

行政院國家科學委員會專題研究計畫 成果報告

先進可選性高鏡像抑制 CMOS 射頻接收機架構 研究成果報告(精簡版)

計畫類別：個別型

計畫編號：NSC 97-2221-E-009-171-

執行期間：97年08月01日至98年07月31日

執行單位：國立交通大學電信工程學系(所)

計畫主持人：孟慶宗

處理方式：本計畫可公開查詢

中 華 民 國 98 年 10 月 19 日

行政院國家科學委員會補助專題研究計畫 ☒ 成果報告
☐ 期中進度報告

先進可選性高鏡像抑制 CMOS 射頻接收機架構

計畫類別：☒ 個別型計畫 ☐ 整合型計畫

計畫編號：NSC 97-2221-E-009-171

執行期間：97 年 8 月 1 日至 98 年 7 月 31 日

計畫主持人：孟慶宗 國立交通大學電信系

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執行單位：國立交通大學電信系

中 華 民 國 98 年 10 月 19 日

Abstract—This project demonstrates single-band and dual-band low-IF image rejection receivers using Weaver and Hartley architectures. A dual-band reconfigure receiver is achieved by changing the polarity of the quadrature LO signals; thus, both the first- and second-stage mixers can be highly reused. On the other hand, at a higher operating frequency, the microwave quadrature signal generation methods are included to generate highly accurate high-frequency quadrature signals.

Index Terms — Image rejection, low-IF, dual-band, Weaver, Hartley.

I. Introduction

Compared with the two (or multiple)-separate-system solutions for multi-band applications, the dual-band (or multi-band) reconfigurable system has a relatively high integration level. For dual-band transceivers, the direct-conversion architectures were demonstrated in the past. Besides, some dual-band dual-conversion systems were demonstrated while the signal was down-converted to the baseband ($IF_2=0$) at the second stage; thus, there is no second image problem but the flicker noise and dc offset problems arise and need more efforts to solve. Most of these zero- IF multi-band receivers require more than one local oscillator. Since $IF_2=0$, it is difficult to use one VCO covering both bands. The extra VCO may cause substrate cross-talk and spurious signals.

II. Weaver-Hartley Image Rejection Architecture

First, for a single-band application, the trigonometric analysis can be applied. Weaver architecture consists of two cascaded complex mixers. Received signals are twice down-converted to a low-IF band by the first LO and second LO signals. The angular frequencies of the desired, first image, and second image, first LO, and second LO signals are denoted as ω_{RF} , ω_{IM1} , ω_{IM2} , ω_{LO1} and ω_{LO2} , respectively. The angular frequencies of the IF signal after the first and second down-conversions are defined as ω_{IF1} and ω_{IF2} , respectively. The relations of the signals defined above can be expressed as:

$$\omega_{RF} - \omega_{LO1} = \omega_{LO1} - \omega_{IM1} = \omega_{IF1} \quad (1a)$$

$$\omega_{RF} - (\omega_{LO1} + \omega_{LO2}) = (\omega_{LO1} + \omega_{LO2}) - \omega_{IM2} = \omega_{IF2} \quad (1b)$$

$$\omega_{IF1} = \omega_{LO2} + \omega_{IF2} \quad (1c)$$

The wire connection of the Weaver system with detailed mathematical analyses is shown in Fig. 1. The desired RF signal, first image signal, and second image signal are denoted as $a_{RF}\cos(\omega_{RF}t)$, $a_{IM1}\cos(\omega_{IM1}t)$, and $a_{IM2}\cos(\omega_{IM2}t)$, respectively. Figure 1(a) indicates the results at each node of the Weaver system when the input signals are RF and first image signals. The first image signal and desired RF signal are converted to the same IF_1 frequency (ω_{IF1}) but with opposite signs of the quadrature signals after the first down-conversion. The high-frequency term ($2\omega_{LO1}+\omega_{IF1}$) can be eliminated by the low-pass or band-pass nature of the first-stage mixer. The other two signals entering the second-stage mixers are down-converted to ω_{IF2} and $(2\omega_{IF1}-\omega_{IF2})$ bands respectively. Therefore, the shifted-out image signal can be easily filtered-out by the low-pass filter at IF_2 stage.

On the other hand, the RF and second image signals are down-converted to ω_{IF1} and $(\omega_{IF1}-2\omega_{IF2})$, respectively, after the first down-conversion, as shown in Fig. 1(b). Although the two signals do not locate at the same frequency, they are still very close and hard to be separated by a narrow-band filter. After the second conversion, the two signals locate at the same frequency (ω_{IF2}) but with opposite signs of quadrature signals. That is, the Weaver system can reject the first image signal but it has no ability to reject the second image signal. To solve this problem, a polyphase filter is employed after the Weaver system. Because the polyphase filters can reject the negative-frequency signal but pass the positive-frequency signal. As a result, the image signals located at the negative spectrum can be highly rejected. The second-stage complex mixers with the subsequent polyphase filter can be called the Harley system.

Instead of the conventional trigonometric analysis, the complex-signal diagrammatic explanation shown in Fig. 2 gives a more direct insight to the RF system architecture.

1) Rejection of the First Image Signal

The RF signal and quadrature LO signals can be denoted as $\cos \omega_{RF}t$ and $e^{-j\omega_{LO1}t}$, respectively. On the other hand, the first

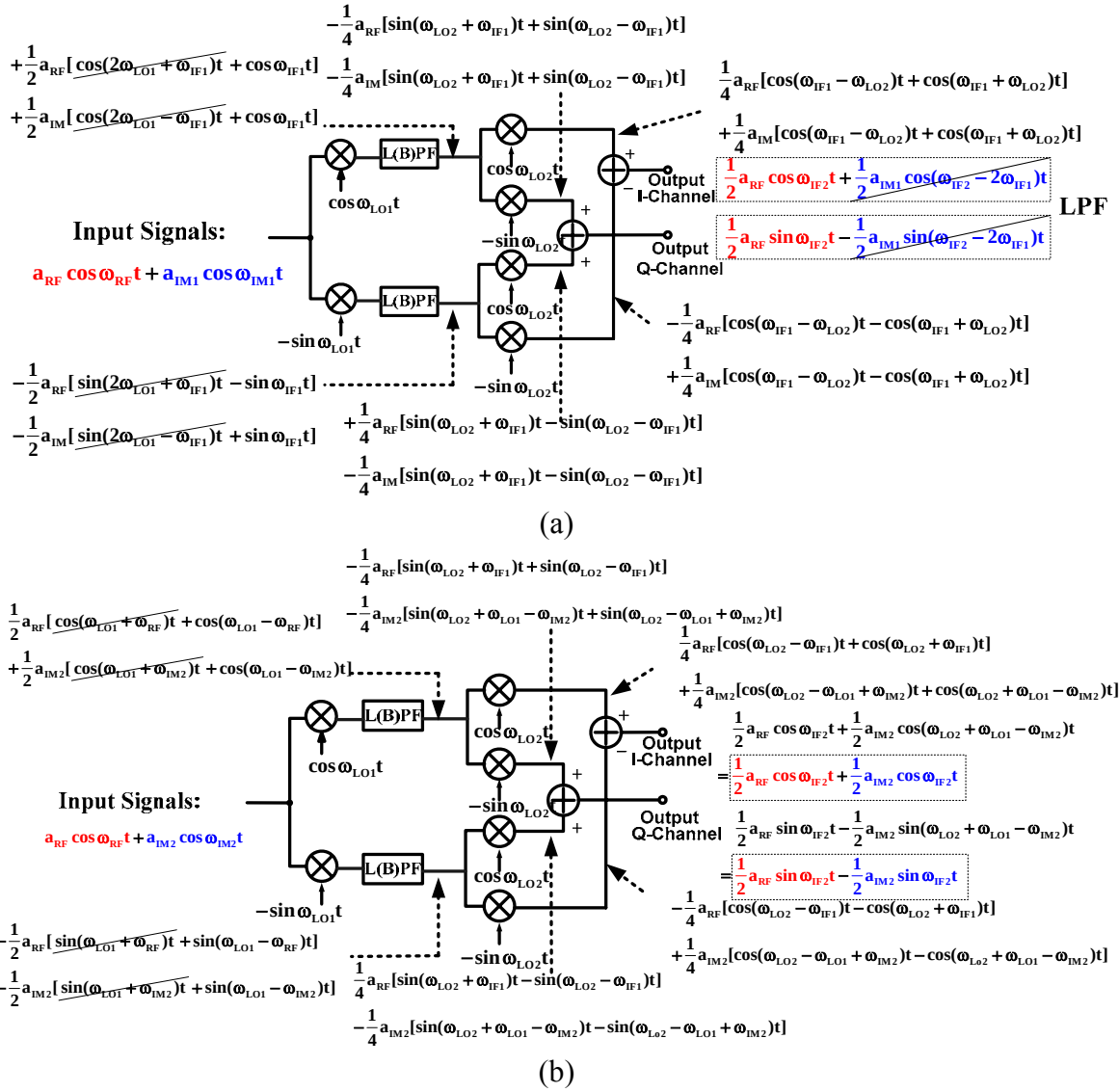


Fig. 1. Block diagram of the Weaver architecture (a) when input signals are desired RF signal and first image signal (b) when input signals are desired RF signal and second image signal.

image signal and secondary image signal can be expressed as $\cos \omega_{IM1}t$ and $\cos \omega_{IM2}t$, respectively. Figure 2(a) shows the corresponding spectra of the ω_{RF} , ω_{IM1} , and ω_{IM2} signals.

Since the quadrature LO_1 signal ($e^{-j\omega_{LO1}t}$) is located at the negative spectrum, the RF signal shifts to the left after down-conversion mixing. In other words, after the complex LO_1 down-conversion, the positive RF complex frequency shifts to the positive ω_{IF1} complex frequency and the negative RF complex frequency shifts to the negative frequency of $(\omega_{LO1} + \omega_{RF1})$ as shown in Fig. 2(b). Thus, the negative frequency spectrum in Fig. 2(b) can be omitted in the analysis for the sake of simplicity.

Next, the signals in Fig. 2(b) are down-converted further by the subsequent second-stage complex mixer. Similarly, the quadrature LO_2 ($e^{-j\omega_{LO2}t}$) signal of the second-stage complex mixer is located at the

negative spectrum. Therefore, the signals in Fig. 2(b) shift to the left in the spectrum again, as shown in Fig. 2(c). This process can be expressed by the complex signal multiplication as:

$$\cos \omega_{RF}t \times e^{-j\omega_{LO1}t} \times e^{-j\omega_{LO2}t} = \frac{e^{j\omega_{IF2}t} + e^{-j(\omega_{RF} + \omega_{LO1} + \omega_{LO2})t}}{2}, \quad (2a)$$

Similarly, the shifted IM_1 and IM_2 signals can be represented as:

$$\cos \omega_{IM1}t \times e^{-j\omega_{LO1}t} \times e^{-j\omega_{LO2}t} = \frac{e^{-j(2\omega_{IF1} - \omega_{IF2})t} + e^{-j(\omega_{IM1} + \omega_{LO1} + \omega_{LO2})t}}{2} \quad (2b)$$

$$\cos \omega_{IM2}t \times e^{-j\omega_{LO1}t} \times e^{-j\omega_{LO2}t} = \frac{e^{-j\omega_{IF2}t} + e^{-j(\omega_{IM2} + \omega_{LO1} + \omega_{LO2})t}}{2}. \quad (2c)$$

The RF signal is down-converted to the

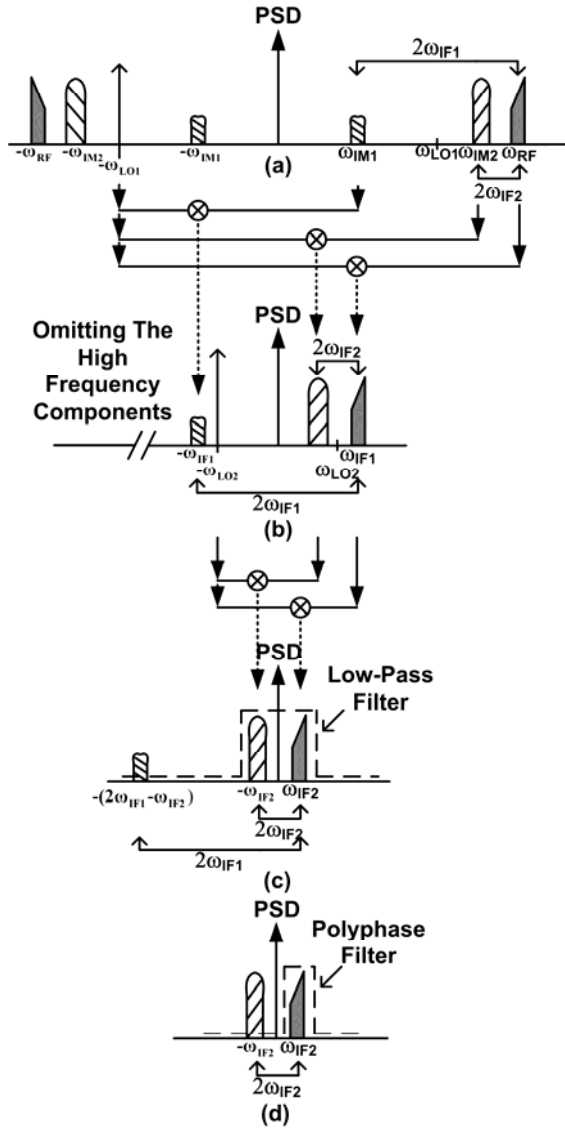


Fig. 2. (a) The RF, IM_1 , and IM_2 signal spectra; (b) the signals after down-converting by the first stage complex mixer; (c) the signals after down-converting by the second stage complex mixer; and (d) the signals filtered by the final IF polyphase filters.

positive ω_{IF2} complex frequency as shown in Fig. 2(c). On the other hand, the first image signal, IM_1 , is down-converted to the negative $(2\omega_{IF1}-\omega_{IF2})$ complex frequency. Because of the dual conversion phenomena, the first image signal can be easily filtered out by the low-pass filter as shown in Fig. 2(c). Normally, the frequency response of the IF circuits can serve as the low-pass filter as long as the ω_{IF1} is large enough.

2) Rejection of the Secondary Image Signal

The dual-conversion down-converter still suffers from the secondary image signal problem, and the frequency of the secondary image signal is $2\omega_{IF2}$ away from the RF frequency as shown in Fig. 2(c). The secondary image signal is also shifted downward twice by the two complex mixers of the Weaver system. However, the

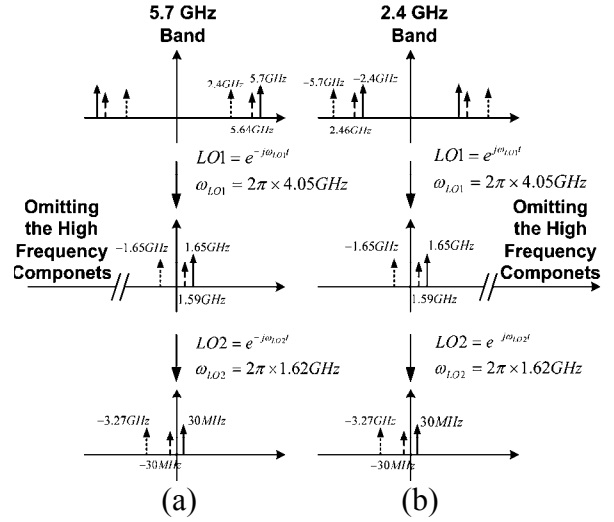


Fig. 3. All the high frequency harmonics are neglected. (a) The frequency shifting by the negative LO_1 and negative LO_2 complex signals when the RF frequency is around 5.7-GHz band; and (b) the frequency shifting by the positive LO_1 and negative LO_2 complex signals when the RF frequency is around 2.4-GHz band.

down-converted secondary image signal is not shifted outside the frequency response of the low-pass filter as with the first image signal, IM_1 . Instead, the frequency of the down-converted secondary image signal is shifted to the negative ω_{IF2} spectrum, while the down-converted RF signal is located at ω_{IF2} , as shown in Fig. 2(c). This down-converted secondary image signal cannot be filtered out by the low-pass filter and disturbs the down-converted desired RF signal. On the other hand, the polyphase filter, which is an asymmetrical complex notch filter, can filter out the down-converted secondary image signal by eliminating the signals in the negative spectrum, as illustrated in Fig. 2(d).

3) Band Selection

The band selection at the 5.7-GHz band is depicted in Fig. 3(a). The solid-line, dotted line, and the dashed-line arrows represent the RF signal, the IM_1 signal, and the IM_2 signal, respectively. The RF signal, as well as the undesired IM_1 and IM_2 signals, are located at the positive spectrum, and down-converted by the $e^{-j\omega_{LO1}t}$ and $e^{-j\omega_{LO2}t}$ LO signals, respectively. As discussed above, the Weaver-Hartley architecture can reject both the first and secondary image signals. On the other hand, when the desired signals are around 2.4-GHz, the signals located at the negative spectrum are down-converted by the $e^{j\omega_{LO1}t}$ and $e^{-j\omega_{LO2}t}$ LO signals, as shown in Fig. 3(b).

III. Circuit Design and Measurement Results

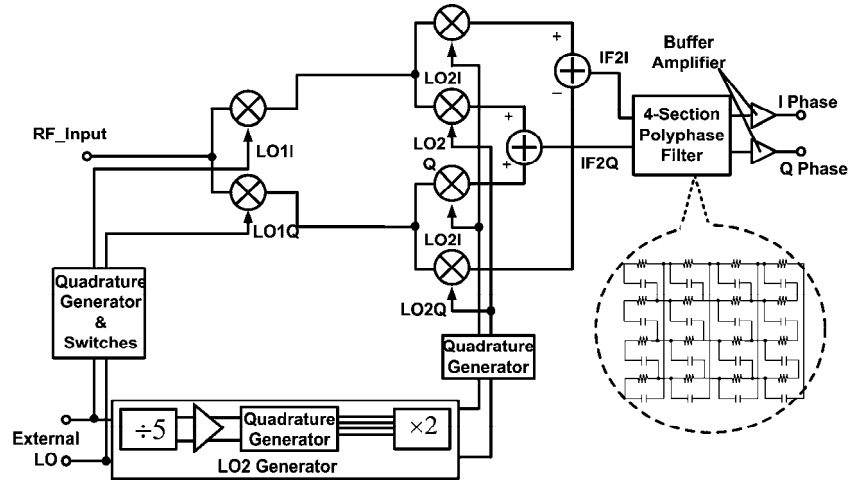


Fig. 4. Block diagram of the demonstrated Weaver-Hartley double-quadrature down-converter.

(A)

A 2.4/5.7-GHz dual-band low-*IF* system for WLAN 802.11 a/g applications is demonstrated. Since the low-*IF* architecture keeps the *IF* frequency flexible, the LO_1 and LO_2 can be set as the fractional multiple of each other. It is noteworthy that the spurious issue should still be taken into account. When the desired signal is at 5.7-GHz, the 2.4-GHz band is set at the image signal of the Weaver structure and vice versa. The LO_1 frequency is chosen to be 4.05-GHz, which is halfway between 2.4-GHz and 5.7-GHz. The 2.4/5.7-GHz band selection is achieved by changing the polarization of the complex LO_1 (the first local oscillator) signals. Since a low-*IF* architecture is employed in this work, it is very easy to perform a frequency arrangement that only needs one on-chip VCO. An LO generator consisting of a new 50% duty cycle divide-by-five circuit and a truly balanced multiplier is employed to generate the 1.62-GHz LO_2 (the second local oscillator) signal for the Weaver system and to increase the image rejection performance of the entire system. Thus, the *IF* frequency of this low-*IF* down-converter is around 30-MHz. As a result, the extra VCO is not needed, and its complexity is reduced. Compared with previous literatures, the demonstrated Weaver-Hartley system has the competitive rejection performance of the first image and secondary image signals because of the double-quadrature Hartley system and the coherence between the LO_1 and LO_2 signals. Besides, the mixer stages can be reused in our Weaver-Hartley structure.

The block diagram of the demonstrated Weaver-Hartley down-converter is shown in Fig. 4. The first-stage and second-stage

complex mixers contain two and four Gilbert multipliers, respectively. The 4.05-GHz LO_1 signal is fed externally and the 1.62-GHz LO_2 signal is generated from the LO_1 signal using the divide-by-five circuit and the frequency doubler. The quadrature generators shown in Fig. 4 are two-section polyphase filters to generate differential-quadrature signals from differential signals. A four-section polyphase filter and an *IF* buffer amplifier are located at the end of the down-converter. Each sub-circuit is discussed in the following sections.

1) LO Frequency Generator

The image rejection depends on the magnitude and phase accuracies of LO and RF signals. The imperfect LO and RF signals are inevitable in the circuit fabrication, and thus the image rejection ratio degrades. Many methods such as off-chip or electronic tuning have been developed to increase the image rejection performance. The coherence between the LO_1 and LO_2 signals improves the image rejection.

In order to achieve coherent LO signals, an accurate LO generator consisting of a 50% duty cycle divide-by-five circuit and a truly balanced frequency doubler is employed to generate coherent signals with less phase errors. As a result, the frequency of the LO_1 signal is 2.5 times that of the LO_2 signal, and thus only one VCO is needed.

As shown in Fig. 4, the LO generator consists of a divide-by-five circuit, a buffer amplifier, a quadrature generator, and a frequency doubler. The block diagram of a 50% duty cycle divide-by-five circuit and its associated timing diagram are shown in Fig. 5(a) and (b), respectively. Non-50% duty

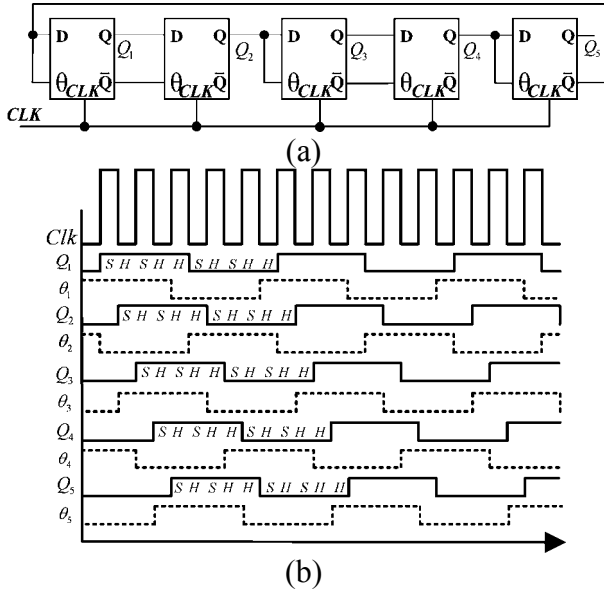


Fig. 5. (a) The block diagram of the 50% duty-cycle divide-by-five circuit and (b) the timing diagram.

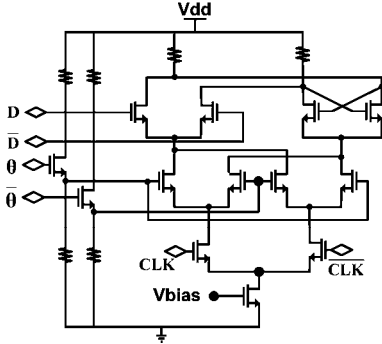


Fig. 6. The schematic of the current switchable source-coupled logic D flip-flop.

cycle signals preceding a polyphase filter (quadrature generator) generate high-order harmonics and lead to large phase error, especially when the following circuit is a hard-switched Gilbert mixer (second-stage mixer). Therefore, the 50% duty cycle is critical for the Weaver-Hartley image rejection system. The divide-by-five circuit consists of five current switchable source-coupled logic D flip-flops and employs the SHSHH (sample-hold-sample-hold-hold) scheme to achieve the 50% duty cycle. The schematic of the current switchable source coupled logic D flip-flop is shown in Fig. 6. Two extra differential pairs with control inputs θ and $\bar{\theta}$ are inserted between the sample-and-hold stage and the clock stage of the traditional D flip-flop to commutate the clock current signal and switch the triggering edge of the clock for sample-and-hold. The input data is sampled when clock and θ are the same; on the other hand, the data is held when they are opposite. Therefore, this current switchable D flip-flop can be triggered at both the positive edge and the negative edge.

A conventional Gilbert multiplier has a

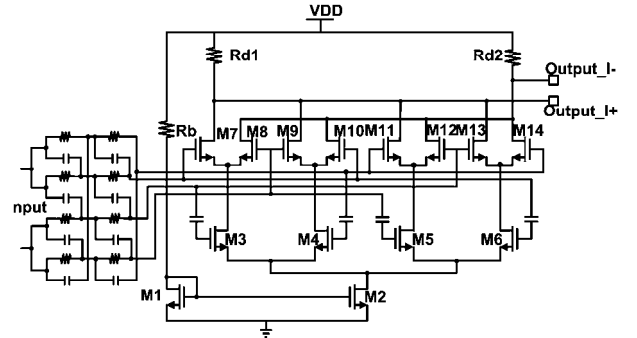


Fig. 7. The truly balanced multiplier and its input quadrature generator.

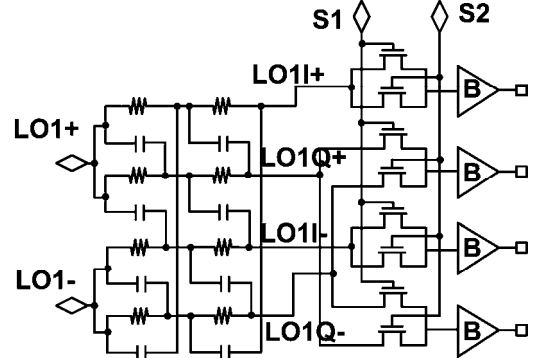


Fig. 8. The LO₁ signal polarization switching circuit used to perform the band selection.

time-delay between the signal through the bottom differential pair and the signal through the top differential pair and thus gives rise to asymmetrical output waveforms. To remedy this problem, a fully symmetrical multiplier circuit with equal delay paths between two input signals is adopted in this work. Figure 7 shows a truly balanced frequency doubling multiplier, consisting of two conventional Gilbert multipliers. The in-phase and quadrature-phase input signals connect to the top and bottom stages of one multiplier, respectively. On the other hand, the input signals of the other multiplier are connected in the opposite way. Therefore, one multiplying path contains a phase lead and the other path contains a phase lag. By summing the outputs of the two constituent multipliers, the phase delay is cancelled in the combined signals. The two-section polyphase filter in Fig. 7 generates quadrature signals needed by the truly balanced frequency doubler.

2) LO Switching Circuit

The 2.4/5.7-GHz band selections are achieved by changing the polarity of the LO₁ signals. Figure 9 shows the band-selection switching circuits. When S1 is low and S2 is high, the 5.7-GHz band is selected. The LO₁ signals are in the clockwise sequence of I+, Q-, I-, and Q+, and can be represented by $e^{-j\omega_{LO1}t}$. On the other hand, the 2.4-GHz band is chosen when S1 is high and S2 is low. The

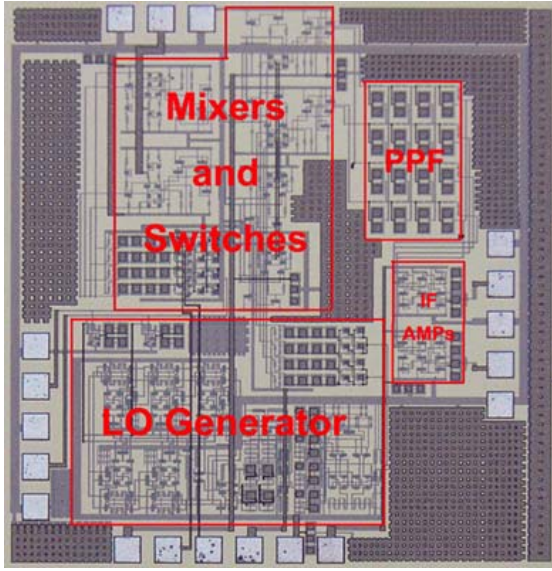


Fig. 9. The die photo of the Weaver-Hartley dual-band low-IF down-converter.

LO₁ signals are in the counterclockwise sequence of I+, Q+, I- and Q- and thus can be represented by $e^{j\omega_{LO1}t}$. The buffer B is a common-drain-configured output buffer to drive the LO port of the mixer.

The die photo of the Weaver-Hartley down-converter is shown in Fig. 9. The supply voltage is 1.8 V and the power consumption of the main mixer cores are only 25 mW while the LO generator consumes 40 mW. However, the common-drain buffers following the active mixers totally consumes 20 mW and this dc power consumption is over designed. The buffer current can be reduced drastically because both IF₁ and IF₂ frequencies are low. Besides, the IF buffer amplifiers, consuming 13 mW for each I/Q output path, are employed mainly to drive the 50-Ω spectrum analyzer for the on-wafer measurement; however, it can be eliminated from our receiver architecture for a practical application due to the high input impedance of the following circuits. Figure 10 shows the conversion gain as a function of LO power for the 2.4-GHz band. The conversion gain reaches 9 dB when the LO power is larger than 5 dBm. On the other hand, the peak conversion is 8 dB when the RF frequency is 5.7-GHz and the LO power is larger than 7 dBm.

When the RF frequency is 5.7-GHz, the demonstrated down-converter has an IP_{1dB} of -13 dBm and IIP_3 of 3 dBm, as shown in Fig 11. When the RF frequency is 2.4-GHz, the IP_{1dB} and IIP_3 are -11 dBm and 2.5 dBm, respectively.

Figure 12 shows the conversion gain as a function of RF frequencies. LO₁ frequency is

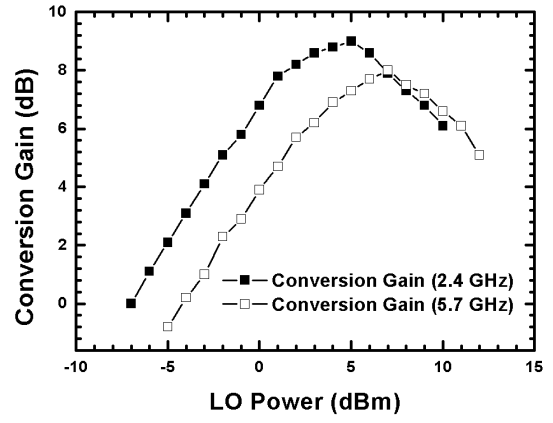


Fig. 10 The conversion gain as a function of LO power of the demonstrated Weaver-Hartley dual-band low-IF down-converter.

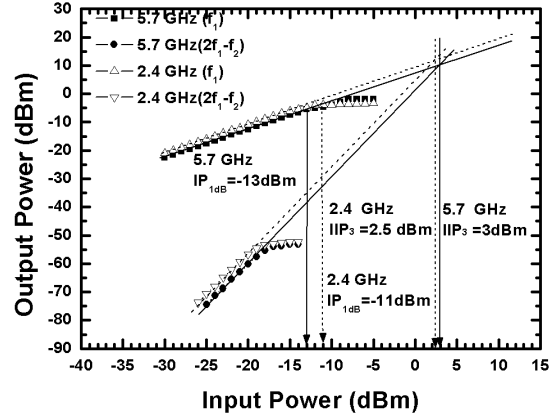


Fig. 11. The power performances of the demonstrated Weaver-Hartley dual-band low-IF down-converter.

fixed at 4.05-GHz and thus LO₂ is equal to 1.62-GHz. The RF frequency is swept to obtain the gain difference between the desired signal and images. The resulting IF frequencies are employed to represent the RF and image frequencies by properly folding the RF and image frequencies into the IF axis.

When the desired RF signal is around 5.7 GHz, the frequencies of the desired IF signal, the IF signals caused by the first image signal, and the second image signal can be calculated by:

$$IF_{desired} = RF - LO_1 - LO_2 \quad (3a)$$

$$IF_{IM1} = LO_1 - IM_1 - LO_2 \quad (3b)$$

$$IF_{IM2} = LO_1 + LO_2 - IM_2 \quad (3c)$$

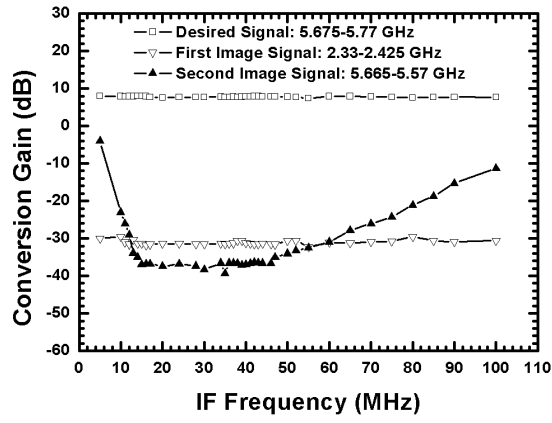
On the other hand, when the desired RF signal is around 2.4 GHz, the frequencies of the desired IF signal, the IF signal caused by the first image signal, and the second image signal can be calculated by:

$$IF_{desired} = LO_1 - RF - LO_2 \quad (4a)$$

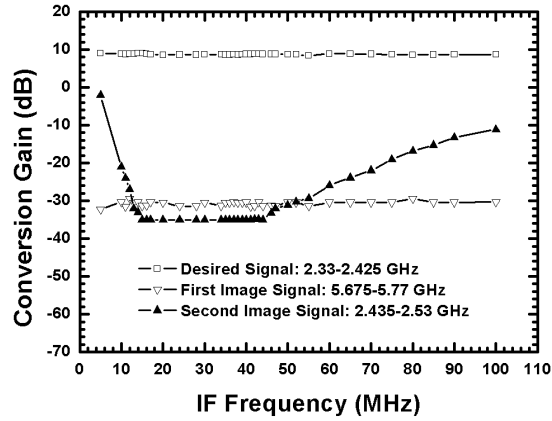
$$IF_{IM1} = IM_1 - LO_1 - LO_2 \quad (4b)$$

$$IF_{IM2} = IM_2 - LO_1 + LO_2 \quad (4c)$$

As shown in Fig. 12(a), the gain of the



(a)



(b)

Fig. 12. The conversion gain as a function of RF and image frequencies of the demonstrated Weaver-Hartley dual-band low-IF down-converter. IF frequencies are employed to represent the RF and image frequencies by properly folding the RF and image frequencies into the IF axis (a) RF near 5.7 GHz. (b) RF near 2.4 GHz.

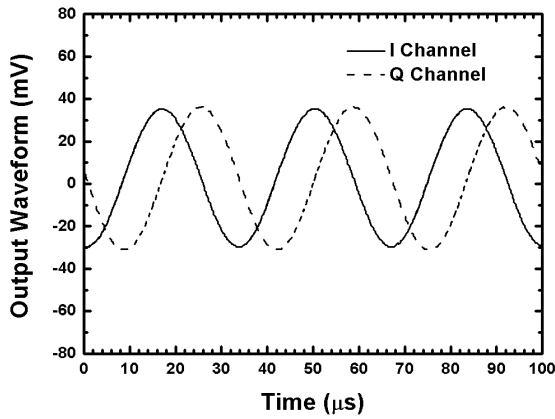


Fig. 13. The measured IQ waveform (with 90.55° phase difference) of the demonstrated Weaver-Hartley dual-band low-IF down-converter.

desired signal is 8 dB when the RF signal frequency is around 5.7 GHz. On the other hand, the first image signal near 2.4-GHz has a conversion loss of 32 dB, and the conversion loss of the secondary image signal (5.64-GHz) is 38 dB. Similarly, the gain is 9 dB when the desired RF signal frequency is around 2.4-GHz, as shown in Fig. 14(b). The first image signal (near 5.7 GHz) and the secondary image signal (near 2.46-GHz) have a conversion loss of 31 and 35 dB,

respectively.

In other words, when the RF frequency is 5.7-GHz (2.4-GHz), the image rejection ratios of the first and secondary image signals are about 40 dB (40 dB) and 46 dB (44 dB), respectively. The frequency response of the secondary image signal is a band-stop shape because of the complex filtering behavior of the polyphase filter. The IF bandwidth of the measured image rejection ratio of the secondary image signal starts from 12-MHz to 48-MHz. On the other hand, the image rejection ratio of the first image signal has a flat frequency response because the image rejection mechanism of the Weaver system comes from dual LO frequency shifting, as discussed in Fig. 2.

Figure 13 shows the waveforms of I/Q output ports. The measured result indicates that the output signals are quite balanced. The in-phase and quadrature-phase waveforms also demonstrate that the circuit is highly balanced, including the signal phase and the signal magnitude.

The measured RF-to-IF, LO-to-RF and LO-to-IF isolations are better than 60 dB, 64 dB and 62 dB, respectively.

(B)

The block diagram of the 2.4/5.7-GHz dual-band dual-conversion downconverter using another LO generator is shown in Fig. 14. The frequency multiplier with the multiple of 2.5 consists of a frequency doubler ($\times 2$) and a frequency divider ($\div 2$). After the mixing operation of a single-sideband (SSB) up-converter, the LO_1 ($2.5 \times LO_2$) signal is thus generated. Both differential-quadrature signals of the LO_1 and LO_2 are generated by a two-section poly-phase filter with the center frequency of 4.05 and 1.62 GHz, respectively. Figure 15(a) shows the block diagram of the static divider consisting of two D-latches realized by emitter-coupled logic. The schematic of the D-latch, consisting of the sample and hold stages, is shown in Fig. 15(b). Instead of using a simple cross-coupled pair, the common-collector configuration is inserted into the positive feedback loop at the hold stage to achieve a wider output swing.

Figure 16(a) shows the function block of the multiplier while the schematic of the multiplier is shown in Fig. 16(b). If $f_1 \neq f_2$, the multiplier is an SSB up-converter while the $(f_1 - f_2)$ frequency component is cancelled if the two input signals have the perfect quadrature

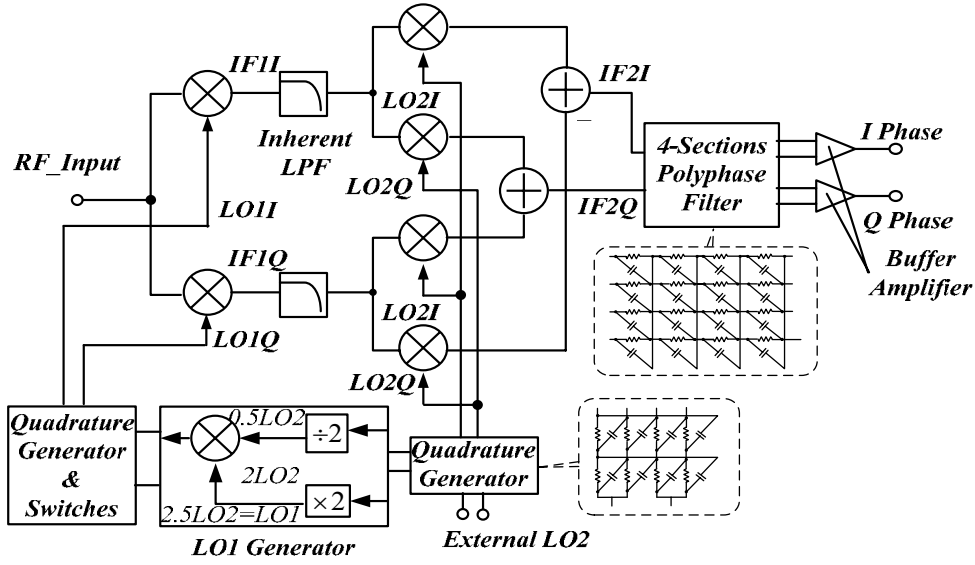


Fig. 14. Block diagram of the 2.4/5.7-GHz dual-band dual-conversion downconverter with correlated LO signal generators.

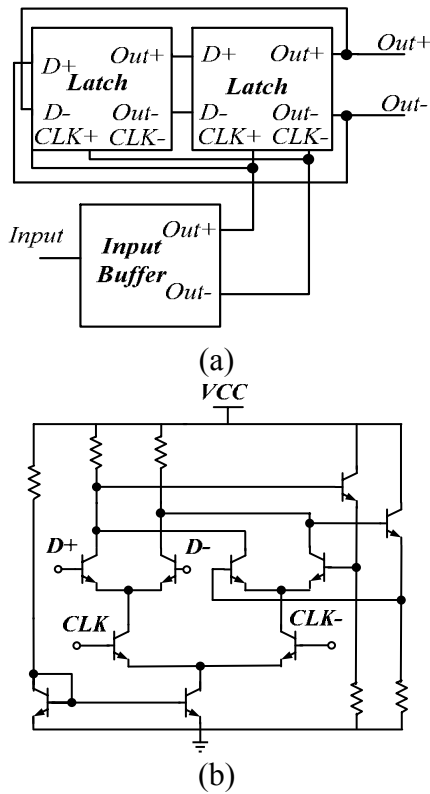


Fig. 15. (a) Block diagram of the static divide-by-two divider (2) the D-latch cell used in the divider.

phase. On the other hand, if $f_1=f_2=f_0$, the circuit becomes a frequency doubler with the output frequency of $2f_0$. In the complex mixer topology of the first stage, the mixing operation leads to the frequency spectrum right-shifting, i.e., $f_{IF1}=f_{RF}+f_{LO1}$, if the differential-quadrature LO_1 has the positive output sequence (0° , 90° , 180° and 270°). On the other hand, if the polarity of the LO_1 is reversed, the output spectrum is left-shifting, i.e., $f_{IF}=f_{RF}-f_{LO1}$. By setting the LO_1 frequency at 4.05 GHz, halfway between 2.4 GHz and 5.7 GHz, the 2.4-GHz receiving mode

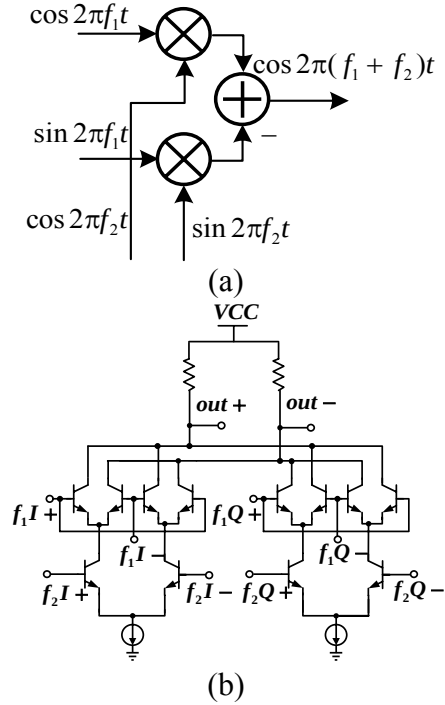


Fig. 16. (a) Function block of the multiplier which can be utilized as a single-sideband upconverter and a frequency doubler (b) schematic of the multiplier.

employs the positive LO_1 sequence (0° , 90° , 180° and 270°) while the 5.7-GHz receiving mode employs the negative LO_1 sequence (0° , 270° , 180° and 90°). As a result, the outputs at the first stage are both down-converted to 1.65 GHz with positive I/Q output sequence when the desired signal is received for both modes. Therefore, the dual-band operation can be achieved.

Figure 17 shows the microphoto of the 2.4/5.7-GHz dual-band dual-conversion downconverter with correlated LO signal generators and the die size is $1.63 \times 1.52 \text{ mm}^2$. On-wafer measurement is employed for the

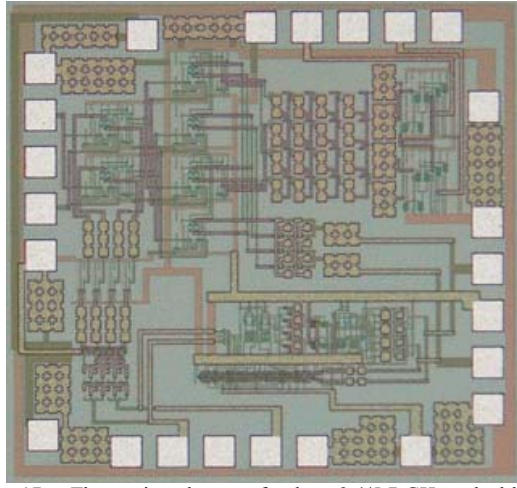


Fig. 17. The microphoto of the 2.4/5.7-GHz dual-band dual-conversion downconverter with correlated LO signal generators.

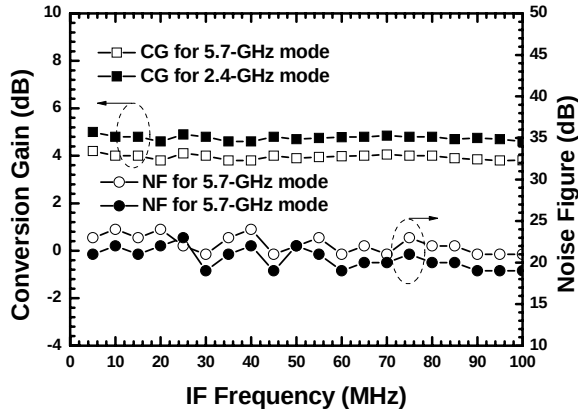


Fig. 18. Conversion gain and noise figure of the 2.4/5.7-GHz dual-band dual-conversion downconverter with correlated LO signal generators.

RF performance. Figure 18 shows the CG and the SSB-NF of 2.4/5.7-GHz bands with a supply voltage of 3 V. The CG is 5/4 dB while the NF is about 20 dB for 2.4/5.7-GHz band while the *IF* frequency is below 100 MHz and the *LO* power is only 2 dBm. Besides, the *IRR*s of both the first and second image signals for 2.4-GHz band are shown in Fig. 19. The *IRR*₁ is above 40 dB and is flat due to the one-way frequency shifting. Compared with the *IRR*₁, the *IRR*₂ is 44-dB within a narrow band from 15 to 45 MHz due to the frequency response of the four-section poly-phase filter following the second-stage mixers. On the other hand, the *IRR*₁ and *IRR*₂ are 39 and 46 dB within the *IF* bands from 15 to 45 MHz for the 5.7-GHz mode as shown in Fig. 20. Figure 21 shows the power performance of both 2.4/5.7-GHz bands. The *IP*_{1dB} is -12/-9 dBm while the *IIP*₃ is 2/6 dBm for 2.4/5.7-GHz band when *IF*=30 MHz. The output waveform of both *I/Q* channels are shown in Fig. 22 and the figure shows the 0.46 dB magnitude mismatch and the 0.62° phase error.

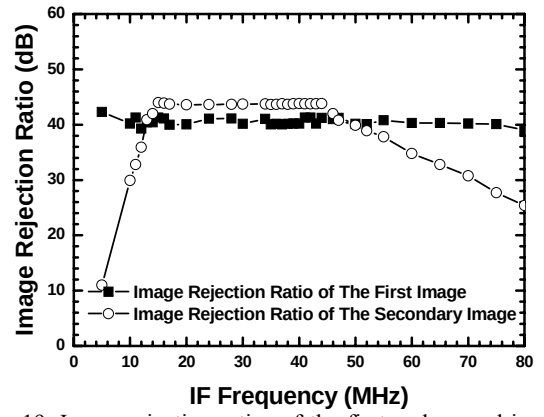


Fig. 19. Image rejection ratios of the first and second image signals at 2.4-GHz mode of the dual-band dual-conversion downconverter.

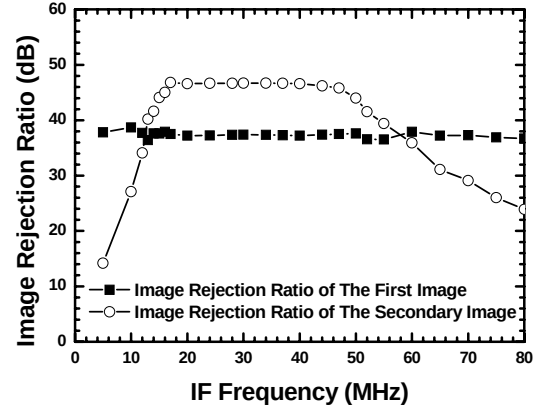


Fig. 20. Image rejection ratios of the first and second image signals at 5.7-GHz mode of the dual-band dual-conversion downconverter .

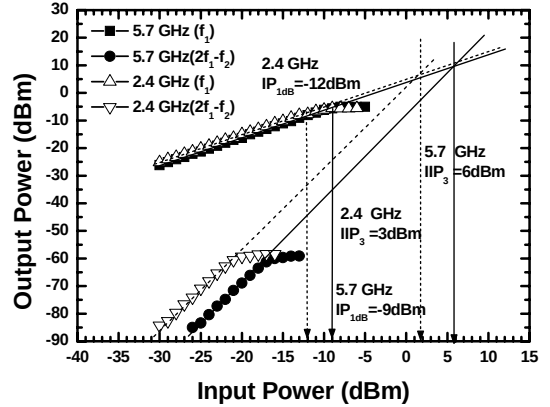


Fig. 21. Power performance of the 2.4/5.7-GHz dual-band dual-conversion downconverter with correlated LO signal generators.

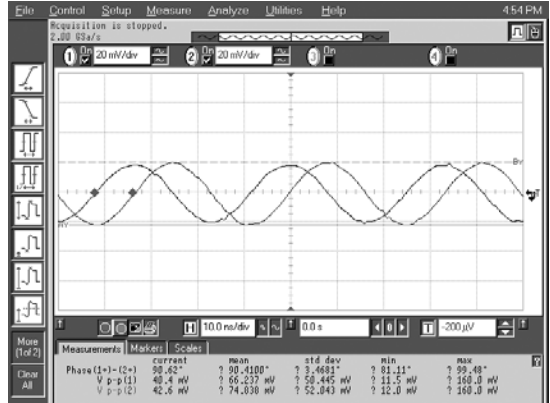


Fig. 22. Output *I/Q* waveforms of the 2.4/5.7-GHz dual-band dual-conversion downconverter with correlated LO signal generators.

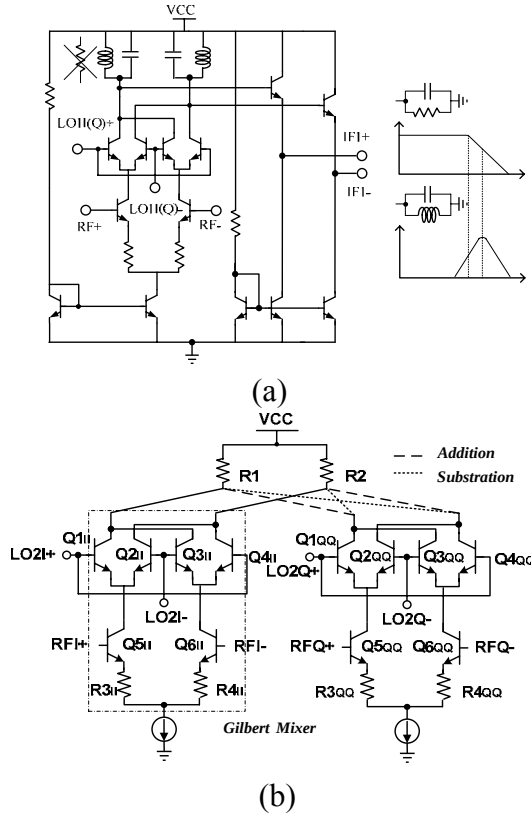


Fig. 23. Schematic of (a) the first-stage Gilbert mixer with a resonant LC load (b) the second-stage mixer with a resistive load.

(C)

Conventionally, the load resistance in parallel with the load capacitance, including parasitic capacitances, determines the IF output low-pass bandwidth. Therefore, the transistor sizes should be optimized and the output load resistance should be sufficiently low to increase the IF bandwidth at the cost of the conversion gain. In this work, a parallel LC tank resonated at ω_{IF1} (around 4 GHz, which is hard to be achieved by a low-pass response of an RC load) is utilized at the first-stage mixer to increase the conversion gain at a high IF frequency, as shown in Fig. 23(a). Figure 23(b) shows the second-stage mixers with output in-phase/anti-phase connections for an addition or subtraction function to realize the complex mixing operation. The IF buffer amplifier is employed facilitating 50- Ω measurements. In addition, both quadrature LO_1 and LO_2 are generated by a two-stage polyphase filter.

A die photo of the 10-GHz Weaver-Hartley low-IF downconverter is shown in Fig. 24 and the die size is $1.9 \times 1.73 \text{ mm}^2$. The dc power consumption is 75 mW at a 3.3-V supply. A conversion gain reaches a flat region when the LO_1 and LO_2 power are larger than 6 and 0 dBm, respectively. Figure 25 shows the conversion gain and single-sideband noise

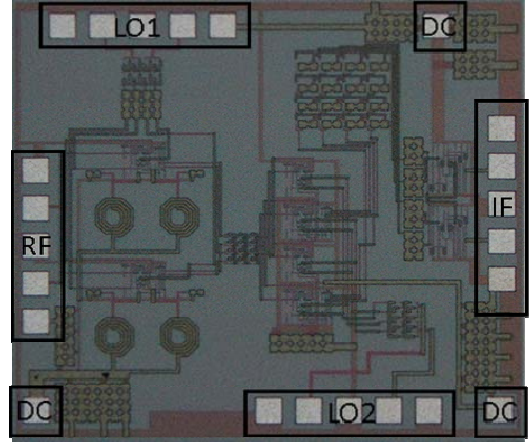


Fig. 24. Die photo of the SiGe HBT Weaver-Hartley low-IF downconverter.

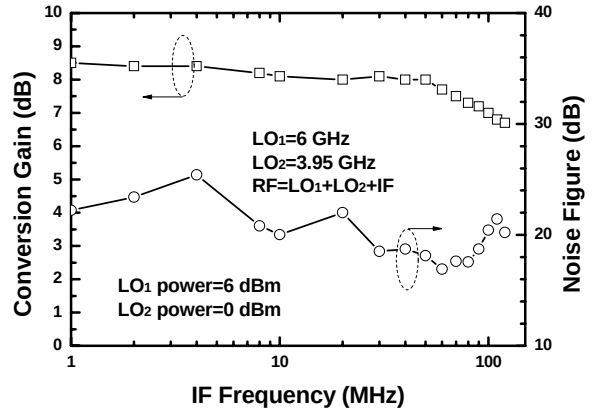


Fig. 25. Conversion gain and single-sideband noise figure with respect to IF frequency of the SiGe HBT Weaver-Hartley low-IF downconverter.

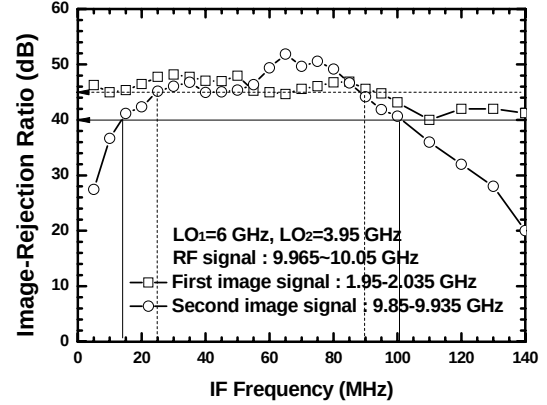


Fig. 26. Image-rejection ratio of the first and second image signals of the SiGe HBT Weaver-Hartley low-IF downconverter.

figure with respect to IF frequency. The conversion gain is 8 dB and noise figure is around 18 dB when IF frequency (ω_{IF2}) ranges from 15 to 100 MHz, which is also the image-rejection band. Figure 26 shows the IRR for the first and second image signals, respectively. The IRR_1 is around 43 dB (maximum: 48 dB) within 100 MHz. The IRR_2 is better than 40 dB (maximum: 52 dB) within 15 to 100 MHz and is better than 45 dB within 25 to 90 MHz. The IP_{1dB} and IIP_3 are -9 and 0 dBm when $RF_1=10 \text{ GHz}$, $RF_2=10.001 \text{ GHz}$, $LO_1=6 \text{ GHz}$, and $LO_2=3.95 \text{ GHz}$,

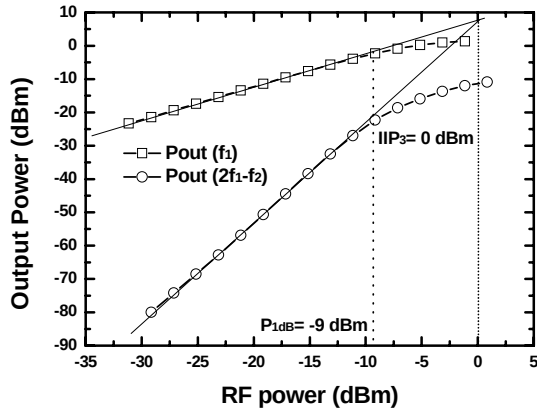


Fig. 27. Power performance, including IP_{1dB} and IIP_3 of the SiGe HBT Weaver-Hartley low-IF downconverter.

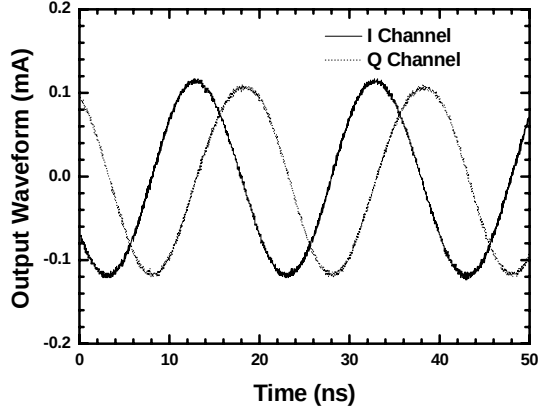


Fig. 28. I/Q output waveforms of the SiGe HBT Weaver-Hartley low-IF downconverter.

as shown in Fig. 7. Figure 8 shows I/Q output waveforms of the downconverter with 0.1° phase error and 0.3 dB amplitude mismatch.

(D)

The passive mixer has a low flicker noise corner since the transistors operate at the off-overlap region, but the severe loss forfeits the capability of suppressing the subsequent noise contributions. Thus, a preceding low-noise amplifier with a much higher gain is necessary. On the other hand, the conversion gain of the active Gilbert mixer performs a good suppression of the noise contributions of the following stage. The dynamic-injection method can reduce the flicker noise contribution to the output through the direct mechanism if the RF frequency is not high. However, as the operating frequency becomes higher, the flicker noise issue becomes much serious since the indirect mechanism, resulting from the parasitic capacitances, dominates. Moreover, the inductor is placed at the source nodes of the Gilbert core to resonate the parasitic capacitances; however, the performance is still limited.

Therefore, the low- IF architecture is

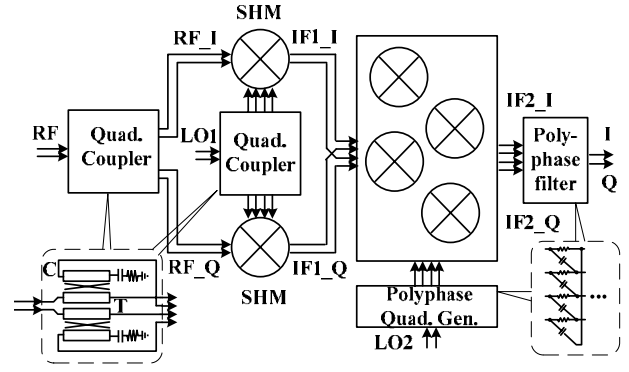


Fig. 29 Block diagram of the 10-GHz Weaver-Hartley dual-conversion downconverter with microwave and analog quadrature generators.

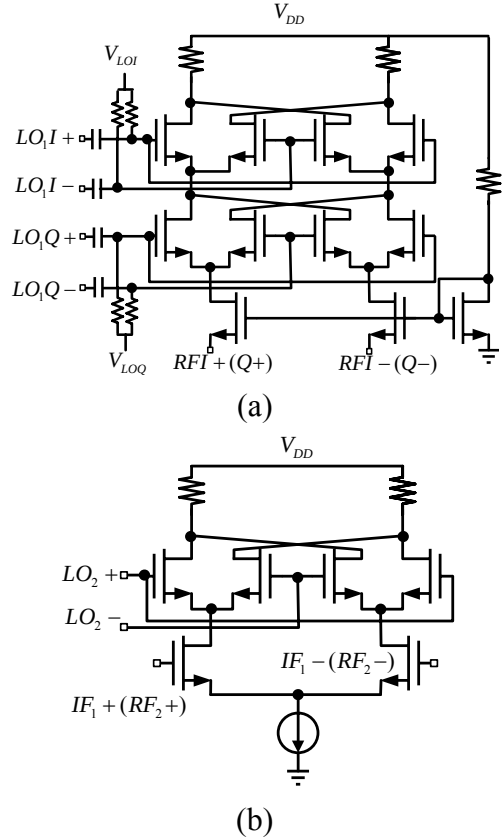


Fig. 30. (a) Schematic of a sub-harmonic mixer with the common-gate input stage for the first down-conversion (b) schematic of a Gilbert mixer utilized for the second down-conversion.

preferred for a 10-GHz receiver because the several tens MHz IF band is far away from the flicker noise corner. Besides, the dc offset problem can be lessened by placing a large dc-blocking capacitor at the output.

On the other hand, the extra dc power consumption is needed to realize a frequency divider or a quadrature voltage-controlled oscillator (QVCO) for the LO quadrature signal generation. A poly-phase filter can generate accurate I/Q signals within a narrow bandwidth; however, the parasitic capacitances and the process variations lead to the shifting of the center frequency and the

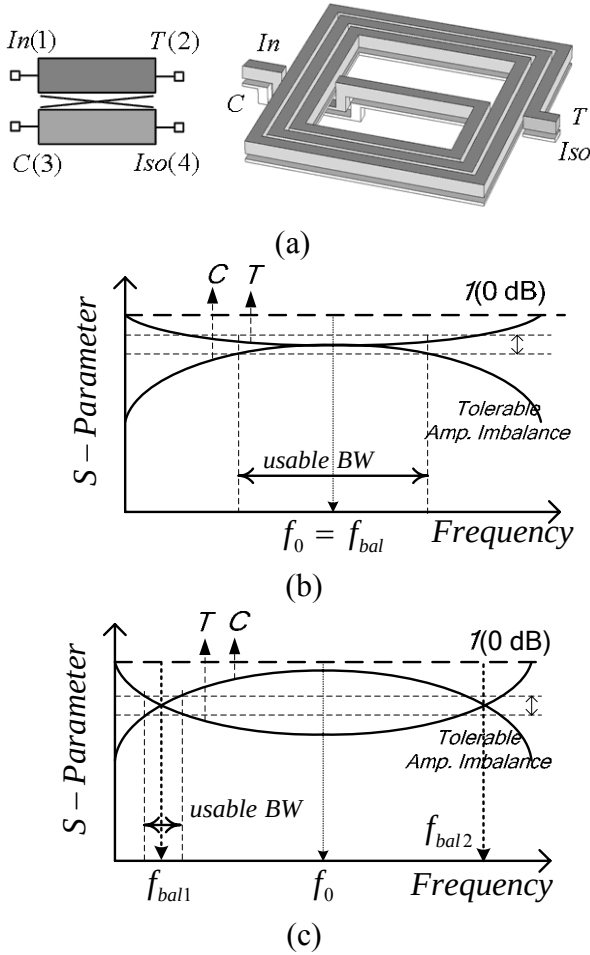


Fig. 31. (a) Symbol and layout of a broadside-coupled quadrature coupler (b) description of the S parameters for an ideal 3-dB coupler (c) description of the S parameters for a high-coupling coupler.

degradation of the phase accuracy, especially at high frequencies. Therefore, the microwave passive quadrature coupler is introduced in this work to generate high-frequency quadrature signals. However, at low frequencies, the poly-phase filter still has accurate quadrature phase difference and balanced amplitudes for the quadrature signal generation.

Figure 29 shows the block diagram of the 10-GHz Weaver-Hartley dual-conversion downconverter. The first down-conversion system consists of two stacked-LO sub-harmonic mixers (SHMs) [Fig. 30(a)] with quadrature RF inputs while four Gilbert mixers [Fig. 30(b)] are employed for the second down-conversion.

Because the SHM is employed, the effective LO_1 frequency is twice of the LO_1 frequency. In this work, $f_{LO1}=4.175$ GHz, $f_{LO2}=1.62$ GHz while f_{IF} ranges from 10 to 60 MHz to avoid the effects of the flicker noise. As a result, f_{RF} , f_{IM1} and f_{IM2} are about 10 GHz, 6.7 GHz and 9.94 GHz, respectively.

The conventional quadrature coupler is implemented by a quarter-wavelength coupled

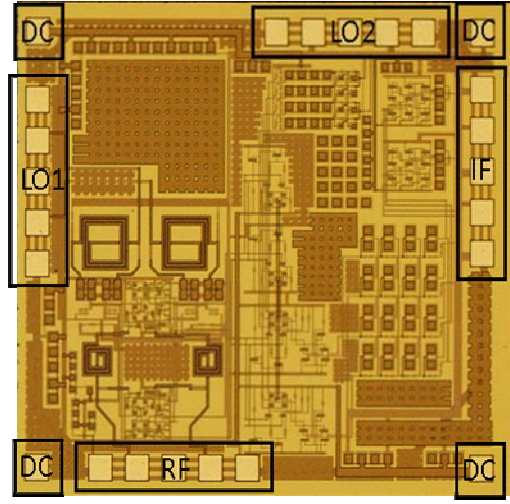


Fig. 32. Die photo of the 10-GHz Weaver-Hartley dual-conversion downconverter with microwave and analog quadrature generators.

line. As a result, the 3-dB coupler is realized with a coupling factor of 0.707 and the description of the S parameters is shown in Fig. 31(b). To widen the bandwidth, the coupling factor can be chosen larger while the amplitude imbalance between the coupled port and the through port is still tolerable. It is noteworthy that the phase difference of the quarter-wavelength coupled-line coupler between these two ports is always 90° as described in (5) without considering the metal loss and the substrate loss of the coupler.

$$\frac{S_{31(C)}}{S_{21(T)}} = j \sin\left(\frac{\pi f}{2 f_0}\right) \frac{k}{\sqrt{1-k^2}} \quad (5)$$

$$f_{bal} = \frac{2}{\pi} \sin^{-1}\left(\frac{\sqrt{1-k^2}}{k}\right) f_0$$

$$\text{or } 2 f_0 - \frac{2}{\pi} \sin^{-1}\left(\frac{\sqrt{1-k^2}}{k}\right) f_0, 0 < f_{bal} < 2 f_0 \quad (6)$$

where k is the coupling factor and f_0 is the center frequency of the quarter-wavelength coupled-line coupler. In this work, the broadside-coupled 10-GHz and 4-GHz quadrature couplers are implemented as shown in Fig 31(a). The high coupling factor (~ 0.8) results in a significant amplitude difference at the center frequency (f_0) as described in Fig 31(c). On the contrary, the perfect amplitude balance occurs at two frequencies within twice the center frequency, one is higher and the other is lower than the center frequency, as described in (6). For our application of the fixed LO frequency without the criterion of a wide operating bandwidth, the high coupling factor makes the usable frequency much lower. That is, for a given operating frequency, the coupler can be very

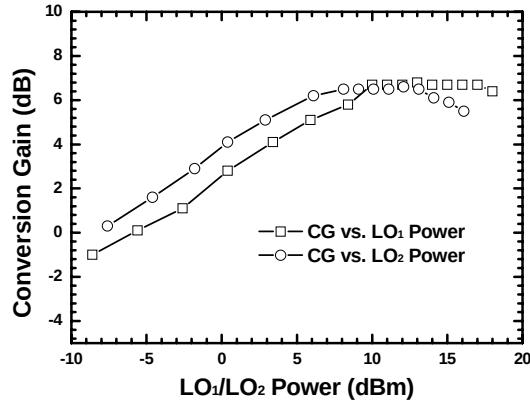


Fig. 33. Conversion gain with respect to the LO_1/LO_2 power of the 10-GHz Weaver-Hartley dual-conversion downconverter with microwave and analog quadrature generators.

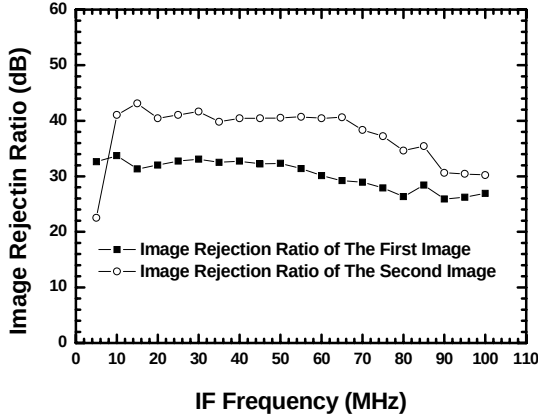


Fig. 34. Image-rejection ratios with respect to IF frequency of the 10-GHz Weaver-Hartley dual-conversion downconverter with microwave and analog quadrature generators.

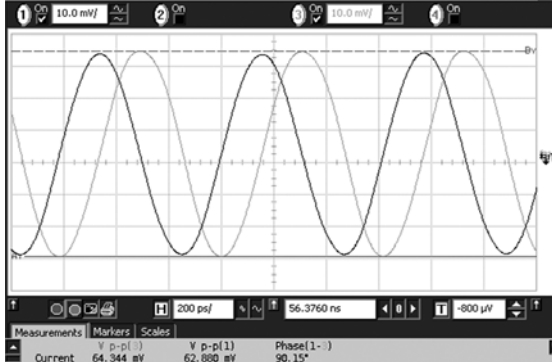


Fig. 35. Output I/Q waveforms of the 10-GHz Weaver-Hartley dual-conversion downconverter with microwave and analog quadrature generators.

compact since the center frequency of the coupler is several times the operating frequency, e.g., $f_0 = 1.85f_{\text{ball}}$ if $k=0.8$; however, the multiple will be even larger when considering the loss effects.

Since the RF coupler needs $50\text{-}\Omega$ loads to maintain the quadrature outputs and thus the common-gate configuration is chosen to achieve a wideband $50\text{-}\Omega$ matching for the 10-GHz operation.

The die photo of the 10-GHz Weaver-Hartley dual-conversion down-converter is shown in Fig. 32 and the die size is $1.92 \times 1.7 \text{ mm}^2$. On-wafer measurement is employed for

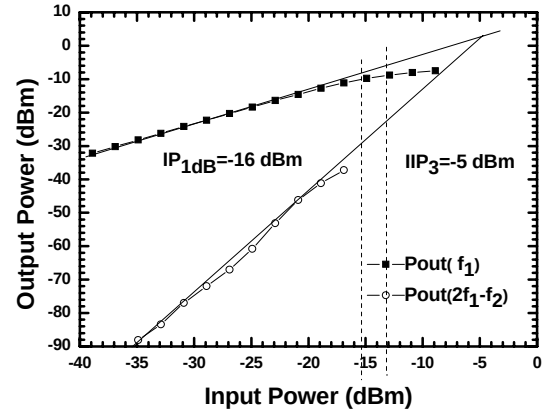


Fig. 36. Power performance, including $IP_{1\text{dB}}$ and IIP_3 , of the 10-GHz Weaver-Hartley dual-conversion downconverter with microwave and analog quadrature generators.

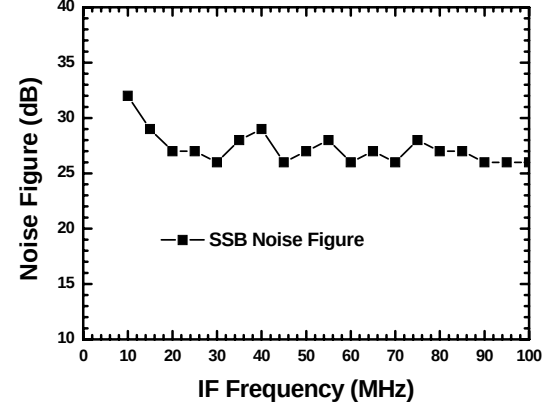


Fig. 37. Single-sideband noise figure of the 10-GHz Weaver-Hartley dual-conversion downconverter with microwave and analog quadrature generators.

the RF performance. The supply voltage is 1.8 V with the power consumption of 80 mW. In Fig. 32, the spiral-typed broadside coupler is realized using metal 6 and metal 5. The line width, line spacing and the outer diameter of the 10-GHz coupler are 10, 3 and $120 \text{ }\mu\text{m}$, respectively. On the other hand, the 4-GHz coupler has $10\text{-}\mu\text{m}$ line width, $3\text{-}\mu\text{m}$ line spacing and $200\text{-}\mu\text{m}$ outer diameter. From simulation, the center frequency of each 10-GHz/4-GHz broadside-coupled coupler is 25 and 13 GHz, respectively.

Figure 33 shows the conversion gain with respect to the LO_1 and LO_2 power; thus, LO_1 power is chosen to be 10 dBm while the LO_2 power is 8 dBm for the fully current commutation of both the first- and second-stage active mixers.

The image-rejection performances are shown in Fig. 34. The image-rejection ratio (IRR) of the first image signal (from 6.67 to 6.72 GHz) is 33 dB covering the IF_2 frequency from 10 to 60 MHz. On the other hand, the IRR of the second image signal (from 9.91 to 9.96 GHz) is over 40 dB. The output I/Q waveforms are shown in Fig. 35 when f_{RF} , f_{LO1} , f_{LO2} , and f_{IF} are 10, 4.175, 1.62 and 0.03 GHz, respectively. Figure 7 shows a 0.2-dB amplitude mismatch

and a 0.15° phase error. IP_{1dB} is -16 dBm and IIP_3 is -5 dBm as shown in Fig. 36. The LO_1 - and $2LO_1$ -to- RF isolations are 40 and 55 dB, respectively. Besides, the SSB-NF is 26 dB as shown in Fig. 37.

Conclusion

In this project, single-band and dual-band low-IF image rejection receivers using Weaver and Hartley architectures are demonstrated. Two dual-band reconfigure receivers using different correlated LO generators are achieved by changing the polarity of the quadrature LO signals. Moreover, microwave quadrature signal generators are applied to achieve highly accurate high-frequency quadrature signals at a high frequency.

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