-3dB point)	4.8	4.8 - 9.0	MHz
Equivalent input noise	47	45 - 48	nV/ $\sqrt{\text{Hz}}$
THD differential input voltage ($V_{in}=400\text{mV}_{pp}$)	0.14	0.14 - 1.4	%
THD single-ended input voltage ($V_{in}=400\text{mV}_{pp}$)	0.4	0.04 - 1	%
CMRR $V_{common} = V_{common\ dc} + V_{ac}(100\text{mV}_{tt}, 100\text{Hz})$	58	26 - 58	dB
PSRR $V_{dd} = 3V_{dc} + V_{ac}(100\text{mV}_{tt}, 100\text{Hz})$	48	25 - 49	dB
Supply current	300	300 - 400	μ A
Conditions: $V_{dd} = 3\text{V}$, $I_{dc} = 8\mu\text{A}$, $V_{in} = 100\text{mV}_{pp}$, 1kHz; unless stated otherwise			

Table 1 Simulated performance characteristics of the transconductor.

V. Conclusions

A low-voltage linear CMOS transconductor with a rail-to-rail common-mode input range and a large signal transconductance which is independent of the common-mode input voltage has been presented. The circuit can be used for differential signals with varying common-mode levels or single-ended signals. In both cases the transconductance value will remain constant within 3%. It can operate on a supply voltage of 2.2 Volts or higher.

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