A Class AB Monolithic Mixer for 900-MHz Applications

Keng Leong Fong, Christopher Dennis Hull, and Robert G. Meyer, *Fellow, IEEE*

Abstract— A Class AB downconversion mixer for 900-MHz applications is presented. The circuit is implemented in a 25-GHz f_T bipolar process and consumes 10.2 mA total current from a 3-V supply. The design has a power gain of 7.5 dB and operates with supply voltages from 2.7 to 5 V. A single-sideband noise figure of 7.5 dB is achieved by design simplicity, use of bond-wire degeneration in the common-emitter driver stage, and optimization of the device sizes and bias currents. An input 1-dB compression point of -1.5 dBm is achieved by exploiting the class AB behavior of the common-emitter driver stage.

Index Terms— Analog circuits, bipolar analog integrated circuits, circuit noise, circuit optimization, high-speed integrated circuits, integrated circuit noise, microwave integrated circuits, microwave mixers, mixer noise, mixers, nonlinear circuits, time-varying circuits.

I. INTRODUCTION

THE rapid growth of portable wireless communication systems has led to a demand for low-power and high-performance front-end downconverters. The downconversion mixer is an important building block of a typical front-end as shown in Fig. 1. The downconversion mixer is used to convert the radio frequency (RF) signal down to an intermediate frequency (IF) by mixing the RF signal from the low-noise amplifier (LNA) with the local oscillator (LO) signal. This allows channel selection and gain control at lower frequencies where high quality-factor (Q) filters and variable-gain amplifiers can be constructed economically. Instead of using an IF filter with tunable passband frequency, an IF filter with fixed passband frequency is used, and the LO frequency is tuned to select the desired channel. The LNA is used to amplify the RF signal to reduce the noise contribution from the mixer. The RF and image-rejection filters are used to reject undesired out-of-band signals. In this paper, a class AB downconversion mixer for 900-MHz applications is described. The emphasis is on design techniques that achieve a high 1-dB compression point and a low noise figure.

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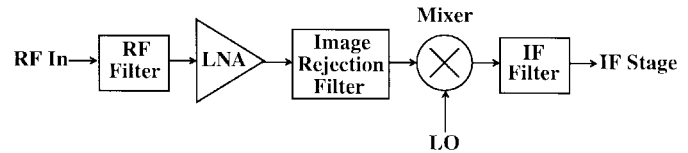
Fig. 1. RF front-end.

II. MIXER SPECIFICATION DISCUSSIONS

The system noise figure (in linear scale) for the downconverter shown in Fig. 1 is

$$
\begin{aligned}
\mathrm{NF} &= \frac{1}{L_{\mathrm{RF}}} + \frac{\mathrm{NF}_{\mathrm{LNA}} - 1}{L_{\mathrm{RF}}} + \frac{1}{L_{\mathrm{RF}} G_{\mathrm{LNA}}} \left(\frac{1}{L_{\mathrm{IM}}} - 1 \right) \\
&\quad + \frac{\mathrm{NF}_{\mathrm{MIX}} - 1}{L_{\mathrm{RF}} G_{\mathrm{LNA}} L_{\mathrm{IM}}} \\
&= \frac{1}{L_{\mathrm{RF}}} \left(\mathrm{NF}_{\mathrm{LNA}} + \frac{\mathrm{NF}_{\mathrm{MIX}} - L_{\mathrm{IM}}}{G_{\mathrm{LNA}} L_{\mathrm{IM}}} \right)
\end{aligned}
\tag{1}
$$

where L_{RF} and L_{IM} are the insertion losses of the RF filter and the image-rejection filter, respectively, $\mathrm{NF}_{\mathrm{LNA}}$ and $\mathrm{NF}_{\mathrm{MIX}}$ are the noise figures of the LNA and the mixer, respectively, and G_{LNA} is the power gain of the LNA. This equation assumes that the noise figures of the filters are the same as their insertion losses. As shown in (1), the LNA needs to have sufficient power gain to reduce the noise contribution from the mixer. Hence, a mixer with low noise figure (NF) is highly desirable in order to relax the gain requirement of the LNA. Most of the low-noise active mixers (in silicon technology) currently available have a single-sideband noise figure greater than 10 dB. The goal of this design is to obtain a mixer with significantly lower noise figure, without sacrificing linearity.

A downconversion mixer should provide sufficient power gain to compensate for the IF filter loss and to reduce the noise contribution from the IF stage. Typically, a power gain greater than 5 dB is desirable. However, this gain should not be too large since saturation at the IF output port may limit the 1 dB compression point (P_{1dB}).

A strong undesired signal (known as a blocker or interferer) can overload a mixer and cause gain compression of the small desired signal if the mixer does not have sufficiently high input P_{1dB}. In some applications, a blocker as strong as -5 dBm can appear at the input of the mixer. In order to avoid increasing the noise contribution from the IF stage significantly, this blocker should not reduce the gain of the mixer with respect to the small desired signal by more than 1 dB. Unfortunately, there is no simple relationship between

the gain compression of the small desired signal and that of the large undesired signal. The relationship derived in [1] assumes a weakly nonlinear condition where the gain compression is solely caused by the third-order term in the transfer function of the mixer. This describes many practical mixers, but higher-order terms can also be important in the presence of a large-amplitude blocker. The small desired signal can be viewed as amplitude modulation on top of the blocker which functions like a carrier. Typically, the modulation signal (the small desired signal) is compressed more than the carrier (the large blocker). Using this intuitive knowledge, an input P_{1dB} specification greater than -5 dBm (-3 dBm is used in this design to provide 2-dB margin) for the mixer is used as a starting point for this design. In the actual design, SPICE simulation is used to verify that the gain compression of the small desired signal in the presence of a -5-dBm blocker is less than 1 dB. The actual value of P_{1dB} is not the true design criterion.

Due to the third-order nonlinearity of the mixer, two strong signals at the adjacent channels can generate third-order intermodulation (IM_3) product at the desired channel. If the IM_3 product is strong enough, it can corrupt the weak desired signal. Hence, the mixer should be linear enough to avoid this situation. The linearity requirement, characterized by third-order intercept point (IP_3), depends on the power difference between the desired signal and the undesired signals at the adjacent channels.

The impedances of the RF and LO input ports are typically matched to 50 Ω, while the impedance of the IF output port is matched to that of the IF filter. Impedance matching at the RF and IF ports is necessary to avoid signal reflection from the ports and excessive passband ripples in the frequency responses of the filters. Typically, return losses of less than -10 dB are required. On the other hand, the return loss specification on the LO port can be more relaxed. However, excessive LO port return loss requires increased LO drive level, which increases the power consumption of the overall system.

If the mixer is in a different package from the LNA, the isolation between LO and RF ports of the mixer is important, since LO-to-RF feedthrough results in LO signal leaking through the antenna. The amount of feedthrough that is allowed depends on the reverse isolation of the LNA and the stopband attenuation of the RF and image-rejection filters at the LO frequency. If the LNA is in the same package as the mixer, the LO feedthrough to RF input port of the LNA may become more significant. LO-to-IF and RF-to-IF isolations are not important because the high-frequency feedthrough can be easily rejected by the high-Q IF filter.

III. CIRCUIT DESIGN

The basic topology of the class AB mixer is shown in Fig. 2. It comprises a common-emitter driver stage (Q_1) and a differential switching pair (Q_2 and Q_3). The driver stage amplifies the RF signal to compensate for the attenuation due to the switching process and to reduce the noise contribution from the switching pair. The switching pair performs the mixing function which converts the RF
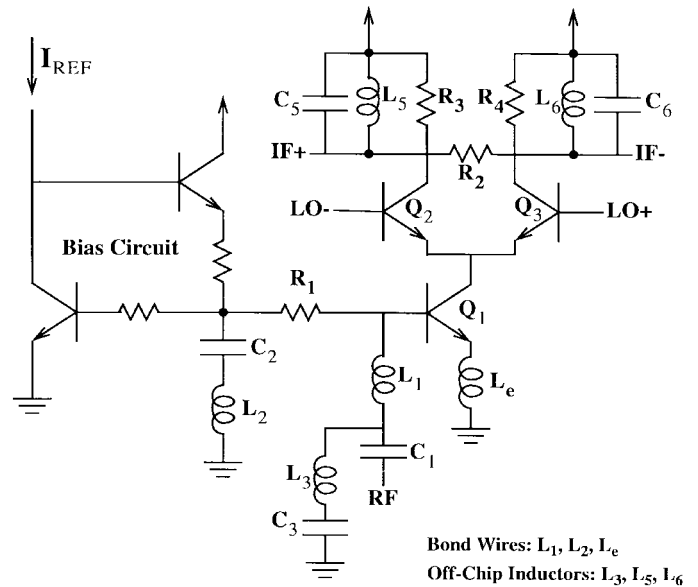

Fig. 2.   Class AB mixer.

signal down to the IF as illustrated in the following equation:

$$
\begin{aligned}
I_O &= V_{RF} \cos \omega_{RF} t \times G_M \\
&\quad \times \left(\frac{2}{\pi} \cos \omega_{LO} t - \frac{2}{3\pi} \cos 3\omega_{LO} t + \cdots \right) \\
&= \frac{1}{\pi} G_M V_{RF} \cos(\omega_{LO} - \omega_{RF}) t \\
&\quad + \frac{1}{\pi} G_M V_{RF} \cos(\omega_{LO} + \omega_{RF}) t + \cdots
\end{aligned}
\tag{2}
$$

where I_O is the signal current flowing through the differential output resistance R_2 ($R_3 + R_4$ is matched to R_2), ω_{RF} and ω_{LO} are the RF and LO frequencies, respectively, V_{RF} is the RF input signal, and G_M is the transconductance of the driver stage. This equation assumes instantaneous switching (multiplying the RF signal with square wave) of Q_2 and Q_3. The ($\omega_{LO} - \omega_{RF}$) term is the IF signal, while the ($\omega_{LO} + \omega_{RF}$) term is an unwanted signal. The $1/\pi$ factor is due to the power lost in the ($\omega_{LO} + \omega_{RF}$) term and other higher frequency terms.

To improve the linearity of the common-emitter driver stage, it is degenerated by the bond wire L_e (modeled as a high-Q inductor) instead of a resistor. In contrast to resistive degeneration, inductive degeneration does not create an additional noise source. To reduce the noise contribution from Q_1, a large device with small base resistance ($r_b \sim 3 \ \Omega$) is used. The bias current is also optimized to reduce the sum of base and collector shot noise contributions. The input-referred noise due to the collector shot noise decreases with bias current, while that due to the base shot noise increases with bias current. Ideally, the gain of the driver stage should be maximized (by minimizing the bond-wire inductance) to minimize the noise contribution from the switching pair. However, linearity (third-order intermodulation and spurious response) and head room considerations (1-dB compression point) set the lower limit on the degeneration inductance.

Capacitor C_2 is a bypass capacitor used to prevent noise of the bias circuit from entering the base of Q_1. Series tuning between C_2 and a bond wire (L_2) provides an ac ground at the RF (900 MHz). The impedance looking into the base of Q_1 is given by

$$2\pi f_T L_e + r_b + sL_e + \frac{1}{sC_\pi} \tag{3}$$

where f_T is the unity current gain frequency of Q_1. This equation neglects the effect of collector-base junction capacitance (C_μ) of Q_1. The parallel combination between the real part of this equation and the resistance of R_1 (150 Ω) provides 50 Ω for impedance matching [2]. The imaginary part of this equation is cancelled by the reactance of the bond wire L_1 in series with the external dc blocking capacitor C_1.

Since there is little resistance (dc feedback) between the emitter of Q_1 and ground, the common-emitter driver stage exhibits class AB behavior. This phenomenon is due to the exponential collector current (I_C) versus base emitter voltage (V_{BE}) characteristics of the bipolar transistor. Using

$$V_{BE} = V_0 + V_1 \cos\omega t + V_2 \cos 2\omega t + V_3 \cos 3\omega t + \cdots \tag{4}$$

$$I_C = I_S \exp\left(\frac{V_{BE}}{V_T}\right) \tag{5}$$

where V_n is the amplitude of the nth harmonic, we have

$$\begin{aligned} I_C = I_S \exp\left(\frac{V_0}{V_T}\right) &\left[I_0\left(\frac{V_1}{V_T}\right) + 2I_1\left(\frac{V_1}{V_T}\right)\cos\omega t + \cdots\right] \\ &\times \left[I_0\left(\frac{V_2}{V_T}\right) + 2I_1\left(\frac{V_2}{V_T}\right)\cos 2\omega t + \cdots\right] \\ &\times \left[I_0\left(\frac{V_3}{V_T}\right) + 2I_1\left(\frac{V_3}{V_T}\right)\cos 3\omega t + \cdots\right]\cdots \end{aligned} \tag{6}$$

where I_0 and I_1 are modified Bessel functions of order zero and one, respectively. Without any feedback, the I_1 term causes gain expansion. Due to the I_0 term, the average current increases. With dc (resistive) feedback, the increase of dc current can compress V_{BE} and the transconductance of the driver stage. To avoid gain compression due to dc feedback, R_1 and the parasitic resistance at the emitter of Q_1 (R_E) should be minimized. Without any dc feedback components, the transconductance of the driver stage is not reduced even at very high input power level (>10 dBm). However, R_1 should be large enough to avoid significant loading at the RF port which would cause impedance mismatch and noise-figure degradation.

For a mixer with 5 dB of power gain, an RF signal of -3 dBm produces 2 dBm of IF output power. The current required to drive 2 dBm of power into the 1-kΩ differential resistor (R_2) is 1.78 mA. Without the class AB phenomenon (as in the case of a differential-pair driver stage with constant current source bias), the bias current required is

$$1.78 \text{ mA} \times \pi \times 2 = 11.2 \text{ mA}. \tag{7}$$

The π factor is due to the current lost in the switching (mixing) process. The factor of two is due to half of the current being lost in the filter-matching resistor $(R_3$ and $R_4)$. On the other hand, a class AB mixer can be biased at a much lower
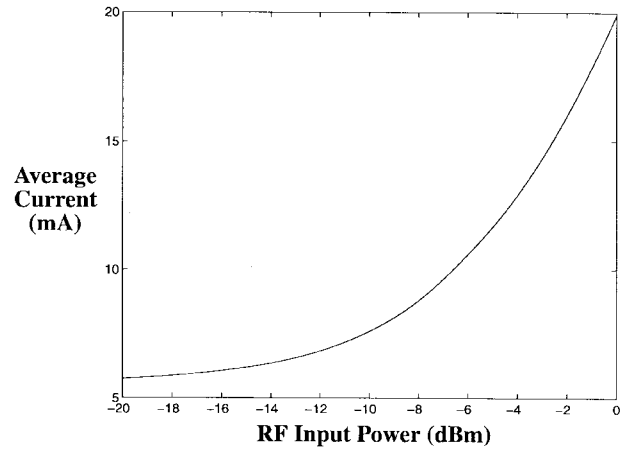

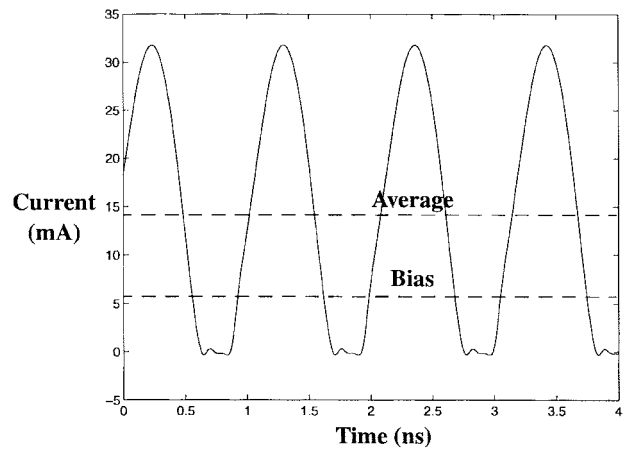Fig. 3.   Average current versus RF input power.

Fig. 4.   Simulated output current of the driver stage.

quiescent current level. Fig. 3 shows the simulated average current of the driver stage (biased at collector current of 5.6 mA) as a function of RF input power. Fig. 4 shows the simulated output current waveform of the driver stage with a -3-dBm sinusoidal signal applied to its input. Due to the class AB phenomenon, the average current increases to 14.4 mA.

To further investigate this Class AB phenomenon, a small signal and a large blocker are applied to the input of the driver stage. Fig. 5 shows the normalized transconductances G_M (normalized to the small-signal transconductance) of these two signals, as a function of the blocker input power. Due to the class AB phenomenon, the driver stage has an input P_{1dB} of about 4 dBm. With a -5-dBm blocker, the transconductance of the small desired signal is compressed by only 0.5 dB. In this mixer design, gain compression is dominated by saturation of Q_2 and Q_3 due to large signal swings at the IF output port. With 2 dBm of IF output power applied to the 1 kΩ differential resistor R_2, the resulting differential output amplitude is 1.8 V (0.9 V at each IF output terminal). Hence, sufficient head room is required at the collectors of Q_2 and Q_3 to avoid limiting the IF output swings. In this design, the bases of Q_2 and Q_3 are biased at 1.1 V below the power supply voltage. Assuming the bipolar transistor saturates at a base-collector voltage of about 0.4 V, this allows 0.6 V (= 1.1 V + 0.4 V − 0.9 V)

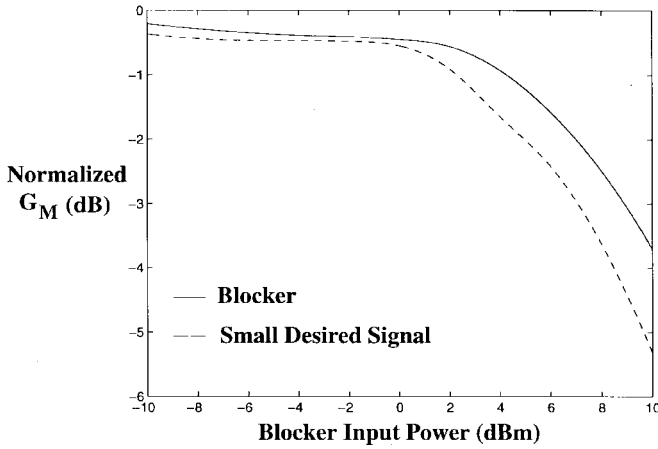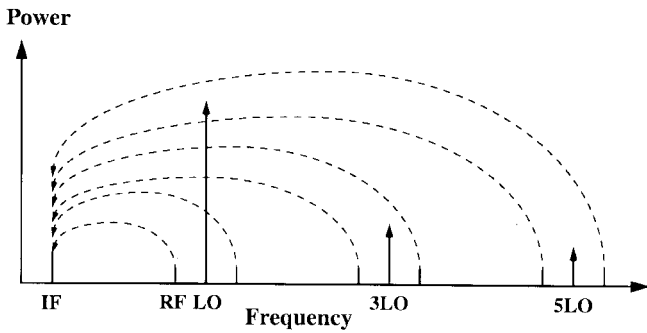Fig. 5. Normalized G_M versus blocker input power.

Fig. 7. Mixer noise figure versus LO amplitude.

Fig. 6. LO mixes noise to the IF.

head room for LO swing to drive the bases of Q_2 and Q_3 and high-frequency feedthrough at the IF output port, as well as bias shift due to component variations.

If the gain of the driver stage and its output noise were constant across all frequencies, the instantaneous switching process would increase the noise contribution from the driver stage by a factor of $(\pi/2)^2$ (or 3.9 dB) as illustrated in Fig. 6 [3]. The LO and its harmonics mix noise at various frequencies down to the IF. In this case, the overall input-referred noise power (in linear scale) of the mixer would be

$$\text{noise of gain stage} \times \left(\frac{\pi}{2}\right)^2 + \text{noise from switching pair.} \tag{8}$$

With inductive degeneration, the gain of the driver stage decreases with frequency [2]. If high-side mixing (LO has higher frequency than RF) is used, the RF signal (and associated noise) has higher gain than the noise at the image frequency. Also, noise at higher frequencies is attenuated by the degeneration inductance. In this case, the mixing process increases the noise of the driver stage by a factor of less than $(\pi/2)^2$.

When one transistor $(Q_2$ or $Q_3)$ of the switching pair is on, the mixer functions like a cascode amplifier. Noise from the driver stage at the IF can feed through to the IF output terminals. Since the driver stage has high gain at the IF (due to inductive degeneration), the output noise of the mixer would increase significantly if a single-ended output
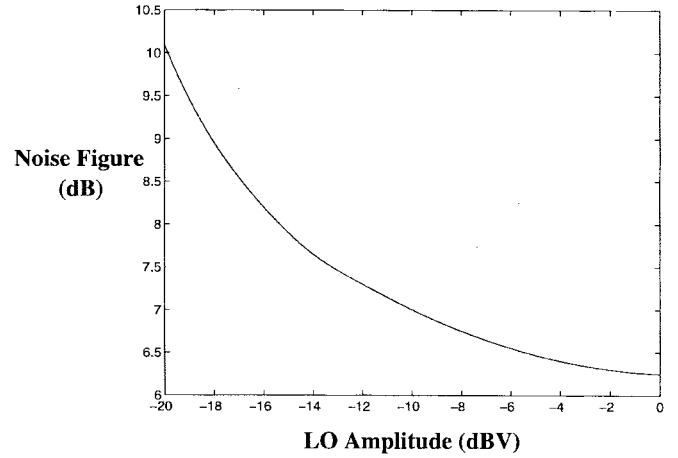
were taken. However, since this noise is common-mode, it can be suppressed by taking the IF output differentially.

Single-balanced mixers do not reject LO and RF feedthrough at the IF output port. Normally, this is not a problem since the IF filter has high enough stopband attenuation to filter out unwanted signals at both LO and RF frequencies (and their harmonics). However, this high-frequency feedthrough can produce large signal swings at the IF output port of the mixer and degrade the P_{1dB} and the spurious performance of the mixer. Capacitors C_5 and C_6 are used to attenuate the LO and RF feedthrough without affecting IF signal.

The switching pair should be driven by a large LO signal to minimize its noise contribution. The switching pair contributes noise to the mixer output when both transistors $(Q_2$ and $Q_3)$ are on [3], and a large LO amplitude is needed to reduce the duration of this condition. Fig. 7 shows the simulated noise figure (NF) of the mixer (LO buffer is not included) as function of differential LO amplitude. For LO amplitudes above -10 dBV (1 V $= 0$ dBV), the noise figure of the mixer decreases slowly as the LO amplitude increases because the overall input-referred noise is dominated by the noise contribution from the driver stage. Linearity, head room, LO feedthrough, and power consumption considerations set the upper limit on the LO amplitude. A very large LO amplitude results in excessive current being pumped into the common-emitter point of the switching pair through the base-emitter junction capacitance (C_{je}), and thus generates additional third-order intermodulation [3]. Large LO amplitudes also decrease the head room at the collectors of Q_2 and Q_3 and increase the LO feedthrough to the RF port (through C_μ of Q_1). Another disadvantage of using a large LO amplitude is that large power consumption is required.

Considering the tradeoffs between noise figure and other performance parameters mentioned above, a differential LO amplitude of -10 dBV (630 mV peak-to-peak) is used in this design. Using this LO amplitude and devices with high f_T, the switching pair is switched very rapidly and thus generates relatively little noise and IM_3 at the mixer outputs [3], [4]. Reasonably large devices are used to reduce the r_b noise contribution. However, if the base-emitter junction capacitance
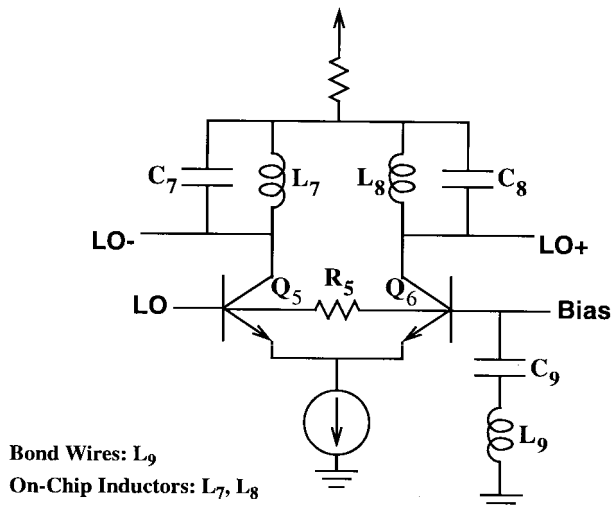
Fig. 8.　LO buffer.

Fig. 9.　Chip micrograph.

(C_{je}) is too large, the switching of C_{je} can pump additional current into the common emitter point [3]. With proper choices of LO swing and device sizes in this design, the noise and nonlinearity contributions from the switching pair are small relative to those from the driver stage.

If the switching pair were driven directly by an external LO, 0 dBm of LO power would be required. In the absence of an on-chip LO buffer, it might take up to 10 mA of bias current in an external LO driver to supply this LO power. Hence, a LO buffer with a gain of 10 dB is included in this design to reduce the LO input power requirement to -10 dBm. Fig. 8 shows the basic topology of the LO buffer. One side of the differential pair (Q_5 and Q_6) accepts the LO input while the other side is ac grounded by series tuning between capacitor C_9 and a bond wire (L_9). Resistor R_5 is used for impedance matching. The noise contribution from the LO buffer is minimized by using large devices for the differential pair and using a low-noise current source. Further reduction of the LO input power requirement (by increasing the gain of the LO buffer) is not desirable because the LO buffer would become very noisy. Like the switching pair in the mixer, the differential pair (Q_5 and Q_6) contributes noise to the outputs of the LO buffer when both transistors are on.

For a single-balanced design, the noise from the LO buffer at the IF can feedthrough to the IF output port [3]. Hence, on-chip inductors L_7 and L_8 (short circuit at the IF) are used at the output of the LO buffer to remove this noise. Capacitors C_7 and C_8 are used for parallel tuning with the inductors at the LO frequency in order to increase the load impedance of the LO buffer.

IV. DESENSITIZATION BY A BLOCKER

A blocker with a large amplitude can desensitize the mixer by reducing the conversion gain of the mixer with respect to the small desired signal. Further desensitization may occur due to an increased noise floor in the presence of the large blocking signal [1]. A blocker at 10 MHz away from the RF is used to illustrate this phenomenon.
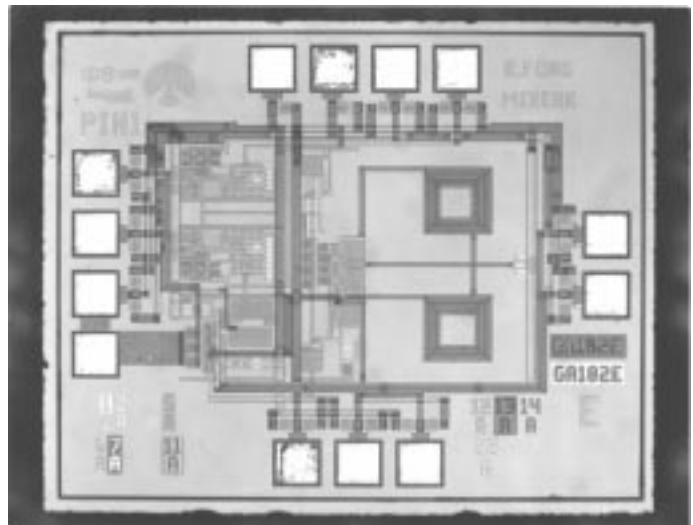
The blocker can be seen as functioning as a second LO signal that mixes low-frequency noise from the bias circuit (in Fig. 2) to the RF [1]. For instance, the blocker at 10 MHz away from the RF mixes with the noise from the bias circuit at 10 MHz and shifts it to the RF. To remove this low-frequency noise, a low-frequency trap (C_3 and L_3) as shown in Fig. 2 is placed at the base of Q_1. The trap should have low enough impedance at 10 MHz to filter out the low-frequency noise from the bias circuit. At low frequencies, L_3 appears to be short. Hence, C_3 can also filter out other low-frequency noise at frequencies higher than 10 MHz, in case there are blockers at more than 10 MHz away from the RF signal. At RF, the trap appears open (due to L_3) and does not affect the impedance matching.

The blocker also contributes noise through reciprocal mixing [5]. Once amplified by the driver stage, the blocker mixes the phase noise from the LO buffer to the IF. For instance, the blocker at 10 MHz away from the RF mixes with phase noise at 10 MHz offset from the LO carrier. Hence, the LO buffer needs to have low phase noise to reduce the desensitization effect by the blocker.

V. IMPLEMENTATION

The class AB mixer is implemented in a 25-GHz f_T bipolar process. Fig. 9 shows the die micrograph. The pads are electrostatic discharge (ESD) protected and the die is housed in a TQFP32 package (for prototyping). Transistor Q_1 is placed close to the bonding pads to which the bond wires used for degeneration are connected. This is to minimize parasitic capacitance and resistance at the emitter of Q_1. The degeneration inductor L_e ($\sim$2.4 nH) is implemented by using two adjacent pins and bond wires in parallel. The LO pin is perpendicular to the RF and IF pins to reduce the inductive coupling of the LO power to the RF and IF ports. The LO pin and ac ground pin (L_9) of the LO buffer are placed next to each other to minimize the cross-sectional area of the resulting current loop. This technique is also applied to the RF pin and the ground pin (L_2) for the bypass capacitor C_2. The

TABLE I
PERFORMANCE SUMMARY

| Parameters	Simulations	Measurements
Power Supply	2.7 V to 5V	2.7V to 5V
Current Consumption	11.4 mA	10.1 mA
LO Input Power	-10 dBm	-10 dBm
Conversion Gain (Power)	7.7 dB	7.5 dB
Gain Variation (@-10 dBm Blocker)	-0.4 dB	-0.25 dB
Gain Variation (@-5 dBm Blocker)	-0.6 dB	-0.2 dB
Gain Variation (+/- 20 MHz)	+/- 0.2 dB	+/- 0.25 dB
Input IP$_3$	4.5 dBm	2.5 dBm
Input P$_{1dB}$	-0.5 dBm	-1.5 dBm
Single Sideband (SSB) NF	6.9 dB	7.5 dB
SSB NF (@-10 dBm blocker)	12.3 dB	12.1 dB
SSB NF (@-5 dBm blocker)	16.6 dB	16.7 dB
LO to RF Feedthrough	-44.8 dBm	-47.5 dBm
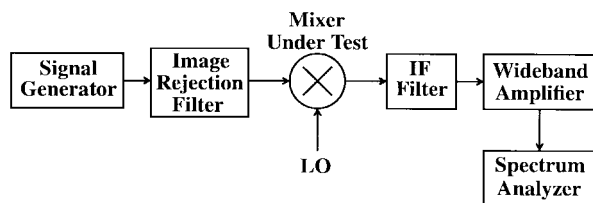
Fig. 10. Noise measurement under blocking condition.

two IF pins are placed next to each other on the opposite side of the package from the RF pin. The supply and ground pins for the bias circuit are placed on the fourth side of the package.

VI. MEASUREMENT RESULTS

Table I summarizes the simulation and measurement results. Noise figure is simulated without ESD and package model. The measurements were performed at 25°C and 3 V supply. The RF, LO, and IF frequencies used are 900 MHz, 1150 MHz, and 250 MHz, respectively. The LO input power used is -10 dBm. The input return loss of the RF port is less than -14 dB (600 MHz to 1.5 GHz), using only one off-chip blocking capacitor (C_1) for impedance matching. With -5 dBm and -10 dBm blockers (at 890 MHz), the current consumption increases to 14.6 mA and 11.4 mA, respectively. Due to the class AB behavior of Q_1, the input 1 dB compression point (P_{1dB}) and the input third-order intercept point (IP_3) differ by less than the theoretical value (weak nonlinear condition) of 9.6 dB. To achieve a comparable P_{1dB} in a conventional class A mixer would require much higher bias current. A single-sideband noise figure of 7.5 dB is achieved in this design. When the LO input power is increased to -5 dBm, the noise figure is improved to 6.9 dB. As discussed in the previous section, the desensitization effect by the -5 dBm and -10 dBm blockers increases the noise figure to 16.7 dB and 12.1 dB, respectively.

Fig. 10 shows the experimental setup to measure the mixer noise figure under the blocking condition. The signal generator is used to apply the blocker at 890 MHz. The image-rejection filter is needed for single-sideband noise figure measurement. The IF filter is used to filter out the blocker (at 260 MHz after the mixing process) to prevent it from desensitizing the spectrum analyzer. The wideband amplifier (24 dB gain) is used to increase the sensitivity of the spectrum analyzer. A typical spectrum analyzer has a noise floor of about -150 dBm/Hz. The output noise from this setup should be much larger than that of the spectrum analyzer so that it can be displayed on the spectrum analyzer. To calculate the noise figure of the mixer under the blocking condition, the noise contributions from the image rejection filter, the IF filter, the wideband amplifier, and the spectrum analyzer have to be subtracted.

VII. CONCLUSIONS

A class AB downconversion mixer for 900-MHz applications has been designed. Class AB behavior of the common-emitter driver stage improves the 1 dB compression point and the blocking performance, but not the third-order intercept point. Low noise figure is achieved by using bond-wire degeneration and optimum device sizes and bias currents.

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Christopher Dennis Hull was born in San Francisco, CA, on November 10, 1963. He received the B.S. degree in electrical engineering from the University of California, San Diego, in 1987. He received the M.S. and Ph.D. degrees from the University of California, Berkeley, in 1989 and 1992, respectively.

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