# DC to 8 GHz 11 dB gain Gilbert micromixer using GaInP/GaAs HBT technology

C.C. Meng, S.S. Lu, M.H. Chiang and H.C. Chen

A wideband GaInP/GaAs heterojunction bipolar transistor (HBT) micromixer from DC to 8 GHz with 11 dB single-ended conversion gain is demonstrated. The input return loss is better than 10 dB for frequencies up to 9 GHz. The single-to-differential input stage in a Gilbert micromixer renders good wideband frequency response and eliminates the need for common-mode rejection. $IP_{1dB} = -17$ dBm and $IIP_3 = -7$ dBm are achieved for a small local oscillator power of $-2$ dBm when $LO = 5.35$ GHz and $RF = 5.7$ GHz.

*Introduction:* The micromixer proposed by Gilbert [1] is the ideal circuit for an RF high-frequency wideband mixer. Fig. 1 is a circuit schematic diagram of the designed GaInP/GaAs HBT micromixer. The biased current source in a conventional Gilbert mixer contributes noise and deteriorates rapidly the common-mode rejection ratio at high frequency. However, the single-to-differential input stage in a Gilbert micromixer renders good wideband frequency response and eliminates the need for common-mode rejection that is necessary in a conventional Gilbert mixer [2, 3]. In this Letter, a wideband GaInP/GaAs HBT Gilbert micromixer with 11 dB conversion gain from DC to 8 GHz is demonstrated. A single-to-differential stage is constructed with $Q_5$, $Q_6$, $Q_7$ and two resistors as shown in Fig. 1. The common-base-biased $Q_5$ and common-emitter-biased $Q_7$ provide equal but out of phase transconductance gain when $Q_6$ and $Q_7$ are connected as a current mirror. The common base configuration possesses good frequency response while the speed of common-emitter-configured $Q_7$ is improved drastically by adding the low impedance diode-connected $Q_6$ at the input of common-emitter-configured $Q_7$. Thus, the single-to-differential stage in Fig. 1 is suitable for wideband high-frequency operation.

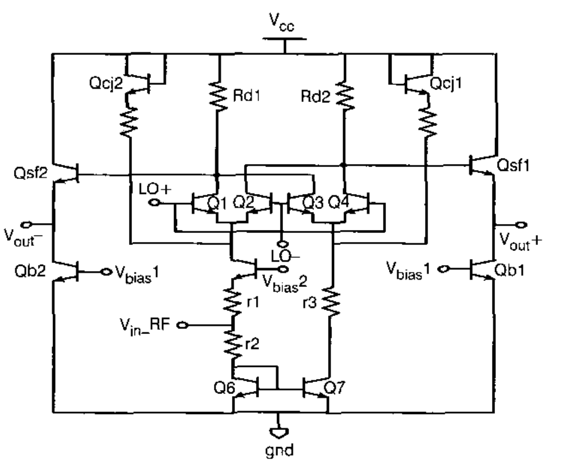


**Fig. 1** *Schematic diagram of GaInP/GaAs HBT Gilbert micromixer*

*Circuit design:* The single-to-differential input stage is not only used to turn an unbalanced signal into balanced signals but also facilitates wideband impedance matching. The two resistors in the single-to-differential input stage can improve linearity at the cost of gain reduction and noise figure degradation. The resistance seen in RF input in Fig. 1 is equal to the parallel combination of the resistance seen in the up branch and the resistance seen in the down branch. When $r_1$ and $r_2$ are equal to 60 Ω, we can make the resistance seen in the RF input be 50 Ω by biasing the $Q_5$, $Q_6$ and $Q_7$ under the condition that the three transistors all have a transconductance value close to 25 mS.

A simple method called current-injection-bias is applied to enhance the conversion gain of the mixer [4]. A current source which is also a diode-connected transistor $Q_{cj1}$ ($Q_{cj2}$) as shown in Fig. 1 is added as another branch current, i.e. the tail current of differential pair in the LO stage joining with the current flowing in the added branch is the DC current of the collector current of the single-to-differential stage. By making the current of the added branch greater, we can lower the DC current of the LO differential pair and further make the load resistor of the LO stage larger and thus conversion gain higher at the cost of lower IF bandwidth. Another single-to-differential stage with load resistors, not shown in Fig. 1, is also used as an LO buffer amplifier and converts the single-ended LO signal into balanced signals required by the Gilbert cell core in the fabricated circuit.

*Experimental results:* Fig. 2 is a photograph of the fabricated heterojunction bipolar transistor (HBT) micromixer. The micromixer is implemented with 1.4 μm emitter width GaInP/GaAs HBT process. On-wafer measurement was performed at 5 V supply voltage and a single-ended IF output was used in the measurement. Single tone and two tone measurements were performed when $LO = 5.35$ GHz and $RF = 5.7$ GHz. $IP_{1dB} = -17$ dBm, $IIP_3 = -7$ dBm and the maximum conversion gain of 11 dB was obtained for a single-ended output when the LO power is $\sim -2$ dBm.

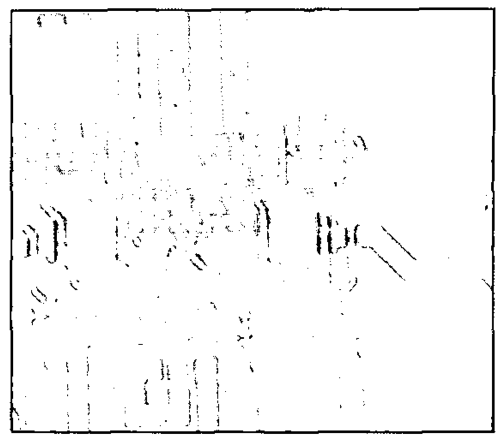

**Fig. 2** *Photograph of GaInP/GaAs HBT Gilbert micromixer*

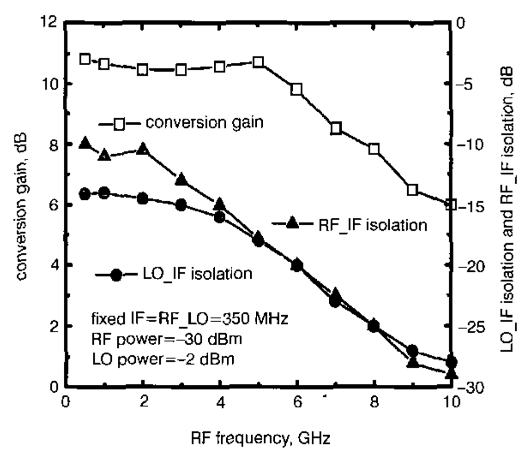


**Fig. 3** *Measured wideband downconverter GaInP/GaAs HBT micromixer performance when IF frequency fixed at 350 MHz*

The single-to-differential stage in a micromixer has intrinsically very wideband frequency response. Fig. 3 shows the conversion gain against RF frequency of the GaInP/GaAs HBT micromixer when IF is fixed at 350 MHz and RF power is $-30$ dBm. The RF-IF and LO-IF isolations are also shown in Fig. 3. LO-IF isolation is $<-15$ dB and RF-IF isolation is $<-10$ dB for all the measured frequencies. The LO-IF isolation of the Gilbert cell core is much better than $-15$ dB if the effect of the LO buffer amplifier is ignored. RF-IF isolation and conversion gain can be further improved if the IF output signal is taken differentially [4]. The input return loss is measured when the IF signal is fixed at 350 MHz, LO power is $-2$ dBm and RF power is $-30$ dBm. The measured input return loss is better than 10 dB for frequencies up to 9 GHz. It is obvious that the micromixer has 8 GHz RF 3 dB bandwidth. Fig. 4 shows the conversion gain against IF frequency when LO is fixed at 5.35 GHz and LO power is −2 dBm. The results show that the GaInP/GaAs HBT micromixer has 850 MHz IF 3 dB bandwidth.

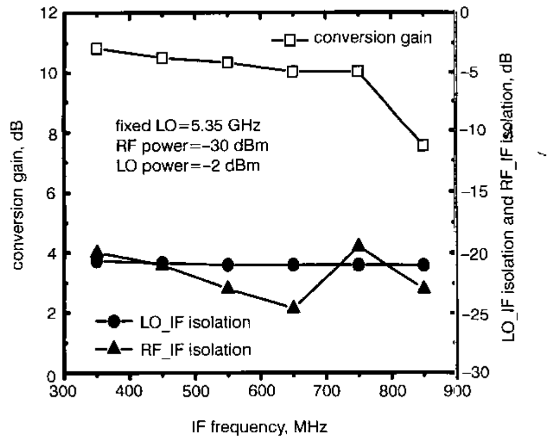


**Fig. 4** *Measured wideband downconverter GaInP/GaAs HBT micromixer performance when LO frequency fixed at 5.35 GHz*

*Conclusion:* A wideband GaInP/GaAs HBT Gilbert micromixer has a small die size and is demonstrated in this Letter. The fabricated GaInP/GaAs HBT Gilbert micromixer has 11 dB single-ended conversion gain, 8 GHz RF 3 dB bandwidth and 850 MHz IF 3 dB bandwidth. The linearity of this micromixer is also very good. $IP_{1dB}$ is −17 dBm and $IIP_3$ is −7 dBm.

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## Enhancing output voltage swing in low-voltage micro-power OTA using self-cascode

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It is shown how a self-cascode configuration can be profitably used in a micro-power operational transconductance amplifier (OTA), to enhance the output voltage swing, which eventually results in a power consumption reduction. A practical design example is proposed and used in order to discuss and quantify the circuit performance.

*Introduction:* Shrinking of supply voltages, associated to the channel length scaling in modern CMOS technologies, tends to reduce the available voltage headroom for signal swing, therefore circuit solutions that are able to grant maximum voltage swing are highly desirable in low-voltage (LV) and low-power (LP) applications. From this perspective, a current mirror like the one formed by M4, M6 and M10 in the circuit of Fig. 1, is attractive, since it offers the advantages of a cascode configuration, requiring only a limited additional voltage headroom for the cascoding transistor (M10). This cascode configuration has been previously presented in the literature as 'self-cascode': in [1] devices with different threshold voltages were exploited in order to properly bias the circuit, while in [2] short channel effects allowed the same goal to be achieved. Alternatively, M6 can be forced to work in the linear region [3, 4].

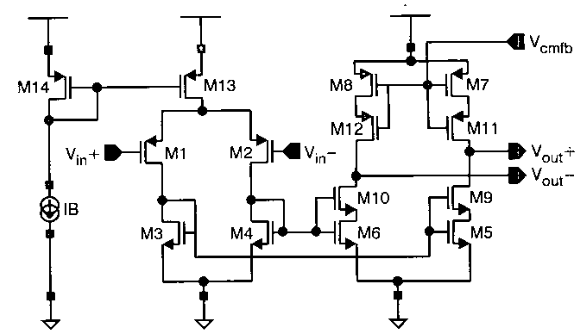


**Fig. 1** *Circuit schematic diagram of OTA*

In this Letter we propose a different approach for the biasing of the self-cascode, consisting in forcing the current mirror couple (M4–M6) to work in strong inversion regime, while M10 is biased in weak inversion. Such an approach allows one to optimally bias the self-cascode circuit. The proposed circuit is suitable for very LP applications, where biasing devices in the weak inversion regime are very common. The example of the operational transconductance amplifier (OTA) of Fig. 1, designed for a switched capacitor (SC) Sigma-Delta converter and included in a cardiac pacemaker sensing stage, is used to quantify the self-cascode performance and to discuss some of its advantages. In particular, we show how the proposed self-cascode can be an effective solution to enhance the output voltage swing, eventually achieving a significant saving in the OTA power consumption. Furthermore, we demonstrate that performance in terms of insensitivity to device mismatch is improved.

*Self-cascoded micro-power OTA:* The cardiac pacemaker poses design constraints typical of LV applications and, more importantly, requires very low power consumption. A micro-power mirrored OTA is well suited for such an environment, since it has a high-swing output stage and requires no frequency compensation for stability. The main drawback of such an architecture is that the first stage gain is very close to unity and hence the total gain of the amplifier is potentially lower than the gain of a classical two-stage amplifier. In this case, the self-cascode configuration is a useful means to raise the total gain to an acceptable value. The resulting circuit is shown in Fig. 1, which is identical to a standard mirrored amplifier, with the addition of the four cascoding transistors (M9, M10, M11, M12).

Considering the circuit schematic diagram of Fig. 1, the gate of M6 is biased at a voltage $V_{g6}$, which equates the voltage drop across the gate and source of transistors M4 and M6 and, assuming that these devices are in strong inversion, can be expressed as:

$$V_{g6} = V_{gs6} = V_{th} + V_6^{sat} \tag{1}$$

where $V_{th}$ is the n-MOS threshold voltage and $V_6^{sat}$ is the saturation voltage of transistors M4 and M6. The bias voltage of the drain of M6, $V_{d6}$, is related to the voltage drop across the gate and source of M10 according to

$$V_{d6} = V_{th} + V_6^{sat} - V_{gs10} > V_6^{sat} \tag{2}$$