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A High-sensitivity 135 GHz Millimeter-wave Imager by Compact Split-ring-resonator in 65-nm CMOS

Nan Li^{1,2}, Hao Yu^{1*}, Chang Yang¹, Yang Shang¹, Xiuping Li², Xiong Liu³

¹School of Electrical and Electronic Engineering Nanyang Technological University, Singapore

²School of Electronic Engineering, Beijing University of Posts and Telecommunications, Beijing, China

³Electrical Engineering Department, Northwestern Polytechnic University, USA

*Email: haoyu@ntu.edu.sg

Abstract: A high-sensitivity 135 GHz millimeter-wave imager is demonstrated in 65 nm CMOS by on-chip metamaterial resonator: a differential transmission-line (T-line) loaded with split-ring-resonator (DTL-SRR). Due to sharp stop-band introduced by the metamaterial load, high-Q oscillatory amplification can be achieved with high sensitivity when utilizing DTL-SRR as quench-controlled oscillator to provide regenerative detection. The developed 135 GHz mm-wave imager pixel has a compact core chip area of 0.0085 mm² with measured power consumption of 6.2 mW, sensitivity of -76.8 dBm, noise figure of 9.7 dB, and noise equivalent power of 0.9 fW/ $\sqrt{H_z}$ Hz. Millimeter-wave images has been demonstrated with millimeter-wave imager integrated with antenna array.

Index terms: metamaterial resonator, CMOS 65 nm, imaging system, super regenerative receiver (SRX)

1. Introduction

Millimeter-wave (mm-Wave) radiation above 100 GHz has a moderate wavelength that not only allows penetration into nonconductive materials with nonionizing nature, but also enables high spatial resolution. It shows great potential for imaging applications, including both active and passive imaging, in various areas such as pharmacy and security [1]–[3]. With the recent advance of CMOS technology, CMOS-based mm-wave (60–300 GHz) signal transmitting and receiving components have been recently developed [4]–[5], [23]–[25]. The major advantage offered by CMOS technology over III–V approaches is the possibility of constructing low power and low cost full 2D mm-Wave imaging arrays, as illustrated in Fig. 1(a). However, for active imaging system, the signal strength of mm-wave radiation above 100 GHz is usually weak when illuminated by CMOS source; for passive imaging system, the received radiation from the object itself, such as black body radiation, or the reflection due to other illumination sources that are much lower. In addition, large path loss will be introduced by absorption and diffraction during the propagation. As such, one primary challenge of a CMOS mm-wave imager is to design a high-sensitivity receiver to generate sufficient margin in the link budget. As illustrated in Fig. 1(a), each mm-wave image pixel can consist of a receiver with an antenna. Schottky barrier diodes (SBDs) as shown in Fig. 1(b) have been used to implement the mm-wave imaging detector traditionally. For example, Excellent NEP performance for III-V semiconductor SBD has been reported in [6] (20 pW/Hz at 800 GHz). A Shallow Trench Separated (STS) Schottky barrier diode with 1.5 THz cut-off frequency can be fabricated in 130 nm digital CMOS [7]. Other than the direct-conversion receiver with SBD, the recent super-regenerative receiver (SRX) is also explored [8] with high sensitivity, which is optimum for mm-wave imaging [9]. Note that even though the data rate of the SRX is usually limited by the bandwidth, it is still more than enough to carry a VGA video stream. As depicted in Fig. 1(c), the core of the SRX is a quench-controlled oscillator, which consists of a resonator with positive feedback to realize an oscillatory amplification. When a periodic quench-control signal is applied, the average of the detected signal envelope is amplified.

However, the Q-factor degradation of conventional CMOS LC-tank in mm-wave range can significantly degrade the performance of one SRX.

Fig. 1. (a) Large-arrayed super-pixel CMOS millimeter-wave imaging system, (b) Schottky barrier diode based detector, (c) Super regenerative receiver

Recently, a metamaterial-based resonator has been explored in [10]-[11] to improve the Q with compact area at mm-wave frequency region. A split-ring-resonator (SRR) can be designed in CMOS process with top-metal layer. When loading SRR to a host transmission-line (TL-SRR), the integrated structure becomes a non-transmission medium with single negative property ($\mu \cdot \epsilon < 0$) in the vicinity of resonance frequency. A sharp stop-band is thereby formed such that the incident mm-wave can be perfectly reflected at SRR load with a stable standing-wave established in the host T-line. Compared to the traditional LC-tank based resonator, TL-SRR has enhanced EM-energy storage capability within a compact area by stacking multiple SRR cells, which results in much higher Q factor. As such, it becomes relevant to study the CMOS on-chip SRR for the compact and high sensitivity SRX design of the mm-wave imager.

2. Overview of CMOS Millimeter-wave Imager

In this section, the two typical detector design, Schottky barrier diode and super regeneration, for mm-Wave imaging by CMOS technology is described and compared in brief firstly. For mm-wave imaging system with both of active imager and passive imager, the metrics, such as sensitivity, noise equivalent power (NEP), noise equivalent temperature difference (NETD), are applied to evaluate the performance of variety of imager. Among these specifications, NEP is the most critical specification relevant to the sensitivity of power detectors in both active and passive imagers, other specifications, such as NETD, is related to other system specification, such as integrating time interval. The analysis in this section is focusing on the NEP of each type of detector, qualitative comparison is given based on analytical calculation of NEP for each type of detector.

2.1 Direct Conversion with Schottky-barrier Diode

Block diagram of Schottky-diode-based detector is shown in Fig. 2(a). Connected with an antenna, a DC-biased Schottky diode serves as envelope detector. The equivalent circuit of Schottky diode is also illustrated in this figure. The series resistance is

R_s , junction capacitance is C_j and R_j represents the dynamic resistance ($1/g_m$) of diode. NEP of the diode detector with modulation frequency higher than the $1/f$ noise corner frequency is denoted [7] [12]

$$NEP \approx \frac{4nt_{\omega}^{1/2} \cdot (kT)^{3/2}}{q} \cdot \frac{1 + \frac{R_s}{R_j} + \frac{R_j}{R_s} \left(\frac{\omega}{\omega_c} \right)^2}{R_j^{1/2}} \quad (1)$$

where ω_c is cut-off frequency of the diode which agree to $1/(R_s C_j)$. R_s, C_j scale inversely and linearly with the number of unit cells. At operating frequency of 150 GHz, a bunch of NEPs are plotted versus scales of R_j in Fig. 2(b) with typical diode ideality factor n of 1.35 and unit diode R_s of 10 Ω [13]. From Fig. 2(b), one can observe, for Schottky diode converter in CMOS process, NEPs of Schottky detector array with different number of Schottky diodes located roughly at $\text{pW}/\sqrt{H_z}$ level. NEP degrades as the increasing of cell numbers which limits performance in large array applications.

Fig. 2. (a) Equivalent circuit model of Schottky barrier diode detector. (b) Calculated NEP of detectors with different diode junction resistance and the number of unit cells.

2.2 Super Regeneration with Quench Oscillator

The block diagram of super regenerative power detector is shown in Fig. 3(a). It consists of a quench-controlled oscillator injected by an external signal and an envelope detector. This type of imager merges the both amplifier and detector into one module. An equivalent circuit model of super regenerative amplifier (SRA) is shown in Fig. 3(b). The resonator is modeled by an RLC block, and its oscillation is quench controlled by a time-dependent negative resistance $-1/G_m(t)$, where $G_m(t)$ is the equivalent conductance determined by the associated active devices. The external signal injected is modelled as a time-dependent current source $I_i(t)$. $V_o(t)$ is the output voltage. The resonance frequency is $\omega_o = 1/\sqrt{LC}$; the quality factor is $Q_o = R/Z_o = 0.5\zeta_o$; ζ_o and Z_o are quiescent damping factor and characteristic impedance, respectively. Assuming $G_m(t)$ varies much slower than ω_o such that a quasi-static condition holds in the system to have a time-varying transfer function in the s -domain by:

$$\frac{V_o(s, t)}{I_i(s)} = \frac{Z_o \omega_o s}{s^2 + 2\zeta(t) \omega_o s + \omega_o^2} \quad (2)$$

where is $\zeta(t) = \zeta_o [1 - G_m(t)R]$ the instantaneous damping factor. By varying $\zeta(t)$, the pole can be shifted between left and right sides of the s -plane periodically. In other words, the oscillation starts in SRA when $\zeta(t)$ is negative, and stops when $\zeta(t)$ is positive. Note that (3) is only valid when SRA works in the linear mode such that $V_o(s, t)$ is small enough to prevent significant

distortion in each quench cycle. Generally, SRA working in the linear mode is preferred in the application of millimeter-wave imaging since it has a better sensitivity than that in the logarithmic mode [14].

The sensitivity of SRA is defined as the minimum detected power that means the induced output signal power is the same as its variance.

$$S_{SRA} = P_{\min} |_{I_x^2 = \sigma_x^2} = \frac{I_o^2 R}{2} |_{I_x^2 = \sigma_x^2} \quad (3)$$

where I_x is the equivalent current induced in SRA in response to the ac input I_i , σ_x^2 is the variance of I_x . As discussed in [15], for a typical ramp-damping function with a normalized ramping slope of k , we have

$$I_x = \frac{I_o \omega_o \sigma_s}{2}, \sigma_x^2 = \frac{N \omega_o^2 E_g}{2} \quad (4)$$

where $\sigma_s = \sqrt{2Q_o/\omega_o k}$ is the SRA time constant with a unit of $s/\sqrt{rad}$, k is the normalized ramping slope of time variant conductance G_m with the unit of $1/s$ [14]. $E_g = \sigma_s \sqrt{\pi}$ is the energy of density function, and N is the noise power density with $N = 4KTF/R$. Note that K and F denote the Boltzmann constant and noise factor of SRA contributed by active devices, respectively. As such, the noise equivalent power (NEP) can be calculated by $S/\sqrt{3_{dB} Bandwidth}$:

$$NEP = 1.38KTF \sqrt[4]{\frac{k \omega_o}{\pi Q_o}} \quad (5)$$

Fig. 3. (a) Block diagram of super regenerative imager, (b) Equivalent circuit model of super regenerative amplifier (c) Calculated NEP versus the Q-factor of resonator

The relation between NEP and Q-factor of resonator is analytically calculated by (5) and shown in Fig. 3(c). The noise figure 10dB introduced by MOSFET is determined by simulation at 140 GHz. The rising time of quench signal is set from 20ns to 80ns with 10ns step size. As shown in Fig. 3(c), the Q-factor of resonator has a major impact on the minimum NEP which the SRA can achieve. With improved Q-factor of resonator, the NEP of super regenerative receiver can be decreased dramatically.

3. Millimeter-wave Imager by Metamaterial Resonator

Metamaterial with a negative refraction index was first demonstrated in 2001 with split ring resonator (SRR) showing $\mu < 0$ at resonance frequency [16]. A planar SRR structure can be considered as a magnetic dipole excited by the magnetic field (H-field)

along the ring axis and originally designed as high-Q sub-wavelength resonators [17]. In this section, high-Q compact stacked SRR working in millimetre wave is analysed and designed by CMOS technology. The performance comparison is performed to illustrate the advantage of SRR over conventional LC tank.

3.1 On-chip SRR

In order to design a compact SRR with on-chip implementation, stacked SRR configuration is utilized to lower the resonant frequency with compact dimension. The on-chip SRR can be implemented in a stacked configuration with an on-chip multi-layer back end of line (BEOL) [10]. More stacked layers result in a lower resonant frequency, but suffer from lower Q simultaneously. The larger sizes of SRR result in the lower resonant frequency. The proposed SRR unit-cell is realized by the top two metal layers stacked alternatively, considering the trade-off among resonant frequency, area and loss. The distance between the two rings and the gap of the ring are reduced to lower the resonant frequency. Then, the target resonant frequency can be achieved with high area-efficiency and low loss.

3.2 On-chip DTL-SRR

For mm-Wave imaging purpose, a stacked compact SRR loaded by differential transmission line (DTL-SRR) is design for 140 GHz. As shown in Fig. 4(a), the DTL-SRR is designed by stacked SRRs with the same dimensions of $24 \times 24 \mu\text{m}^2$ in four metal layers (M5 to M8). All SRRs are closely coupled to the same host T-line implemented in the top most metal layer (M8). The overall size of the proposed DTL-SRR is $35 \times 34 \mu\text{m}^2$. The property of DTL-SRR can be analyzed by the T-line model [18] with distributed series impedance (Z) and shunt admittance (Y), which are determined by ϵ and μ , respectively. It can be shown that the $\epsilon > 0$ and $\mu < 0$ condition is satisfied by the DTL-SRR in the frequency range $\sqrt{\frac{1}{L'_s C'_s}} < \omega < \sqrt{(L + L'_s)/LL'_s C'_s}$, where $L'_s = C_s M^2 \omega^2$ and $C'_s = L_s / (M^2 \omega^2)$ are the equivalent series inductance and capacitance of the SRR [19]. Note that M needs to be sufficiently high for a negative μ . As such, a differential host T-line is deployed in the design with SRRs placed in between as close as possible.

For the purpose of comparison, a traditional LC-tank resonator is designed in the M8 metal layer, as shown in Fig. 4(b), which has the same resonance frequency of 135 GHz. The S-parameters of both structures are also verified by EMX with the same parasitic capacitance of 18 fF.

The performance of both types of resonator is extracted by EM simulation. The differential impedance analysis is used for Q-value extraction. An ideal capacitor is connected in parallel to resonator to imitate the capacitance introduced by cross-coupled pair to shift the frequency to 140 GHz. Differential one-port impedance analysis to the resonator network is used to correctly characterize the Q factor of proposed resonators. First, the proposed resonators are operating in differential mode in the oscillator; secondly, differential mode characterization will not be affected by the common mode shunt elements in the network. An ideal capacitor (C) is connected in parallel to the resonator to shift the resonance frequency to the targeted value. Note that the

additional ideal capacitor ($Q_C \rightarrow \infty$) will not affect the Q_{total} result of resonator ($\frac{1}{Q_{\text{total}}} = \frac{1}{Q_{\text{Resonator}}} + \frac{1}{Q_C}$). As such the Q factor can be obtained by:

$$Q_{\text{Resonator}} = Q_{\text{Total}} = \omega_o / BW_{3dB} \quad (6)$$

where ω_o is the resonance frequency, BW_{3dB} is the 3dB bandwidth of total impedance of resonator and shunt C. Alternatively, the same Q value can also be obtained by phase based method:

$$Q = \frac{\omega_o}{2} \frac{d\angle(Y(j\omega))}{d\omega} \quad (7)$$

where $\angle Y(j\omega)$ is the phase of resonator admittance. Note that for the same resonance frequency, the equivalent inductance of DTL-SRR or DTL-CSRR resonator should be the same as LC-Tank.

Fig. 4. (a) Proposed stacked SRR at 140 GHz with equivalent circuit model. (b) Conventional LC-tank resonator at 140 GHz with equivalent circuit model.

Fig. 5(a) (b) shows the comparison of designed DTL-SRR and LC tank in 65 nm CMOS process with the same equivalent inductance of 81 nH. Note that the trace widths in both resonators are the same for a fair comparison. As shown in Fig. 5(a), the Q factors of DTL-SRR extracted from the differential impedance (Z_{Diff}) between P1 and P2 (Fig. 4) is 40, which is more than two times higher than the 18.5 of LC-Tank. The enhanced Q of SRR can also be illustrated by phase characteristics of differential admittance (from (7)) in Fig. 5(b). The higher Q factors of SRR based resonator is mainly due to the enhanced energy storage in the resonators coupling and reduced energy loss. Note that a SRR based resonator with stacked structure will help to confine the EM energy into the silicon dioxide (SiO_2) and reduce the less energy loss in the lossy silicon substrate (Si), resulting the resonators with higher Q. Moreover, the DTL-stacked SRR resonator layout area ($1190 \mu\text{m}^2$) is less than half of the LC-tank resonator ($2500 \mu\text{m}^2$). Such Q factor enhancement effect can also be explained by the strong phase non-linearity in the frequency range closed to SRRs resonance. Fig. 5(c) shows the impedance diagram of both DTL-SRR and LC-Tank without any capacitor loading. A resonance generated by the SRR loadings is observed at 167 GHz for DTL-SRR. Such resonance causes non-linear phase shift at 140 GHz. As shown in Fig. 5(d), DTL-SRRs has much stronger phase non-linearity around 140 GHz than that of LC-Tank.

Fig. 5. (a) Normalized magnitude of differential impedance of LC-tank and DTL-SRR. (b) Phase diagram of differential impedance of LC-tank and DTL-SRR. (c) Impedance diagram of DTL-SRR and LC-Tank in 65 nm CMOS process. (d) Phase diagram of DTL DTL-SRR and LC Tank in 65 nm CMOS process

4. Experimental results

4.1 DTL-SRR based SRX

To demonstrate the high-Q advantage in mm-Wave imaging, a 140GHz super regenerative detector is design based on designed stacked compact SRRs in CMOS.

Fig. 6.(a) Schematic of proposed 140 GHz SRX with stacked DTL-SRRs (b) Die micrograph of fabricated mm-wave SRX with stacked DTL-SRR.

Fig. 6 (a) depicts the schematic of a 140-GHz DTL-SRR-based SRX. Firstly, a transformer based matching network is applied to the input matching for M1 for the electrostatic discharge (ESD) protection when integrating with the antenna; secondly, the virtual ground formed by two T-lines is replaced by the high metal–oxide–metal (MOM) capacitor to further reduce the chip area; thirdly, the detected envelope signal is directly averaged by an on-chip low-pass filter formed by R3 and C3 at the output.

The proposed SRR-based SRX is implemented in 65 nm CMOS RF process, and its chip micrograph is shown in Fig. 6(b). It has a total die area of $570 \times 460 \mu\text{m}^2$, and a core area of 0.0085 mm^2 . The receiver chip is firstly measured on probe station with a VDI signal source (135GHz, 0dBm) directly injected by GSG probe. The input RF signal level is calibrated by R&S spectrum analyzer with harmonic mixer. A 12.5 MHz sinusoid quench-control signal is applied from a function generator (Agilent 33250a) with voltage sweep-range of 0~400 mV. The receiver consumes 6.2 mW under 1 V supply. The frequency response of the receiver is measured when the input power is -51.5dBm. A narrow bandwidth (BW) of 530MHz is obtained around the center frequency of 135 GHz. Receiver sensitivity (S) is measured as -76.8 dBm from the normalized V_{out} to P_{in} at 135 GHz; and the noise equivalent power (NEP) is $0.9 \text{ fW}/\sqrt{H_z}$.

TABLE I: PERFORMANCE COMPARISON OF STATE-OF-THE-ART RECEIVERS FOR IMAGING APPLICATION

	[4]	[21]	[22]	This work
Technology	CMOS 65nm	CMOS 65nm	BiCMOS 0.18 μm	CMOS 65nm
Frequency (GHz)	144	183	103	135
Sensitivity (dBm)	-74	-72.5	-56	-76.8
Noise Figure(dB)	10.2	9.9	15.0	9.7
Bandwidth (GHz)	0.94	1.4	20	0.53
NEP (fW/Hz ^{0.5})	1.3	1.51	17.8	0.9
Power (mW)	2.5	13.5	225	6.2
Core Area (mm ²)	0.021	0.013	0.75	0.0085

The performance of the measurement results of the proposed SRXs is summarized in Table I as well as the previous state-of-the-art receiver designs. Compared to the direct conversion receiver [22], the SRXs have a 16-22 dB better sensitivity due to a narrower receiver bandwidth. Compared to the traditional SRX designs with the *LC*-tank resonator [4], [21], the proposed SRXs are showing 30%~50% reduced NEP, 2.8~4 dB better sensitivity, and 60% area reduction, which makes it more suitable for the portable millimeter-wave imaging with a large sensor array. However, more metal layers are required to form a DTL-SRR structure, which will also increase the chip cost.

Fig. 7.(a) Simulated radiation pattern of 2X4 antenna array (b) Printed circuit board (PCB) integration of CMOS 135-GHz SRX with antenna array mounted on evaluation board.

4.2 Integrated mm-Wave imager with Antenna array

A 2×4 antenna array using hybrid series/parallel feeding is designed and fabricated in Roger RT5880 with size of $8\text{mm} \times 8\text{mm}$. The antenna has 15.4 dBi simulated gain at 135 GHz; the simulated radiation pattern is shown in Fig. 7(a). Its input is matched to

50 Ω with measured S11 below -10dB from 124 to 139GHz [21]. Wire-bonding is applied to integrate the antenna and receiver chip on PCB. The effect of wire-bonding on the input matching is also considered during the antenna design. The whole integrated mm-Wave imager with antenna array mounted on demonstration board is illustrated in Fig. 7(b).

4.3 Mm-Wave Imaging

Based on the designed CMOS detector chip and antenna array, the entire millimeter-wave imaging demonstration is set up and shown in Fig. 8(a). The 135-GHz radiation from a VDI source (0 dBm) is received by proposed imager after propagating through the objects under test, which is held by an X–Y moving stage (STANDA) placed in the middle. Generally, the frequency tuning range (FTR) of the source needs to cover the variation of the designed SRX operation frequency at advanced CMOS process nodes. Moreover, the tuning step size has to be sufficiently small to enable an accurate matching to the operation frequency of the designed SRX. Although a substantial portion of the object is illuminated due to the divergent beam from the source antenna, only the power propagating to the direction of receiver is detected. As such, a high-resolution image can be obtained without a focus lens. The resulting V_{OUT} at each X – Y stage position is recorded into a 2-D matrix, which can be plotted in a color image by MATLAB with a JET color map. It is worthy mention that there exists the frequency calibration (alignment) issue between transmitter and receiver, due to the PVT variation. For our SRA-based receiver part, varactor or switch capacitor array (SCA) can be utilized to be merged into the SRR based resonator design to achieve a tuneable resonator to decouple the influence of PVT variation on resonance frequency. In real-world environment, automatically calibration procedure can be applied by close-loop feedback based on measurement of maximum detected power.

Fig. 8.(a) Millimeter-wave imaging measurement setup with the proposed receiver chip integrated on PCB and object under test fixed on an X–Y moving stage. (b) Various types of oils images captured by imaging system with the proposed CMOS 135-GHz SRX.

The imaging results by the proposed CMOS millimeter-wave imaging system is presented by Fig 8(b). Fig. 8(b) shows the imaging of various types of eaten oil including sunflower, olive, fresh soybean, and soybean that has been used once. Note that four types of oil samples are held by petri dishes with the same inner diameter of 35 mm, in which the oil surfaces are perpendicular to the propagation path between transmitter and receiver. The thickness of each oil sample is controlled through the volume measurement of a syringe pump with an accuracy of 5% before being injected to the petri dish.

5. Conclusions

In this paper, a high-sensitivity millimetre-wave imager has been demonstrated by the compact high-Q quench-controlled metamaterial resonator at 135 GHz in 65-nm CMOS technology. Compared to the conventional SRX design with an LC-tank resonator at a similar frequency, the proposed SRXs have 2.8~4 dB improved sensitivity and 60% reduced area. The proposed 135 GHz SRX is integrated in a millimeter-wave imaging system with various demonstrated imaging applications. It has great potential to be utilized for the future large-arrayed transmission-type millimeter-wave imaging system.

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