W-band Mixer With High Image Rejection by Mismatch Compensation Using Buffer Amplifier

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\textbf{ABSTRACT} In this paper, a W-band mixer integrated circuit (IC) with high image rejection ratio (IRR) is presented, compensating for the amplitude and phase mismatches between the I and Q channels in the image rejection mixer (IRM) using RF buffer amplifiers. It is shown by analysis and simulation that the signal coupling between the I and Q channels in the IRM can generate mismatches which can severely degrade the IRR, even though other circuit components are symmetrically designed so as not to induce mismatches. The coupling between two channels can become serious, especially in millimeter-wave IRM ICs where the circuit components are laid out in close proximity to reduce the chip size. It is also shown that poor isolation of millimeter-wave couplers can seriously degrade IRR. In this work, we employ an RF buffer amplifier at the RF port of the resistive mixer to compensate for the amplitude and phase mismatches. The designed W-band IRM IC is fabricated in a 0.1-\textmu m GaAs pHEMT process. Measurements show that the bias tuning of the RF buffer amplifiers can minimize the mismatches and improve the IRR by up to 35 dB at RF from 91 to 95 GHz at an IF of 50 MHz. The IRM exhibits an IRR of 19.2–47.9 dB with conversion loss of 7.9–9.2 dB, which belongs to the highest IRR among the reported IRMs in the W-band.

\textbf{INDEX TERMS} IC, image rejection, millimeter-wave, mixer.

\section{I. INTRODUCTION}
The millimeter-wave band (30–300 GHz) is being actively studied for various applications because it provides a wide bandwidth for high-speed wireless communications. It also allows a short wavelength for high-resolution radar and imaging applications such as security, medical imaging, and virtual reality (VR) applications [1], [2]. High performance millimeter-wave transceivers are required for these applications.

A mixer is an essential circuit component in millimeter-wave transceivers. In the receiver, the mixer down-converts radio frequency (RF) input signal to intermediate frequency (IF) by multiplying it with a local oscillator (LO) signal [3], [4]. The image problem is one of the biggest issues with the practical use of the mixer. The image is located at a frequency that is in either the upper sideband (USB) or lower sideband (LSB) of the LO frequency by IF, and is an unwanted signal in single sideband (SSB) applications. The signals and noise at the image frequency are both downconverted to the same IF as the signal at RF, so that the receiver noise is increased unless it is properly suppressed. An image reject filter can be used in front of the mixer to suppress the image but is not preferred in millimeter-wave frequencies due to its high loss and poor suppression capability [5]. Instead, image rejection mixers (IRM) using a phase cancellation technique are widely used in millimeter-wave frequencies.

Fig. 1 shows a block diagram of the IRM consisting of core mixers, low pass filters (LPFs), local oscillator (LO) signals, and 90° phase shifters at RF and IF. The 90°-phase shifted RF signal is downconverted by the mixer in the Q channel. It is then shifted by 90° in phase at the IF and combined with the other IF signal from the I channel to cancel out the image at the LSB [6]. The image rejection ratio (IRR) is one of the key performance parameters of the IRMs, representing the power ratio of the IF signal produced by the wanted RF to that by an image frequency.
There are several IRM ICs reported in the W-band exhibiting typical IRRs of 16–40 dB [7]–[13]. These values are lower than the results reported for lower frequency IRMs, because of the increased phase and amplitude mismatches between the Q and I channels at high frequencies. Sideband separating mixers have been reported at millimeter-wave and terahertz frequencies for radio astronomy in multi-chip waveguide modules which can entail severe amplitude and phase mismatches between I and Q channels [14], [15]. Several methods to reduce the mismatches and thus improve the IRR have been investigated. In [9], the imperfections were reduced by proposing a symmetrical layout of Lange couplers which are one of the key components in IRMs generating quadrature signals. However, the measured IRR is still lower than the value predicted from the performance of the designed couplers. In [7], an excellent IRR of around 40 dB was obtained by empirically varying RF and LO frequencies across a very wide IF bandwidth (1–12 GHz). However, this experimental method cannot provide the theoretical origin of high IRR or a design guide, and thus cannot be used in practical applications where the IF range is prespecified. In [14], the operating points of the amplifiers in the multi-chip W-band sideband separating mixer were experimentally optimized to reduce the amplitude mismatches and achieve high IRR. In summary, the previous techniques to enhance the IRR are mostly focused on reducing the mismatches of circuit components or the experimental optimization. Therefore, there has been a limit to the IRRs that can be achieved from the millimeter-wave IRMs.

The main cause of IRR degradation is known to be the amplitude and phase mismatches between the I and Q channels in the IRM. Mismatches can be produced by the imperfect performance of RF, LO and IF couplers, variations in the circuit components such as transistors, resistors and capacitors, and asymmetrical layout. In this paper, we introduce another cause of mismatches, the signal coupling between the I and Q channels which can be induced by the substrate coupling between the closely placed circuit components and the poor isolation of the couplers in the millimeter-wave ICs. We will show that this coupling can increase the amplitude and phase mismatches and significantly degrade IRR. Unfortunately, it is somewhat difficult to accurately predict the amount of mismatch due to signal coupling. Therefore, we introduce an RF buffer amplifier in the IRM to minimize the mismatches and restore high IRR via the adjustment of gate bias voltage.

In section II, the effect of amplitude and phase mismatches of IRM on IRR is analyzed. It is shown through simulations that the signal coupling between the I and Q channels in the IRM can generate the mismatches and degrade IRR. In section III, the IRM is designed and the IRR is predicted using full-wave electromagnetic (EM) simulations. A compensation technique for the mismatches induced by the coupling is also presented. The fabrication of the designed IRM using a 0.1-μm GaAs pseudo-morphic high electron mobility transistor (pHEMT) technology and its experimental results are presented in section IV.

II. ANALYSIS OF COUPLING EFFECT ON IMAGE REJECTION

It is known that the IRR is primarily degraded by amplitude and phase mismatches between the I and Q channels of the IRM, as shown in Fig. 1. Let us refer to the amplitude and phase errors in $x_A(t)$ in Fig. 1 as $\varepsilon$ and $\Delta \phi^\circ$, respectively. The IRR can then be expressed as:

$$
IRR = \frac{\gamma^2 + 2 \gamma \cos \Delta \phi + 1}{\gamma^2 - 2 \gamma \cos \Delta \phi + 1},
$$

where $\gamma$ is an amplitude mismatch, or $\gamma = 1 + \varepsilon$ [6]. Fig. 2 shows the calculated IRR using (1) according to amplitude and phase mismatches, demonstrating that the imbalances should be minimized to achieve high IRR.

Mismatch can be produced by the RF, LO, and IF couplers. In millimeter-wave IRM ICs, 90° couplers generating quadrature signals are commonly implemented using the Lange coupler due to its good performance and small size. This coupled-line coupler should be carefully designed to provide good amplitude and phase matches between two output ports. The in-phase power divider at the RF input of Fig. 1 can be designed using a Wilkinson power divider which allows excellent amplitude and phase matches between two output ports thanks to its symmetrical structure. The IF 90° coupler can be implemented off-chip to reduce the IC size. It has been reported in several publications that millimeter-wave IRM ICs exhibit a measured IRR much lower than the predicted value, even though every circuit component in the
IRM is well-designed to have minimal amplitude and phase mismatches [8], [9].

We introduce another cause of IRR degradation in IRMs. There inherently exists a coupling between the circuits in the two channels of the IRM ICs. It can be more severe when they are laid out in proximity in order to reduce the chip area in the millimeter-wave IRM ICs. The signal in one channel can be coupled to the other channel via the substrate with high permittivity. Coupling can also be generated by imperfect isolation of the couplers. It will be shown that the signal coupling can produce amplitude and phase mismatches, even if there are no mismatches caused by the circuit components such as transistors and couplers. Fig. 3(a) shows two network As connected with an ideal Lange coupler which is used at the RF port in the designed IRM as shown in Fig. 1. The network A represents the circuits in the I or Q channels of the IRM. For simplicity of the analysis, we assume that the two-port network A is perfectly matched at each port and has a transmission coefficient of $\alpha \angle 0^\circ$. The coupling coefficients are assumed to be $\beta \angle \theta^\circ$ between the ports on the opposite sides and 0 between the ports on the same sides. The network A is also assumed to be lossless so that $\alpha^2 + \beta^2 = 1$.

Suppose that an RF signal ($V_1^+$) is applied to port 1. Then, the output signals $V_2^-$ and $V_3^-$ are given as follows:

$$V_2^- = \frac{1}{\sqrt{2}} V_1^+ (\alpha e^{-j\frac{\beta^2}{2}} + \beta e^{j\theta}),$$

$$V_3^- = \frac{1}{\sqrt{2}} V_1^+ (\alpha + \beta e^{j(\theta - \frac{\pi}{2})}).$$

In this derivation, the coupler is also assumed to be ideal with perfect isolation. Each output signal is the vector sum of two signals: the transmitted signal from one input port through network A and the coupled signal from the other input port. The amplitude and phase mismatches between $V_2^-$ and $V_3^-$ can be derived from (2) and (3) as follows:

$$\gamma = \sqrt{\frac{1 - 2\alpha \beta \sin \theta}{1 + 2\alpha \beta \sin \theta}},$$

$$\Delta \phi = \tan^{-1} \frac{2\alpha \beta \cos \theta}{\alpha^2 - \beta^2}.$$
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III. DESIGN OF W-BAND IMAGE REJECTION MIXER

Fig. 5 shows a block diagram of the designed W-band IRM consisting of resistive mixers, Wilkinson divider, Lange coupler, RF buffer amplifiers, and IF 90° hybrid. Apart from the IF 90° hybrid, all the circuits are integrated in a 0.1 - μm GaAs pHEMT process which has a typical maximum oscillation frequency (fmax) of 180 GHz. The GaAs substrate has a thickness of 100 μm. The target RF (fRF) and LO (fLO) frequencies are determined to be 91.00-95.00 GHz and 90.95-94.95 GHz with a fixed IF of 50 MHz, respectively, for military radar applications.

A. RESISTIVE MIXER

The resistive mixer in Fig. 6(a) allows high linearity and a wide bandwidth [14]-[18]. The large LO signal is applied to the gate of the transistor, varying the channel conductance of the transistor to generate IF current as a result of mixing with the RF signal applied from the drain. The IF signal is then extracted at the drain. Drain bias voltage is set to 0 V with gate bias near the threshold voltage for high conversion gain. It is difficult to design a compact IF matching circuit, because the IF is as low as 50 MHz. Therefore, the transistor size (0.1 × 100 μm²) is carefully determined to have an output impedance close to 50 Ω at IF, while a DC blocking coupled line is used at the drain to allow a very high impedance at the IF. Fig. 6(b) shows the simulated conversion gain of the designed resistive mixer as a function of RF from 91 to 95 GHz at a fixed IF. The LO and RF powers are 10.0 and -10.0 dBm, respectively. The simulated USB and LSB conversion gains are nearly identical to -9.7 dB.

B. RF BUFFER AMPLIFIER

The RF buffer amplifiers are employed at the RF input of the resistive mixers, and are used to compensate for the amplitude and phase mismatches. The details will be discussed in a later section. They can also improve the conversion gain and the LO-to-RF isolation. They are designed in a single-stage common-source using a transistor of 0.1 × 100 μm² as shown in Fig. 7(a). Fig. 7(b) shows the simulated S-parameters of the RF buffer amplifier at the gate and drain bias voltages of -0.3 and 2.0 V, respectively. The simulated gain of the RF buffer amplifier is greater than 3.5 dB at 91-95 GHz. In the same frequency range, the RF buffer amplifier allows the reverse gain (S12) lower than -14.0 dB, effectively improving LO-to-RF isolation.

C. COUPLERS

The Wilkinson power divider provides LO signals to the two resistive mixers with equal magnitude and phase. The modified Wilkinson power divider is designed in order to reduce the circuit area, where the isolation impedance is moved from the output ports to the middle of the quarter-wave
long line [21]. It exhibits a simulated insertion loss of 0.12 dB and return losses greater than 15 dB with isolation of 18.0 dB at 93 GHz. It does not exhibit any imbalances between the two output ports, since they have a symmetrical structure.

The RF signal is applied to the two mixers through a Lange coupler which generates the quadrature RF signals. The designed Lange coupler has a simulated insertion loss of about 0.55 dB and return losses greater than 15 dB with isolation of 15.6 dB at 93 GHz. The amplitude and phase imbalances between the two output ports at a frequency of 93 GHz are 0.05 dB and 0.26°, respectively.

In order to save the chip area, a commercial IF 90° hybrid (JSPQW-100 by Mini-circuits) is utilized to combine the two IF signals. It will be wire-bonded to the IRM ICs. It exhibits measured amplitude and phase mismatches of 0.16 dB and 0.16° at the target IF, respectively.

D. SIMULATION OF IRM

Fig. 8(a) shows the layout of the IRM IC using the designed resistive mixers, RF buffer amplifiers, LO and RF couplers. The chip size is 1740 μm × 1550 μm. Note that the circuits in the I and Q channels are identical. They are simulated using a full-wave EM simulator (Momentum in Advanced Design System (ADS)) as shown in Fig. 8(b).

To investigate the effect of the coupling on IRR, two EM simulations are performed on the circuits in the I and Q channels. In the half circuit simulation, the circuit in one channel is EM-simulated and used to represent the circuit in the other channel. Therefore, the half circuit simulation does not account for the coupling between the I and Q channels. Conversely, the full circuit simulation simulates the entire circuit of both the I and Q channels at the same time, and thus includes the effect of the coupling between the two channels. The EM simulated data is given in S-parameters as a function of frequency, and is combined with the non-linear transistor models which express the time-domain behavior of the transistor. Finally, the LO, RF, and IF couplers are connected for the harmonic balance simulation in ADS to predict the performance of the IRM.

In order to explain this fact, the IRR is calculated in the half circuit simulation dependent on the isolation of the Lange and Wilkinson coupler. In this simulation, both couplers are assumed to have perfect matches in amplitude, phase, and input impedance. Fig. 11 shows the simulated IRR as a function of the isolation of the couplers. It can be seen that the isolation of the Lange coupler has a more dominant effect on the IRR degradation than the Wilkinson coupler. The IRR is degraded from 61.3 dB to 27.2 dB as the isolation of both couplers decreases from 70 dB to 10 dB. The designed Lange coupler exhibits a 15.6 dB isolation which produces an IRR of 32.4 dB as shown in Fig. 11. By contrast, the simulated IRR is only 19.5 dB when the couplers are assumed to have perfect isolation in the full circuit simulation. Therefore, it can be concluded that the signal coupling between two channels has a more dominant effect on the IRR degradation than the isolation of the couplers.

E. MISMATCH COMPENSATION

The signal coupling is highly dependent on the circuit layout and factors including the distance between the lines, the line
length, and the matching components (open or shorted stubs). It is also affected by the isolation performance of the real Lange couplers. It is therefore somewhat difficult to accurately predict the value of the coupling and compensate for the mismatches in the design stage. We propose a compensation technique that can be applied after the circuit fabrication, namely the gate voltage adjustment of the RF amplifier which is located at the RF port of the resistive mixer, as shown in Fig. 5.

Fig. 12 shows the simulated gain and phase variation of the RF buffer amplifier at a frequency of 93.0 GHz as a function of gate bias voltage ($V_{gg}$) with a fixed drain bias voltage $V_{dd} = 2$ V. As $V_{gg}$ increases from $-0.7$ to $-0.3$ V, the gain...
and phase are adjusted from 1.2 to 3.9 dB and from 10.7° to 21.5°, respectively. Therefore, the mismatches (the amplitude by 2.8 dB and phase by 11.9° in the full circuit simulation above) can be compensated for by applying different gate biases to the RF buffer amplifiers in the I and Q channel. We set $V_{gg1}$ to be $-0.3$ V for the RF buffer amplifier 1 (Q channel). The gate bias voltage ($V_{gg2}$) of the RF buffer amplifier 2 (I channel) is then altered from $-1.0$ to $0.0$ V, and Fig. 13 shows the conversion gain and IRR. It can be seen that the IRR is greatly improved from 17.2 to 31.1 dB at the cost of an increased conversion loss from 7.7 to 9.5 dB, when $V_{gg2}$ is tuned to $-0.7$ V. It is found from the simulation that at this bias condition, the amplitude and phase mismatches of the RF signals are greatly reduced to 0.05 dB and 1.45°, respectively.

**IV. EXPERIMENTAL RESULTS**

The designed IRM was fabricated using a 0.1-µm GaAs pHEMT process. The diced chip and IF coupler were mounted on the printed circuit board (PCB) in a CER-10 substrate and connected by bonding wires. Fig. 14 shows photographs of the fabricated chip and IRM module. The size of the entire module is 6.5 cm × 8.5 cm.

Fig. 15 shows the measurement setup of the IRM module. The W-band RF and LO signals are generated using low-frequency signal generators, frequency multipliers, and power amplifier modules, and applied to the chip via an on-wafer probe. The powers of the applied LO and RF signals are measured using directional couplers and power meters. The IF signals in the I and Q channels are combined by the IF $90^\circ$ hybrid through bond wires, and its output power and frequency are measured by the spectrum analyzer.

Fig. 16 shows the measured conversion gain and IRR of the IRM module. RF and LO frequencies were swept from 91.0 to 95.0 GHz and from 90.95 to 94.95 GHz, respectively, with an IF of 50 MHz. The powers of LO ($P_{LO}$) and RF ($P_{RF}$) signals were fixed at 10.0 and $-10.0$ dBm, respectively.
TABLE 1. Comparison of the reported W-band image rejection mixers.

| Reference | Technology | Topology | RF (GHz) | LO (GHz) | IF (GHz) | $P_{1dB}$ (dBm) | Conversion gain (dB) | IRR (dB) | LO-to-RF isolation (dB) | Size (mm×mm) |
|-----------|------------|----------|----------|----------|----------|----------------|---------------------|----------|------------------------|---------------|
| [7]       | 0.15-μm GaAs pHEMT | IRM | 71–76   | 81–86   | 70–90   | 1–12   | 4               | -9                      | 10–40   | n/a                     | n/a           |
| [8]       | GaAs HEMT  | IRM  | 93.7–93.9 | 94.15 | 0.25–0.45 | 10 | -11 | 12–16 | n/a | 2.1×1.2 |
| [9]       | 0.15-μm GaAs pHEMT | IRM | 90–112 | 85–105 | 2–8 | 9.4–11.3 | >-10 | 10–29.2 | 28.9–31.5 | 2×2 |
| [10]      | 0.1-μm GaAs HEMT | LNA1+IRM | 93–95 | 93–94 | 0.02–0.2 | 9 | 7–9 | 16–26 | n/a | 5×2 |
| [11]      | GaAs mHEMT | IRM | 91.5 | 94 | 2.5 | n/a | -15 | 23.2 | n/a | 1.3×1.5 |
| [12]      | 0.1-μm GaAs pHEMT | LNA+IRM+LO buffer | 82–87 | 81 | 1–6 | 4 | 1–9 | 21–32 | n/a | 2×2.5 |
| [13]      | 0.1-μm GaAs pHEMT | LNA1+IRM+LO AMC2 | 70–98 | 75–92 | 1–8 | 5–7 | 3–6 | 15–28 | n/a | 3×1.8 |
| [14]      | Multi-chip modules3 | SSM4 | 91–99 | 47.5–51.5 | 0.8–4.2 | 12.2–15.3 | 5–11 | 10–34 | n/a | 25×50 |
| This Work | 0.1-μm GaAs pHEMT | IRM with buffer | 91.0–95.0 | 90.95–94.95 | 0.05 | 10.0 | -8.7 | 19.2–47.9 | >25.3 | 1.7×1.6 |

1 LNA: low noise amplifier. 2 AMC: amplifier multiplier chain.
3 Consists of LNA ICs (70-nm GaAs mHEMT), mixer ICs (GaAs Schottky diode), and off-chip RF/LO couplers.
4 SSM: sideband separating mixer.

The measured IRR is higher than 30 dB across the RF bandwidth of 1.2 GHz which is limited by the variation of the signal coupling with RF frequency. Fig. 16 (b) demonstrates that the proposed gate voltage adjustment can improve the IRR across a bandwidth wider than 4 GHz (91–95 GHz) compared with the case without the gate voltage adjustment. The measured input 1-dB compression point was as high as 11.0 dBm. The isolation characteristics of the IRM were also measured to have LO-to-RF and RF-to-LO isolations greater than 25 and 19 dB, respectively.

Fig. 17 shows the measured conversion gain and IRR of the IRM module as a function of IF with an LO frequency of 93 GHz. The conversion loss and IRR were measured to be around 7.7 dB and 5.9–12.3 dB, respectively, at the same gate voltage of −0.3 V for the two RF buffer amplifiers. The IRR was significantly improved by tuning the gate bias voltages of the RF buffer amplifiers. The maximum IRR of 50.6 dB was obtained at an IF of 70 MHz with $V_{gg1}$ and $V_{gg2}$ of −0.3 and −0.7 V, respectively. The measured IRR is greater than 33.6 dB across the very wide IF bandwidth from 20 to 80 MHz. The IF bandwidth of the IRR was limited by the amplitude and phase mismatches of the off-chip IF coupler.

Table 1 shows a comparison of the published W-band IRMs which were fabricated using transistor technology with a high $f_{max}$ such as pHEMT or mHEMT. Most of them utilized passive mixers as a mixer core such as resistive or diode mixers. Some of them have a low noise amplifier (LNA) and LO amplifier to increase conversion gain and reduce LO power level. This table demonstrates that...
employing the RF buffer amplifiers preceding the core mixers. The proposed approach allows an optimum bias tuning to achieve high image rejection ratio after the chip is fabricated. We believe that the designed mixer with high image rejection ratio can be effectively applied to high performance W-band wireless transceivers.

V. CONCLUSION

In this work, we showed that amplitude and phase mismatches can be generated by the signal coupling between the I and Q channels in the image rejection mixer. This coupling can severely degrade the image rejection ratio, especially in millimeter-wave image rejection mixer ICs, even though no particular circuit component introduces the amplitude and phase mismatches. It is also shown that the imperfect isolation of the couplers can degrade the image rejection ratio. To the best of the authors’ knowledge, the effect of the isolation of two channels on the image rejection ratio has not been addressed in previous publications.

In order to achieve a high image rejection ratio, the two channels of the image rejection mixer should be sufficiently separated in the layout and by designing the couplers (especially 90° hybrid) with high isolation. However, we found from the simulation that separating the two channels in the layout has a limitation of the IRR improvement (due to the substrate coupling).

We demonstrated that mismatches due to the coupling as well as circuit components can be compensated for by employing the RF buffer amplifiers preceding the core mixers. The proposed approach allows an optimum bias tuning to achieve high image rejection ratio after the chip is fabricated. We believe that the designed mixer with high image rejection ratio can be effectively applied to high performance W-band wireless transceivers.

REFERENCES

[1] N. Kukutsu and Y. Kado, “Overview of millimeter and terahertz wave application research.” NTT Tech. Rev., vol. 7, no. 3, pp. 1–6, Mar. 2009. [Online]. Available: https://www.ntt-review.jp/archive/ntttechnical.php?contents=ntr200903s1.pdf&mode=show_pdf
[2] Rajiv. (Jul. 23, 2017). Applications of Millimeter Waves and Future. RFpage. [Online]. Available: https://www.rfpage.com/applications-of-millimeter-waves-future/
[3] W. Mohyuddin, I. B. Kim, H. C. Choi, and K. W. Kim, “Design of a compact single-balanced mixer for UWB applications,” J. Electromagn. Eng. Sci., vol. 17, no. 2, pp. 65–70, Apr. 2017.
[4] Y. Jeon and S. Bang, “Front-end module of 18–40 GHz ultra-wideband receiver for electronic warfare system,” J. Electromagn. Eng. Sci., vol. 18, no. 3, pp. 188–198, Jul. 2018.
[5] B. C. Henderson and J. A. Cook, “Image-reject and single-sideband mixers,” WJ Commun. Inc., vol. 12, no. 3, pp. 1–6, Jun. 1985. [Online]. Available: https://www.rfcafe.com/References/articles/wj-tech-notes/ImageRej_n_SSB_mixers.pdf
[6] B. Razavi, “Transceiver architectures,” in RF Microelectronics, 2nd ed. Upper Saddle River, NJ, USA: Prentice-Hall, 2011, pp. 157–250.
[7] M. Gavell, M. Ferndahl, S. E. Gunnarsson, M. Abbasi, and H. Zirath, “An image reject mixer for high-speed E-band (71–76, 81–86 GHz) wireless communication,” in Proc. Annu. IEEE Compound Semiconductor Integ. Circuit Symp., Greensboro, NC, USA, Oct. 2009, pp. 1–4.
[8] T. N. Ton, T. H. Wang, K. W. Chang, G. S. Dow, G. M. Hayashibara, B. Allen, and J. Berenz, “A W-band monolithic InGaAs/GaAs HEMT Schottky diode image reject mixer,” in Proc. 14th IEEE Gallium Arsenide Integ. Circuit (GaAsIC) Symp., Oct. 1992, pp. 63–66.
[9] Y. Wu, S. Lin, C. Chiong, Z. Tsai, and H. Wang, “A W-band image-reject mixer for astronomical observation system,” in IEEE MTT-S Int. Microw. Comp. Dig., Baltimore, MD, USA, Jun. 2011, pp. 1–4.
[10] H. Wang, K. W. Chang, T. N. Ton, M. Biedenbender, S. T. Chen, J. Lee, G. S. Dow, K. L. Tan, and B. R. Allen, “High-yield W-band monolithic HEMT low-noise amplifier and image rejection downconverter chips,” IEEE Microw. Guided Wave Lett., vol. 3, no. 8, pp. 281–283, Aug. 1993.
[11] J. Gong, M. Lei, Y. Wang, and Y. Li, “The design of image rejection mixer in W-band,” in Proc. 16th Int. Conf. Electron. Packag. Technol. (ICEPT), Changsha, China, Aug. 2015, pp. 22–24.
[12] J. Zhang, Y. Ye, R. Tong, and X. Sun, “A highly integrated direct conversion receiver for E-band wireless communication,” in Proc. IEEE Int. Wireless Symp. (IWS), Xi’an, China, Mar. 2014, pp. 1–4.
[13] M. Ferndahl, M. Gavell, M. Abbasi, and H. Zirath, “Highly integrated E-band direct conversion receiver,” in Proc. IEEE Compound Semiconductor Integ. Circuit Symp. (CSICS), La Jolla, CA, USA, Oct. 2012, pp. 1–4.
[14] D. Monasterio, C. Jarufe, D. Gallardo, N. Reyes, F. P. Mena, and L. Bronfman, “A compact sideband separating downconverter with excellent return loss and good conversion gain for the W band,” IEEE Trans. THz Sci. Technol., vol. 1, no. 2, pp. 572–580, Nov. 2019.
[15] P. J. Sobis, A. Enrich, and J. Stake, “A low VSWR 2SB schottky receiver,” IEEE Trans. THz Sci. Technol., vol. 1, no. 2, pp. 403–411, Nov. 2011.
[16] S. Maas, “A GaAs MESFET mixer with very low intermodulation,” IEEE Trans. Microw. Theory Techn., vol. MTT-55, no. 4, pp. 425–429, Apr. 1987.
[17] S. Peng, “A simplified method to predict the conversion loss of FET resistive mixers,” in IEEE MTT-S Int. Microw. Symp. Dig., vol. 2, Denver, CO, USA, Jun. 1997, pp. 857–860.
[18] A. R. Barnes, P. Munday, R. Jennings, and M. T. Moore, “A comparison of W-band monolithic resistive mixer architectures,” in IEEE MTT-S Int. Microw. Symp. Dig., Jun. 2002, pp. 1867–1870.
[19] J.-C. Kao, K.-Y. Lin, C.-C. Chiong, C.-Y. Peng, and H. Wang, “A W-band high LO-to-RF isolation triple cascade mixer with wide IF bandwidth,” IEEE Trans. Microw. Theory Techn., vol. 62, no. 7, pp. 1506–1514, Jul. 2014.
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[20] Z.-M. Tsai, J.-C. Kao, K.-Y. Lin, and H. Wang, “A 24–48 GHz cascode HEMT mixer with DC to 15 GHz IF bandwidth for astronomy radio telescope,” in Proc. Eur. Microw. Integ. Circuits Conf. (EuMIC), Rome, Italy, Sep. 2009, pp. 5–8.

[21] W. Choe and J. Jeong, “Compact modified Wilkinson power divider with physical output port isolation,” IEEE Microw. Wireless Compon. Lett., vol. 24, no. 12, pp. 845–847, Dec. 2014.

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