Citation	Kaizhe Guo, Patrick Reynaert A 0.59THz Beam-Steerable Coherent Radiator Array with 1mW Radiated Power and 24.1dBm EIRP in 40nm CMOS ISSCC, Feb, 2020.
Archived version	Author manuscript: the content is identical to the content of the published paper, but without the final typesetting by the publisher
Published version	https://ieeexplore.ieee.org/document/9063139
Journal homepage	https://ieeexplore.ieee.org/xpl/conhome/9046640/proceeding
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29.2 A 0.59THz Beam-Steerable Coherent Radiator Array with 1mW Radiated Power and 24.1dBm EIRP in 40nm CMOS

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In THz imaging systems, signal sources are needed to provide illumination of the object. In several reported imaging systems working at 0.3THz, 0.62THz and 0.81THz, an output power around 1mW is required to achieve an acceptable dynamic range [1]. In these systems, the THz signal is generated using III-V devices and waveguide assemblies, which are expensive and bulky. In contrast, CMOS technologies have the advantages of compact size, low cost and high integration. Until today, 1mW radiated power has not been achieved at frequencies higher than 0.35THz in silicon technologies. Besides high output power, beam-steering ability of a signal source is also essential as it can replace the bulky and slow mechanical beam-steering approaches often found in THz imaging systems [1]. In this paper, a 0.59THz beam-steerable coherent radiator array is presented in 40nm CMOS. Combining 36 coherent radiators on 0.68mm², it achieves 1mW radiated power, 24.1dBm EIRP and 0.8% DC-to-THz efficiency at 586.7GHz. Besides, a beam-steering range of 30° is achieved in the E-plane.

As shown in Fig.1, the coherent array has 36 oscillator cores operating at 147GHz (f_0) which generate an in-phase signal at the 4th harmonic frequency of 0.59THz ($4f_0$). Each oscillator core is a cross-coupled pair with microstrip line between the gate of one transistor and the drain of the other transistor. At $4f_0$, this microstrip line transforms the low impedance at the gate to a high impedance at the drain, and stops the 4th harmonic power generated at the drain of one transistor going into the gate of the other transistor. The 4th harmonic power is delivered to the antenna through a capacitive L-type impedance matching network with compact layout. The DC supply network provides an open circuit at $4f_0$. At $2f_0$, it provides a reactance required for maximizing the 4th harmonic generation. The area of one oscillator core together with the antenna is 0.01mm². CMOS technology allows metal lines with much smaller width and thickness compared to III-V technologies. This ability is utilized in this work to design a compact coupling structure in the array with extremely low impact on the oscillator core itself. As shown in Fig.1, the coupling structure has a via resistor in the middle, and two 0.36μm-wide microstrip lines on both sides connected to the gates of the transistors of adjacent oscillator cores. The resistor uses parallel-series connection of 120 vias to reduce the impact of process variation. For differential-mode excitation, a virtual ground is created at the center of the coupling structure. The 24Ω via resistor is transformed to an impedance equivalent to a 50.8Ω resistor in parallel with a 66.4pH inductor at f_0 . This parallel resistor stops any differential-mode oscillation. For common-mode excitation, a virtual open is created at the center of the coupling structure. As the parasitic parallel capacitance of the via resistor is only 0.5fF and the width of the microstrip line is very narrow, the impedance of the coupling structure at f_0 is equivalent to a small capacitor in parallel with a big resistor (4.7fF and 29kΩ). The small capacitor only contributes to 6.7% effective capacitance in the oscillator tank, and the big resistor has negligible impact on the overall quality of the oscillator tank. At $4f_0$, the coupling structure shows an impedance of 3.8-j*22Ω connected in parallel to the transistor gate. In simulation, only 0.1dB reduction of the output power at $4f_0$ is caused due to the coupling structure. Two dummy coupling structures are connected to both edges of the array. The dummy coupling structure is half of the real coupling structure with a floating end point which can perfectly duplicate the virtual open in the real coupling structures, and will not cause any imbalance in the radiator array.

The coupling structure also provides beam-steering capability in the radiator array. As shown in Fig.2, in the radiator array, a bulk voltage difference (ΔV_b) can be created between different oscillator rows to create a difference in the free-running oscillation frequency. By doing this, a gradual phase progress will arise from one end of the oscillator chain to the other end of the oscillator chain, due to the induced injection power through the coupling structure between neighboring oscillator cores [2]. The phase difference among the 6 rows of the array results in a radiation beam that can be steered in the E-plane of the radiator array. The simulated phase progress in the radiator array and the beam-steering radiation patterns for different ΔV_b are shown in Fig.2.

In the topology shown in Fig.1, the 2nd harmonic is also delivered to the antenna. The radiation of the 2nd harmonic should be avoided, because this harmonic should be reflected back to the oscillator core to enhance the 4th harmonic output through self-mixing. In this work, a harmonic-selective antenna is designed to suppress the 2nd harmonic radiation without degrading the 4th harmonic radiation. The structure of the antenna is developed from a conventional monopole antenna shown in the left side of Fig.3. The monopole antenna has compact layout and high cross-polarization ratio. However, when it is designed to radiate at $4f_0$, the gain of a 6×6 monopole antenna array at $2f_0$ is only 8.7dB lower than that at $4f_0$, as shown in Fig.3. In the harmonic-selective

antenna shown on the right side of Fig.3, the monopole is now extended, folded and connected to ground. At $4f_0$, the standing wave on the folded metal line changes its current direction at the point with a distance of $\lambda/4$ to the ground point. Then, the current on the two halves of the folded metal line has the same phase, and the power can be radiated. At $2f_0$, the length of the folded metal line is shorter than $\lambda/4$, so the current on the two halves of the folded metal line is out of phase. As a result, the radiation is cancelled. As shown in Fig.3, the gain of the harmonic-selective antenna array at $2f_0$ is 21dB lower than that at $4f_0$.

The chip is fabricated in 40nm CMOS. The backside of the chip is attached on a hemispherical (radius=5mm) high-resistivity silicon lens for backside radiation. The output frequency and radiation pattern are measured using a spectrum analyzer with a WR1.5 harmonic mixer. The EIRP at $4f_0$ is measured using a PM5 power meter. The PM5 power meter also detects the heat radiated from the chip. The impact of the heat is measured by rotating the powered sample by 90° which preserves the heat radiation but blocks the linear-polarized THz signal. The EIRP at f_0 , $2f_0$ and $3f_0$ are too low to be measured directly by the PM5 power meter, so they are measured using the spectrum analyzer with its power value calibrated using the PM5 power meter and VDI source modules [3]. Fig.4 shows the measured output spectrum, the phase noise at $4f_0$, and the EIRP at f_0 , $2f_0$, $3f_0$ and $4f_0$ for a 0.9V supply voltage (V_D). The measured radiation patterns in the E-plane and H-plane are also shown in Fig.4. To measure the antenna directivity accurately, the 2-D radiation pattern of the array is measured and shown in Fig.5. By integrating the measured 2-D radiation pattern, the directivity is calculated as 24dB. The radiated power for 0.9V V_D is calculated as 0.1dBm using the measured EIRP and directivity. The power consumption of the chip for 0.9V V_D is 1.278W. Fig.5 also shows the EIRP and the output frequency of the array for different V_D . The bottom of Fig.5 shows the measured beam-steering radiation patterns normalized to the maximum value in Fig.4. A beam-steering range of 30° is achieved in the E-plane. Fig.6 compares this work with the state-of-the-art silicon-based signal sources with similar output frequencies.

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Figure 29.2.1: Architecture of the 0.59THz beam-steerable coherent radiator array.

Figure 29.2.2: Mechanism of beam steering in the radiator array. Simulated phase progress and beam-steering radiation patterns for different ΔV_b .

Figure 29.2.3: Layout and simulated normalized radiation patterns of the conventional monopole antenna and the harmonic-selective antenna.

Figure 29.2.4: Measured output spectrum, phase noise at $4f_0$, EIRP at f_0 , $2f_0$, $3f_0$ and $4f_0$, and radiation pattern for a 0.9V supply voltage (V_b).

Figure 29.2.5: Measured 2-D radiation pattern, EIRP and output frequency for different V_b , and beam-steering performance.

	This Work	JSSC19 [3]	ISSCC14 [4]	ISSCC16 [5]	JSSC15 [6]
Frequency (GHz)	586.7	531.5	530	553	498
EIRP (dBm)	24.1*	2.3	25	N/A	N/A
Directivity (dBi)	24	14.3	37	N/A	N/A
Radiated Power (dRm)	0.1*	-12	-12**	-27	-16.6***
Tuning Range (%)	0.7	0.9	3.2	3.8	5.1
DC Power (W)	1.278	0.26	0.156	0.172	0.425
DC-to-THz Efficiency (%)	0.8	0.24	0.4	0.01	0.051
Phase Noise at 1MHz (dBc/Hz)	-82	N/A	N/A	-74	-87
Beam Steering (°)	30	60	No	No	No
Array Size	36 (6×6)	4 (1×4)	16 (4×4)	1	1
Area (mm ²)	0.68	2.5	4.2	2.79	0.5
P_{rad} Per Element (dBm)	-15.5	-18	-12	-27	-16.6
P_{rad} /Area (μW/mm ²)	1505	25	238	0.7	44
Technology (f _{max})	40nm Bulk CMOS (300 GHz)	40nm Bulk CMOS (300 GHz)	130nm SiGe (500 GHz)	65nm Bulk CMOS (240 GHz)	90nm SiGe (350 GHz)

* An extra 1.5dB output power can be achieved by increasing the supply from 0.9V to 1V.

** Radiated power of a single source pixel in a non-coherent oscillator array.

*** Measured on wafer using probe.

Figure 29.2.6: Performance comparison between this work and the state-of-the-art silicon-based signal sources with similar output frequencies.

Figure 29.2.7: Die micrograph of the 0.59THz beam-steerable coherent radiator array.

Figure 29.2.8: Measurement setup used to measure the output frequency and radiation pattern.

Figure 29.2.9: Measurement setup used to measure the EIRP at $4f_0$.

Figure 29.2.10: Radiated power of silicon-based signal sources with output frequencies from 0.3THz to 0.7THz.

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