

Design of a 2.4 GHz High-Performance Mixer with Current Bleeding Topology

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Abstract—The important characteristics of a mixer including high conversion gain, low noise, high dynamic range, low third-order intermodulation distortion and low unwanted leakage signals between ports have reconsidered. A double balanced Gilbert-cell with a class-A amplifier bleeding mixer (DBGC CAAB mixer) is designed and implemented. The voltage gain of 14 dB at class-A amplifier is used to magnifier the local oscillator (LO) signal. Current of class-A amplifier is injected into the Gibert mixer as a bleeding current to improve the performance of the trans-conductor stage. The DBGC CAAB mixer achieves a conversion gain of 17.5 dB at -14 dBm LO power, and the noise figure is suppressed from 45 dB to 10.7 dB. It is important to stress that the new configuration has a third-order intermodulation (IIP₃) at -9 dBm which is considered to be insignificant. The DBGC CAAB mixer is implemented by using 0.18- μ m RFCMOS technology and operates at the 2.4 GHz ISM application with 10 MHz intermediate frequency. The power consumption is 12 mW at 1.5 V supply voltage.

Keywords— Double balance Gilbert-cell, Current Bleeding CMOS Mixer.

I. INTRODUCTION

The key characteristics of mixers are the conversion gain, noise figure and linearity [1]. Many research scholars made lots of contributions to enhance the conversion gain and reduced the noise performance [2-3]. Among all new designs, Gilbert-cell [4] is one of the most successful topology with high gain, low noise figure, good linearity and good isolation between IF output and the inputs from local oscillator and RF signal.

In general, improvements of certain aspects of performance lead to some degradation in others. From Figure 3-7 (a) the conversion gain (CG) and third-order intermodulation distortion (IM₃) are proportional to the trans-conductance g_m of the transistor M1 [5-6].

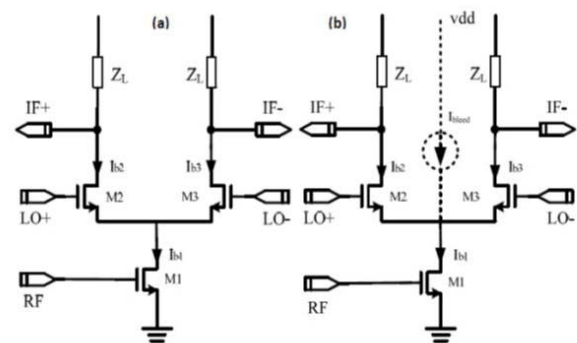


Figure 1 (a) conventional single-balanced mixer; (b) Single-balanced mixer with current bleeding.

A simple way to enhance it is to increase the currents I_{b2} and I_{b3} by reducing the load R_L . As M2 and M3 are working as an on-off switches, reduce R_L can minimise the voltage drop at load and drive the switches at saturation mode. In return, the noise contributions from M2-M3 are increased [7]. Moreover, the reduction in loading resistance R_L causes conversion gain to reduce too. Thus, bias adjustments at MOSFET alone will not be used to improve the overall performance of mixer.

Current bleeding topology is a new design to retain the benefits of increasing the bias current (I_{b1}) without degrading other performances. From Figure 1 (b), a current source is used to inject a current I_{bleed} into M1 [8]. It is of interest to observe that current bleeding is the brilliant method that preserves both the higher conversion gain and the high load R_L because only part of the current I_{b1} comes from Transistor M1 and M2. Thus, the compact size of transistors M2-M3 can be reduced and a small driving voltage V_{Lo} can turn M2-M3 on and/or off. This is a very important feature because additional noise from switch pair will be added to the mixer output if the switching transistors M2-M3 are not turn on or off instantaneous.

In this design, a double balanced Gilbert-cell class-A amplifier bleeding mixer (DBGC CAAB mixer) was proposed

by inserting a class-A amplifier in parallel with M2-M3. Thus, a large local oscillator signal was expected to switch M2 fully on and M3 fully off instantaneously or vice versa. The bias current of the class-A amplifier would be treated as the injected current source of the mixer, thus improving the overall performance.

II. GILBERT CELL MIXER

The double-balanced mixer provides a good isolation of RF signal, local oscillator signal LO to the IF output. The circuit diagram of double-balanced Gilbert-quad mixer, is as shown in Figure 3-5.

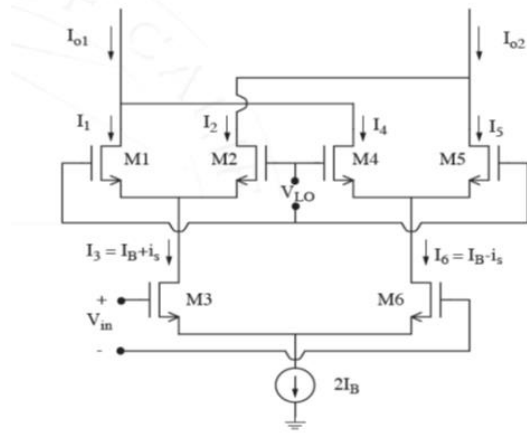


Figure -2 Gilbert Double balanced mixer.

It is assumed that M₁-M₂, M₄-M₅, M₃-M₆ are worked in pairs. In Figure 2, the RF signal is input to a differential amplifier M₃-M₆. The local oscillator signal is applied to the cross connected set of four MOSFET M₁-M₂, M₄-M₅. Assume that v_{LO} is large enough to switch M₂ and M₄ to be fully turn on and M₁ and M₅ to be fully turn off or M₂ and M₄ fully turn off and M₁ and M₅ fully turn on. The current path will be steered from one side to the other side periodically. In other words, the output is equal to the v_{in} multiple with v_{LO} . In order to prevent saturation, the input RF signal is adjusted to be less than the 1 dB compression point.

The RF input voltage V_{RF} is given as

$$v_{in} = v_{RF} = V_{RF} \cos(\omega_{RF} t) \quad (1)$$

$$v_{IF} = (v_{RF} \times v_{LO}) g_m R_L \\ = v_{RF} \times \frac{4}{\pi} (\cos \omega_{LO} t - \frac{1}{3} \cos 3 \omega_{LO} t + \frac{1}{5} \cos 5 \omega_{LO} t \dots) g_m R_L \quad (2)$$

The high frequency term are filtered by the IF filter. Thus

$$v_{IF} = V_{RF} \frac{2}{\pi} [(\cos \omega_{LO} t + \cos \omega_{RF} t) + (\cos \omega_{LO} t - \cos \omega_{RF} t)] g_m R_L \quad (3)$$

It can be found that the signal of local oscillator (LO) is rejected completely at matching constraints and the LO-IF

leakage is eliminated. The linearity is improved as half of signal is processed by each side. Figure 3 shows the output frequency spectrum.

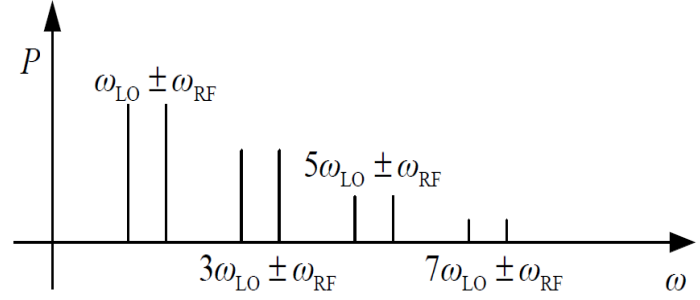


Figure-3 Output frequency spectrum of Gilbert double-balanced mixer.

III MIXER NOISE ANALYSIS

In practice, all RF double sideband signals are down converted into IF frequency for amplification. It means that both noises are combined at IF band. The total noise at IF consists of the noise at desired RF band, the noise at image RF band and the noise that coming from the mixer circuit itself. Assume that the noise at upper and lower bands are identical. The double-side band signal and noise at output can be written as:

$$S_{out} = S_d G_{Mixer_d} + S_{Image} G_{Mixer_im} \quad (4)$$

$$N_{out} = N_d G_{Mixer_d} + N_{im} G_{Mixer_im} + N_{mix_d} G_{Mixer_d} + N_{mix_im} G_{Mixer_im} \quad (5)$$

Where S_d is the desired signal, S_{Image} is the image signal, N_d is the noise in the desired band, N_{im} is the noise in the image band, G_{Mixer_d} and G_{Mixer_im} are the mixer gain at both desired and image frequencies, respectively. N_{mix_d} and N_{mix_im} are the desired and image band noise due to the mixer circuit itself referred to its input.

To simplify the analysis,

$$S_d = S_m, N_d = N_m, G_{Mixer_d} = G_{Mixer_im}, N_{mix_d} = N_{mix_im}$$

$$N_{out} = 2N_d G_{Mixer_d} + 2N_{mix_d} G_{Mixer_d}$$

$$\frac{S_{out}}{N_{out}} = \frac{2S_d}{2N_d + 2N_{mix_d}} = \frac{S_d}{N_d} \left[\frac{1}{1 + \frac{N_{mix_d}}{N_d}} \right] \quad (6)$$

$$\text{The noise factor } F_{DSB} = 1 + \frac{N_{mix_d}}{N_d} \quad (7)$$

It is found that the noise factor of double sideband mixer depends on the mixer noise (N_{mix_d}) itself. From Figure 1 (b), when M₂ is on and M₃ is off, M₂ acts as a cascode amplifier and does not contribute any additional noise. Thus when M₃ is off, zero noise is achieved. The same case applies when M₃ is on and M₂ is off.

Terrovitis [9] and Darabi [2] model the noise contributed when both switching transistors are at ON state at interval Δ . Moreover, any noise in the local oscillator is also magnified during this active interval.

The mean noise voltage due to M_2 and M_3 can be expressed as:

$$\overline{v_{n-m2,3}} = \sqrt{2\gamma \cdot \frac{4kT}{g_{m2,3}}} \quad (8)$$

Where k — Boltzmann's constant,

T — Absolute temperature

γ — noise coefficient ($= 2/3$ for long channel device)

$g_{m2} \cdot g_{m3}$ —Trans-conductance of M_2 - M_3

The total noise contribution due to switches M_2 and M_3 are found as [37].

$$\overline{i_{noise-M2,3}^2}(\omega_{IF}) = \left(\frac{1}{\left(\frac{T_{LO}}{2} \right)} \right) \cdot G_{m0}^2 \cdot \Delta T \cdot \overline{v_{n-m2,3}^2} \quad (9)$$

Assume the trans-conductance of M_2 and M_3 are equal and $V_{LO}(t)$ is denoted as the local oscillator voltage.

$$G_{m0} = g_{m2} = g_{m3} = g_{m2,3} = \frac{2I_{DC,mix}}{\Delta V} \quad (10)$$

$$V_{LO}(t) = A_{LO} \cos(\omega_{LO}t) \quad (11)$$

The slope of change in $V_{LO}(t)$ can be expressed as:

$$\text{Slope} \Big|_{\omega_{LO}t=90} = \left[\frac{dV_{LO}(t)}{dt} \right]_{\omega_{LO}t=90} = A_{LO}\omega_{LO} \quad (12)$$

$$\begin{aligned} \overline{i_{noise-M2,3}^2}(\omega_{IF}) &= \frac{1}{T_{LO}/2} \cdot G_{m0}^2 \cdot \Delta T \cdot \overline{v_{n-m2,3}^2} = \frac{1}{T_{LO}/2} \cdot G_{m0}^2 \cdot \Delta T \cdot \left(2\gamma \cdot \frac{4kT}{g_{m2,3}} \right) \\ &= \frac{1}{T_{LO}/2} \cdot G_{m0} \cdot \Delta T \cdot (2\gamma \cdot 4kT) = \frac{1}{T_{LO}/2} \cdot \frac{2I_{DC,mix}}{\Delta V} \cdot \Delta T \cdot (2\gamma \cdot 4kT) \\ &= \frac{2I_{DC,mix}}{T_{LO}/2} \cdot (2\gamma \cdot 4kT) \cdot \frac{\Delta T}{\Delta V} = \frac{2I_{DC,mix}}{T_{LO}/2} \cdot (2\gamma \cdot 4kT) \cdot \frac{1}{A_{LO}\omega_{LO}} \\ &= 4\gamma 4kT \left(\frac{I_{DC,mix}}{\pi A_{LO}} \right) \end{aligned}$$

$$= 16kT\gamma \frac{I_{DC,mix}}{\pi A_{LO}} \quad (13)$$

The term $I_{DC,mix}$ is the bias current of M_2 and M_3 . It is seen from (13) that minimising the simultaneous conduction interval reduces this degradation, so a large LO drive is proposed to make the switching function approximate a square wave as shown in Figure 4. Furthermore, a fast switching transistor with low bias current ($I_{DC,mix}$) is definitely an advantage in this design

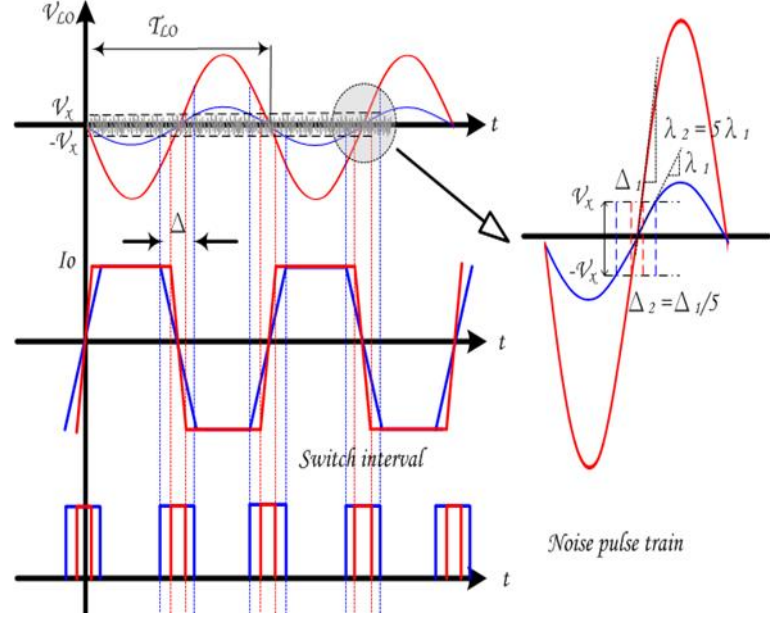


Figure -4 $V_{LO}(t)$ amplitude that affect the noise performance during switching.

The conversion gain (CG) can be analysed [10].

$$CG = \frac{2}{\pi} \left(\frac{\sin\left(\frac{\pi\Delta}{T_{LO}}\right)}{\frac{\pi\Delta}{T_{LO}}} \right) g_m R_L \quad (14)$$

For a very small Δ . The conversion gain will approach its maximum value.

$$CG_{max} = \frac{2}{\pi} g_m R_L \quad (15)$$

where a high trans-conductance transistor in M_1 is definitely an advantage to enhance the conversion gain.

IV PROPOSED NEW TOPOLOGY

A Gilbert-cell double-balanced with class-A bleeding mixer (DBGC CAAB) was proposed as shown in Figure 5. The mixer was in a symmetry manner such that the right-hand side and the left-hand side were perfectly identical.

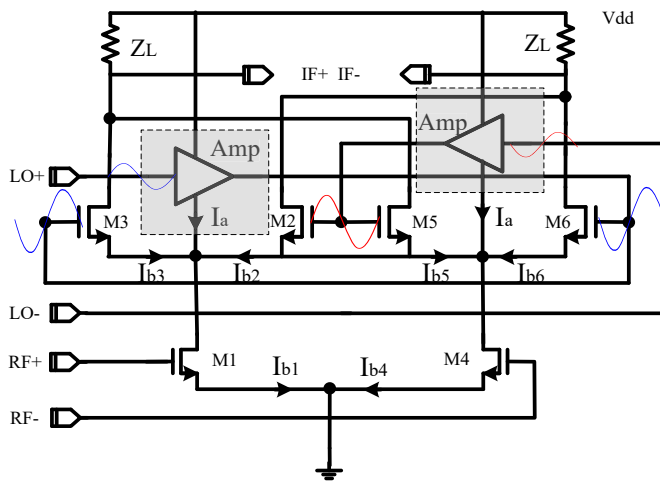


Figure-5 Schematic of DBGC CAAB mixer.

As discussed in section I, current bleeding could improve the overall performance of mixer. In Figure 5, a class A amplifier was used to magnify the local oscillator signal and hence suppressed the noise in IF stage. The class-A amplifier supply current I_a could also be used as an addition bleeding current source at transistor M1 and M4. (i.e. $I_{b1}=I_{b2}+I_{b3}+I_a$, or $I_{b4}=I_{b5}+I_{b6}+I_a$). Similar arrangement was applied to the image part transistors M5 and M6. Thus, a low noise (due to large A_{LO}) and high conversion gain (due to large $g_{m1,4}$ and small Δ) mixer was expected. The DC bias current I_a was set to 4.5 mA. The gain of the class A amplifier was 14 dB, five time in magnification.

The DBGC CAAB mixer was realised by a 0.18 μm 1-poly 6-metal RFCMOS technology. A microphotograph of the device is shown in Figure 6.

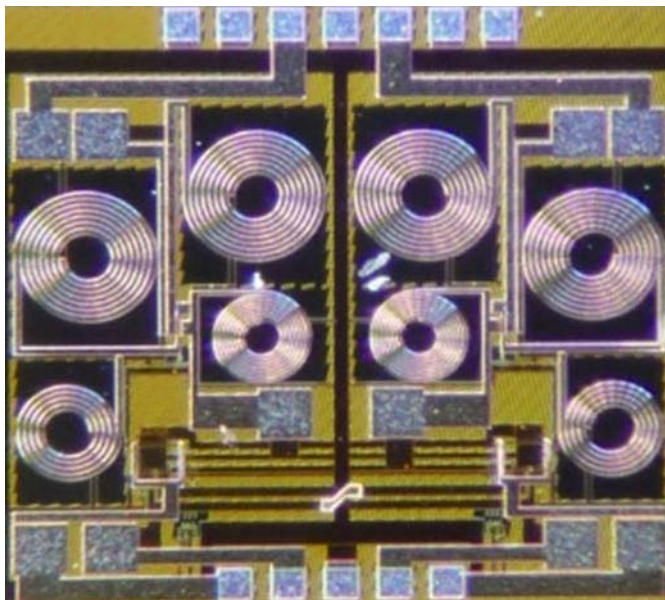


Figure-6 Micro-photograph of implemented DBGC CAAB mixer

The chip area was 1.2x1.3 mm². The mixer supply current was 12 mA from the 1.5 V supply. The operating frequency was 2.4 GHz with 10 MHz (IF) as the output frequency. As the amplitude of Lo signal (A_{LO}) is magnified 5 times, it could be concluded that the noise was reduced by a factor of “Five” from (13).

Conversion gain in the DBGC CAAB mixer was directly related to the interval Δ . From (15), when Δ was very small due to large LO signal (A_{LO}), conversion gain would approach its maximum limit where $CG = \left(\frac{2}{\pi}\right)g_m R_L$. The gain reduced when $\left(\frac{\Delta}{T_{LO}}\right)$ increased. As the $\left[\left(\frac{\Delta}{T_{LO}}\right)=0.5\right]$, the gain would be 64% with respect to the maximum value. The proposed LO amplifier with a gain of 5 in class-A amplifier would certainly help to relieve the inflect of Δ function on conversion gain. Table 1 showed the reduction of CG at five different Δ/T_{LO} cases with or without class-A amplifier. It was of interest to observe that the switching loss is reduced significantly.

Table -1 The quantized switching loss reduction due to the LO amplification in linear and logarithm scale.

	$\left(\frac{\sin\left(\frac{\pi\Delta}{T_{LO}}\right)}{\frac{\pi\Delta}{T_{LO}}}\right)$		
Δ/T_{LO}	Without Amp	With Amp	Loss Improvement (dB)
0.1	0.9836	0.9993	0.04
0.2	0.9354	0.9973	0.27
0.3	0.8584	0.9940	0.636
0.4	0.7568	0.9895	1.164
0.5	0.6366	0.9836	1.88

The original DC bias current of M1 and M2 (Figure 5) was at 1.5 mA. The Class A amplifier fed an additional current of 4.5 mA to M1 and M2. Thus, g_{m1} was improved by a factor of 4, (12 dB) by the current bleeding circuit. In conclusion, our new design would boost g_{m1} , and CG. Assumed that the LO port was fed by a signal of -14 dBm, the maximum gain achieved was 17.5 dB. Figure 7 showed the gain performance of the DBGC CAAB mixer with respect to the LO input power. Figure 8 showed the noise performance at various LO power. The total noise figure of the mixer is measured to be between 8.7 and 12.5 dB at a LO power range of -17 to 10 dBm, respectively. The performance is competitive compared to other related work as shown in Table 2.

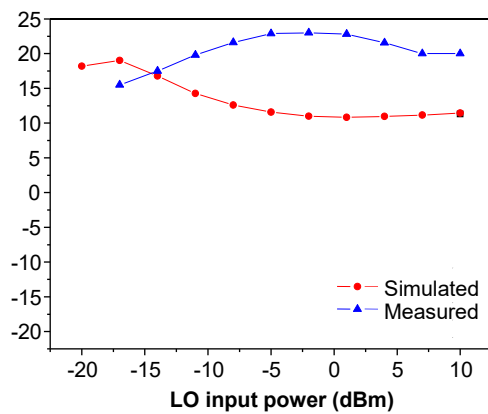


Figure-7 conversion gain comparison between DBGC CAAB mixer and other researchers' works.

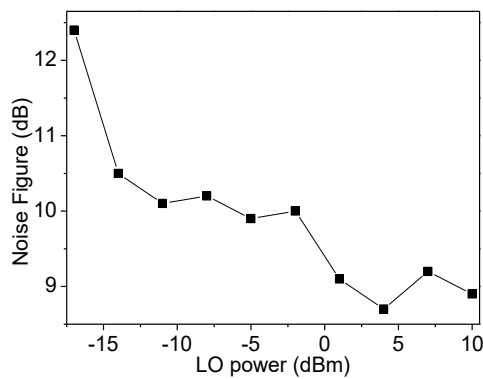


Figure -8. Measured noise figures at different LO power levels at 10 MHz intermediate frequency.

Figure of merit (FOM) is one of a benchmarking process to determine the performances of a mixer. The FOM incorporates important parameters such as Gain, DC power, NF, IIP3, LO power and operating frequency. The FOM is defined as:

$$FOM = 10 \cdot \log \left\{ \frac{10^{(Gain - 2NF + IIP3 - 10 - P_{LO})/20} \cdot f_0 / 1KHz}{P_{DC} / 1mW} \right\} \quad (16)$$

The overall performance of this design and other researchers' results are listed for comparison.

Table 2, showed that the DBGC CAAB mixer had the lowest LO power requirement (-17 dBm) and the highest gain (15.7 dB), while maintaining a relatively low noise figure of 9.7 dB.

Table -2 The performance of the DBGC CAAB mixer w.r.t. other works.

Various works	CMOS Technology	VDD(v)	RF GHz	P _{DC} mW	P _{LO} dBm	Gain dB	NF dB	P _{1dB} dBm	IM ₃ dBm	FMO dB
Darabi[2]	0.13μm	1.2	2	2.4	-	0.5	13.5	-1.5	10.5	-
Vidojkovic[12]	0.18μm	1.8	2.4	8.1	-2	15.7	12.9	-	1	46.2
Hermann [13]	0.13μm	0.6	2.5	1.6	-1	5.4	14.8	-9.2	-2.8	44
Park[12]	0.18μm	-	5	7	-1	16.2	9.8	-	-5	46.6
Sullivan[15]	0.8μm	5	1.9	13.3	-3	9.7	7.8	-	-1	34.6
Test Chip	0.18μm	1.5	2.4	18	-17	15.7	10.7	-10	-9	48.3
Test Chip	0.18μm	1.5	2.4	18	-14	17.5	10.5	-12	-8	48
Test Chip	0.18μm	1.5	2.4	18	-5	23	9.7	-10	3.5	47.6

From the comparison, it was found that the DBGC CAAB mixer had the highest conversion gain (17 dB), the lowest LO power requirement (-14 dBm), and maintained a relatively low noise-figure of 10.5 dB.

V CONCLUSIONS

The relationship of noise, conversion gain and linearity of a Gilbert mixer was derived and analysed. One of the best solution to tackle this problem was to magnify the local

oscillator's output together with injecting a bleeding current into the mirror transistors pair. The proposed double balanced Gilbert-cell class-A amplifier bleeding mixer (DBGC CAAB mixer) was designed using 0.18 μm CMOS technology. The power supply voltage was at 1.5 V with 12mA. From Table -2, it was observed that this design provided highest conversion gains from 15.7 to 23 dBm at a very low LO power feeding of -14dBm. Moreover, the low noise figures and high FOM indicated that this design was good for practical implementation at 2.4 GHz band.

ACKNOWLEDGMENT

The support of the Electronic and Information Engineering, The Hong Kong Polytechnic University is gratefully acknowledged.

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