This letter presents a wideband and high conversion gain mixer based on Gilbert-cell in InP DHBT process. Capacitive degeneration is introduced to increase the bandwidth of the mixer with no voltage drop and few cost of area. Current bleeding technique is employed to improve conversion gain. Utilization of device $f_t$ is achieved to 0.919. The measurement demonstrates a conversion gain of 14.5 dB at 26 GHz with -3 dB bandwidth of 49 GHz. BW/$f_t$ of 0.306 and GBP of 260 GHz are obtained, which are believed to be among the best compared to mixers fabricated with the same size process.

**key words:** InP DHBT, Gilbert-cell, mixer, monolithic microwave integrated circuit (MMIC)

**Classification:** Microwave and millimeter-wave devices, circuits, and modules

### 1. Introduction

A number of active mixers have been reported so far in various topologies and processes for different applications. SiGe-based heterojunction bipolar transistor (HBT) mixers [1, 2], SiGe-based BiCMOS mixers [3, 4, 5, 6, 7], CMOS-based mixers [8, 9, 10, 11] and mixers based on InP HBT technology [12, 13, 14, 15] were presented. Particularly, InP double heterojunction bipolar transistor (DHBT) has the advantages of outstanding high frequency performance, high electron mobility and high breakdown voltage [16, 17], which indicates that InP DHBT is an attractive choice for wideband mixers. Several enhancement techniques have been employed to extend the bandwidth of mixers. The designed Marchand baluns were used as input matching networks to achieve a wider operating bandwidth [18, 19, 20, 21]. But they are not suitable for system level design. Liu et al. proposed a mixer which adopts a common-gate stage with magnetic coupling. The mixer obtained a wide radio frequency (RF) bandwidth without extra input matching components, which is beneficial for being integrated in a receiver [22]. Nevertheless, the common-gate stage introduces voltage drop leading to a higher voltage supply. Zhang et al. demonstrated a mixer using a $\pi$-network technique to increase bandwidth without the need for higher power consumption [23]. However, these approaches all demand a considerable area for implementing inductor. An active mixer that supports wideband conversion gain without the extra cost of area is few reported. For a receiver front-end system, it is also beneficial from an active mixer with high conversion gain due to canceling intermediate frequency (IF) buffer and being compact in size [24]. This letter presents a wideband and high conversion gain mixer based on Gilbert-cell. In this paper, a capacitive degeneration technique which occupies no voltage drop and few additional area is employed to extend bandwidth. The current bleeding technique is used to improve the conversion gain of the mixer. Utilization of device $f_t$ is analyzed for the purpose of making full use of the device performance. The Gilbert-cell mixer obtains a peaking conversion gain of 14.5 dB. The -3 dB bandwidth is from 0.1 GHz to 49 GHz. The bandwidth utilization (BW/$f_t$) [25] and gain-bandwidth product (GBP) [26] achieve 0.306 and 260 GHz respectively, which are excellent performance for wideband mixers.

### 2. Mixer design

The proposed mixer was realized in our in-house 0.8 $\mu$m InP DHBT process with peak $f_t$ of 160 GHz and peak $f_{\text{max}}$ of 350 GHz. The total circuit is powered by -6 V, which consists of RF input buffer, local oscillator (LO) input buffer, bias circuit and mixer core.

#### 2.1 Input buffer and bias circuit of the mixer

The LO input buffer is shown in Fig. 1. A single-ended input signal can be converted to a differential signal by means of this topology. Two-stage differential pairs are used to guarantee high isolation of the mixer by suppressing signal leakage of LO port. RF input buffer and bias circuit are shown in Fig. 2 (a). The emitter followers and diodes are applied to the RF input buffer. The diodes which are formed by transistors with
Fig. 1. LO input buffer of the mixer.

their bases connected to the collectors provide level shift to ensure a appropriate quiescent work point for transistors in next stage. Since the current source in this mixer has multiple outputs, β-helper current mirror is employed to bias circuit [27]. The β-helper transistor \( Q_{20} \) reduces the gain error from the input to each output by a factor of \( (\beta_F + 1) \), where \( \beta_F \) is the forward current gain of transistor \( Q_{20} \).

\[ G_{cnv} = \frac{2}{\pi} g_m R_S \] (1)

where \( g_m \) is the transconductance of the transistors \( Q_1 \) and \( R_S \) is the load resistor.

Considering capacitive degeneration, the expression of the circuit transconductance \( G_m \) can be written by a mathematical derivation on the basis of conversion matrix analysis [31] as

\[ G_m = \frac{g_m}{1 + \frac{C_{be3}}{g_m}} \left( 1 + \frac{R_E C_E s}{2} \right) \] (2)

where \( R_E, C_E \) represent the degenerative resistor and capacitor, \( C_{be1}, C_{be3} \) are the base-emitter capacitance of transistors \( Q_1, Q_3 \), \( g_m, g_m \) are the transconductance of transistors \( Q_1, Q_3 \), respectively. The influence of emitter resistance, base resistance, collector resistance, base-collector capacitance and collector-substrate capacitance have been ignored. As IF output is fixed at a low frequency in measurement of this work, only the poles and zero with regard to the RF frequency are considered. The poles and zero are then found as follows:

\[ \omega_z = -\frac{1}{R_E C_E} \] (3)

\[ \omega_{p1} = -\frac{g_m}{C_{be3}} \] (4)

\[ \omega_{p2} = -\frac{2 + g_m R_E}{2 R_E C_E + R_E C_{be1}} \] (5)

\( G_m \) contains a zero \( \omega_z \) and two poles with dominant pole \( \omega_{p1} \) and subdominant pole \( \omega_{p2} \). Via the appropriate values assignment to \( R_E, C_E \), the dominant pole can be cancelled by the zero when \( 1/(R_E C_E) \leq g_m/C_{be3} \). Then the bandwidth of Gilbert-cell is extended to the subdominant pole.

As the location of the zero \( \omega_z \) is determined by the values of \( R_E, C_E \) and the low frequency gain decreases with the introduction of \( R_E \), there is a trade-off between low frequency
gain and the enhancement factor $\omega_{m1}/\omega_c$ [29]. The values of circuit parameters are finally assigned as $R_s$, $R_E = 550 \Omega$, $R_E = 40 \Omega$, $C_E = 75 \text{ fF}$. As shown in Fig. 3, the simulation results demonstrate that the bandwidth of the mixer with capacitive degeneration can be extended compared to the mixer without degenerative capacitor. The conversion gain is compensated by capacitive degeneration in the high frequency.

![Simulated conversion gain swept over the RF frequency with and without degenerative capacitor.](image)

Fig. 3.

3. Analysis for maximum utilization of device $f_t$

The maximum operating frequency of a circuit is limited by the device cut-off frequency $f_t$, which means the bandwidth of the mixer can not exceed a section of device $f_t$ [25]. It may become a considerable limitation for the bandwidth of the mixer if transistors are biased at a point with much lower $f_t$ than the peak $f_t$ ($f_{t_{\text{max}}}$). The performance of a Gilbert-cell mixer is mainly determined by the mixer core. As a result, a proper bias point for mixer core which fully utilizes the $f_t$ characteristics of the device is essential for the mixer to drive high frequency signal.

Fig. 4 depicts the $f_t$ versus $I_C$ characteristics of the device under different collector-emitter voltage $V_{ce}$ which takes the values of 0.7 V, 1.2 V, 1.7 V, 1.9 V, 2.2 V. The results which were measured by Agilent N5230C VNA show that as $I_C$ increases, the $f_t$ curve rises until the kirk effect is exhibited. In addition, it can be recognized that the growth of collector-emitter voltage $V_{ce}$ enhances the device $f_t$. The peak $f_t$ of 160 GHz is reached when $V_{ce}$ is 1.7 V and $I_C$ is greater than 11 mA. According to the analysis for the $f_t$ characteristics, the bias of the transconductance stage transistors $Q_1$, $Q_2$ can be adjusted to a point which is closer to the peak $f_t$. As the switches are biased with relatively low current in the analysis of current bleeding technique, the $f_t$ utilization of the switching transistors has a certain reduction.

Taking account of the supply voltage of -6 V and the consumption of dc power, $I_C$ and $V_{ce}$ of the transconductance stage transistors are finally set to be 7.2 mA and 1.7 V respectively. As shown in Fig. 5, the device $f_t$ of 147 GHz which is highly close to $f_{t_{\text{max}}}$ (160 GHz) has been achieved. The transconductance stage transistors work at almost optimal state while the utilization of device $f_t$ ($f_t/f_{t_{\text{max}}}$) reaches 0.919.

4. Experiment results

4.1 Measurement setup

The photograph of the fabricated chip is shown in Fig. 6. The DC power consumption of the mixer is 504 mW with the supply voltage of -6 V. The chip size is 0.5 x 0.6 mm$^2$ including all testing pads.

The measurement setup is shown in Fig. 7. The InP Gilbert-cell mixer chip was measured with on-wafer probing for RF, LO and IF ports using ground-signal-ground (GSG) probes. The signal generators which consist of AV1464 and Keysight-N5247A can output LO and RF input signals from 250 kHz to 67 GHz and from 10 MHz to 67 GHz respectively. The LO and RF input signals can be transmitted to the chip via a 50 GHz cable and a 50 GHz GSG probe. The IF output signal was measured by a 43 GHz spectrum analyzer (Agilent-N9030A) through the same cable, GSG probe and DC-blocking capacitors.
4.2 Measurement results
Fig. 8 shows the IF port output spectrum with RF and IF frequency of 20 GHz and 100 MHz, respectively. The IF output power of 3.02 dBm is obtained while RF input power is -11.36 dBm.
As shown in Fig. 9, conversion gain and RF port return loss swept over RF and LO frequency with a fixed IF frequency of 100 MHz. The mixer achieves a peaking conversion gain of 14.5 dB. In order to eliminate the impact of insertion loss on differential cables, the error of signal phase-shifting and facilitate setup of the probe, single-ended input and output rather than double-ended in traditional were applied. The IF output signal was led out from a single emitter follower to the GSG probe which means that the measured conversion gain demonstrates only a half capacity of the proposed mixer. In other words, the measured conversion gain is 3 dB lower than measurement with differential output.
Fig. 10 presents the measured IF output power which swept over RF input power with a fixed RF frequency of 49 GHz.

The input 1-dB compression point of the mixer is -9.17 dBm. All of measurements were performed under the condition that the loss of cable and probe have been calibrated in available frequency range.
As shown in Table I and Fig. 11, the performance and comparison are summarized with some advanced results of wide-band mixers based on InP HBT and SiGe BiCMOS process. By comparison with other mixers, the proposed Gilbert-cell mixer demonstrates higher gain-bandwidth product (GBP)
and higher bandwidth utilization (BW/f₀) which indicate that the capacitive degeneration and current bleeding technique are effective to the enhancement of bandwidth and conversion gain, respectively.

5. Conclusion

In this letter, a wideband and high conversion gain mixer based on Gilbert-cell is implemented in 0.8 µm InP DHBT process. The bandwidth of the mixer is extended by means of capacitive degeneration with no voltage drop and few cost of area. The current bleeding technique is employed to improve conversion gain. Utilization of device f₁ is analyzed to fully release the potential ability of the device. The maximum utilization of device f₁ reaches 0.919. The measurement results show that -3 dB bandwidth from 0.1 GHz to 49 GHz is obtained with a peaking conversion gain of 14.5 dB. With peak f₁ of 160 GHz, BW/f₀ of 0.306 and GBP of 260 GHz are obtained, which are believed to be among the best compared to mixers fabricated with the same size process.

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Table I. Comparison with some advanced mixers

| Reference | Process | Conversion Gain (dB) | Power DC (mW) | Chip size (mm²) | Chip size (mm²) | BW/f₀ (GHz) | GBP (GHz) |
|-----------|---------|---------------------|--------------|----------------|----------------|------------|-----------|
| [5]       | SiGe    | 12                  | 240          | 1.1            | 0.18           | 0.221      | 105.5     |
| [6]       | InP     | 12                  | 140          | 1.6            | 0.18           | 0.093      | 78.7      |
| [7]       | SiGe    | 12                  | 1860         | 0.65           | 0.18           | 0.15       | 66.0      |
| [14]      | InP     | 15                  | 860          | 0.055          | 0.18           | 0.171      | 17.0      |
| This work | InP     | 15                  | 860          | 0.055          | 0.18           | 0.171      | 17.0      |

Fig. 11. Comparison of BW/f₀ and GBP.
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