

# An Active Mixer Topology for High Linearity and High Frequency Applications

Mingquan Bao and Yinggang Li

Microwave and High Speed Electronics Research Center, Ericsson Research, Ericsson AB,  
Mölndal, SE-43184, Sweden

Mingquan.bao@ericsson.com

**Abstract**—A new mixer topology is proposed which combines in parallel two single-balanced mixers, one Gilbert mixer and the other a so-called transconductance mixer. Such a mixer is supposed to improve linearity performance without degradation in noise figure and conversion gain. Based on the proposed topology, an active mixer operating at 23 GHz has been designed and fabricated in InGaP/GaAs technology. Under a power consumption of 52.6 mW, the measured IIP3 is +12 dBm, with 4 dB conversion gain and 11.5 dB double sideband noise figure.

**Index Terms**—mixer topology, active; linearity; InGaP/GaAs HBT

## I. INTRODUCTION

Though passive (resistive) mixers are attractive in terms of their linearity and noise performance [1]-[3], active mixers are preferred in cases where power boosting is expansive, because they normally have positive conversion gain and power-wise do not add extra burden to their pre- and/or post-stages in a receiver or transmitter chain.

A popular topology used for active mixers is Gilbert mixer [4]-[5]. A common challenge of using this topology is to meet the tough linearity requirements in, for instance, modern wireless communication systems. Approaches that may improve the linearity performance of Gilbert mixers include i) applying large dc current [4]; ii) utilizing a nonlinear pre-distortion circuit [5]; and iii) deploying various dynamic biasing techniques in which bias condition alters when RF input power varies [6]. These approaches indeed improve the linearity to some extent but suffer from either large dc power consumption or high noise figure or low conversion gain.

The so-called Micromixer was an attempt to improve the linearity of Gilbert mixers from topology point of view [7]-[10]. In essence it couples two single-balanced Gilbert mixers together. Each of the single-balanced Gilbert mixers contains a transconductance stage, except for one using common-base and the other using common-emitter configuration. Improvement in terms of linearity has been demonstrated with Micromixer [10]. For high frequency applications, however, obtaining simultaneously low noise figure and positive gain seems to be difficult [9]-[10].

In this paper a different mixer topology is proposed<sup>1</sup> that enables high linearity, relatively low noise and positive conversion gain simultaneously. As an example, a mixer based on the proposed topology has been designed and fabricated in InGaP/GaAs HBT technology. Experimental results are

presented and attractive features of the proposed mixer are demonstrated.

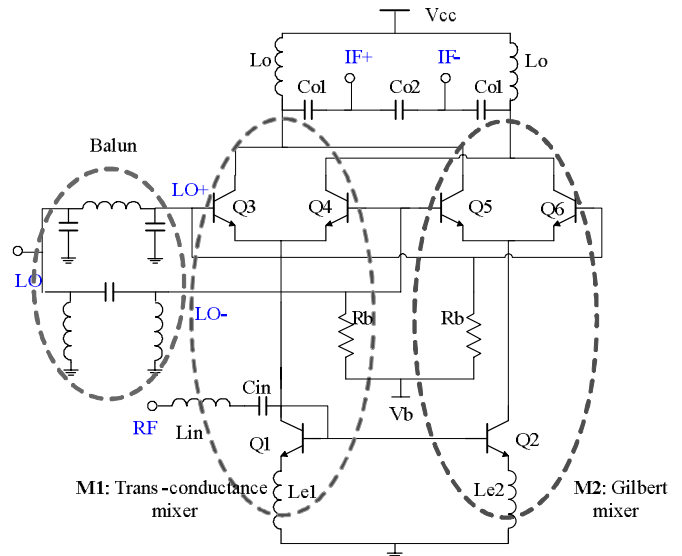


Fig. 1 Schematic of the proposed mixer

## II. CIRCUIT TOPOLOGY AND ANALYSIS

A schematic of the proposed mixer is shown in Fig.1. Like a Micromixer, the proposed mixer combines in parallel two single-balanced mixers, *M1* and *M2*, as indicated in the figure. *M2* is an ordinary Gilbert mixer using the common-emitter configuration as its transconductance stage to convert RF input voltage signal to current signal. The current signal is then fed into the switch core of *M2* which turns in an ideal case on and off at a frequency of LO drive, resulting in the desired IF current. Unlike a Micromixer, however, the *M1* in the proposed mixer does not contain any RF transconductance stage. Mixing is achieved by applying LO and RF voltage signals directly to the base and emitter of transistor  $Q3^2$ , respectively. The dynamic base-emitter voltage depends on the instantaneous difference between RF and LO. The nonlinear relationship between collector current and base-emitter voltage, i.e., the nonlinear transconductance of  $Q3$ , generates an IF current component [11]. For this reason *M1* is called transconductance mixer in this paper. *M1* and *M2* are dc-coupled in the RF input portion and, through this coupling, the dc components of the base-emitter voltage of  $Q3$  in *M1* and

<sup>1</sup> Patent application pending

<sup>2</sup> All discussions for  $Q3$  are equally valid for  $Q4$  as well. Therefore, discussions for  $Q4$  are omitted

$Q2$  in  $M2$  are related to each other. Therefore,  $M1$  and  $M2$  are interactive in response to dc-bias variation caused by increase of RF input power. In particular, the partition of the dc base-emitter voltages of  $Q3$  and  $Q2$  depends strongly on the emitter load impedance, i.e.,  $Le1$  in  $M1$  and  $Le2$  in  $M2$ . Consequently, best linearity performance of the proposed mixer is obtained by optimizing the partition via appropriate choice of  $Le1$  and  $Le2$ .

It is instructive to show some numerical results at this point to illustrate what is stated in previous paragraph. For simplicity, simulation is carried out for a circuit shown in Fig.2 which is basically the same as Fig.1 but the IF output parts are de-coupled now in order to study the individual gain of each single-balanced mixer. To further simplify,  $Le2$  is set to be zero. This does not alter the fundamental features to be discussed, since it is the ratio of  $Le1$  to  $Le2$  that really matters. Ideal inductors are used for  $Le1$  and the IF loads.

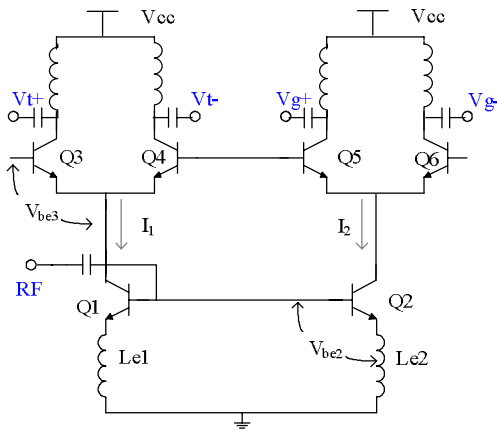


Fig. 2 Simplified schematic for illustration purpose

Fig. 3-5 show how the dc component of the base-emitter voltage of  $Q3$  and  $Q2$  (i.e.,  $V_{be3}$  and  $V_{be2}$ ), the dc currents of the  $M1$  and  $M2$  (i.e.,  $I_1$  and  $I_2$ ) and the individual conversion gain of  $M1$  and  $M2$  change when RF input power ( $P_{in}$ ) increases. For comparison, they are calculated for two values of  $Le1$ , 0 and 0.9 nH.

When  $Le1=0$  nH,  $V_{be3}$  increases while  $V_{be2}$  decreases with increasing  $P_{in}$  (cf. Fig.3), resulting in gain expansion for  $M1$  and gain compression for  $M2$  (cf. Fig.5). The dc currents, almost equal in magnitude, increase as  $P_{in}$  increases (cf. Fig.4), even though  $V_{be2}$  is reduced at large  $P_{in}$ . This is because the rectification effects of  $Q2$  and the diode  $Q1$  [6]. If proper IF phase relationship is maintained when coupling  $M1$  and  $M2$  together for IF output, the gain expansion of  $M1$  is expected to compensate the gain compression of  $M2$  and, consequently, extend the linear region of the whole mixer.

The situation for  $V_{be3}$  and  $V_{be2}$  is reversed when  $Le1=0.9$  nH, namely,  $V_{be3}$  decreases while  $V_{be2}$  increases slightly as  $P_{in}$  grows (Fig.3). This is because in this case  $Q2$  is in class-B when  $P_{in}$  is large and the rectified dc voltage increases. The two currents differ, with  $I_2$  being much larger and increasing much faster than  $I_1$ . This large boost in dc component of the current and the increase in  $V_{be2}$  give rise to a gain expansion of  $M2$  as  $P_{in}$  increases, while at the same time the gain of  $M1$  is almost constant and much smaller than that of  $M2$ . As a result, the IF components and the linearity are dominated by  $M2$ . In this

case,  $M1$  functions mainly as a dynamic bias to  $M2$  [6] and therefore, improves the linearity of  $M2$ .

It should be noted that the results shown above are exaggerated in the sense that ideal inductors and zero  $Le2$  have been used. Also a constant dc bias of LO input ports (i.e., dc base voltage of transistors  $Q3$ - $Q6$ ) is used, to get rid of the complication because of the dc voltage drop of this bias due to the increase of  $P_{in}$ . When realistic components and bias are used, the gain of  $M1$  and  $M2$  are compressed at large  $P_{in}$ . However, the fundamental features discussed already do not alter.

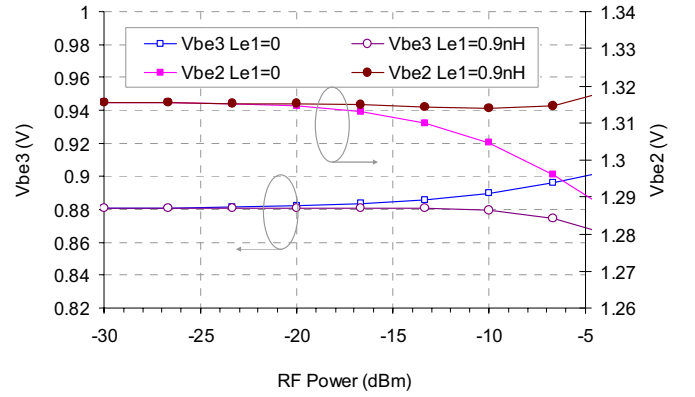


Fig. 3  $V_{be3}$  and  $V_{be2}$  versus  $P_{in}$

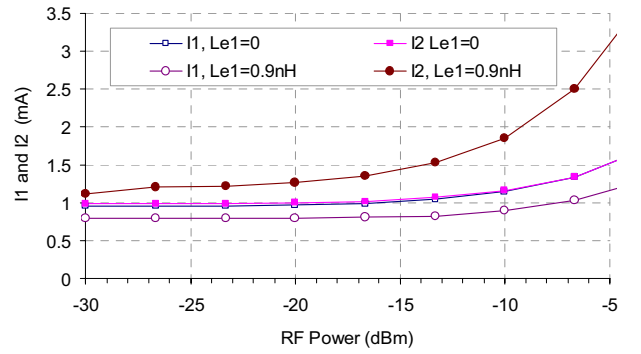


Fig. 4 dc currents of  $M1$  and  $M2$  versus  $P_{in}$

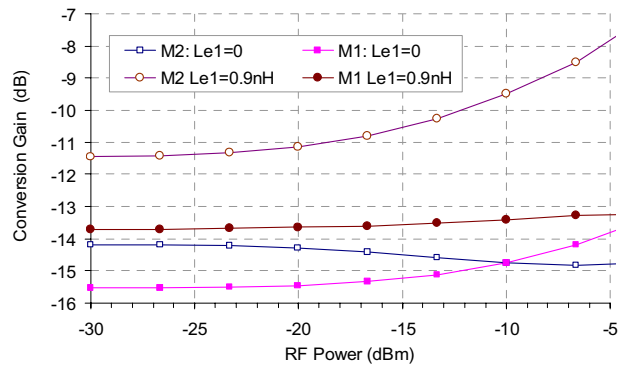


Fig. 5 Conversion gain of  $M1$  and  $M2$  versus  $P_{in}$

As compared with Micromixer, the proposed mixer does not have the common-base stage and therefore should have better noise performance. It also allows lower supply voltage by roughly a diode voltage drop.

### III. IMPLEMENTATION OF THE PROPOSED MIXER

A 23 GHz mixer is designed based on the proposed topology. As shown in Fig. 1, the circuit receives a single-ended RF input and delivers differential IF output. It also contains an on-chip, lumped-element balun to generate differential LO input. A LC tank resonating at IF frequency is used as IF load which except for providing large IF impedance also filters out RF and LO signals. Tapped capacitors,  $C_{o1}$  and  $C_{o2}$ , are applied for impedance matching.

At RF input port, an inductor  $L_{in}$  and a capacitor  $C_{in}$  are used for impedance matching and for reducing the leakage IF signal into RF port. Both  $Le1$  and  $Le2$  are used to facilitate input impedance match and to optimize linearity performance.

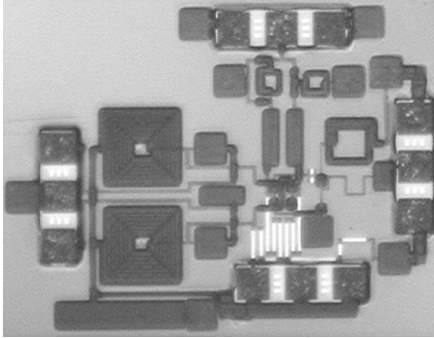


Fig. 6 Mixer chip photograph, size  $1.0 \times 1.35 \text{ mm}^2$ .

The mixer is manufactured in InGaP/GaAs HBT technology provided by KnowledgeOn. The HBTs have 60 GHz  $f_T$  and 110 GHz  $f_{max}$ . The final chip layout has a size of  $1.00 \times 1.35 \text{ mm}^2$  including pads, IF filter and LO balun, as shown in Fig. 6.

### IV. MEASURED MIXER PERFORMANCE

On-chip probing is carried for the mixer chip. One-tone measurement using spectrum analyzer is performed for RF frequency of 23GHz; LO frequency of 22 GHz and LO power of 5 dBm, IF frequency of 1GHz and dc supply voltage  $V_{CC}$  of 4.8V. The measured conversion gain and dc current as function of input RF power are plotted in Fig. 7. The conversion gain is 4 dB in the linear region. 1-dB compression point,  $P_{1dB}$ , is -5.6 dBm. The dc current increases dramatically when  $P_{in}$  is larger than -10 dBm. This is due to the fact that transistor Q2 operates in class-B and the rectified current gives rise to a large dc component of the current which increases with  $P_{in}$ . This is reproduced in simulation.

Two-tone measurement is carried out to determine IIP3. The frequency spacing between the two tones is 1MHz. Fig. 8 shows the output power of fundamental and 3<sup>rd</sup>-order inter-modulation. IIP3 is determined to be +12 dBm. The corresponding OIP3 is thus +16 dBm.

Noise figure is measured using Noise Figure Analyzer (N8975A). The measured results versus RF frequencies are plotted in Fig. 9. Double sideband (DSB) noise figure is 11.5 dB at 23 GHz, which is much less than that of the Micromixers reported in [9]-[10]. DSB conversion gain is also obtained in the noise measurement using the same Noise Figure Analyzer. The results are included in Fig. 9. The mixer can operate up to

25 GHz beyond which a significant gain drop occurs. The DSB gain is 3-5 dB larger than the gain obtained in single sideband (SSB) measurement using spectrum analyzer.

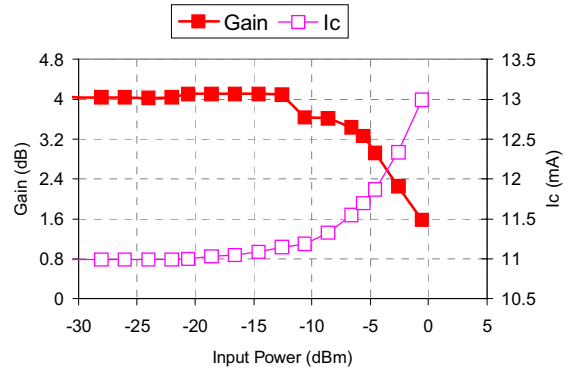


Fig. 7 conversion gain and dc current versus input RF power

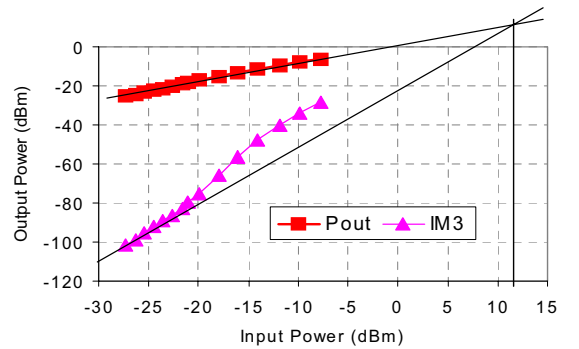


Fig. 8 Fundamental and 3<sup>rd</sup>-order inter-modulation versus input power at 23GHz; LO power = 5 dBm

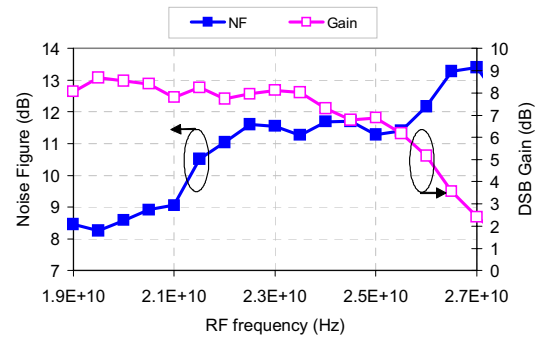


Fig. 9 DSB gain and noise figure versus RF frequency: IF = 1GHz, LO power = 5 dBm,  $V_{CC} = 4.8 \text{ V}$

Mixer performance as function of LO power is also examined and the obtained IIP3 and conversion gain are plotted in Fig. 10. IIP3 versus LO has a bell shape and peaks around 5 dBm. The SSB conversion gain increases monotonously with LO power.

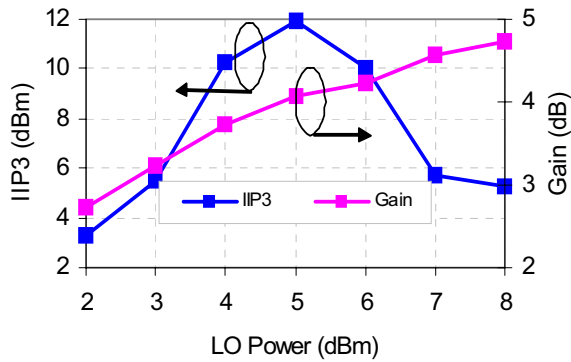


Fig. 10 IIP3 and conversion gain versus LO power.

The DSB noise figure as function of the LO power is measured and the results are given in Fig. 11. When LO power is swept from 2 to 8 dBm, the DSB noise figure is almost a constant even though the dc-current increases considerably. This is because the conversion gain grows rapidly as well when the current increase. (cf. Fig. 10).

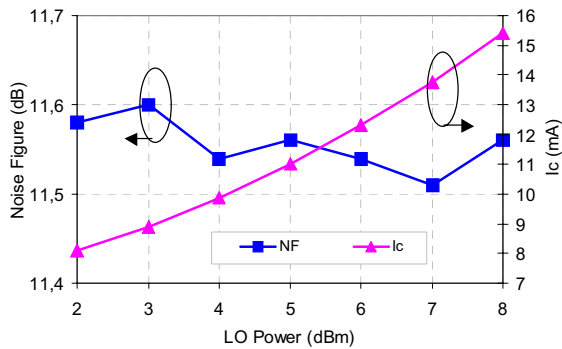


Fig. 11 Noise figure and dc bias current versus LO power:  $V_{cc}=4.8V$

## V. CONCLUSION

A new mixer topology is proposed in this paper. Based on the topology a 23 GHz active mixer has been designed and implemented in InGap/GaAs HBT technology. Experimental results show attractive features of the mixer and demonstrate the capability of the proposed mixer to simultaneously offer high linearity, relatively low noise figure and positive gain at high frequency operation.

## ACKNOWLEDGMENT

The authors are grateful to Dr. T. Lewin, Prof. H. Zirath, and Dr. J. Grahn for their generous and consistent support.

## REFERENCES

- [1] S. Gunnarsson, K. Yhland, and H. Zirath, "pHEMT and mHEMT ultra wideband millimetre wave balanced resistive mixers", *IEEE 2004 MTT-S International Microwave Symposium Digest*, pp. 1139 – 1142, June 2004.
- [2] F. Ellinger, "26.5-30-GHz resistive mixer in 90-nm VLSI SOI CMOS technology with high linearity for WLAN", *IEEE Trans. Microwave Theory and Techniques*, Vol. 53, No. 8, pp.2559-2565, Aug. 2005.

- [3] M. Bao, H. Jacobsson, L. Aspemyr, A. Mercha, and G. Carchon, "A 20GHz sub-1V low noise amplifier and a resistive mixer in 90nm CMOS technology", *Proc. Asia-Pacific Microwave Conference*, 2005.
- [4] S. Hackl and J. Böck, "31 GHz monolithic integrated quadrature demodulator in SiGe bipolar technology", *Proc. of Asia-Pacific Microwave Conf.*, vol. 2, pp.846-848, Nov. 2002.
- [5] M. Bao, Y. Li, and A. Cathelin, "A 23GHz active mixer with diode linearizer in SiGe BiCMOS technology", *Proc. of 33<sup>rd</sup> European Microwave Conference*, pp. 391-393, 2003.
- [6] T. Yoshimasu, M. Akagi, N. Tanba, and S. Hara, "An HBT MMIC power amplifier with an integrated diode linearizer for low-voltage portable phone applications", *IEEE J. Solid-State Circuits*, vol. 33, pp.1290-1296, Sept. 1998.
- [7] B. Gilbert, "The MICROMIXER: A highly linear variant of the Gilbert mixer using a bisymmetric class-AB input stage", *IEEE J. Solid-State Circuits*, vol. 32, pp.1412-1423, Sept. 1997.
- [8] S. C. Tseng, C. C. Meng, C. H. Chang, C. K. Wu, and G. W. Huang, "Monolithic broadband Gilbert micromixer with an integrated Marchand balun using standard silicon IC process", *IEEE Trans. Microwave Theory and Techniques*, Vol. 56, No. 12, pp.2559-2565, Dec. 2006.
- [9] M. Bao and Y. Li, "A 17 to 26 GHz Micromixer in SiGe BiCMOS Technology", *Proc. of 35<sup>th</sup> European Microwave Conference*, 2005.
- [10] M. N. Do, D. Dubuc, A. Coustou, E. Tournier, P. Ancey and R. Plana, "Highly linear 20GHz-Micromixer in SiGe Bipolar Technology" *Proc. of 35<sup>th</sup> European Microwave Conference*, 2005.
- [11] T. H. Lee, *The design of CMOS radio-frequency integrated circuits*, pp. 318, Cambridge University Press, 1998