A 20–32-GHz Wideband Mixer With 12-GHz IF bandwidth in 0.18-μm SiGe Process

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Abstract—This paper presents a 20–32-GHz wideband BiCMOS mixer with an IF bandwidth of 12 GHz. The mixer utilizes an inductive peaking technique to extend the bandwidth of the downconverted IF signal. To our knowledge, the proposed mixer achieves the widest IF bandwidth using silicon-based technologies in K-band. Analytical expressions for the conversion gain and output noise of the proposed mixer are presented. The wideband mixer is implemented using 0.18-μm BiCMOS technology and occupies an area of 0.19 mm². It achieves a conversion gain of 3 dB, a noise figure between 10.5 and 13.0 dB, and an IIP3 higher than 0.5 dBm with a power consumption of 18 mW from a 1.8-V supply.

Index Terms—BiCMOS, microwave, millimeter-wave, mixer, wideband.

I. INTRODUCTION

ADVANCED silicon-based technologies, such as CMOS and BiCMOS, have demonstrated the potential for implementing transceivers operating at high microwave (μ-wave), and millimeter-wave (mm-wave) ranges. Employing silicon-based technologies enables the design of low-cost transceivers targeting a variety of applications within K-band including ultra-wideband (UWB) automotive radar for 22–29 GHz and phased arrays for 24-GHz Industrial Scientific Medical (ISM) band [1]–[6]. However, implementing silicon-based μ-wave, mm-wave circuits is challenging due to the low quality factor of passive elements, coupling between various components, high noise figure, and parasitics that limit the operating frequency. Wideband operation is another important challenge that needs to be addressed in this frequency range to allow for the design of ultra-wideband or multi-band receivers.

A wideband μ-wave/mm-wave downconversion mixer is one of the important building blocks in wideband silicon-based receivers. This mixer should provide high conversion gain, while achieving a moderate noise figure and linearity for a wide frequency range. Several wideband techniques have been employed in the literature to increase the bandwidth of operation at K-band [7]–[12]. Verma et al. proposed a Gilbert-cell based mixer operating at 19 GHz [7]. The mixer employs two LC sections to provide a 1.4-GHz 3-dB intermediate frequency (IF)-bandwidth. This approach relies on a limited IF bandwidth extension technique. Lin et al. demonstrated that the conventional mixer can be used for wideband operation through increasing the power consumption (97 mW) [8]. However, the measured conversion gain shows that a flat gain is difficult to achieve for input frequencies higher than 20 GHz due to the parasitics at the intermediate nodes. To overcome this limitation, Ellinger used low-Q tank circuit placed at the output of the mixer to provide the IF selection [9]. For wideband operation, the low-Q tank reduces the conversion gain leading to a loss of 2.6 dB for a 26–34-GHz frequency range. A similar technique is applied for a resistive mixer to cover the 26.5–30-GHz frequency range [10]. The reported IF bandwidth in [9] and [10] is limited to 0.5 and 3.5 GHz, respectively. Another approach is employing a passive mixer with a pi-network as the load [11]. Measurements show a conversion loss of at least 8 dB with a 9–31-GHz operating frequency range; however, the 3-dB IF bandwidth is around 2 GHz. Finally, Yang et al. used a multi-layer balun to design a 16–46-GHz passive mixer with a conversion loss higher than 12 dB and an IF bandwidth of 400 MHz [12]. All the above techniques show a tradeoff between increasing the IF-bandwidth, power consumption, and conversion gain (loss). An active μ-wave/mm-wave mixer that supports a wideband flat conversion gain for UWB signals within K-band is important for silicon-based receivers to relax the noise figure and gain requirements of the low noise amplifier (LNA) and the following blocks.

In this paper, a new technique is introduced to increase the flat 3-dB IF-bandwidth of an active K-band μ-wave/mm-wave mixer to 12 GHz. The approach allows increasing the IF-bandwidth without the need for higher power consumption and is based on employing a C–L–C pi-network in the mixer topology. A similar technique was reported to extend the bandwidth of low-pass amplifiers [13] and tuned μ-wave/mm-wave LNAs [14]. The mixer follows an LNA in a silicon-based receiver, and hence, the input matching is not an important requirement as in the case of a stand-alone mixer. The proposed mixer is employed in the architecture of a μ-wave/mm-wave dual-band switchable harmonic receiver and is described very briefly in [15].

The paper is organized as follows. In Section II, the basic idea of the proposed wideband mixer is presented. Section III shows the noise analysis. Section IV demonstrates simulated and measured results. Finally, Section V concludes the paper.
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is found by decomposing the mixing square wave into sine waves and selecting the coefficient corresponding to the IF frequency. As a result, 

where 
and
is the IF frequency. Assuming a low side injection of LO
, 

where 
is the mixing frequency. Note that 
due to the frequency doublers.
The conversion gain is found by evaluating the following expression:

This model, small signal analysis is used for signals before and after the switch, i.e., in RF and IF sections indicated in Fig. 3. The time-variant analysis is considered for the term 
where 
is the RF current passing through switches and 
is the downconverted RF current to the IF frequency, as shown in Fig. 3. Using Fourier series analysis, 
is found by decomposing the mixing square wave into sine waves and selecting the coefficient corresponding to the IF frequency. As a result, 

where 
is the IF frequency, and 
is the RF frequency. Assuming a low side injection of LO
, 

where 
is the mixing frequency. Note that 
due to the frequency doublers.
The conversion gain is found by evaluating the following expression:

(3) 

(4) 

(5) 

(6)
Equation (6) highlights that the amount of peaking and 3-dB bandwidth extension depends on the value of $L_r$, $C_{s1}$, $C_{s2}$, and $R_m$. Substituting (1), (2), (5), and (6) into (3) shows that the transfer function has one real pole and two complex poles at the RF frequency side, and another real pole at the IF frequency side. The real IF pole is given by $\omega_1$, and two RF complex poles are defined by $\omega_0$ and $Q$. The real IF pole is given by $\omega_{LOd} = 1/(C_{LOd}R_{LOd})$. The two real poles reduce the gain at frequencies higher than their values, while the complex poles can be adjusted to provide peaking to reduce this decrease in the gain. This is the basic idea behind the pi-network. To achieve maximum 3-dB IF-bandwidth, both real poles should be placed at the highest possible frequency. Extending the bandwidth using this approach introduces in-band ripples. As a result, there is an optimum value for $\omega_0$ and $Q$ to maximize the bandwidth for a given in-band ripple. Fig. 4 shows the simulated frequency response of the conversion gain versus IF frequency for different ripple values. In this example, $\omega_1$ and $\omega_{LOd}$ are assumed to be 40 and 5 GHz, respectively. As depicted in Fig. 4, the 3-dB bandwidth varies between 12.3 to 14.3 GHz for ripples changing from 0.1 to 1 dB, respectively. For 0.1 dB in-band ripple, $Q = 4.05$ and $\omega_0 = 2\pi \times (30,5) \text{Grad/sec}$, while for 1 dB in-band ripple, $Q = 4.9$ and $\omega_0 = 2\pi \times (34) \text{Grad/sec}$. This example shows that both $Q$ and $\omega_0$ increase as the in-band ripple requirement is relaxed. Increasing $Q$ and $\omega_0$ helps in increasing the 3-dB bandwidth as indicated in Fig. 4. This result is more clarified by Fig. 5, where the term $i_{s,RF}(s_{RF})/V_{in}(s_{RF})$ is simulated for several values of IF in-band ripples. As depicted, increasing $\omega_0$ shifts the frequency at which $i_{s,RF}(s_{RF})/V_{in}(s_{RF})$ starts to increase to a higher frequency. For the same $\omega_{LOd}$ and $\omega_1$, increasing $\omega_0$ is associated with an increase in the value of $Q$, which increases in-band ripples. The highest bandwidth is achieved for a 1-dB IF in-band ripple in this example.

Another important design parameter for the pi-network is the location of $\omega_1$. As mentioned earlier, placing the two real poles at higher frequencies increases the bandwidth. Placing the IF
pole, \( \omega_{LD} \), at a higher frequency requires decreasing the conversion gain if \( C_{LD} \) is limited by the load capacitance. On the other hand, \( \omega_1 \) should be placed at the highest possible frequency. Hence, the minimum value of \( \omega_1 \) is limited by the location of the IF pole when it is transferred to the RF frequency, i.e., \( \omega_{1,\text{min}} > \omega_{LD} + \omega_{LOd} \). If \( \omega_1 \) is decreased below \( \omega_{LD} + \omega_{LOd} \), the IF bandwidth decreases significantly. The maximum value of \( \omega_1 \) is limited by the technology. This appears if one approximates \( \omega_1 \) in (6) as \( \omega_1 \approx 1/R_{\text{min}} C_{Q2} \), where \( R_{\text{min}} \) is the inverse of the transconductance of \( Q_1 \), and the maximum value of \( C_{Q2} \) is limited by base-emitter capacitance of \( Q_1 \). Hence, the upper value for \( \omega_1 \) is limited by the cut-off frequency of BJT transistor, i.e., \( \omega_{1,\text{max}} = 2\pi f_{T,\text{BJT}} \). However, \( \omega_{1,\text{max}} \) is lower than the cut-off frequency because of the effect of additional parasitics. In this design, \( f_{1,\text{max}} = \omega_{1,\text{max}}/2\pi \) is limited to 60 GHz (\( f_{T,\text{BJT}} < 70 \) GHz).

Fig. 6(a) shows the resultant maximum 3-dB IF-bandwidth versus \( \omega_1 \) for different values of \( \omega_{LD} \) and in-band ripples of 0.1 and 1 dB. Reducing the value of in-band ripple and/or \( \omega_{LD} \) reduces the resultant maximum IF-bandwidth, as shown in Fig. 6(a). Note that increasing \( \omega_{LD} \) by 2 GHz increases the maximum IF bandwidth by almost 3.5 GHz. The extra 1.5-GHz increase is because of the drop in the value of \( Q \) as a result of having higher \( \omega_{LD} \), i.e., in-band gain is lower. The simulated conversion gain versus \( \omega_1 \) for the same test case is shown in Fig. 6(b). Also, increasing \( \omega_1 \) slightly changes the overall IF bandwidth and conversion gain.

In the schematic-level simulation, \( f_{LD} = 5 \) GHz and in-band ripple lower than 0.3 dB result in a maximum IF-bandwidth of 13.4 GHz for \( f_1 = 40 \) GHz. Fig. 7 shows the simulated conversion gain with and without the additional pi-network. Without the pi-network, \( C_{Q1} \) and \( C_{Q2} \) are parasitic capacitances, and \( I_d = 0 \). As depicted, the 3-dB bandwidth is limited to 6.2 GHz without the pi-network, and the gain is decaying with frequency. Including the pi-network increases the flat bandwidth by 115% when compared to the conventional case on the cost of decreasing the DC gain by almost 1 dB. Another advantage of using the pi-network is that out-of-band attenuation decays faster than the conventional gain, hence providing filtering of the out-of-band signals. The difference between the DC gain for two cases is because the DC gain in IF band is a signal around \( f_{LOd} \) in RF band which is reduced by 1 dB when the pi-network is introduced. Adding multiple pi-networks does not help to increase the bandwidth. This is because multiple pi-networks have higher order that is not suitable to compensate the 20-dB/dec decay in the IF domain (first order output pole). Using a single pi-network section helps to compensate this 20-dB/dec decay in the IF pole and hence maximize the bandwidth and reduce the in-band ripples. However, the IF bandwidth can be increased by reducing the value of \( R_{\text{ld}} \), which determines the location of IF pole as mentioned earlier in this section.

It is important to note that the designed mixer targets maximizing the IF bandwidth. Accordingly, the proposed designed methodology is derived for this target. If this mixer is used for
a constant IF receiver such that the LO tracks the changes of the RF frequency, then a different design methodology should be used to size the pi-network. In such a case, the effect of output pole does not appear (constant IF), and peaking should be avoided to have a constant gain versus the RF frequency. In this design, peaking in the RF domain is adjusted to compensate the effect of the IF pole, hence increasing the overall IF bandwidth.

2) Frequency Doubler: The frequency doubler architectures is shown in Fig. 8(a). The basic operation of the frequency doubler can be summarized as follows: Two out-of-phase LO signals with the same frequency ($f_{LO}$) are applied to gates of $M_{N1}, N2$. The output voltage $V_o$ is taken from sources of $M_{N1}, N2$. Due to the differential configuration, the fundamental component of LO does not appear at the output. However, because of the nonlinearity of the MOSFET, twice the operating frequency ($2f_{LO}$) appears at the output. The inductor $L_r$ is placed at the output to resonate with the parasitic capacitance, hence increasing the output swing. Yang et al. used the same architecture without the capacitor $C_b$ to self-bias the doubler [17]. However, in this design, $C_b$ is added to increase the output swing at $V_o$. The main reason is providing another signal path, through $C_{biast}$, that increases the output swing at $V_o$, as shown in Fig. 8(b). The signal passing through path 2 is added coherently to the original signal passing through path 1 at the node of $V_o$. The pMOS transistor performs self-biasing of the doubler and provides a high impedance node to generate path 2. If the nMOS drain is directly connected $V_{dd}$, path 2 will not appear and the doubler output voltage will be reduced. Fig. 9 shows the schematic-level simulated output amplitude at $2f_{LO}$ versus the value of $C_b$ for different values of $V_{LO}$. As indicated for $V_{LO} = 500$ mV, there is an optimum value of 60 fF for $C_b$, resulting in a maximum output voltage amplitude of 235 mV at $2f_{LO} (f_{LO} = 10$ GHz, and $I_{biast} = 1$ mA). Reducing the amplitude of $V_{LO}$ below 500 mV changes the value of optimum capacitor.

The mixing stage, shown in Fig. 2, requires a differential signal to drive BJT switches. This differential signal is generated using two frequency doublers. These doublers require a quadrature LO signal. Under the ideal condition, a virtual ground appears at the middle of the inductor $L_r$ and two identical tank circuits appear at the output of the two doublers. These tank circuits are designed to resonate at the required mixing frequency, i.e., $2f_{LO}$. A simple LC phase shifter can be used to generate the quadrature signal from an external differential LO signal (Fig. 10). $L_{ph}$ and $C_{ph}$ are adjusted to obtain an exact 90° phase shift, have input matching of 50 Ω, and make the voltage amplitudes of $LO$ and $LO_{180}$ the same. These conditions are fulfilled as following: $R_{ph} = 50$ Ω, $\sqrt{L_{ph}/C_{ph}} = R_{ph}$, and $1/\sqrt{L_{ph}C_{ph}} = \omega_{LO} = 2\pi \cdot 10.25$ GHz. Due to phase mismatches between the generated LO quadrature signals, the resulting output voltage amplitude of the doubler circuit is lower than the ideal case. Using the doubler equivalent circuit in Fig. 11, the differential output voltage $\Delta V_m$ between the two driving voltages $V_{m,0}$ and $V_{m,180}$ is found by decomposing the half-sine waveform using Fourier series analysis. Then, differential output voltage at $2f_{LO}$ is found by solving the passive circuit in Fig. 11:

$$\Delta V_m(2f_{LO}) = V_{m,0}(2f_{LO}) - V_{m,180}(2f_{LO}) = i_{2}(2f_{LO}) \cdot R \cdot \left( 1 - \frac{i_{1}(2f_{LO})}{i_{2}(2f_{LO})} \right)$$  (7)

where $R = 1/gm_{N1}$ is the resistance seen at the output, and $gm_{N1}$ is the transconductance of $M_{N1}$. In the above analysis, it is assumed that $s^2L_rC_p/2 = -1$, which is the resonance condition at $2f_{LO}$. This assumption divides the circuit into two half circuits as a virtual ground appears at the middle of the inductor at $2f_{LO}$. In the ideal case, there is a time shift of $1/4f_{LO}$ leading
are the amplitude and phase of \( V_{LO} \), the

differential output amplitude decreases the differential output amplitude, and \( C_{th} \) is the channel noise factor. The first term in (9) is due to mismatches at the output because it is upconverted around \( \omega_{LOd} \), \( 3\omega_{LOd} \), \( 5\omega_{LOd} \), etc., after the mixing operation [18]. Therefore, the output noise voltage spectral density due to the RF transistor white noise \( \frac{v_{2n,RF}}{\Delta f} = 4kT\gamma g_{m,RF} \) is given by

\[
\frac{v_{2n,RF}}{\Delta f}(\omega_{IF}) = \frac{4kT\gamma}{g_{m,RF}}
\]

\[
\times \left( \frac{V_{IF,LO}(S_{IF})}{V_{in}(S_{IF})} \right)^2 + \left( \frac{V_{IF,LO}(S_{IF})}{V_{in}(S_{LOD} - S_{IF})} \right)^2
\]

where \( k \) is Boltzmann constant, \( T \) is the temperature in Kelvin, and \( \gamma \) is the channel noise factor. The first term in (9) is due to downconverted upper side-band noise (above LO frequency), and second term is due to the lower side-band noise (below LO frequency). Equation (9) indicates that increasing \( g_{m,RF} \) increases the output noise; however, the input-referred noise is decreased.

### III. Noise Analysis

#### A. RF Transistor Noise

The thermal noise of the RF transconductance appears at the output IF frequency. Due to the mixing operation, noise around \( \omega_{LOd}, 3\omega_{LOd}, 5\omega_{LOd}, \ldots \) etc., fold back into the baseband frequency of interest [18].

\[
\text{Due to amplitude and phase mismatches between generated LO quadrature signals, the differential output amplitude is changed as indicated by (8). Amplitude mismatches between } i_{10} \text{ and } i_{20} \text{ may increase or decrease the output amplitude, while phase mismatches between } \theta_1 \text{ and } \theta_2 \text{ decreases the differential output amplitude (} |\text{Re}(e^{j(\theta_1 - \theta_2)})| < 1 \text{). This result is verified using schematic-level simulations and presented in Fig. 12 for different values of } V_{LO} \text{. In the ideal case for } V_{LO} = 500 \text{ mV, the maximum differential amplitude is 466 mV (} I_{bias} = 1 \text{ mA, and } f_{LO} = 10 \text{ GHz). For a phase mismatch of } \pm 20^\circ, \Delta V_{m} \text{ reduces to 454 mV. This shows only a variation of 2.5% in the amplitude. Similar conclusion is observed for the different values of } V_{LO} \text{. As a result, mismatches due to process variation do not affect the functionality of this differential frequency doubler noticeably.}

#### B. Load Resistance Noise

The load noise appears at the output directly. In this design, a resistive load is used for the mixer. Since the output pole is lower than the operating bandwidth, the thermal noise of the resistance is attenuated at frequencies above the output pole \( \omega_{LD} \). Hence, the output voltage spectral density due to \( R_{LD} \) is given by

\[
\frac{v_{2n,R_{LD}}}{\Delta f}(\omega_{IF}) = 2\cdot\frac{2}{\Delta f} \cdot \frac{Z_L(\omega_{IF})^2}{1 + (\omega_{IF} C_{LD} R_{LD})^2}.
\]

The multiplication by 2 that appears in (10) is due to the presence of two resistive loads at the output of the mixer.

#### C. Switching Transistors Noise

The noise in switching transistors is due to base and collector noise currents \( \frac{\Delta I_{b,c}}{\Delta f} \). The noise voltage is

\[
\frac{v_{2n,c,b}}{\Delta f} = \frac{V_T}{2} \cdot \frac{2}{\Delta f} \cdot |Z_L(\omega_{IF})|.
\]

The parameter \( V_T = kT/q \) is the thermal voltage, \( q \) is the electron charge constant, and \( \beta \) is the current gain. The parameters \( g_{m,Q1}, I_{b,Q1}, \) and \( I_{c,Q1} \) are...
the transconductance, base, and collector currents of $Q_1$, respectively.

The noise appearing at the output due to the switching transistors comes from two major mechanisms. The first one appears when both switches are on and the differential LO amplitude is crossing the zero point. Similar approach reported in [18] is used to find the amount of output noise in case a BJT is used instead of a MOS transistor. Assuming an LO signal with a sine wave shape, the output voltage noise spectral density is given by

$$\frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF}})}{\Delta f} = \frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF}})}{\Delta f} \cdot |Z_L(\omega_{\text{RF}})|^2,$$

$$\frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF}})}{\Delta f} = \frac{2kT \cdot I_c \cdot Q_1}{\pi A_{\text{LO}} \cdot L_d} \cdot \frac{R_{\text{f}}^2}{1 + (\omega_{\text{RF}} C_{\text{Ld}} R_{\text{Ld}})^2}$$ (11)

where $\frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF}})}{\Delta f}$ is the output current noise power spectral density when both switches are on and $A_{\text{LO}}$ is the amplitude of the sine wave driving the switches. The calculated output noise in (11) is due to the collector noise current only. Base noise does not appear at the differential output because it is injected into the common node of switches. Hence, it appears at both outputs with the same amount and is differentially removed.

The second mechanism appears only at $\mu$-wave/mm-wave frequencies. This is because, under a hard switching condition, transistor $Q_1$ is not degenerated with a high impedance when one of the switching transistors is on while the other one is off. Hence, the noise due to $Q_1$ appears at the output, and the overall output noise is increased. This situation is not the case for low gigahertz mixers, because the effect of parasitic capacitance at the emitter of $Q_1$ can be neglected, and the large degeneration resistance prevents the noise from appearing at the output. Fig. 13 shows the equivalent circuit used to find the contribution of $Q_1$ to the output noise. In this case, $Q_1$ is degenerated with the impedance $Z_s(\omega_{\text{RF}})$, consisting of $L_s$, $C_{s1}$, $C_{s2}$, and $r_{\text{fRF}}$ (output resistance of RF transistor). Since the circuit switches on and off periodically with a duty cycle of 50%, the output noise that is shaped by conversion gain transfer function is convoluted with delta functions appearing at $\omega_{\text{LO}}, 2\omega_{\text{LO}}, 3\omega_{\text{LO}}, \ldots$ in the frequency domain, and the noise appearing at these frequencies is folded back to baseband (Fig. 14).

Superposition is used to find the noise contribution of each noise source at the output IF frequency independently. The RF noise current, $i_{\text{RF}}$, in Fig. 13, is initially found by solving the equivalent noise small signal model for three different noise sources of $Q_1$, assuming $Z_s(\omega_{\text{RF}}) \ll \beta / g_{\text{mQ}}$ as follows:

$$\frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF}})}}{\Delta f} = \frac{1}{1 + g_{\text{mQ}} Z_s(\omega_{\text{RF}})} \cdot \frac{2}{\mathcal{N}_{n_{\text{RF}}}^2},$$

$$\frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF}})}}{\Delta f} = \frac{2}{\mathcal{N}_{n_{\text{RF}}}^2} \cdot \frac{g_{\text{mQ}}}{1 + g_{\text{mQ}} Z_s(\omega_{\text{RF}})} \cdot \frac{2}{\mathcal{N}_{n_{\text{RF}}}^2},$$

$$\frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF}})}}{\Delta f} = \frac{2}{\mathcal{N}_{n_{\text{RF}}}^2} \cdot \frac{g_{\text{mQ}}}{1 + g_{\text{mQ}} Z_s(\omega_{\text{RF}})} \cdot \frac{2}{\mathcal{N}_{n_{\text{RF}}}^2}. \quad (12)$$

Then, the downconverted noise is obtained by convolving each output noise current in (12) with the delta functions appearing due to the switching in the frequency domain. This approach results in the downconverted noise in the IF bandwidth. In this analysis, both the upper side and the lower side noise should be considered in the final expression which means $Z_s(\omega_{\text{RF}})$ needs to be evaluated at both $\omega_{\text{RF},1} = n \omega_{\text{LO}} + \omega_{\text{IF}}$ and $\omega_{\text{RF},2} = n \omega_{\text{LO}} - \omega_{\text{IF}}$ $(n = 1, 3, 5, \ldots)$. The output noise voltage due to each noise source of $Q_1$ is then given by

$$\frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF})}}{\Delta f} = \sum_{n_{\text{LOd}}} \frac{4}{\pi^2 n^2} \cdot |Z_L(\omega_{\text{RF}})|^2 \cdot \left( \frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF}})}}{\Delta f} \cdot \sum_{\omega_{\text{LOd}}} \frac{4}{\pi^2 (n \omega_{\text{LOd}} + \omega_{\text{IF}})} \right) \cdot \left( \frac{\mathcal{N}_i^{\text{noise}}(\omega_{\text{RF})}}{\Delta f} \cdot \sum_{\omega_{\text{LOd}}} \frac{4}{\pi^2 (n \omega_{\text{LOd}} - \omega_{\text{IF}})} \right) \quad (13)$$
where the index $X$ stands for either, $i_c$, $i_b$, or $r_b$. The multiplication by two that appears before the summation in (13) is because of two switching transistors; each is operating in a different time window. The term $4/\pi^2 n^2$ appears due to the effect of the delta functions in Fig. 14. The total output voltage noise spectral density due to the second mechanism is found by summing the three output voltage noise spectral densities resulting from (13) for $X = i_c$, $i_b$, and $r_b$.

### D. Total Output Noise

The total output noise is found by summing all the noise terms in (9), (10), (11), and (13). The noise analysis presented in this section is verified using schematic-level simulations. Fig. 15 shows the schematic-level simulated output noise voltage spectral density versus the IF frequency and the total output noise predicted by (9), (10), (11), and (13). As indicated, the theory matches closely the simulation results. The noise within the IF bandwidth is mainly due to the noise generated by switches. This analysis proves that the noise of switches due to the second mechanism is an important factor for active mixers operating at $\mu$-wave/mm-wave frequencies.

### IV. SIMULATION AND EXPERIMENTAL RESULTS

Table I shows the circuit element values for the implemented mixer. The wideband mixer is fabricated using 0.18-\mu m BiCMOS technology provided by Jazz Semiconductor. The die micrograph is shown in Fig. 16, where the total area is 0.19 mm$^2$, excluding pads and output buffer. The chip includes the wideband mixer, phase shifter, and output buffer shown in Figs. 2, 10, and 17, respectively. The $\mu$-wave/mm-wave input, output, LO signal, and DC biasing signals are applied and monitored using on-wafer probing to reduce the losses and mismatches introduced by the measurement setup. A ground–signal–ground (GSG) probe is used to apply the input signal, and ground–signal–ground–signal–ground (GSGSG) differential probes are used to derive the LO signal and to measure the IF signal at the output. Since the RF input port is not matched to 50 \Omega, the effective RF input power is estimated using the measured reflection coefficient ($S_{11}$) of the RF input port. An 8-pin DC probe is used to apply the required DC biasings. The external LO is running at 10 GHz (effective mixing frequency of 20 GHz) with an input power of 5 dBm. The effect of the output buffer is de-embedded from the mixer+buffer measurements. The buffer is added at the output of the mixer to drive the 50-\Omega input impedance of the spectrum analyzer. The schematic of the buffer circuit is shown in Fig. 17.

The conversion gain is measured by applying an input RF signal using Agilent N5230A network analyzer and measuring the output using the spectrum analyzer while the RF input frequency varies from 20–35 GHz. The simulated and measured conversion gain after de-embedding the buffer and cable loss effects are shown in Fig. 18. The measured conversion gain is 3 dB with a 3-dB IF bandwidth of 12 GHz. The post-layout simulated conversion gain is 4 dB with a bandwidth of 13 GHz. The difference between the simulated bandwidth and measured one could be due to the inaccurate models of the transistors that may lead to extra capacitance, and/or the effect of process variation.

The simulated and measured noise figure versus the IF frequency are shown in Fig. 19, where the noise figure varies from 10.5 to 13.0 dB. As indicated, the noise figure initially decreases as the IF frequency increases. This is because the pi-network impedance increases with the RF frequency to provide the required peaking. As a result, the noise contribution of the switches is reduced. The measured RF to IF and LO to IF isolation are better than 15 and 35 dB over the entire IF band, respectively.

A two-tone $II_{P3}$ measurement is performed for the mixer and the results are shown in Fig. 20 for the 1–12-GHz IF-frequency range. The two tones are applied with the same amplitude and a frequency offset of 50 MHz. The measured $II_{P3}$ changes from 0.5 to 2.6 dBm across the entire frequency range. Also, simulation results showed that the $II_{P3}$ reduced by 0.3 dB when the tone spacing is increased to 2 GHz.

### TABLE I

| Circuit Element Values for the Implemented Mixer |
|---|---|---|---|---|
| $L_s$ (pH) | $C_{s1}$ (IF) | $C_{s2}$ (IF) | $L_n$ (pH) |
| 220 | 130 \footnote{a including parasitics of $Q_1$ and $M_{RF}$.} | 300 \footnote{b including the parasitics of $M_{N1}$ and $M_{N2}$.} | 450 |
| $R_d$ (\Omega) | $C_d$ (IF) | $R_b$ (\Omega) | $C_b$ (IF) |
| 150 | 300 | 1.7 | 170 |

Fig. 15. Schematic-level simulated and predicted output noise voltage spectral density.

Fig. 16. Die-photo of the wideband mixer with an area of 0.19 mm$^2$ (pads and buffer are not included).
the mixer consumes 7 mA from a 1.8-V supply, and each doubler consumes 1.5 mA from the same supply. The buffer consumes 4 mA from a similar supply. The performance of the proposed wideband mixer and comparison with other existing mixers around the same frequency range are summarized in Table II. The proposed mixer achieves the widest IF bandwidth, and comparable $NF$ and $IIP_3$ compared to the state-of-the-art.

V. CONCLUSION

A wideband BiCMOS mixer for downconverting a 12-GHz wideband signal (20–32 GHz) was implemented in this paper. The mixer utilizes an inductive peaking technique to extend the IF bandwidth and a frequency doubler with enhanced output swing to reduce the LO power consumption. Analytical expressions for the conversion gain and output noise voltage were developed to highlight the design tradeoffs. The mixer was implemented using 0.18-μm BiCMOS technology with an area of...
0.19 mm². Measurements showed that the mixer is able to operate across a 12-GHz signal, while keeping the gain almost constant. A conversion gain of 3 dB, a 3-dB IF-bandwidth of 12 GHz, a noise figure between 10.5 and 13.0 dB, and a linearity higher than 0.5 dBm with a power consumption of 18 mW from a 1.8-V supply are achieved.

**REFERENCES**


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