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Design Techniques and Measurement Methods for Broadband Millimeter-Wave and THz Systems in Silicon

by

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Abstract

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Short impulses in millimeter-wave (mm-wave) and THz regimes (30 GHz - 30 THz) have a potentially large bandwidth that can be exploited for various applications, for example, high-resolution 3D imaging, high-speed wireless communication, broadband spectroscopy, etc. Existing methods for impulse generation have the following drawbacks: First, photonics solutions are usually not compatible with silicon technologies, i.e. CMOS and BiCMOS, impeding higher level SOC designs and increasing cost; Second, electronic oscillator-based solutions usually require phase-locked loop (PLL) and delay-locked loop (DLL) to ensure coherency of generated impulses, which increases system complexity, power consumption, and die area; Third, existing electronic digital-to-impulse solutions can be further improved by generating shorter impulses, reducing late-time ringing, and achieving amplitude modulation. In addition, high demands on using silicon technology to generate picosecond or sub-picosecond impulses impose challenges on standard chip characterization methods in both time domain and frequency domain.

This dissertation demonstrates three chip designs and one chip characterization method to resolve the aforementioned drawbacks and chal-
lenges. The first chip design is to use a CMOS-compatible silicon photonics process technology to design a THz photoconductive antenna (PCA) chip, which can radiate 1.14 ps impulses. The prototype silicon photonics chip enables easier integrations with other photonics and electronics devices on a single chip. The second chip design is to implement an asymmetric-VCO-based impulse radiator without requiring any PLL or DLL in a 130 nm SiGe BiCMOS. With on-chip antennas, it radiates 60 ps impulses with less power consumption, system complexity, and die area than conventional oscillator-based solutions. The designed impulse radiator has also been applied for 3D imaging. The last chip design is to apply a new circuit technique, nonlinear Q-switching impedance, to implement a 4 ps impulse radiator with pulse amplitude modulation in a 130 nm SiGe BiCMOS. An optoelectronics-based time-domain characterization method was invented to test the 4 ps impulse radiator, and this new measurement technique shows a significant accuracy improvement compared with standard time-domain methods.

The demonstrated techniques in this dissertation show that silicon technology is a promising solution to generating picosecond and even sub-picosecond impulses, exploiting the benefits of easier integration with control electronics and low cost with mass production. The demonstrated prototype chips are approaching to the performance of photonics devices. Ultra-broadband silicon-based impulse radiators can be characterized using optoelectronics technology to achieve better measurement accuracy.
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Chapter 1

Introduction

This chapter discusses the background of this dissertation, including the advantages and importance of using silicon technology to create broadband millimeter-wave (mm-wave) and THz systems, and challenges of characterizing broadband silicon chips with conventional methods. This chapter consists of four sections. First, an overview of mm-wave and THz impulse is given. Second, existing techniques for mm-wave and THz impulse generations are discussed. Third, challenges of using standard characterization methods to test ultra-broadband silicon-based impulse radiators are demonstrated. Finally, the organization of the remaining context of this dissertation is listed.

1.1 Overview of Mm-wave and THz Spectrum

Mm-wave and THz spectrum is typically referred to the frequency range from 30 GHz to 30 THz, as shown in Fig. 1.1. Technologies exploiting mm-wave and THz spectrum exhibit several advantages compared with their counterparts using the neighboring electromagnetic (EM) spectrum, i.e. conventional microwave (<30 GHz) and infrared regions. First, wavelength of mm-wave and THz spectrum is shorter than that of con-
ventional microwave spectrum, which enables higher imaging resolution and smaller feature size. Second, mm-wave and THz radiation can penetrate through many non-metallic mediums that are very lossy or even opaque to infrared and visible light wavelengths, expanding human being’s vision. For example, THz radiation can be transmitted through plastics, clothing, paper products and other non-conductive materials [1, 2]. In addition, millimeter-wave radiation can be used in fog condition that normally limits the visibility of both infrared and visual sensors [3]. Third, many hazardous compounds, such as explosives and illicit drugs, have characteristic transmission or reflection spectra in the mm-wave and THz regimes rather than conventional microwave or infrared wavelengths.

![Figure 1.1: Mm-wave and THz spectrum lies in the frequency range from 30 GHz to 30 THz.](image_url)

These unique properties of mm-wave and THz spectrum demand techniques to generate signals in this EM spectrum. As shown in Fig. 1.2, there are two kinds of signals that can be generated directly in this spectrum. One is continuous wave (CW), the other is picosecond or sub-picosecond impulse. For continuous wave, there is, ideally, a Dirac delta function in its frequency spectrum. For picosecond or sub-picosecond impulse, because of its very short timing duration, its frequency spectrum have a large bandwidth and its SNR > 1 bandwidth is usually greater than 100 GHz. Because of this unique property of ultra-large bandwidth, various applications have
been proposed and demonstrated using very short impulses, such as high-resolution 3D imaging [4], high-speed and secure communication [5, 6], medical diagnostics, and broadband spectroscopy for detecting gas and hazardous material [7, 8, 9, 10]. As a result, there exists a high demand on generating broadband mm-wave and THz impulse.

1.2 Existing Technical Solutions for Impulse Generation

Existing technical solutions for generating mm-wave and THz impulses can be divided into two categories: silicon-based solutions and non-silicon-based solutions. For non-silicon based solutions, laser technologies and optical methods are the main representatives. However, both silicon-based and non-silicon based solutions have drawbacks that demand improvements.
1.2.1 Silicon-based Solutions

Thanks to the rapid development of silicon technologies, maximum oscillation frequency ($F_{\text{max}}$) of silicon transistors has surpassed 500GHz [11, 12]. However, silicon transistors in the most advanced technology nodes (CMOS and BiCMOS) have low power gain in the mm-wave spectrum. This shortcoming has made most of existing silicon solutions focus on narrowband signal generation in the mm-wave and THz spectrum. They are usually based on an "up-conversion" principle: silicon circuits extract high-order harmonics of fundamental oscillations to act as a mm-wave or THz CW signal source.

Traditionally, in silicon technology, short impulses can be generated by modulating the produced CW signal with a short digital time-gating signal. Because CW signal generation usually requires oscillators, this type of implementation can be called "oscillator-based solution". For many applications, such as time-of-flight distance measurement and broadband spectroscopy, it is desired to generate coherent impulses, which requires a synchronization scheme between the time gating mechanism and the CW source. Fig. 1.3 presents an example of silicon-based impulse generator with a synchronization scheme between the oscillator and the switching mechanisms, which was applied in [13] and [14]. In this system architecture, the time-gating signal is applied to switch a power amplifier and an on-chip antenna to generate short impulses. To ensure the coherency of produced impulses, a phase-locked loop (PLL) and a delay-locked loop (DLL) are co-designed to synchronize the switching operations and the on-chip oscillator. As a result, this implementation increases system complexity, power consumption, and die area significantly. In the Chapter 3 of this dissertation, a circuit technique is introduced to resolve this drawback by achieving the coherency of generated impulses without using any PLLs or DLLs in an oscillator-based design.

A new silicon-based technique, called digital-to-impulse (D2I), was proposed in
Figure 1.3: A conceptual illustration of oscillator-based impulse radiator with a synchronization scheme between the oscillator and the switching operations.

2014 [15]. As shown in Fig. 1.4 [15], in this solution, a digital trigger signal is firstly converted into an ultra-fast transient current by a switch transistor. With an impulse matching network, the ultra-fast transient current is delivered to an on-chip antenna, which radiates a short impulse into free space. Consequently, this direct conversion principle ensures the coherency of the radiated impulse without using any PLLs or DLLs. Compared with the previously described oscillator-based solution, this technique exhibits huge improvements on system complexity, power consumption, and die area. Apart from that, measurement results have shown that the D2I solution can generate 8 ps radiated impulse, which is shorter than those by state-of-the-art oscillator-based designs. Despite of these advantages, there are still some improvement requirements for D2I solutions. First, smaller late-time ringing in the impulse waveform is preferred so that less ambiguity and inter-symbol interference will occur. Second, impulse amplitude reconfiguration and modulation are needed for
impulse-based high-speed communication. In the Chapter 4 of this dissertation, a circuit technique is presented to accomplish these two improvements.

![Digital-to-Impulse Solution Diagram]

Figure 1.4: A conceptual illustration of digital-to-impulse solution.

1.2.2 Non-Silicon-based Solutions

Non-silicon-based solutions were invented much earlier than silicon-based solutions mainly due to the inferior performance of silicon technologies decades ago. Even now, many non-silicon-based solutions for mm-wave and THz impulse generation are widely used in both academia and industry for superior performance. Commonly-used techniques include free-electron lasers (FELs) [16], quantum cascade lasers (QCLs) [17], electro-optic rectification [18], and photoconductive antennas using ultrafast charge transport [19, 20]. In FELs, a magnetic undulator drives an accelerated electron beam to produce high power coherent pulsed THz radiations. The average radiated power can reach to nearly 20 W [16]. In QCLs, electrons undergoes a series of intersubband
transitions with the help of multiple serial quantum-well heterostructures, radiating photons and causing EM coherent radiations. Quantum well heterostructures are usually formed by sandwiching a material, like gallium arsenide between two layers of a material with a wider bandgap, such as $Al_xGa_{1-x}As$. QCLs have been widely used for infrared and THz radiations, and there are challenges of designing a QCL that can generate mm-wave coherent radiations [21]. Both electro-optic rectification and photoconductive antennas convert femtosecond ($\approx 100\text{fs}$) visible/near-infrared laser pulses into THz pulses via nonlinear optical crystals and ultrafast photoconductive semiconductor materials, respectively. In electro-optic rectification solutions, when a femtosecond visible/infrared laser pulse with a large peak electric field is incident onto an appropriate nonlinear optical crystal, the second-order nonlinear susceptibility of the crystal induces mixing of the frequency components of the incident broadband femtosecond laser pulses, resulting in broadband polarizations and thus radiating a broadband THz pulse [18]. In photoconductive antennas, a femtosecond visible/infrared laser is incident onto an ultrafast photoconductive semiconductor material. If the bandgap of the semiconductor material is smaller than the energy of the incident photons, incident photons are absorbed and electron-hole pairs are created in the semiconductor. When a dc voltage is applied across the material, the generated electrons and holes are drifted and collected by electrodes before photocarrier recombinations occur, inducing an ultrafast transient current, which excites an antenna and radiates a THz pulse [19].

Non-silicon-based solutions can generate a much shorter impulse with a much larger peak energy than silicon-based counterparts, but they suffer from the following limitations. First, they are not compatible with conventional silicon technologies, like CMOS or BiCMOS. Silicon-based DSP units can not be integrated with those non-silicon-based devices to form a mm-wave or THz system-on-chip, which has a higher
level of integration and functionalities. Second, fabrication cost of non-silicon-based solutions is higher than that of silicon-based counterparts. Therefore, it is beneficial to use silicon technologies to achieve the aforementioned working mechanisms in the non-silicon-based solutions. In the Chapter 2 of this dissertation, an example of using a silicon photonics process technology to design a THz photoconductive antenna is presented.

1.3 Challenges of Ultra-Broadband Silicon Chip Characterizations

Standard silicon chip characterizations are usually conducted in either frequency domain or time domain using electronic equipment. However, both methods suffer from limitations when applied to characterize ultra-broadband silicon chips whose SNR > 1 bandwidth is greater than 100 GHz.

1.3.1 Frequency-Domain Methods

Existing mm-wave [22] and THz silicon chips [23] are usually characterized using frequency-domain techniques. Fig. 1.5(a) demonstrates an example of frequency-domain characterization technique for mm-wave and THz silicon chips. A diagonal horn antenna is used to collect the incident mm-wave and THz radiations, and then the collected signal is delivered to a harmonic mixer through a chain of waveguides and adapters. A local oscillator and the harmonic mixer work together to down-convert the collected mm-wave and THz signals to a much lower frequency band (usually smaller than 50 GHz) that a standard spectrum analyzer can process. To further increase the sensitivity of the whole measurement chain, RF amplifiers are sometimes used to amplify the down-converted signal. This type of measurement
chain works well for narrowband signals for the following reasons. First, calibrations on loss and phase distortions induced by the passive waveguides and adapters are straight-forward in a narrow frequency band. Second, only one local oscillator is required for this measurement.

![Diagram of frequency-domain characterization methods](image)

Figure 1.5: A conceptual illustration of frequency-domain characterization methods for (a) narrowband signal and (b) broadband picosecond impulse.

However, this frequency-domain method exhibits several limitations when it is used for picosecond impulses. As shown in Fig. 1.5(b), a picosecond impulse has a large frequency bandwidth and it demands broadband calibrations on the antenna and the chain of waveguides and adapters. In addition, harmonic mixers will generate complicated nonlinear components from both self-mixing of the picosecond impulse and mixing between the picosecond impulse and the local oscillator. The relationship
between the output of the mixer and the incident picosecond impulse will be very complicated. Even though only the first-order mixing output between the picosecond impulse and the local oscillator is filtered out, shown in Fig. 1.5(b), due to the limitation on the spectrum analyzers’ bandwidth, multiple local oscillators covering several microwave bands are needed to obtain the full-frequency spectrum of the picosecond impulse. In this situation, there is a strict requirement on the RF amplifier’s nonlinearity performance because the down-converted signal is broadband.

1.3.2 Time-Domain Methods

The critical component in the time-domain characterization method is electronic oscilloscopes. Sampling oscilloscopes are usually used for testing short impulses because of its much smaller effective sampling interval than real-time oscilloscopes. However, the main obstacle of electronic sampling oscilloscope for measuring picosecond impulse is its limited bandwidth. The best of existing off-the-shelf electronic sampling oscilloscopes has a bandwidth of 70 GHz [24], which is not large enough to sample picosecond impulses with enough accuracy [25]. Of course, a down-conversion scheme can be applied to overcome the bandwidth limitation of sampling oscilloscopes, but the same challenges will happen as in the frequency-domain methods. Therefore, a direct time-domain characterization technique is preferred to test silicon-based picosecond impulse radiators. In the Chapter 5 of this dissertation, details of an optoelectroncis-based direct time-domain characterization technique are described.

1.4 Organization

The remaining context of this dissertation is organized as follows. Three design techniques for impulse generation using silicon technologies are described from Chapter 2
to 4. These techniques resolve the aforementioned limitations in the existing technical solutions for impulse generation. Chapter 5 presents a direct time-domain measurement technique for ultra-broadband silicon chips. Conclusions are given finally in the Chapter 6.
This chapter presents an integrated germanium-based optical waveguide coupled THz photoconductive antenna (PCA) in a low-cost SOI process with potentials to perform THz beam-steering. Some contents in this chapter have been reported in [26]. The designed THz PCA chip works with 1550 nm femtosecond lasers. The radiated THz pulses achieve a full-width at half maximum (FWHM) of 1.14 ps and a bandwidth of 1.5 THz. The average radiated power is $0.337 \mu W$ [26]. Compared with conventional THz PCAs, this design exhibits several advantages: first, it uses silicon-based technology, which reduces the fabrication costs. Second, it works with 1550 nm wavelength that is widely used for optical-fiber communications. This THz PCA design is compatible with various low-cost lasers sources in this field. Third, in this design, a waveguide-coupled photoconductive switch is implemented to enable a monolithic and "horizontal" interaction between the excitation laser propagating out of an on-chip optical waveguide and the photoconductive material. This configuration allows to integrate silicon photonics modules that can manipulate the excitation laser with the
THz PCAs on a single chip so as to build a system-on-chip (SoC) with a higher-level of functionality.

2.1 Introduction

In the past several decades, photoconductive antennas (PCAs) have been used for emitting and detecting THz waves [27]. Fig. 2.1 presents a conventional THz PCA emitter. In a conventional THz PCA, two parts of metal contacts are fabricated on a semiconductor substrate and there is a gap between the two contacts. Part of the two metal contacts also act as an on-chip THz antenna. Conventional THz PCAs are usually triggered by a free-space femtosecond laser pulse, which is incident onto the semiconductor substrate through the gap between the metal contacts. The semiconductor material has an appropriate bandgap energy to absorb the incident light, and hence, photocarriers are generated in the semiconductor substrate. A DC voltage is applied across the gap through the metal contacts, forming an electric field in the semiconductor material below the gap. Consequently, photocarriers are drifted and collected by the metal contacts before photocarrier recombinations occur. The collected photocarriers form an ultrafast transient current, which drives the on-chip THz antenna that produces THz pulse radiations. A silicon lens is usually attached to the back-side of the semiconductor substrate to increase radiation efficiency.

There are some limitations on the semiconductor substrates used in conventional THz PCAs. First, in order to push the bandwidth of radiated THz pulses, the semiconductor substrate must be an ultrafast photoconductive material that should have short photocarrier lifetime and high carrier mobility. Commonly-used materials are radiation-damaged silicon-on-sapphire (RDSOS) and low-temperature-grown GaAs (LT-GaAs) [28]. Because these kinds of materials are specially designed for THz PCA applications, their demands are quite limited and much less than silicon tech-
technologies, and consequently, their costs are higher than electronics devices nowadays. Another limitation is that these materials are not compatible with 1550 nm laser sources. 1550 nm regime is widely used for fiber-optic communications and various 1550 nm laser sources for fiber-optic communications are low-cost and well-developed. Therefore, it is a wise strategy to design a THz PCA chip that operates with 1550 nm wavelength to further reduce its total cost. Due to the aforementioned two limitations, the first technical challenge of this work is to use a low-cost and ultrafast photoconductive semiconductor material that operates with 1550 nm wavelength.
Conventional PCAs also have limitations on the free-space optical excitation scheme. First, this excitation scheme requires accurate optical alignments to ensure that the femtosecond laser is shine onto the tiny gap between the metal contacts. Optical alignment requires high system stability and therefore is not suitable for portable applications. Second, in order to adjust the timing of the optical excitation signal, which is the free-space femtosecond laser, manipulations on the optical excitations are usually performed in the free space by using a mechanical translation stage with retroreflector mirrors. However, this technique requires bulky devices and cannot be integrated with THz PCA modules for SoC purposes. The ultimate form of SoC here should consist of laser sources, photonics control units, and THz PCA units integrated on a single chip to build a fully-integrated miniaturized and portable photonics THz device. Therefore, the second technical challenge of this work is to design an on-chip optical excitation scheme for THz PCA chips.

It is necessary to give a brief overview on CMOS-compatible silicon photonics technology. The goal of this technology is to implement optical functionalities using CMOS-compatible silicon-based technologies and to integrate photonics and electronics devices on a single silicon wafer so as to build a highly-integrated system with mass production capabilities. Therefore, this technology will reduce the cost of photonics devices by tens to hundreds times with mass production. In this work, I used a CMOS-compatible silicon photonics process technology to investigate the possibility of designing THz PCAs. The silicon photonics process technology is provided by the Institute of Microelectronics (IME), Agency for Science Technology and Research (A*STAR), Singapore. As shown in Fig. 2.2 [29], this process has a silicon-on-insulator substrate, and it provides many pre-designed passive devices, such as single-mode optical waveguides and optical grating couplers in 1550 nm regime. Apart from the passive devices, Ge-based active devices, for example, ring modula-
tors and photodetectors, are available in this process that are optimized for operating with 1550 nm wavelength.

Figure 2.2: Illustration of the CMOS-compatible silicon photonics process technology used in this work.

The remaining context of this chapter is organized as follows. Design techniques and system architectures of the prototype PCA chip are presented in the Section 2.2, followed by measurement results described in the Section 2.3. Finally, conclusions are given in the Section 2.4.

2.2 Design Techniques and System Architectures

2.2.1 Ultrafast Germanium Thin Films

The silicon photonics process technology used in the work provides Germanium as the photoconductive material for 1550 nm wavelength. As discussed in the Section 2.1, there are two requirements for the photoconductive materials of THz PCAs.
One is that the photocarrier lifetime of the material should be on the order of sub-picoseconds. The other is that photocarriers’ mobility should be high. A previous work [30] has demonstrated that photocarrier lifetime of Germanium thin films can be reduced to the order of sub-picoseconds by implanting O\(^+\) ions without sacrificing a reasonably high mobility, which was about 100\(cm^2/(Vs)\). Luckily, this investigation can be applied into this work. The process technology can fabricate Germanium thin films with 500 nm thickness, and it also provides Phosphorus implant with a dose of \(4 \times 10^{15} cm^{-2}\). Therefore, the first technical challenge of this work, using a low-cost and ultrafast photoconductive semiconductor material that operates with 1550 nm wavelength, is resolved.

### 2.2.2 Waveguide-Coupling THz Photocondutive Switch

The second technical challenge is to design an on-chip optical excitation scheme for THz PCA chips. Fig. 2.3 demonstrates a conceptual illustration of the proposed waveguide-coupling excitation scheme. In this scheme, the free-space 1550 nm femtosecond laser is firstly coupled into the on-chip optical waveguides through an on-chip optical grating coupler. Then the femtosecond laser propagates in the optical waveguide and reaches to silicon photonics devices that can manipulate the excitation laser by performing amplitude modulation or phase modulation. The processed femtosecond laser keeps traveling in the optical waveguide until it is absorbed by the Germanium thin film, exciting the PCA switch that drives the on-chip THz antenna to radiate THz pulses.

Fig. 2.4 presents the structure of the implemented Germanium-based waveguide-coupling photoconductive switch. Femtosecond laser as an optical excitation signal travels to the PCA switch through on-chip optical waveguides. A tapered transition is designed between the optical waveguide and the PCA switch to reduce undesired
optical reflections. A Phosphorus-doped Germanium thin film is grown on the silicon layer. The N++ (Phosphorus) implant layer is split into two parts to prevent a large dc current produced under dc biasing. In order to increase the transient response speed of the PCA switch, the spacing between the two metal electrodes is set to the minimum value limited by the DRC rule of the process technology. Compared with conventional THz PCA switches, the proposed waveguide-coupling PCA switch exhibits one big advantage. In this design, the femtosecond laser comes from the substrate side rather than through the gap between the metal electrodes as in conventional designs. Therefore, when a small spacing between the metal electrodes is used for increasing response speed of the proposed waveguide-coupling PCA switch, there is no need to worry that the excitation signal will be blocked or plasmonics effects must be taken into account. As shown in Fig. 2.4, on-chip THz antennas are connected to the two metal electrodes.

The Germanium thin film is optimized to achieve both high absorption efficiency and high conversion efficiency in 1550 nm regime. The length of Germanium thin film
Figure 2.4: Structure of the proposed waveguide-coupling THz PCA switch.

is designed to be 20 µm to ensure high absorption efficiency. Fig. 2.5(a) shows that the 20 µm Germanium thin film is long enough to absorb almost all the incident 1550 nm excitation light. Fig. 2.5(b) presents the simulated 1550 nm light mode distributions at different propagation distances within the PCA switch. These simulation results reflect another advantage of using this long Germanium thin film: The traverse size of the light mode is expanded along propagation in the Germanium thin film. As a result, more photocarriers are generated at closer locations to the metal electrodes, and consequently, more photocarriers can be collected by metal electrodes before recombinations occur, which increases conversion efficiency.

The efficiency of THz PCA switches can be potentially enhanced by introducing artificial 3D architectures that exhibit tailored optoelectronic properties. Artificial 3D structures have been proposed and demonstrated for various THz applications, such as THz lasers, THz photodetectors, and THz polarizers [31, 32].
Figure 2.5: (a) Simulated light absorption at 1550 nm within the proposed PCA switch. (b) Simulated 1550 nm light mode distributions at different propagation distances within the proposed PCA switch.

2.2.3 System Architecture

Fig. 2.6 demonstrates the chip micrograph of the proposed Germanium-based optical waveguide coupled THz PCA chip using an SOI-based silicon photonics process technology. It occupies a small die area of 440µm × 680µm. A bowtie antenna is designed on the top metal layer to reduce conductive loss. Metal vias connect the PCA switch and the antenna. An on-chip grating coupler is used to couple the free-space femtosecond laser into the on-chip waveguide, which guides the femtosecond laser to the PCA switch. The main focus of this work is to design the Germanium-based
waveguide-coupled PCA switch, therefore, pre-designed grating couplers and optical waveguides in the process development kit (PDK) are used in the design. These components can be further improved. The chip package is also shown in Fig. 2.6. A highly resistive silicon lens is attached to the backside of the chip to increase radiation efficiency.

![Chip Package](image)

Figure 2.6: Micrograph of the prototype THz PCA chip and chip package.

### 2.3 Measurement Results

The prototype THz PCA chip was characterized in both time domain and frequency domain using an Advantest THz-TDS system (TAS7500TS). The characterization setup is demonstrated in Fig. 2.7. A 50 fs pump laser beam from the Advantest THz-TDS system was firstly coupled to the free space and then focused onto the on-chip grating coupler in the prototype chip. The chip was mounted on a rotation stage, which can be adjusted to achieve maximum coupling efficiency of the on-chip grating coupler. A THz polyethylene lens is used to focus the THz radiation to
the THz detector. To measure average radiated power from the prototype chip, a calibrated pyroelectric detector, which is sensitive from 20 GHz to 1.5 THz, is used with a mechanical chopper modulating the pump femtosecond laser beam in the free space.

![Figure 2.7: Measurement setup for the prototype THz PCA chip.](image)

Fig. 2.8 presents the measured THz pulse radiated by the prototype chip. With a maximum bias voltage of 3.5 V, the designed PCA chip radiates THz pulses with an FWHM of 1.14 ps. Its frequency spectrum, shown in Fig. 2.8(b), is obtained by performing DFT on the measured time-domain waveform. The radiated THz pulse has a peak frequency component at 176 GHz, and it has an SNR>1 bandwidth of 1.5 THz. The measured average radiated power is 0.337 µW and its DC-to-RF conversion efficiency is $3.6 \times 10^{-5}$.

Fig. 2.9 demonstrates the measured effects of bias voltage on the performance of the prototype chip. Theoretically, by increasing the bias voltage, the generated photocarriers in the Germanium thin film have higher drift velocity (before saturation happens), producing stronger transient current, and consequently, generating stronger THz pulse radiations. Measured results confirm the theoretical prediction. When the
bias voltage is increased from 2 V to 3.5 V, the measured THz pulse has a larger peak amplitude [Fig. 2.9(a)], a greater SNR > 1 bandwidth [Fig. 2.9(b)], and a stronger average radiated power [Fig. 2.9(c)]. In this design, the maximum bias voltage that the prototype chip can tolerate is 3.5 V, which is limited by the current capacity of the electrical vias in the THz PCA chip.

2.4 Conclusions

In this work, a CMOS-compatible Germanium-based photoconductive antenna with waveguide coupling scheme was implemented using an SOI-based silicon photonics process technology provided by IME A*STAR, Singapore. In the prototype THz PCA chip, a Phosphorus-doped Germanium thin film is designed to reduce photocarrier’s lifetime and also to allow the THz PCA chip to operate with 1550 nm femtosecond laser. In addition, the proposed waveguide coupling photoconductive switch provides a monolithic on-chip excitation scheme between the incident femtosecond laser and the
photoconductive material. This monolithic interaction configuration enables easier integrations of the THz PCA prototypes and other silicon photonics modules on a single silicon chip. The prototype CMOS-compatible THz PCA chip can radiate 1.14 \( \text{ps} \) THz pulses with an SNR\( >1 \) bandwidth of 1.5 THz and an average radiated power of 0.337\( \mu \text{W} \). The DC-to-RF conversion efficiency of the prototype chip is \( 3.6 \times 10^{-5} \).
Figure 2.9: Measured effects of bias voltage on the performance of the prototype THz PCA chip. (a) Effects on the peak amplitude of the radiated THz pulse. (b) Effects on the SNR > 1 bandwidth of the radiated THz pulse. (c) Effects on the average radiated power of the radiated THz pulse.
3D Radar Imaging based on a Synthetic Array of 30 GHz Impulse Radiators with On-Chip Antennas in 130 nm SiGe BiCMOS

This chapter reports a 30 GHz impulse radiator chip for high-resolution 3D radar imaging, published in [4]. Part of the content is also reported in [33]. In this work, an asymmetrical topology in the cross-coupled pulsed VCO is introduced to minimize timing jitter of radiated impulses to 178 fs, which enables highly efficient spatial combining. The coherent combining over the air has been performed with two widely spaced impulse radiators. The shortest FWHM (Full-Width-at-Half-Maximum) pulse-width of 60 ps is recorded. 3D images of various metallic objects and dielectric objects are produced using a custom-designed synthetic array imaging system. A depth resolution of 9 mm and a lateral resolution of 8 mm at a range of 10 cm have been achieved. The impulse radiator was implemented in a 130 nm SiGe BiCMOS process technology with an area of 2.85 $mm^2$ and an average power consumption of 106 mW. Compared with conventional oscillator-based impulse generator designs, the implemented chip does not require PLLs or DLLs to ensure the coherency of the produced impulse.


3.1 Introduction

Over the past decade, microwave and mm-wave impulse radiators have been reported for short-range high-data rate communication [34], vital-signal monitoring [35], security imaging [36], spectroscopy [37], and short-range automotive radar [38]. Impulse radiators have three main advantages compared with continuous-wave (CW) radiators. First, impulse radiators have larger instantaneous bandwidth (BW) than CW radiators. Second, timing ambiguity can be suppressed without sacrificing image resolution by using repeated ultra-short impulses with a large repetition period. Third, energy consumption of impulse radiators can be easily adjusted by varying the repetition rate of radiated impulses.

Over the last few years, several groups have reported silicon-based mm-wave impulse radiators for imaging applications [39, 40, 15, 41, 42, 43, 44, 45]. Main objectives in designing silicon-based imaging radars are high image resolution, large image range, and short acquisition time. There exists a trade-off between higher image resolution and larger image range: Higher image resolution requires larger bandwidth of shorter impulses, but shorter impulses have lower average power, which degrades SNR and image range. This trade-off can be mitigated by coherent combining of radiated impulses over the air. By synchronizing radiators and applying proper time-delay to each radiator, the radiated impulses can be combined coherently at a certain point in 3D space. As a result, SNR at the combining point is increased with the number of radiators.

The majority of published studies on silicon-based impulse radiators adopt oscillator-based architectures [43, 44, 45]. In [43], co-locked delay-lock-loop (DLL) and phase-lock-loop (PLL) were used to generate coherent impulses, resulting in a significant increase in power consumption, system complexity, and die area. In addition, only distance measurement (1D imaging) was performed in [43]. In [44], impulse-based 3D
imaging was demonstrated by using a 4TX/4RX radar with electronic scanning, but two large high-gain (40 dBi) reflector dishes were used to improve image resolution and range. Off-chip antennas were used to radiate and receive impulses. Therefore, the imaging system in [44] is bulky and lacks potential in further miniaturization. In [45], 2D localization measurement of a metal plate was performed by using a coherent UWB MIMO radar operating at 26 GHz. No imaging was reported in [45]. Furthermore, the size of the MIMO radar was large due to the PCB-based transmitter and receiver antennas.

This chapter presents a silicon-based impulse radiator with on-chip antennas, which eliminate the power loss and phase distortion caused by the connection to off-chip antennas. The designed chip converts an input digital trigger signal to radiated impulses with accurate timing, which is required for coherent impulse combining. In contrast to previous designs [43, 44, 45], the employment of the asymmetric cross-coupled pulsed VCO, without PLL or DLL, ensures accurate timing control of the produced short impulses, benefiting from a significant reduction in power consumption, system complexity, and die area of the core circuitry. It is worthwhile to note that, although a similar asymmetric VCO topology has been reported in [46], in this work, the designed VCO works at a higher frequency and the silicon chip produces much shorter impulses. More importantly, this work investigates the sources of timing jitter in the produced impulses and discusses the optimization of the size ratio between the two transistors in the cross-coupled pair. The reported chip is designed to radiate impulses with a FWHM pulse-width of 74 ps, which matches closely with the measured results. The whole system, including the on-chip antenna, has a 3 dB bandwidth of 11 GHz and a 6 dB bandwidth of 23 GHz. In the measurement, by optimizing the bias voltages, a shortest FWHM pulse-width of 60 ps is recorded. In addition, coherent impulse combining over the air is performed by using two widely
Impulse and ultra-wideband (UWB) technology have been developed for radar applications over the past decade [47]. In this work, the designed impulse chip is used to implement an impulse-based synthetic aperture array for 3D imaging, which does not require any lens or reflector dishes. Digital beamforming is applied in time domain based on time-of-flight information of echo signals. High resolution 3D images have been produced for both metallic and dielectric objects. This achievement not only proves the capability of using silicon technology for short range lens-free high resolution 3D imaging, but also demonstrates the potential of building a highly-integrated silicon-based 2D imaging array for real-time 3D imaging.

The reminder of this chapter is organized as follows. Section 3.2 shows the details of each individual block of the chip, including the 30 GHz asymmetric pulsed VCO, the 30 GHz power amplifier, and the on-chip bowtie antennas. System performance of the chip is also investigated. Section 3.3 presents chip characterization results, which is followed by demonstration of coherent impulse combining in Section 3.4. Section 3.5 reports 3D imaging experiments performed by the designed impulse radiator chip. This chapter is concluded in Section 3.6.

### 3.2 30GHz Impulse Radiator

The 30 GHz impulse radiator designed in this work adopts a pulsed-VCO-based system architecture, as shown in Fig. 3.1. Compared with carrier-based architectures [43, 48], pulsed-VCO-based architectures [45, 46, 49] have some intrinsic advantages: First, pulsed VCO is switched on and off by a trigger signal to generate short impulses. Therefore, imaging sensitivity is enhanced due to zero leakage to RX from pulsed VCO. Second, pulsed VCO can be designed to ensure coherency of generated impulses, eliminating the need for co-locked PLL and DLL.
Figure 3.1: System architecture of the 30 GHz impulse radiator (pulsed-VCO-based architecture).

The designed 30 GHz impulse radiator consists of an edge-sharpening circuitry, a pulsed VCO with asymmetric cross-coupled topology, a power amplifier, and on-chip bowtie antennas (Fig. 3.1). The input trigger signal is a digital square wave, which is fed to the edge-sharpening block. This block is a two-stage digital inverter chain that switches the current source of the pulsed VCO. The input impedance of the edge-sharpening circuitry is designed to be 50 Ω to minimize the reflection of the input trigger signal. The edge-sharpening circuitry consumes 8.4 mW.

### 3.2.1 30 GHz Pulsed VCO

In order to radiate short impulses with large amplitude, the startup time of the pulsed VCO needs to be short [49]. Additionally, generating coherent impulses for efficient beamforming requires that the phase of the generated impulses must be locked with that of the input trigger signal. Therefore, two more requirements need to be achieved: first, initial phase of the impulse signal must be deterministic; second, timing jitter
added by the pulsed VCO must be minimized. In this work, timing jitter is defined as standard deviations of $\Delta \phi_i$ shown in Fig. 3.2,

$$\text{Timing Jitter} = \sqrt{\frac{1}{N} \sum_{i=1}^{N} (\Delta \phi_i - \mu)^2}$$

(3.1)

where $\Delta \phi_i$ is the timing difference in one period between time $t_{ia}$ when the generated impulse is at its minimum amplitude and time $t_{ib}$ when the trigger signal is at 50% of its amplitude on its falling edge, and $\mu = \frac{1}{N} \sum_{i=1}^{N} \Delta \phi_i$.

Figure 3.2: Timing relationship between the generated impulses and the input trigger signal.

Schematic of the pulsed VCO is shown in Fig. 3.3. An asymmetric cross-coupled topology is introduced to minimize timing jitter. The preceding edge-sharpening circuitry drives the current source $T_1$. Emitter length ratio between transistors $T_2$ and $T_3$ is optimized to be 1:3 to minimize timing jitter. Two identical capacitors, $C_1$ and $C_2$, boost the oscillation amplitude and hence phase noise is reduced. Two varactors are used to adjust oscillation frequency by applying tuning voltage at the common anode. An impedance matching network, consisting of the differential transmission line $TL_1$, capacitors $C_3$ and $C_4$, provides the required inductance for resonance from 30 GHz to 35 GHz, and also enables maximum power delivery from the pulsed VCO.
Figure 3.3: Schematic of the 30 GHz asymmetric cross-coupled pulsed VCO.

to the following power amplifier. Bypass capacitors $C_b$ filter out the noise from bias voltages $V_{cc}$, $V_{bias}$, and $V_{tune}$. De-Q blocks, consisting of capacitors $C_d$ and resistors $R_d$ in series, eliminate undesired low-frequency common-mode oscillations caused by wirebonds.

As discussed earlier, startup time of the pulsed VCO needs to be short. An approximate startup transient response for differential output of oscillators can be expressed as [50]

$$v(t) = \frac{2v_0}{\sqrt{1 + \left(\frac{2v_0}{v(0)}\right)^2 - 1}} \cos(\omega_0 t - \phi) e^{-\epsilon \omega_0 t}$$  \hspace{1cm} (3.2)

where $v(0)$ is the initial condition, $2v_0$ is the steady-state oscillation amplitude and $\epsilon$
is a damping factor. From Eq. 3.2, the startup time $t_s$ for the oscillation $v(t)$ to reach 90% of its steady-state oscillation amplitude $2v_0$ is approximated as follows [49, 50],

$$t_s \approx \frac{C}{g_{m,eff}} \left[ 2 \ln \left( \frac{2v_0}{v(0)} - 1 \right) + 1.45 \right]$$

(3.3)

where $g_{m,eff}$ is the effective average transconductance of transistors $T_2$ and $T_3$ in the asymmetric cross-coupled pair and $C$ is the tank capacitance.

![LC Tank](a) ![Voltage Vs Time](b)

Figure 3.4: (a) Initial collector current asymmetry in the asymmetric cross-coupled pulsed VCO. (b) Comparison between simulated oscillations of asymmetric VCO and symmetric VCO.

According to Eq. 3.3, to obtain a shorter startup time $t_s$, a larger initial condition $v(0)$ is required. In conventional (symmetric) cross-coupled pulsed VCOs, the initial condition is introduced by thermal noise. However, in the proposed asymmetric cross-coupled pulsed VCO, unbalanced initial collector current flows are generated when the pulsed VCO is switched on. The unbalanced current flows set the initial condition that is much stronger than thermal noise, as illustrated in Fig. 3.4(a). Therefore, the asymmetric cross-coupled pulsed VCO reduces the startup time [Fig. 3.4(b)].
In addition to reducing startup time, initial phase of the generated impulse signal should be deterministic. As stated above, initial condition of the asymmetric cross-coupled pulsed VCO, \( v(0) \), is set by the unbalanced initial collector current flows, which are deterministic because they are controlled by a certain emitter length ratio of the asymmetric cross-coupled pair. As a result, the initial condition is deterministic, resulting in a deterministic initial phase of the generated impulse signal, which is the second design requirement of the pulsed VCO.

The final design requirement of the pulsed VCO is that the timing jitter added by the pulsed VCO must be minimized. There are two sources of timing jitter: noise-induced perturbation during the startup time and phase noise in the steady-state oscillation. In designing pulsed VCO, these two timing jitter sources are equally important. As stated in [51, 52], a symmetric cross-coupled pair improves phase noise in the steady-state oscillation by achieving high common-mode rejection ratio (CMRR) and high differential gain but it suffers from unsuppressed noise perturbation in the startup time, because the initial condition \( v(0) \) is set mostly by thermal noise. However, for the pulsed VCO with asymmetric cross-coupled pair, the initial condition \( v(0) \) is predominately given by the deterministic unbalanced collector current flows when the pulsed VCO is switched on. The deliberate asymmetry in the cross-coupled pair suppresses the noise-induced perturbation in the startup time but sacrifices the phase noise performance in the steady-state oscillation. Therefore, there exists an optimum emitter length ratio of the two transistors in the cross-coupled pair for minimizing the timing jitter added by the pulsed VCO.

The summation of emitter lengths of transistors \( T_2 \) and \( T_3 \) is set to be \( 20.8 \mu m \), which is limited by the power budget of the pulsed VCO. Timing jitter, as defined in Eq. 3.1, versus different emitter length ratios \( (T_3/T_2) \) is calculated from transient noise simulations. As shown in Fig. 3.5, timing jitter is minimized to be 29.2 fs.
Figure 3.5: Simulated timing jitter added by the pulsed VCO versus different emitter length ratios between transistors in the asymmetric cross-coupled pair.

at the emitter length ratio of 3:1, where the emitter length of $T_2$ is 5.2 $\mu$m and the emitter length of $T_3$ is 5.6 $\mu$m. Fig. 3.5 shows that the timing jitter increases when the emitter length ratio between $T_3$ and $T_2$ approaches to 1, because the symmetric cross-coupled pulsed VCO has phase ambiguity at the startup time.

As discussed earlier, asymmetric cross-coupled VCO has a systematic tiny unbalance in the differential-mode outputs. In addition, unbalanced common-mode noise induced by the asymmetric cross-coupled pair slightly degrades phase noise performance, which is undesired in continuous-wave operations. However, instead of considering phase noise alone, minimizing timing jitter induced by the pulsed VCO is the objective, which is different from designing CW-based VCOs. As a result, an asymmetric cross-coupled pair is designed with an optimum emitter length ratio of 3:1.
3.2.2 30 GHz Power Amplifier

In order to amplify impulses without distortions, the power amplifier is biased in the class A region and adopts a differential common-emitter topology, as shown in Fig. 3.6. 12 $0.12 \, \mu m^2$ HBT npn transistors, $T_1$ and $T_2$, are used as power transistors biased near their peak $f_T$ current density at a quiescent current of $1.7 \, mA/\mu m$.

![Schematic of the 30 GHz power amplifier](image)

Figure 3.6: Schematic of the 30 GHz power amplifier.

Fig. 3.7(a) presents the maximum power gain of the PA core versus frequency by load-pull simulations. In the impulse mode simulation, the input power at each frequency is determined by the power spectrum of the impulse signal delivered to the PA. As shown in Fig. 3.7(a), the maximum power gain decreases from 22.7 dB at 16 GHz to 13 dB at 45 GHz. Because the impulse has the center frequency of 30 GHz, the PA should have a power gain curve that is as flat as possible around 30 GHz. Therefore, the load impedance of the PA is designed to deliberately sacrifice the
power gain of the PA at lower frequencies and meanwhile to approach to the optimum
value at higher frequencies. Two series capacitors, $C_1$ and $C_2$, are placed between
the transmission lines, $TL_2$ and $TL_3$, in order to add one more degree of freedom to
achieve broadband matching. With the optimized matching network, the PA has a 3
dB power gain bandwidth of 29 GHz, ranging from 16 GHz to 45 GHz [Fig. 3.7(a)].
At the center frequency of 30 GHz, the PA core has a 15 dB simulated gain and
the matching network has a simulated insertion loss of 1.5 dB. Within the 3 dB gain
bandwidth, variation of saturated output power ($P_{sat}$) is within 3 dB. Fig. 3.7(b)
shows the simulated output power contours at 30 GHz, the optimum load impedance
($Z_{load,opt}$), and the actual load impedance ($Z_{load}$) provided by the matching network.
At 16 GHz, there is a big difference between $Z_{load}$ and its optimum value, $Z_{load,opt}$.
However, the difference is much smaller at 45 GHz. At 30 GHz, the PA core has a
simulated average output power ($P_{out,ave}$) of 8 dBm with a 33+j17.9 Ω differential
load impedance. To investigate the peak output power, which is related to the peak
EIRP (Equivalent Isotropically Radiated Power), the PA is also simulated with an
input of a 30 GHz sinusoidal signal, whose amplitude is same as the peak amplitude
of the impulse signal delivered to the PA. Fig. 3.7(c) demonstrates the simulation
results. At 30 GHz, the PA has a simulated peak output power ($P_{out,peak}$) of 14.5
dBm. Peak EIRP will be defined later in this paper.

De-Q blocks ($C_d$ and $R_d$) and parasitic resistance of metal interconnects ensure
that both common-mode and differential-mode K-factors are greater than 1 in a wide
frequency range from dc to 200 GHz.

3.2.3 On-Chip Bowtie Antennas

In designing on-chip mm-wave antennas, surface waves are a major concern, because
they degrade radiation efficiency [53]. A silicon lens [15, 48, 54] can be used to mitigate
the surface wave problem, by collecting surface waves and converting them to useful radiation. Unfortunately, attaching a silicon lens significantly increases the directivity of the on-chip antenna and limits the field-of-view. In this work, the impulse radiator chip is designed for short-range wide field-of-view 3D imaging. Therefore, adding a silicon lens is not a good choice for this work.

As suggested in [53], silicon substrate thickness can be optimized to increase radia-
tion efficiency. Because wirebonds may affect topside radiation, the on-chip antennas are designed to radiate through substrate (bottom) side. Additionally, due to the die area constraints, the dimension of the on-chip antennas is limited to 2 mm. Therefore, the on-chip antenna can be considered as a small antenna at 30 GHz. For packaging purpose, the chip is attached onto an undoped silicon slab, which provides negligible conductive power dissipation.

An electromagnetic simulator based on the Method of Moments, HyperLynx 3D, is used to simulate the on-chip antennas with the silicon slab. As shown in Fig. 3.8(b), without grinding the chip substrate, when the total substrate thickness, including the chip substrate and the additional silicon slab, is a half-wavelength ($\epsilon_{r,si} = 11.9$) at 30 GHz, approximately 1440 $\mu$m, the on-chip bowtie antennas have a peak radiation efficiency of 11% for bottom radiation. Fig. 3.8(c) and (d) demonstrate that, with a 1440 $\mu$m silicon substrate, the on-chip bowtie antennas have a 3 dB radiation efficiency bandwidth of 8.5 GHz and a 3 dB peak gain bandwidth of 13 GHz. The peak gain at 30 GHz is -4.3 dB.

In terms of future directions for designing better antennas, new materials have been proposed for high-efficiency antennas, such as carbon nanotubes (CNTs) and CNT bundles [55]. More efforts are needed to integrate CNTs with silicon processes.

![Figure 3.8](image-url)

**Figure 3.8:** (a) The designed on-chip bowtie antennas. (b) Simulated bottom-side radiation efficiency versus substrate thickness at 30 GHz. (c) Simulated input impedance and radiation efficiency of the antennas versus frequency. (d) Simulated peak gain of the backside radiation of the antennas.
### 3.2.4 System Performance

In this design, the input trigger signal is generated by an external arbitrary waveform generator (AWG). Fig. 3.9 shows the measured waveform of the trigger signal used in the measurement, which has a pulse shape with a FWHM pulse-width of 88 ps. By changing the dc offset of this trigger signal using an external bias-tee, the switching-on duration of the pulsed VCO is adjustable, resulting in tuning the FWHM pulse-width of the radiated impulse continuously.\(^{3.1}\) In the simulations (Fig. 3.9), the VCO is tuned to oscillate at 30 GHz. When the dc offset of the trigger signal is adjusted so that the pulsed VCO has an 83 ps switching-on duration in each period, the simulated radiated impulse signal has a FWHM pulse-width of 73 ps, a peak EIRP of 8.7 dBm, and an average EIRP of 2.2 dBm.\(^{3.2}\)

Fig. 3.9 shows the evolution of the generated impulse signal from the asymmetric pulsed VCO to free space in both time domain and frequency domain. At the differential output node of the asymmetric pulsed VCO, the generated impulse signal has a 70 ps FWHM pulse-width, with a 3 dB bandwidth of 8 GHz, from 27 GHz to 35 GHz, and a 20 dB bandwidth of 22.5 GHz, from 20.5 GHz to 43 GHz. Even though even harmonics are generated due to the asymmetric topology of the pulsed VCO, it is not a concern for this impulse radiator.\(^{3.3}\) In terms of common-mode operations, the increased common-mode output from the asymmetric pulsed VCO is suppressed by the following differential PA and the differential bowtie antennas.

---

\(^{3.1}\) The external distortion effects of PCB traces and wirebonds slightly increase the switching-on duration of the pulsed VCO. But these effects can be compensated by tuning down the dc offset of the input trigger signal.

\(^{3.2}\) Including matching loss, the PA delivers 6.5 dBm simulated average power and 13 dBm simulated peak power to the on-chip antennas at 30 GHz. Since the antennas have the simulated peak gain of -4.3 dB at 30 GHz, the simulated average EIRP and peak EIRP at 30 GHz are 2.2 dBm and 8.7 dBm, respectively.

\(^{3.3}\) The 2nd harmonic of the 3 dB band of the generated impulse signal is from 54 GHz to 70 GHz, which lies out of the 20 dB bandwidth of the impulse signal, resulting in negligible distortion effects on the generated impulse. In addition, the range from 54 GHz to 70 GHz lies out of the 3 dB power gain bandwidth of the designed power amplifier following the asymmetric VCO, as shown in Fig. 3.7(a).
Figure 3.9: Simulation details of the impulse signal from the asymmetric pulsed VCO to free space in both time domain and frequency domain.

In Fig. 3.9, the PA core and the following matching network reduce the 3 dB bandwidth of the impulse signal slightly by 0.5 GHz, due to the PA gain peaking at the center frequency of 30 GHz (Fig. 3.7). Because the on-chip antenna is a small antenna at 30 GHz limited by the die area, it has a smaller 3 dB gain bandwidth than that of the PA, resulting in some late-time ringing shown in the simulated radiation waveform. The simulated radiated impulse signal has a 73 ps FWHM pulse-width, with a 3 dB bandwidth of 6 GHz, from 26 GHz to 32 GHz, and a 20 dB bandwidth of 19 GHz, from 21 GHz to 40 GHz. Fig. 3.9 also presents the simulated system bandwidth, including the PA and the on-chip antennas. It has a 3 dB bandwidth of 11 GHz and a 6 dB bandwidth of 23 GHz. The system bandwidth is mainly limited by the relatively small on-chip bowtie antennas. A FDTD-based 3D EM simulator, CST Microwave Studios, is used to simulate the time-domain performance of the on-chip differential bowtie antennas.
3.3 Measurement Results

The 30 GHz impulse radiator was fabricated in a 130 nm SiGe BiCMOS process technology. A micrograph of the chip, which occupies 1.9×1.5 \( \text{mm}^2 \) die area, is shown in Fig. 3.10. The impulse radiator is mounted on a piece of undoped silicon slab that is glued to the backside of a Roger 4350B PCB, which has a cut-out to couple the radiation from the chip substrate to air. Gold wirebonds are used to connect the on-chip pads to PCB traces (Fig. 3.10).

![Figure 3.10: Chip micrograph and chip assembly.](image)

The impulse radiator is measured in both time domain and frequency domain. Fig. 3.11 illustrates the measurement setup. The trigger signal \( T \) is a 1 GHz digital square wave signal, which is generated by the AWG (Tektronix 7122C). The trigger signal travels through a bias-tee (Picosecond Pulse Labs) before being fed into the impulse radiator. Tuning the bias-tee changes the FWHM pulse-width of the radiated impulses. In time-domain characterization, a custom broadband PCB impulse antenna was designed and used as a receiving antenna [15, 41].

Fig. 3.12 presents the measured impulse waveform with a minimum FWHM pulse-width of 60 ps. It was achieved by tuning the oscillation frequency to 32 GHz,
Figure 3.11: Measurement setup for (a) time-domain and (b) frequency-domain characterizations.

sacrificing the gain of PA and the on-chip antenna. Its normalized power spectrum is demonstrated in Fig. 3.12(b). In contrast to the simulation results shown in Fig. 3.9, its center frequency is shifted to 32 GHz and its 20 dB bandwidth is increased to 20 GHz. Fig. 3.12(c) shows that the generated impulses have a measured RMS jitter of 178 fs when the measured RMS jitter of the input trigger signal is 150 fs. An averaging of 64 was used in the measurement for timing jitter to reduce the noise from the sampling oscilloscope (Agilent DCA-X 86110D). The measured RMS jitter
of the radiated impulses is limited by that of the input trigger signal. A more stable input trigger signal will result in a better measured RMS jitter.

![Figure 3.12](image)

Figure 3.12: (a) Measured impulse waveform with 60 ps FWHM pulse-width. (b) Normalized frequency spectrum of the radiated 60 ps impulse. (c) Measured timing jitter of the radiated impulse.

The impulse radiator was designed to provide a center frequency of 30 GHz. Fig. 3.13(a) shows the measured impulse waveform under the designed bias condition. It achieves a FHWM pulse-width of 74 ps. Its EIRP frequency spectrum has been characterized using a standard gain horn antenna (A-Infomw LB-180400-KF, 18-40 GHz, Gain = 15 dB) with a distance of 20 cm [Fig. 3.11(b)]. As shown in Fig. 3.13(c), the radiated 74 ps impulse has an average EIRP of 0.5 dBm at the center frequency of 30 GHz with a 3 dB bandwidth of 4.5 GHz, a 10 dB bandwidth of 10 GHz, and a 20 dB bandwidth of 18 GHz. Because the radiated impulse train has a duty-cycle of nearly 25%, the peak EIRP can be estimated to be 4 times larger (6 dB larger) than the measured average EIRP, equaling to 6.5 dBm. The difference between the measured EIRPs (0.5 dBm and 6.5 dBm) and the simulated values (2.2 dBm and 8.7 dBm) arises from the effects of the package PCB, variations of the additional silicon slab thickness, and process variations. In addition to EIRP characterization, antenna

---

3,4 The definition of peak EIRP in here is different from that in [56, 33]. This definition is more accurate because it is based on the standard frequency-domain characterization method using a standard gain horn antenna.
radiation patterns have been measured at 20 GHz, 30 GHz and 40 GHz, as shown in Fig. 3.13(d). The chip consumes an average dc power of 106 mW.

![Figure 3.13](image)

Figure 3.13: (a) Measured 74 ps radiated impulse using designed bias voltages. (b) Uncalibrated measured signal tone at 30 GHz. (c) Measured EIRP spectrum. (d) Measured radiation patterns at 20 GHz, 30 GHz and 40 GHz.

### 3.4 Coherent Impulse Combining over the Air

In order to perform synthetic array 3D radar imaging, it is essential to achieve highly efficient coherent impulse combing over the air, which has been reported in [56, 33] briefly. Details are provided here to analyze impulse combining accuracy and correlation factor between the individual radiated impulses for combining. These two factors are important, because the first factor interprets the combining error caused by timing jitter of impulses, and the latter factor represents the similarity between
the impulse waveforms, which limits the maximum combined amplitude due to destructive combining.

![Figure 3.14: Measurement setup of coherent impulse combining over the air.](image)

The measurement setup is shown in Fig. 3.14. Due to process variations and unbalanced signal traveling of the two input digital trigger signals ($T_1$ and $T_2$), timing shift between the generated impulse and the input trigger signal can be different between impulse radiators. To compensate this discrepancy in order to achieve highly efficient impulse combining, two digital trigger output channels, $T_1$ and $T_2$, are shifted with a resolution of 1ps. The receiving antenna is placed in the far field. In contrast to the work reported in [15, 41], in this measurement, no millimeter-wave lens has been used.

Since the radiated impulse has an FWHM pulse-width of 60 ps and the measured timing jitter is less than 0.2 ps, the effects of timing jitter on impulse combining should be very small, resulting in an almost ideal impulse combining with negligible errors if all the discrepancies are compensated. As expected, Fig. 3.15(b) shows that the measured combined waveform is almost identical to the algebraic summation of the two individual impulse waveforms, meaning that the impulse combining accuracy is close to 1. Noise deteriorated the accuracy of impulse combining a little. Fig. 3.15(a)
reports the measured impulse radiated from each chip. The correlation factor\textsuperscript{3.5} between them is 0.956, which shows a highly similarity between these two waveforms.

The measured RMS jitter of the combined waveform is 216 fs with an averaging of 64 (Fig. 3.15(c)). The increase in the measured RMS jitter compared with that of a single impulse radiator (178 fs) is caused by the additional noise contributed by the second impulse radiator as well as the second input digital trigger signal.

\[ r = \frac{\int v_1(t) v_2(t) dt}{\sqrt{\int [v_1(t)]^2 dt} \sqrt{\int [v_2(t)]^2 dt}} = 0.956, \]

where \( v_1(t) \) and \( v_2(t) \) are the measured waveforms, respectively.

Figure 3.15: Measurement results of impulse combining over the air. (a) Measured time-domain waveform of the radiated impulse from each radiator chip. (b) Measured time-domain waveform of the combined impulse. (c) Measured timing jitter of the combined impulse.
3.5 3D Radar Imaging with a Synthetic Array

A custom synthetic impulse radiator array is built to produce high-resolution 3D images by adopting an impulse beamforming technique. As stated in the introduction, coherent impulse combining is the solution to the trade-off between higher image resolution and larger image range. In this work, a single radiator is utilized to construct a synthetic aperture radar [57] and DSP-based beamforming is applied to generate 3D images. Since the custom imaging systems built in this work have no RF amplification in the RX path, averaging is needed to reduce the noise from the sampling oscilloscope (Agilent DCA-X 86110D). The data acquisition time at each TX location is about 1s. To increase image resolution, the impulse radiator is set to emit impulses with the minimum FWHM pulse-width of 60 ps. The repetition rate of the radiated impulse train is 300 MHz.

A four-step 3D imaging methodology [58] is proposed using a synthetic impulse radiator array based on the designed impulse radiator chip. After a complete array scan with background cancellation\(^{3.6}\), the direct-coupled signal between the impulse radiator chip and the receiving antenna is eliminated in the processed waveforms, which only contain the desired echo impulses from objects. Based on time-of-flight (ToF) calculation, DSP-based beamforming technique is applied to the processed impulse waveforms in order to reconstruct 3D images. Enlarging the effective aperture size of the synthetic array enhances lateral image resolution. In addition, because beamforming technique has the essence of coherent impulse combining, SNR of reconstructed 3D images can be improved by increasing the element number of synthetic array.

\(^{3.6}\) For each imaging setup, background waveforms, which only contain the direct-coupled signals, were measured and saved at every TX location without placing any objects. This calibration measurement was performed only once for each imaging setup. In the imaging experiments with objects, the background waveforms were subtracted from the received waveforms in the digital domain and therefore the direct coupling effects were eliminated.
3.5.1 3D Radar Imaging of Metallic Objects

A custom 99-element synthetic impulse radiator array is built to produce 3D images of metallic objects, as shown in Fig. 3.16. The impulse radiator is mounted on a 2D traveling stage (Thorlabs LTS-300). The 2D traveling stage moves the radiator to 99 locations to form a 99-element synthetic array. The spacing between the adjacent elements in the synthetic array is 5 mm, and therefore, the effective aperture size of the synthetic array is $40 \text{ mm} \times 40 \text{ mm}$. A 3D coordinate system is built as follows: The top left element in the synthetic array is located at $(0, 0, 10 \text{ cm})$, and the synthetic array is on the XY plane. The Z coordinate of the metallic objects is 25 cm, which means that the distance between the objects and the synthetic array is 15 cm. The custom PCB-based broadband receiving antenna is placed at a fixed location in the 3D space. The entire imaging system is automated and controlled by a laptop with GPIB protocols.

In this work, image resolutions are defined as

\[
\text{Depth Resolution} = \frac{\text{pulse\_width} \times c}{2} \quad (3.4)
\]

\[
\text{Lateral Resolution} = \frac{\lambda \times D}{R} \quad (3.5)
\]

where $c$ is the speed of light in the medium, $\lambda$ is the center wavelength of the impulse signal in the medium, $D$ is the depth range, and $R$ is the aperture size of synthetic array.

Three scenarios are designed to produce 3D radar images of various metallic objects. As shown in Fig. 3.17, in the first scenario, the object is a small aluminum cylinder wrapped on a glass rod; in the second scenario, two spaced small aluminum cylinders are used as objects; in the third scenario, the object is an aluminum ring.
Figure 3.16: Measurement results of impulse combining over the air. (a) Measured time-domain waveform of the radiated impulse from each radiator chip. (b) Measured time-domain waveform of the combined impulse. (c) Measured timing jitter of the combined impulse.

By adopting the proposed imaging methodology [58], 3D radar images of the three scenarios are generated with a depth resolution of 9 mm and a lateral resolution of 2.65 cm at a depth range of 15 cm in the air (Fig. 3.17). When the image plane is on the XY plane with the Z-coordinate of 20 cm, where there is no object, the generated 3D images are completely black in all three scenarios. These images show that the synthetic impulse radiator array detects no reflections from the image plane at Z = 20 cm. When the image plane is set as the XY plane at Z = 25 cm, which is the exact depth range of the objects, the generated 3D images in all three scenarios show clear reflections. Color scales are identical in both radar images at Z = 20 cm and Z = 25 cm for each scenario. In all three scenarios, sizes of the reflections shown in the produced radar images (Z = 25 cm) are close to those of real objects. Furthermore,
in the third scenario, the reflection variations in the generated image at $Z = 25 \text{ cm}$ also coincide with the surface roughness variations of the aluminum ring: The lower part of the aluminum ring is rougher than the upper part. Therefore, more EM waves are reflected to the receiving antenna by the lower part of the aluminum ring than the upper part, which matches with the generated radar image.

Figure 3.17: 3D radar images of the metallic objects.

It is necessary to note that, in the third scenario, the reconstructed object has an elliptical shape rather than a circular shape shown in the optical picture. This is mainly caused by the angle-dependent waveform distortions induced by both the impulse radiator chip and the receiving antenna, which reduce the accuracy of ToF calculation and the quality of image reconstruction. These angle-dependent distortion effects can be calibrated for further improvement.
3.5.2 3D Radar Imaging of Rock Samples Immersed in Oil

The measurement setup of this test is demonstrated in Fig. 3.18. The proposed impulse radiator is fixed at the original point (0, 0, 0). A plastic box containing rock samples is placed on a 2D traveling stage (Thorlabs LTS-300), which is on an XY plane. The plastic box is moved in an XY plane relative to the impulse radiator in order to form a synthetic array with an effective aperture size of 75 mm × 100 mm. An element-to-element spacing of 1 mm is chosen. The top surfaces of rock samples are on an XY plane with Z = 100 mm. The PCB receiving antenna is placed at a fixed position.

![Synthetic array imaging system for 3D radar imaging of rock samples.](image)

In this experiment, each rock cube has an approximate size of 25 mm × 25 mm × 25 mm. These rock cubes have different reflection coefficients because of their various porosity and lithology. One limestone rock cube is drilled and filled with distilled water. Another limestone rock cube, which contains no water, is placed next to it. The rock cubes and marble base are immersed in pure oil, whose relative permittivity is 2 (Fig. 3.19).

By applying the proposed imaging methodology, 3D radar images of rock samples
are generated with a depth resolution of 6.4 mm and a lateral resolution of 5.7 mm at a range of 100 mm in the oil, as shown in Fig. 3.19. In order to compare imaging results with real samples, the generated images are overlapped with the optical pictures of rock samples. When the image plane is on the XY plane at $Z = 92$ mm, where no rock exists, intensities in the generated radar image are close to 0, because no objects are detected on this image plane. By shifting the image plane to the XY plane at $Z = 100$ mm, where the top surfaces of the rock samples are, the produced radar image reveals information about the rock types. There is a significant reflection difference between the two limestone cubes in the radar image ($Z = 100$ mm). This is caused by the high dielectric constant of the distilled water hidden in the right limestone cube. Furthermore, it is confirmed that there is little reflection shown across the stand-off area in the generated radar image at $Z = 100$ mm. This is because the stand-off surface is deeper than the image plane at $Z = 100$ mm.

### 3.6 Conclusion

This chapter reports an integrated 30 GHz impulse radiator implemented in a 130 nm SiGe BiCMOS process technology. An asymmetric cross-coupled pulsed VCO topology is introduced to minimize the timing jitter of radiated impulses, which is crucial in achieving highly efficient coherent combining. With a 106 mW average dc power consumption, the impulse radiator emits impulses with a minimum FWHM pulse-width of 60 ps. The measured RMS jitter of the radiated impulses is 178 fs when the input trigger signal has a 150 fs RMS jitter. The chip achieves an average EIRP of 0.5 dBm and a peak EIRP of 6.5 dBm, when producing impulses with an FWHM pulse-width of 74 ps and a center frequency of 30 GHz. Coherent impulse combining over the air is performed by using two widely spaced impulse radiators. To the authors knowledge, this work, for the first time, presents a proof-of-concept of
Figure 3.19: 3D radar images of the rock samples.

using a synthetic array of a fully integrated impulse radiator in silicon to perform high resolution impulse-based lens-free 3D imaging. 3D radar images of metallic objects and dielectric objects have been successfully demonstrated. In this work, a depth resolution of 9 mm, a lateral resolution of 8 mm at a range of 10 cm in the air have been achieved. Table 3.1 shows the comparison of this work with the prior art.
Comparison with state-of-the-art microwave 3D imaging radiators

<table>
<thead>
<tr>
<th></th>
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<th>[15]</th>
<th>[42]</th>
<th>[43]</th>
<th>[44]</th>
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<td>680 fs</td>
<td>1.2 ps</td>
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</tr>
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<td><strong>Coherent Impulse Impulse Combining</strong></td>
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<td>No</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td><strong>3D Imaging</strong></td>
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<td>Performed c</td>
<td>Not Performed</td>
<td>Performed c</td>
</tr>
<tr>
<td><strong>Imaging Objects</strong></td>
<td>Metallic objects Oil-immersed rocks</td>
<td>N/A</td>
<td>Single metallic object &amp; Breast phantom</td>
<td>N/A</td>
<td>Single metallic object</td>
</tr>
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<td>9 mm</td>
<td>5.4 mm</td>
<td>0.1 cm e</td>
</tr>
<tr>
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<td>N/A</td>
<td>N/A</td>
<td>2.4 cm @ 1 m e</td>
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<td>7.56 mm² f</td>
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<td>TX+Antenna</td>
<td>TX+RX</td>
<td>TX+RX+Antenna</td>
<td>TX+RX</td>
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<td>130nm SiGe BiCMOS</td>
<td>65nm CMOS</td>
<td>130nm SiGe BiCMOS</td>
<td>65nm CMOS</td>
</tr>
</tbody>
</table>

a. The FWHM pulse-width is variable but the shortest measured value is 60 ps.
b. It was calculated based on the peak voltage of the transient impulse waveform. Its definition is different from this work.
c. 3D imaging were performed with external PCB TX/RX antennas and reflector dishes.
d. The values are compared when air is the medium.
e. Depth resolution is obtained by using a modulated time-of-flight algorithm.
f. No on-chip antennas are implemented.

Table 3.1: Comparison with state-of-the-art microwave 3D imaging radiators.
A Nonlinear Q-Switching Impedance Technique for Picosecond Pulse Radiation in Silicon

This chapter presents a nonlinear Q-switching impedance (NLQSI) technique for picosecond pulse radiation in silicon, published in [40]. A prototype chip is designed with four NLQSI-based impulse generation channels, which can produce picosecond pulses with a reconfigurable amplitude. An on-chip impulse-coupling scheme combines the outputs from four channels and delivers the combined signal to an on-chip antenna. In addition, an asynchronous optical-sampling measurement system is used to characterize the radiated picosecond pulses in the time domain. Chapter ??ill describe the details of this measurement technique. The prototype chip can radiate 4 ps pulses with an SNR>1 bandwidth of 161 GHz. Furthermore, pulse amplitude modulation is experimentally demonstrated. The prototype chip is fabricated in a 130 nm SiGe BiCMOS process technology with a die area of 1 mm². Compared with original D2I designs [15, 41], the implemented chip achieves the pulse amplitude reconfiguration and modulation, which are discussed in the Chapter 1.2.1 as the preferred improvements for D2I architectures.
4.1 Introduction

Traditionally, signal generation in the millimeter-wave (mm-Wave) and Terahertz (THz) regimes is performed using continuous-wave (CW) or pulse techniques [59, 7]. As shown in Fig. 4.1, CW sources produce long pulses in time with small bandwidth, while pulse sources generate short pulses in time with large bandwidth. Recently, mm-Wave and THz signal have been used in applications, such as high-resolution 3D imaging [60, 4], biomedical sensing [61], high-speed wireless communication links [62], and broadband spectroscopy [63].

Figure 4.1: Comparisons between continuous-wave (CW) and picosecond/sub-picosecond pulse.

Over the past few decades, photonic techniques have been used for signal generation in mm-Wave and THz regimes. These techniques include fsec-laser gated photoconductive antennas, photo-mixing, and quantum cascade lasers (QCL) [64, 31, 17]. In photonic-based solutions, laser sources are required, which makes the whole system bulky and expensive. Recently, fully-electronic laser-free sources have been reported that produce CW signals in the mm-Wave and THz regimes [65, 66, 43, 48, 67, 68, 69, 70]. These sources are based on high-speed transistors that can achieve $f_t$ (current
gain cut-off frequency) of above 300 GHz and $f_{\text{max}}$ (maximum oscillation frequency) exceeding 400 GHz [65, 71]. The majority of the work published on mm-Wave and THz sources is based on CW techniques that are narrowband and contain only few frequency tones. Compared with these methods, far less research has been performed to produce broadband picosecond pulses in the mm-Wave and THz regimes. It is important to note that the picosecond pulses discussed in this paper should have a large frequency bandwidth (SNR>1 BW is greater than 100 GHz). Pulse-shaping techniques, based on mixing harmonics and nonlinear transmission lines, are beyond this paper’s scope [72, 73].

In 2014, a silicon-based digital-to-impulse (D2I) architecture was reported that radiated sub-10 ps impulses [15, 41]. This technique was based on switching a dc current flowing through a broadband on-chip antenna, which results in converting a stored magnetic energy to a radiated picosecond pulse. In D2I, a pulse-matching circuitry was used to reduce the duration of the radiated pulse. Since then, arrays of D2I radiators have been published in the pursuit of generating short pulses with a large radiated power based on spatial power combining [74, 75].

Apart from source technologies, accurate detection of picosecond pulses via electronic methods has been a major challenge for circuit and microwave communities. Over the past 20 years, researchers have been using electronic oscilloscopes to perform time-domain characterizations of picosecond pulses that are generated by silicon chips [76]. Unfortunately, the fastest off-the-shelf real-time oscilloscope has a bandwidth of 100 GHz and a rise time of 4.5 ps (Teledyne Lecroy LabMaster 10 Zi-A Oscilloscope), which is not good enough to characterize sub-10 ps pulses [25]. In addition, in these methods, broadband calibrations of antennas/probes, waveguides, coaxial connectors, and coaxial cables are required, which makes the measurement process complicated and time-consuming [76].
In this work, a novel circuit based on nonlinear Q-switching impedance (NLQSI), is proposed that produces picosecond pulses with a reconfigurable amplitude. A prototype chip is implemented that contains four NLQSI-based impulse generation channels and an on-chip impulse-coupling scheme, which combines the outputs of the four channels and delivers the combined pulses to an on-chip antenna. The chip is characterized with an asynchronous optical sampling system that provides a measurement bandwidth up to 4 THz. The time-domain measurements demonstrate that the prototype chip radiates 4 ps pulses with an SNR $\geq$1 bandwidth of 161 GHz. In addition, pulse amplitude modulation is experimentally demonstrated. The prototype chip is fabricated in a 130 nm SiGe BiCMOS process technology and occupies a die area of 1 mm$^2$.

The remainder of this chapter is organized as follows: NLQSI technique is discussed in Section 4.2, while Section 4.3 presents the design of the prototype NLQSI-based picosecond pulse radiator chip. The characterization results of the prototype chip are demonstrated in Section 4.4, including a brief introduction of the asynchronous optical sampling time-domain measurement system used in this work for picosecond pulses, as well as the measured NLQSI-induced tunable pulse amplitudes and measured pulse amplitude modulation. Conclusions are provided in Section 4.5.

4.2 Nonlinear Q-Switching Impedance (NLQSI) Technique

The idea of NLQSI originates from the step response of a parallel RLC tank. In order to understand its mechanism, in this section, the step response of a parallel RLC tank is briefly reviewed and then the method of NLQSI is introduced.
4.2.1 Step Response of a Parallel RLC Tank

A parallel RLC tank, as shown in Fig. 4.2, has two types of step responses: over-/critically-damped response and under-damped response. The response of the tank depends on its damping rate and resonant frequency, which are defined as follows:

\[
\alpha \triangleq \frac{1}{2RC} \quad \text{(4.1)}
\]

\[
\omega_0 \triangleq \frac{1}{\sqrt{LC}} \quad \text{(4.2)}
\]

where \( R, L, \) and \( C \) are the resistance, inductance, and capacitance of the tank. Equivalently, the step response of a parallel RLC tank can also be determined by investigating the quality factor of the tank \( Q_{\text{tank}} \), which represents the ratio of the stored energy to the energy dissipated in a circuit. \( Q_{\text{tank}} \) can be expressed as follow:

\[
Q_{\text{tank}} \triangleq \frac{2\pi \text{ maximum energy stored}}{\text{total energy lost per cycle at resonance}} = \frac{\omega_0}{2\alpha} = \frac{RC}{L} \quad \text{(4.3)}
\]

\[\text{A loss-less response is excluded from this discussion.}\]

Figure 4.2: Step response of a parallel RLC tank.
To simplify the analysis, the current source is considered to have an ideal step response.

4.2.1.1 Under-Damped Response \((\alpha^2 < \omega_0^2 \text{ or } Q_{tank} > 0.5)\)

When the damping rate \(\alpha\) is smaller than the resonant frequency \(\omega_0\), equivalently, \(Q_{tank}\) is larger than 0.5, the tank displays under-damped responses. In this case, the voltage across the tank, \(V(t)\), and the current flowing through the inductor \(L\), \(I_L(t)\), both have exponentially decaying oscillatory behaviors. Analytical expressions of \(V(t)\) and \(I_L(t)\) can be obtained by solving the following differential equation:

\[
LCI''_L + \frac{L}{R} I'_L + I_L = 0 \tag{4.4}
\]

with initial conditions of

\[
I_L(0) = I_0 \tag{4.5}
\]

\[
V(0) = LI'_L(0) = 0 \tag{4.6}
\]

where \(I_0\) is the steady-state current flowing through the tank before switching off.

The under-damped response of the parallel RLC tank can be expressed as:

\[
I_L(t) = I_0 e^{-\alpha t} \left( \frac{\omega_0}{\omega_d} \right) \cos \left( \omega_d t - \tan^{-1} \frac{\alpha}{\omega_d} \right) \tag{4.7}
\]

\[
V(t) = -LI_0 e^{-\alpha t} \left( \frac{\omega_0^2}{\omega_d} \right) \sin (\omega_d t) \tag{4.8}
\]

where \(\omega_d\) is the oscillation frequency, defined as

\[
\omega_d \triangleq \sqrt{\omega_0^2 - \alpha^2} \tag{4.9}
\]
Therefore, the oscillation period is \( T_d = 2\pi \omega_d = 2\pi \sqrt{\omega_0^2 - \alpha^2} \) From Eq. 4.8 and 4.9, we can make an observation that the oscillation frequency increases by reducing the damping rate, which extends the timing duration of the decay.

### 4.2.1.2 Over-/Critically-Damped Response \((\alpha^2 \geq \omega_0^2 \text{ or } Q_{\text{tank}} \leq 0.5)\)

When the damping rate \( \alpha \) is larger than or equal to the resonant frequency \( \omega_0 \), equivalently, \( Q_{\text{tank}} \) is smaller than or equals 0.5, the tank presents over- or critically-damped responses. The current flowing through the inductor \( L, I_L(t) \), decays to zero exponentially but does not produce a large voltage response, \( V(t) \), across the tank. In this case, there is no oscillatory behavior in either the current or voltage responses.

### 4.2.2 Switching from Under-Damped to Over-Damped to Produce Short Pulses

To produce a large voltage response in a short time with minimal ringing, a parallel RLC tank is designed such that it can switch from an under-damped response to an over-damped response by switching the load resistance of the parallel tank, equivalently, switching the quality factor of the tank \( Q_{\text{tank}} \). In this case, when the falling edge of the excitation current is arrived, the tank behaves in an under-damped response, producing a half-cycle oscillation, which is the desired oscillation shown in Fig. 4.3. The duration of the first half-cycle is \( \Delta T = \frac{1}{2} T_d = \pi / \sqrt{\omega_0^2 - \alpha^2} \), which can be as short as several picoseconds, according to simulations. The ringing, Undesired Oscillations, also shown in Fig. 4.3, needs to be eliminated by forcing the tank to switch to and stay in an over-damped response by reducing the load resistance \( R \) such that \( \alpha^2 > \omega_0^2 \). In this switching event, \( Q_{\text{tank}} \) is decreased from \( Q_U \) that sustains an under-damped response to \( Q_O \) that supports an over-damped response. Therefore, there exists a Q-switching event around the transition time \( t_0 \), which is shown in Fig.
4.3.

4.2.3 NLQSI Block

4.2.3.1 Overview

Fig. 4.4(a) shows the reported NLQSI block for picosecond pulse generations. A bipolar transistor $Q_1$ acts as a current source. When the base node of the transistor $Q_1$ is applied to 0 V by a voltage falling edge, the parallel tank starts to behave the step responses. A bipolar transistor, $Q_2$, monitors the output voltage $V_{out}(t)$ and autonomously changes the quality factor of the parallel RLC tank ($Q_{tank}$). Parasitic capacitance and resistance of transistors $Q_1$ and $Q_2$ play an important role in the performance of this NLQSI block. During the step responses, transistors $Q_1$ and $Q_2$ can be simplified by a parallel combination of $R_{Q1}$ and $C_{Q1}$ and that of $R_{Q2}$ and $C_{Q2}$, respectively, as shown in Fig. 4.4(a). The proposed NLQSI block is nonlinear because
these parasitic elements, especially $R_{Q2}$ and $C_{Q2}$, vary significantly depending on the output voltage $V_{out}(t)$. Fig. 4.4(b) demonstrates the nonlinearity of these parasitic elements when $V_{out}$ is swept from 0.5 V to 3 V, $V_{cc}$ is 1.3 V, and $V_{bias}$ is 2.5 V. The values of $V_{cc}$ and $V_{bias}$ are chosen for the proper operation of NLQSI blocks, which will be discussed later. During the step responses, transistor $Q_1$ is OFF because its base node is set to 0 V. Therefore, transistor $Q_1$ has a little nonlinearity with $V_{out}$. However, transistor $Q_2$ turns OFF when $V_{out}$ is larger than $V_{bias}V_{BE(on),Q2}$, which leads to the fact that transistor $Q_2$ dominantly contributes to the nonlinearity of the NLQSI block and, therefore, transistor $Q_2$ is crucial in shaping the generated ultra-short pulses.

Figure 4.4: (a) Nonlinear Q-switching impedance (NLQSI) block and (b) nonlinearity of transistors $Q_1$ and $Q_2$ versus $V_{out}$ when $V_{cc}$ is 1.3 V and $V_{bias}$ is 2.5 V.
4.2.3.2 Q-Switching Mechanism

The Q-switching mechanism of the NLQSI block is as follows (Fig. 4.5(a)): In the steady state, a dc current flows through the inductor $L$ and $V_{out}$ equals to $V_{cc}$. The tank is designed to have an over-damped response by biasing the transistor $Q_2$ to be ON. Stage I: when the current falling edge arrives, the tank behaves in an over-damped response, increasing $V_{out}$, and consequently, transistor $Q_2$ is turned OFF.
and $Q_{tank}$ is increased so that it is larger than 0.5. Therefore, the first Q-switching event is triggered at the transition time $t_1$, and the tank behaves in an under-damped regime ($Q_{tank} > 0.5$). Stage II: in the under-damped response, the desired oscillation is generated until $V_{out}$ drops below a certain value, which is when transistor $Q_2$ turns ON. The small ON resistance of transistor $Q_2$ ($R_{Q2}$) reduces $Q_{tank}$. $Q_{tank}$ should be decreased enough to force the tank to switch to an over-damped response ($Q_{tank} < 0.5$). Therefore, the second Q-switching event happens at the transition time $t_2$. Stage III: the tank stays in the over-damped response, and $V_{out}$ decays exponentially to $V_{cc}$, resulting in eliminating the undesired ringing. Eventually, the tank returns back to the steady state, where $V_{out}$ stays at $V_{cc}$. The time-domain evolution of $Q_{tank}(t)$ is illustrated in Fig. 4.5(b). It will be shown later that the shape of the generated pulse is dependent on $Q_{tank}(t)$, which is affected by the bias voltage of transistor $Q_2$ ($V_{bias}$), emitter lengths of both transistors $Q_1$ and $Q_2$, and the inductance of $L$.

### 4.2.3.3 Effects of $V_{bias}$ of Transistor $Q_2$

$V_{bias}$ of transistor $Q_2$ has significant effects on the performance of the NLQSI block. Fig. 4.6 shows the simulation results when $V_{bias}$ is swept from 1.5 V to 2.5 V. In this simulation, the emitter lengths of transistors $Q_1$ and $Q_2$ are 15 $\mu$m and 2.5 $\mu$m, respectively. $V_{BE(on)}$ of transistor $Q_2$, $V_{BE(on),Q2}$, is around 0.9 V. $V_{cc}$ is 1.3 V.

As shown in Fig. 4.6(a), when $V_{bias}$ remains below 2.1 V, which is smaller than the sum of $V_{cc}$ and $V_{BE(on),Q2}$, undesired oscillations do not disappear after the transition time $t_2$. This is because transistor $Q_2$ turns OFF again and the tank stays in the under-damped condition, which is validated by studying the simulated time-domain evolution of $Q_{tank}(t)$, as shown in Fig. 4.6(c, 1-3). For the situations where $V_{bias}$ is smaller than 2.1 V, $Q_{tank}$ is always larger than 0.5, meaning that the tank stays in an under-damped response with time-varying damping rate. At around the transition
Figure 4.6: Simulated effects of $V_{\text{bias}}$ on the performance of NLQSI block. (a) Incorrect Q-switching mechanism. (b) Designed Q-switching mechanism. (c) (1-7) Simulated $Q_{\text{tank}}(t)$ in each $V_{\text{bias}}$ condition and (8) comparison between the incorrect Q-switching mechanism and the designed Q-switching mechanism.
time $t_2$, $Q_{tank}$ is reduced along with the decreasing $V_{out}$, but afterwards it returns back to a high level that sustains the ringing (undesired oscillations). Another observation from the simulation is that the amplitudes of the ringing decrease with increasing $V_{bias}$. This is due to the fact that, with increasing $V_{bias}$, the tank has a smaller $Q_{tank}$, equivalently a larger damping rate $\alpha$, for a longer period of time, as shown in Fig. 4.6(c, 1-3). The amplitude of the ringing depends on the minimum value that $Q_{tank}$ can reach after the transition time $t_2$. In all these cases, transistor $Q_2$ adds a very large OFF resistance to the RLC tank, and consequently, the tank has almost identical $Q_{tank}$ before the transition time $t_2$, which explains why the peak amplitudes remain almost constant. In summary, in situations where $V_{bias}$ is smaller than the sum of $V_{cc}$ and $V_{BE(on),Q2}$, the under-damped response dominates the entire transient response.

When $V_{bias}$ is larger than the sum of $V_{cc}$ and $V_{BE(on),Q2}$, for example, $V_{bias}$ is in the range from 2.3 V to 2.5 V, transistor $Q_2$ turns ON as long as $V_{out}$ is smaller than or equals to $V_{cc}$. Therefore, transistor $Q_2$ keeps turning ON after the transition time $t_2$ and forces the tank to remain in the over-damped condition, eliminating undesired ringing, as shown in Fig. 4.6(b). The simulated $Q_{tank}(t)$ also reflects that the designed $Q$-switching mechanism is achieved when $V_{bias}$ is larger than $V_{cc} + V_{BE(on),Q2}$ (Fig. 4.6(b) and (c, 5-7)). In these situations, transistor $Q_2$ turns ON completely in the steady state and $Q_{tank}$ is smaller than 0.5. When the current falling edge arrives, the tank is in the over-damped response (Stage I), where $V_{out}$ increases and turns OFF $Q_2$, boosting $Q_{tank}$ to be larger than 0.5. The tank enters the stage II and behaves an under-damped response until around the transition time $t_2$, where $V_{out}$ is smaller than $V_{bias}V_{BE(on),Q2}$ and transistor $Q_2$ is turned ON again. At this moment, $Q_{tank}$ becomes smaller than 0.5, and the tank performs an over-damped response, switching from the stage II to stage III. Unlike the previously discussed cases ($V_{bias}$
< 2.1 V), after the transition time $t_2$, $Q_{tank}$ is always smaller than 0.5 and, therefore, the tank stays in the over-damped response (stage III), completely eliminating the ringing. Therefore, a clean pulse is generated, compared with the previously discussed cases. With increasing $V_{bias}$, the first Q-switching event happens later and the tank stays in the over-damped response (stage I) for a longer period of time, causing more energy loss and therefore, reducing the peak amplitude of the generated pulse. The comparison between the $Q_{tank}(t)$ in these two cases ($V_{bias} = 1.5$ V and 2.5 V) is plotted in the Fig. 4.6(c, 8).

When $V_{bias}$ is 2.1 V, as shown in Fig. 4.6(c, 4), the simulated $Q_{tank}(t)$ clearly reflects that this bias condition is in the transition to the proper Q-switching mechanism. In summary, when $V_{bias}$ is larger than the sum of $V_{cc}$ and $V_{BE(on),Q2}$, transistor $Q_2$ is completely ON in the steady state, and the NLQSI block has a desired Q-switching mechanism to generate a clean picosecond pulse with tunable peak amplitude.

4.2.3.4 Choose Emitter Lengths of Transistors $Q_1$ and $Q_2$

The emitter lengths of transistors $Q_1$ and $Q_2$ have crucial effects on the performance of the NLQSI block. Under the proper bias condition ($V_{bias} = 2.3$ V), if the emitter length of transistor $Q_1$, $l_{e,Q1}$, is increased, the steady-state current flowing through the RLC tank is increased as well. Fig. 4.7(a) shows the simulation results where $l_{e,Q1}$ is swept from 2µm to 15µm. A step voltage is applied to the base of transistor $Q_1$ to mimic an ideal pre-driver stage. As expected, a larger transistor $Q_1$ generates a pulse with a greater peak voltage but the duration of the pulse becomes larger as well. In this simulation, by choosing a length of 15µm for transistor $Q_1$, the full-width-at-half-maximum (FWHM) of the generated pulse becomes 2.7 ps. Fig. 4.7(b) demonstrates the simulated $Q_{tank}(t)$ with different $l_{e,Q1}$. Except for the case where $l_{e,Q1}$ is 2µm, the NLQSI block has the proper Q-switching mechanism. With a larger transistor
$Q_1$, the tank is in the under-damped response (Stage II) for a longer period of time, causing the duration of the pulse to be larger. In addition, when $l_{e,Q_1}$ is increased, the tank switches from the stage I (over-damped response) to the stage II (under-damped response) earlier, causing less energy loss, and meanwhile, stronger energy ($E = \frac{1}{2}LI^2$) is initially stored in the tank. As a result, the peak amplitude of the generated pulse is increased with $l_{e,Q_1}$. In practice, a larger transistor $Q_1$ slows down the switching speed due its large base capacitance, but this issue can be mitigated by designing a proper pre-driver stage with strong current-discharging capability. In this work, transistor $Q_1$ is designed to have an emitter length of 15 $\mu$m, which is chosen as a compromise between the pulse width (FWHM) and the pulse amplitude.

Figure 4.7: Simulated effects of the emitter length of transistor $Q_1$ on the performance of NLQSI block. (a) Simulated time-domain waveform of the generated pulse. (b) Simulated $Q_{tank}(t)$.

For transistor $Q_2$, when its emitter length, $l_{e,Q_2}$, is increased, transistor $Q_2$ adds more parasitic capacitance and less parasitic resistance to the tank. Fig. 4.8 shows the simulated effects of $l_{e,Q_2}$ on the performance of NLQSI block. With a larger transistor
Q₂, the NLQSI block generates a wider pulse with a smaller peak amplitude, as shown in Fig. 4.8(a). Fig. 4.8(b) demonstrates that, with \( V_{\text{bias}} = 2.3 \) V, the NLQSI blocks with \( l_{e,Q2} \) ranging from 2.5 \( \mu \)m to 15 \( \mu \)m all have the designed Q-switching mechanism. Furthermore, a larger transistor \( Q_2 \) produces a larger \( Q_{\text{tank}} \) in the stage I (over-damped response) and a smaller \( Q_{\text{tank}} \) in the stage II (under-damped response). The relationship between the pulse peak amplitude and \( l_{e,Q2} \) can be explained as follows: first, Fig. 4.8(c) shows that a larger transistor \( Q_2 \) causes more energy loss during the pulse generation period (stage I and II). Second, with a larger transistor \( Q_2 \) turning
ON in the steady state, less dc current flows through the inductor $L$, and consequently, less initial energy ($E = \frac{1}{2}LI^2$) is stored in the tank. As a result, these two effects cause that the peak and dip amplitudes of the generated pulse are decreased with a larger transistor $Q_2$. Fig. 4.8(d) demonstrates that the simulated oscillation frequency $f_d = \frac{1}{2\pi} \sqrt{\omega_0^2 - \alpha^2}$ in the under-damped response (stage II) is decreased with $l_{e,Q2}$, which explains the observation in Fig. 4.8(a) that the pulse width of the generated pulse is increased with $l_{e,Q2}$. In this work, transistor $Q_2$ is designed to have an emitter length of 2.5µm to enable Q-switching without degrading the pulse amplitude.

### 4.2.3.5 Inductance $L$

According to Eq. 4.8 and 4.9, the peak amplitude and the pulse width of the generated pulse are dependent on the inductance of $L$. Fig. 4.9 demonstrates the simulated effects of sweeping $L$ from 30 pH to 60 pH with the proper bias condition of $V_{bias} = 2.3$ V. As expected, with the increasing $L$, the generated pulse has a larger peak amplitude but also a greater pulse width [Fig. 4.9(a)]. The effects of the inductor $L$ can also be investigated in the viewpoint of $Q_{tank}(t)$. With the proper bias condition, the NLQSI blocks have the desired Q-switching mechanism. According to Eq. 4.1 and 4.3, a larger inductor $L$ reduces $Q_{tank}$ without changing the damping rate $\alpha$. Fig. 4.9(c) and (d) validate this point and show that a larger inductor $L$ tends to provide the tank with a longer under-damped response (stage II), which produces less energy loss in the pulse generation period. In addition, a larger inductor $L$ stores more initial energy ($E = \frac{1}{2}LI^2$) in the steady state, and consequently, the peak and dip amplitudes of the generated pulse are increased with a larger inductor $L$. Eq. 4.2 also indicates that, with a larger inductor $L$, the resonant frequency $\omega_0$ of the tank is reduced. Because the damping rate $\alpha$ is independent on the inductance of $L$, according to Eq. 4.9, the oscillation frequency in the under-damped response is
decreased with $L$, resulting in the increased pulse width.

Figure 4.9: Simulated effects of the inductance $L$ on the performance of NLQSI block. (a) Simulated time-domain waveform of the generated pulse. (b) Simulated $\text{SNR} > 1$ bandwidth of the generated pulse. (c) Simulated $Q_{\text{tank}}(t)$. (d) Simulated damping rate $\alpha(t)$.

The design strategy here is to maximize the $\text{SNR} > 1$ bandwidth, because a broad detectable bandwidth is crucial for spectroscopy and imaging applications. As shown in Fig. 4.9(b), the NLQSI block with a smaller inductor $L$ produces picosecond pulses with a larger $\text{SNR} > 1$ bandwidth. However, the actual measured $\text{SNR} > 1$ bandwidth is also dependent on the sensitivity of the detector. If the peak amplitude of the generated pulse is very weak, the measured $\text{SNR} > 1$ bandwidth will be smaller than the theoretical (simulated) value. Furthermore, practically, an inductor smaller
than 35 pH is difficult to be implemented in the process used in this work because the parasitic effects of the neighboring routings are significant. Therefore, based on the above considerations, in this work, the inductance of $L$ is set to be 40 pH.

## 4.3 Circuit and Antenna Details

The NLQSI technique reported in this chapter is implemented in a 130 nm SiGe BiCMOS process technology. In this section, first, an overview of the system architecture is presented. Second, an impulse-coupling scheme is introduced to resolve the issues caused by the transmission line effects in the NLQSI circuit. Third, the design details of the individual impulse generation channel and the four-way impulse combiner are presented. In addition, bias routings in the NLQSI blocks are specially designed. Finally, the design of the broadband on-chip antenna is discussed.

### 4.3.1 System Architecture

Fig. 4.10 shows the architecture of the radiating chip, which consists of four impulse generation channels. A trigger signal is fed to the chip and distributed to these channels through an H-tree distribution network. Each channel converts the input trigger to a picosecond pulse train with a repetition rate as same as the trigger. An impulse-coupling scheme is introduced to resolve the issues induced by the transmission line effects in implementing NLQSI blocks. Based on this scheme, a four-way microstrip-based impulse combiner is designed to combine the generated picosecond pulses from the four channels and feed the on-chip antenna. A good passive structure design can significantly reduce power loss, increasing power-combining efficiency [77]. The four-way impulse combiner also enables pulse amplitude modulation by controlling the individual impulse generation channels. A highly resistive silicon lens is attached
Figure 4.10: System architecture of the radiating chip.

to the backside of the radiator chip in order to increase the radiation efficiency and directivity of the on-chip antenna.

4.3.2 Impulse-Coupling Scheme

4.3.2.1 Transmission Line Effects in Implementing NLQSI Blocks

The chip uses on-chip metal routings, wirebonds and PCB traces for biasing and power supply, as shown in Fig. 4.11. The transmission line effects of these wirings need to be carefully considered in the transient analysis. For stability purposes, De-Qing blocks, consisting of a capacitor and a resistor in series, are placed along the on-chip metal routings. Since the supply current varies in a short period of time, the
inductance of the metal routings adds a strong low-frequency ringing to the picosecond pulse at $V_{\text{out}}(t)$ (Fig. 4.11). Even though the on-chip antenna is optimized for high-frequency radiation, the strong low-frequency ringing will be still radiated, which may cause interference with other low-frequency channels. Therefore, it is important to eliminate the low-frequency ringing before feeding the impulse to the antenna.

Figure 4.11: NLQSI block with transmission line effects.

4.3.2.2 Impulse-Coupling Block

In this work, an impulse-coupling block is introduced to isolate the input of the on-chip antenna from the strong low-frequency ringing caused by the transmission line effects of the supply wirings. Fig. 4.12(a) shows the NLQSI circuit with the impulse-coupling block, which is based on microstrip inductive coupling, consisting of two adjacent 50 µm long metal lines fabricated on the top metal layer AM, as shown in Fig. 4.12(b). Inductor $L_1$ is implemented by a metal line with a width of 4 µm, while inductor $L_2$ is implemented by another line with a width of 5 µm. The metal layer $M_3$ is used as a ground plane. In order to increase the mutual coupling between the
two metal lines, the spacing between them is set to 2.8 µm, which is the minimum value allowed by the DRC rules for the process technology.

Figure 4.12: NLQSI circuit with impulse-coupling block.

A simplified equivalent circuit of the designed inductive coupler is shown in Fig. 4.12(c). The inductive mutual coupling dominates the coupling mechanism, as the parasitic capacitors between $L_1$ and $L_2$ are very small (less than 2 fF). The parasitic capacitors associated with the ground plane are omitted here. EM simulations show that the designed inductive coupler has a broadband performance. As shown in Fig. 4.12(c), $L_1$, $L_2$, and the inductive mutual coupling coefficient, $k$, are flat from 10 GHz to 200 GHz. At 100 GHz, $L_1$ is 40 pH, $L_2$ is 37 pH, and $k$ is 0.45.\textsuperscript{4,2}

To perform impulse coupling in the NLQSI circuit, the inductive coupler is set to the following configuration: the collector node of transistor $Q_1$ is connected to Port 1, Port 2 connects with the $V_{cc}$ of 1.3 V, the connection of Port 3 is changeable,

\textsuperscript{4,2} The coupling coefficient $k$ is optimized to be broadband, which sacrifices the coupling strength as a cost.
which will be discussed later, and Port 4 connects with the load. The inductive mutual coupling and the small parasitic capacitors between $L_1$ and $L_2$ filter out the low-frequency ringing in the forward-coupling direction, which is from Port 1 to Port 4. Fig. 4.12(d) shows the simulated output voltage of the NLQSI block with the introduced impulse-coupling block when Port 3 is grounded. At the antenna input node, the low-frequency ringing is filtered out and a clean impulse-like waveform is fed to the on-chip antenna. Compared with the generated impulse without the impulse-coupling scheme, as shown in Fig. 4.11, the peak amplitude of the coupled pulse at Port 4 is reduced by around 50% as expected. In addition, there are very few distortion effects induced by this impulse coupler. The small downward pulses at 0 ns and 1 ns, shown in Fig. 4.12(d), are produced by the rising edges of the converted 1 GHz square wave, which switch ON transistor $Q_1$. These parasitic pulses can be eliminated by turning ON the transistor $Q_1$ using a waveform with a slow rising edge, which, however, may limit the repetition rate of the generated pulse train.

4.3.2.3 Impulse Generation Channel

The schematic of an impulse generation channel is shown in Fig. 4.13(a). The input trigger is fed into a digital inverter chain through a one-to-four H-Tree distribution network. The input impedance of the digital inverter chain is designed to be 200 $\omega$, in order to reduce the reflections. The digital inverter chain converts the input sinusoidal trigger to a square wave, which switches transistor $Q_1$ in the NLQSI block. In order to switch off transistor $Q_1$ quickly, an 18$\mu$m bipolar transistor, $Q_3$, is added to provide an additional discharging path from the base node of transistor $Q_1$. The maximum trigger frequency is 1 GHz, which is limited by the bandwidth of the digital inverter chain. $V_{digital}$ of each impulse generation channel needs to be optimized, which will be discussed later. Fig. 4.13(b) shows the simulated transient waveform applied to
the base node of transistor $Q_1$, which has a falling time of 13.6 ps.

Figure 4.13: (a) Schematic of an individual impulse generation channel. (b) Simulated transient voltage at the base node of transistor $Q_1$.

4.3.2.4 Four-Way Impulse Combiner

The four-way impulse combiner designed in this work consists of four impulse couplers stacked in series, as shown in Fig. 4.14(a). The spacing between the neighboring channels is identical. The second winding of the combiner connects with the on-chip antenna at one end and is grounded at the other. The collector nodes of transistor $Q_1$ in the four channels are connected to Ports 1 to 4, respectively. The routings in the NLQSI block in each channel are included in the EM simulations of the impulse combiner.

Fig. 4.14(b) illustrates the design methodology of the impulse combiner: the impulse combiner structure, with routings in the NLQSI blocks, is first simulated in an MoM-based EM simulator (Mentor Graphics HyperLynx Full-Wave Solver) in a wide frequency range from 1 GHz to 400 GHz. The extracted S-parameters are then imported into Cadence Virtuoso as an N-port S-parameter box, which is
connected with Assura-RC-extracted circuits of the impulse generation channels and an S-parameter box of the on-chip antenna. To investigate the performance of the designed impulse combiner, transient simulations are performed in Cadence Virtuoso to examine the combined pulse delivered to the on-chip antenna.

By switching on only one impulse generation channel at a time, the transient voltage at the antenna input is simulated and shown in Fig. 4.15(a). Compared with the simulation result shown in Fig. 4.12(d), the ringing effect in the coupled
Figure 4.15: Simulated coupled pulses from four channels with different biasing conditions.

The pulse is increased a little, which is mainly caused by the different port connection configuration from that in Fig. 4.12(a), equivalently, the different load impedance seen by the NLQSI block. The coupled pulses from Channels 1 and 2 are almost identical and their arrival timings to antennas input port are equal. However, the pulses from Channels 3 and 4 have smaller peak amplitudes and arrive in different timings at the antenna node. This is because the impulse combiner is not fully electromagnetically symmetrical among all four channels. Channels 1 and 2 see a similar load impedance from the impulse combiner. The value of this impedance varies slightly for Channels 3 and 4.\textsuperscript{4,3} Timing mismatch is more detrimental to the performance of the pulse combiner than the mismatch in the peak amplitudes because it will distort the pulse shape of the combined signal. In this design, the timing mismatch can be compensated by tuning the propagation delay of the trigger in each channel with varying the supply voltage, $V_{\text{digital}}$, of the digital inverter chain in each channel. The optimized $V_{\text{digital}}$ values are: $V_{\text{digital},\text{CH1}}$ is 1.2 V, $V_{\text{digital,CH2}}$ is 1.2 V, $V_{\text{digital,CH3}}$ is 1.21 V, and $V_{\text{digital,CH4}}$ is 1.19 V.

\textsuperscript{4,3} Even though channel 4 is closer to the antenna node than channel 3, with identical $V_{\text{digital}}$, the timing of the peak of the generated pulse from channel 4 can still be a little later than that of channel 3 because the generated pulses from these two channels have different pulse shapes.
V, and $V_{\text{digital,} CH4}$ is 1.19 V. Fig. 4.15(b) presents the aligned coupled pulses after adjusting the supply voltages of the digital inverter chains. The differences of their arrival timings are within 0.5 ps. With different ON-OFF combinations of the four channels, the combined pulse can have 16 peak amplitudes, which can be used for pulse amplitude modulation purposes.\textsuperscript{4.4} The simulated peak amplitudes in these 16 combinations are shown in Fig. 4.16.

\begin{figure}[h]
\centering
\includegraphics[width=0.6\textwidth]{fig4.16}
\caption{Simulated pulse peak amplitudes of impulse combiner outputs in all 16 combinations.}
\end{figure}

\textsuperscript{4.4} Pulse amplitude modulation is useful in high speed wireless communication link based on time-interleaving picosecond pulses with modulated peak amplitudes.
4.3.2.5 Bias Routings in the NLQSI Block

Thin and long on-chip bias routings have significant parasitic effects in high frequency. These effects cause the bias nodes of circuit blocks to be no longer ideal ground. To mitigate this problem, in this work, two wide metal planes on M1 and M2 layers are used as $V_{\text{bias}}$ and $V_{\text{cc}}$ planes, respectively, which are placed directly beneath the ground plane (M3) of the impulse combiner, as shown in Fig. 4.17(a). The advantages of this design are: first, self-inductance of a large metal plane is much smaller than that of a long and thin metal line; second, large distributed capacitance is formed between these layers and the ground plane. As a result, the two wide metal planes with the ground plane can be considered as two transmission lines with a small $Z_0 \left( \sqrt{\frac{L}{C}} \right)$. With a modest length, the designed bias routing planes present a broadband low impedance at the bias nodes. Fig. 4.17(b) shows the impedance of the bias routings at the $V_{\text{cc}}$ and $V_{\text{bias}}$ nodes of NLQSI blocks from 1 GHz to 200 GHz; Third, The ground plane on M3 isolates the impulse combiner from the bias routings, eliminating the undesired mutual couplings.

4.3.2.6 On-Chip Antenna

In this work, a single triangular metal sheet on the top metal layer (AM), with a slot on the ground plane (M3) is designed to couple the radiation to the silicon substrate. The triangular shape is used to support broadband radiation [78]. For assembly purposes, a silicon slab is placed between the silicon chip and the silicon lens, as shown in Fig. 4.18(a). All these components are included in the EM simulations during the design phase; the geometric details are noted in Fig. 4.18(b). An FEM-based 3D EM simulator, HFSS v13, is used to simulate the antenna in the frequency domain. The on-chip antenna has a relatively flat input impedance from 50 GHz to 200 GHz (Fig. 4.19(a)). The simulated radiation efficiency is shown in Fig. 4.19(b). It will be shown
later that the measured radiated picosecond pulses have a peak frequency component around 54 GHz. At this peak frequency, the designed antenna has a 19% simulated radiation efficiency and a 16 dBi simulated peak directivity. The designed antenna has greater radiation efficiencies at higher frequencies, which compensates the weak...
4.4 Chip Characterization

Conventionally, electronic oscilloscopes (real-time sampling or equivalent-time sampling) with antennas or probes are used to sample short pulses in time-domain. As discussed in Section 4.1, this method has major drawbacks. First, current off-the-shelf electronic oscilloscopes have a shortest rising time of 4.5 ps [80], which is not sufficiently fast to measure picosecond pulses accurately. Second, in this method, the picosecond pulses received by antennas/probes have to be transferred to electronic oscilloscopes through waveguides, coaxial cables, and coaxial adapters. Therefore, these blocks need to be accurately de-embedded by performing a broadband calibration, which is complicated, time-consuming, and prone to error [76].

In this work, a time-domain measurement system based on asynchronous optical sampling (ASOPS) was built for characterizing the radiated picosecond pulses by the designed silicon chip. ASOPS has been historically introduced in the THz
Time-Domain Spectroscopy (THz-TDS), where ultrashort EM pulses are generated by the ASOPS system and used to perform spectroscopy analysis of passive or active samples [81]. However, in this work, the generated picosecond pulse is produced by the designed silicon chip rather than the ASOPS system, and this demands technical solutions to ensure that the sample (silicon chip) is synchronized with the ASOPS system. Additionally, in the conventional ASOPS-based THz-TDS, the repetition rate of the generated ultrashort pulse is close to the sampling rate. Instead, in this work, the repetition rate (1 GHz) of the radiated picosecond pulse from the designed silicon chip is much higher than the sampling rate (50 MHz + 5 Hz) of the ASOSP system, which will be discussed later.
Figure 4.20: Simulated time-domain waveform of the radiated picosecond pulse using CST Microwave Studio.

In this section, first, the ASOPS system is briefly reviewed. Second, technical challenges of using this technique to characterize the designed chip are addressed. Finally, measurement results are presented.

4.4.1 Overview of Asynchronous Optical Sampling

Fig. 4.21(a) illustrates the schematic of an ASOPS system. Two femtosecond laser sources generate a pump beam and a probe beam, respectively. The pump beam excites a photoconductive antenna (PCA) emitter that radiates a THz pulse, while the probe beam excites a PCA detector that samples the received THz pulse. The sampled data is then transferred to data acquisition electronics. The repetition rates of these two beams \( f_{r1} \) and \( f_{r2} \) are slightly different, which enables the PCA detector to sample the entire pulse waveform quickly, as shown in Fig. 4.21(b). The two femtosecond laser sources are controlled by two PLLs that share a common frequency
reference for frequency stability purposes. A typical rising time of a PCA detector is on the order of 100 fs [82, 83], which is fast enough to measure picosecond pulses. In addition, because the PCA detector samples the THz pulse right at the antenna, calibration requirements are relaxed significantly compared with the conventional method of using electronic oscilloscopes.

4.4.2 Measurement Setup for Characterizing the Prototype Chip in the Time Domain

In this work, I used a commercial ASOPS system (TAS7500TS) from Advantest Corp. It is capable of measuring a 380 fs THz pulse with an SNR > 1 bandwidth of more than 4 THz [29]. The \( f_{r1} \) and \( f_{r2} \) of this system are 50 MHz and 50 MHz + 5 Hz, respectively. According to the working mechanism of asynchronous optical sampling, the prototype chip needs to be synchronized with the pump femtosecond laser.

4.4.2.1 50 MHz Synchronization Configuration

To test this measurement technique, a straightforward synchronization configuration is first examined, as shown in Fig. 4.22(a). A photodetector is used to convert the 50 MHz pump laser to an electrical trigger, which is fed to the prototype chip. A PCA detector is placed in the far-field region. With the four impulse generation channels ON, the measurement setup captures a 4 ps (FWHM) radiated pulse. Its normalized power spectrum is obtained by performing DFT on the time-domain waveform. The measured radiated pulse has a peak frequency component at 58 GHz, a 10 dB bandwidth of 60 GHz, and an SNR > 1 bandwidth of 161 GHz [Fig. 4.22(b)]. It is necessary to note that the commonly-used relation between pulse width (\( T_p \)) and bandwidth (\( BW \)), which is \( BW = \frac{2}{T_p} \), is not valid in this case, because the generated picosecond pulses are not obtained by modulating a sinusoidal carrier signal with a
Figure 4.21: (a) Asynchronous optical sampling system and (b) its sampling mechanism.

square wave.
Figure 4.22: (a) 50 MHz synchronization configuration and (b) measurement results.

4.4.2.2 A Custom Synchronization Setup

One of the drawbacks of the TAS7500TS system is that the repetition rate of the pump femtosecond laser is fixed at 50 MHz. However, the prototype chip can radiate picosecond pulses with a repetition rate as high as 1 GHz. Therefore, another configuration is designed to generate an adjustable and synchronized trigger, as shown in Fig. 4.23(a). In the synchronization circuitry, a broadband divide-by-five frequency divider is used to extract a 10 MHz sinusoidal signal from the 50 MHz pump fem-
tosecond laser. Then, an RF signal generator is locked with the 10 MHz signal and used to generate synchronized triggers with tunable frequencies. RF filters and LNAs are used in the synchronization circuitry to achieve a low-noise locking with the RF signal generator. Meanwhile, a PCA detector is placed in the far-field region. With a 1 GHz trigger, this measurement setup captures a 4.8 ps (FWHM) radiated pulse. Similar to the simulated result (Fig. 4.20), some ringing appears after the main pulse. The ringing before the main pulse is caused by the multiple round-trip reflections between the chip and the PCA detector. The maximum distance between them is 4 cm, limited by the low sensitivity of the PCA detector. Its normalized power spectrum has a peak frequency component at 54 GHz, a 10 dB bandwidth of 53 GHz, and an $\text{SNR} > 1$ bandwidth of 144 GHz (Fig. 4.23(b)). Fig. 4.24(a) presents the radiated time-domain waveforms in different angles. This measurement shows that the pulse duration remains small in a wide range of angles. In Fig. 4.24(b), the measured radiation patterns of the peak-to-peak amplitude of the radiated pulse waveform are a little tilted compared with the simulation results, which is mainly caused by the tiny misalignment of the silicon lens.

It is important to note that the Advantest TAS7500TS ASOPS system is designed for THz-TDS applications, which measures the changes caused by the sample under test. Therefore, the PCA detector and its internal amplifiers are not fully calibrated for their gains and distortion effects. As a result, all the reported time-domain waveforms and spectrums are normalized and the difference between the measured pulse width of the radiated pulse and the simulated result is mainly due to the non-ideality of the PCA detector. Radiated power characterizations cannot be performed using this system.
Figure 4.23: (a) Custom synchronization setup and (b) measurement results with a 1 GHz trigger.
Figure 4.24: (a) Measured time-domain radiation waveforms using a 1 GHz trigger at different angles. (b) Measured radiation patterns of the peak-to-peak amplitude of the radiated pulse waveform using an 1 GHz trigger.

4.4.3 Demonstrations of NLQSI Effects on Pulse Amplitudes and Pulse Amplitude Modulation

Using the measurement setup shown in Fig. 4.23(a), pulse amplitude modulation based on NLQSI effects is demonstrated. Fig. 4.25 shows the results of this measurement. As discussed in Section 4.2, when the bias voltage of transistor $Q_2$ is set to 2.5 V, the peak of the measured radiated picosecond pulse reduces by 35% compared
with that when $V_{\text{bias}}$ is 2.3 V. This relative change\textsuperscript{4,5} is close to the simulated value (38\%) shown in Fig. 4.20.

Figure 4.25: Measured NLQSI-induced tunable pulse peak amplitudes.

The same setup is also used to measure pulse amplitude modulation by turning on/off the impulse generation channels. Fig. 4.26 presents the measured peak amplitudes of the radiated combined pulses in all 16 combinations. Due to the limited sensitivity of the measurement setup, the combined pulses of the first two combinations are too weak to be detected. The differences between the measured and simulated results are due to the non-idealities of the on-chip antenna and the PCA detector.

\textsuperscript{4,5} Due to the non-ideality of the PCA detector and its internal amplifiers, it is accurate to perform relative change value comparisons.
4.4.4 EIRP (Equivalent Isotropically Radiated Power) and Frequency-Domain Radiation Pattern

To characterize the EIRP spectrum, a frequency-domain measurement setup is utilized, as shown in Fig. 4.27(a). Four OML harmonic mixers and four standard-gain horn antennas are used to measure EIRP from 50 GHz to 220 GHz. The RF signal generator provides an 1 GHz trigger for the prototype chip. Fig. 4.27(b) shows the measured average EIRP spectrum of the radiated picosecond pulse train with an 1 ns period. It has a peak frequency component at 54 GHz with an average EIRP value of -9.4 dBm, which is close to the simulated value. The measured and simulated EIRP spectrums have similar decay trends. However, in the mm-Wave regime, the simulated EIRP values are larger than the measured results, which is mainly caused
by the inaccurate extrapolated transistor models. The highest detectable frequency component is at 197 GHz, which is limited by the sensitivity of the measurement setup. Meanwhile, the peak frequency in the measured EIRP spectrum is identical to that obtained using the ASOPS system, as shown in Fig. 4.23. Compared with the measured power spectrum in Fig. 4.23, the decay rates after the peak frequency components differ in two figures, which is due to the gain effects of the PCA detector and internal amplifiers in the ASOPS system.

Figure 4.27: (a) EIRP measurement setup and (b) measured average EIRP spectrum.

By using the setup in Fig. 4.27(a), the radiation patterns at the peak frequency
of 54 GHz are measured, as shown in Fig. 4.28. The measured main lobes are a little tilted compared with the simulation results, which is due to the chip package and the tiny misalignment of the silicon lens.

Finally, chip micrograph is shown in Fig. 4.29. The chip is fabricated in a 130 nm SiGe BiCMOS process and occupies a die area of 1 mm × 1 mm. The chip consumes a dc power of 170 mW.

![Normalized E-Plane Pattern at 54 GHz](image1)

![Normalized H-Plane Pattern at 54 GHz](image2)

- **Figure 4.28:** Measured frequency-domain radiation patterns at 54 GHz.

### 4.5 Conclusions

In this work, an NLQSI technique is reported for the generation of picosecond pulses with tunable peak amplitudes. A prototype chip is implemented that comprises four NLQSI-based impulse generation channels, an on-chip impulse combiner, and an on-chip antenna. In addition, an on-chip impulse-coupling scheme is introduced to eliminate the undesired ringing caused by the transmission line effects of the supply...
routings. The on-chip impulse combiner provides a single-chip solution for radiating picosecond pulses with amplitude modulation capability. For the first time, an asynchronous optical sampling system is used to characterize the picosecond pulses directly radiated by a silicon chip in the time domain. Based on the measurements, the prototype chip radiates 4 ps pulses with an $SNR > 1$ bandwidth of 161 GHz. The performance of the chip is compared with state-of-the-art silicon-based picosecond impulse radiators in Table 4.1.

Figure 4.29: Chip micrograph of the NLQSI-based impulse radiator prototype chip.
Comparison with state-of-the-art picosecond impulse radiators in silicon

<table>
<thead>
<tr>
<th></th>
<th>This Work</th>
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<td>Measured SNR&gt;1 Bandwidth</td>
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<td>Approx. 170GHz (Freq. Domain)⁴</td>
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<td>13 dBm @ 50 GHz (peak EIRP)⁵</td>
<td>12 dBm @ 94 GHz (CW mode)</td>
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a The power level of the radiated pulse and the sensitivity level of the receivers in these two papers are different from this work.

b It is average EIRP value of a 4.8 ps pulse train with an 1 ns period.

⁴ Peak EIRP is the estimated CW-mode EIRP based on the measured average EIRP.

⁶ It includes the die area of a receiver.

Table 4.1: Comparison with state-of-the-art picosecond impulse radiators in silicon.
This chapter presents a direct time-domain characterization method for silicon-based integrated picosecond impulse radiators using a femtosecond laser-gated optoelectronic sampling technique, which was reported in [84]. In the proposed measurement system, a 1550 nm femtosecond laser source is used to generate an electrical trigger signal fed to a picosecond impulse radiator, and another synchronized 1550 nm femtosecond laser source is used to gate a photoconductive detector. Technical challenges are addressed to synchronize the silicon radiators with the optoelectronic sampling system. This chapter describes the details of the proposed technique and characterization of 4.8 ps impulses radiated by a custom silicon chip.

5.1 Introduction

Silicon technologies (CMOS and BiCMOS) provide transistors with current gain cutoff frequency ($f_T$) and maximum oscillation frequency ($f_{MAX}$) values exceeding 300 GHz and 400 GHz, respectively [65, 71]. Recently, researchers reported silicon-based THz
integrated electronics, radiating sub-5ps pulses without using any femtosecond lasers [39]. These ultra-short electromagnetic pulses have many applications in biology and medical sciences [61], high-resolution 3D imaging [56, 4], non-destructive evaluation [27], and environmental monitoring [7, 58]. Compared with laser-based THz pulse radiators (photoconductive antenna) [64], silicon-based THz impulse radiators have the advantages of low-cost, high scalability, and low power consumption. Over the past several years, researchers reported picosecond impulse generators and radiators using silicon technologies [40, 15, 75]. In these work, researchers used fully electronic devices to characterize picosecond impulses.\footnote{5.1 These devices include antennas, waveguides, coaxial cables, adapters, and high-speed electronic sampling oscilloscopes.}

The conventional oscilloscope-based characterization methods have insufficient bandwidth in characterizing picosecond pulses [76]. The existing off-the-shelf electronic sampling oscilloscopes are not sufficiently fast to measure picosecond pulses. In an electronic oscilloscope, rise time and bandwidth are related by [85, 25],

$$\text{BW} = \frac{k}{\text{Rise Time}_{(10\%-90\%)}}$$ \hspace{1cm} (5.1)

where \(\text{BW}\) is the bandwidth of oscilloscopes. \(k\) varies from 0.4 to 0.45 for oscilloscopes with a bandwidth of larger than 1 GHz. As reported in [25], oscilloscopes must have sufficiently small rise time in order to accurately capture the details of rapid transitions. The requirement of the rise time is shown below,

$$t_{\text{rise,oscilloscope}} \leq \frac{t_{\text{rise,signal}}}{5}$$ \hspace{1cm} (5.2)

where \(t_{\text{rise,oscilloscope}}\) is the 10\% - 90\% rise time of oscilloscopes and \(t_{\text{rise,signal}}\)

\footnote{5.1 Frequency-domain methods based on spectrum analyzer and harmonic mixers are narrowband and not suitable for characterizing picosecond impulses with an instantaneous bandwidth value exceeding 100 GHz.}
is that of the signal of interest. For example, a pulse with a 10 ps full-width-at-half-maximum (FWHM) has an approximate 5 ps rise time. In order to measure this 10 ps pulse accurately, according to Eq. 5.2, an oscilloscope should have a rise time of less than 1 ps. However, the highest bandwidth that current off-the-shelf electronic sampling oscilloscopes\(^5\) can achieve is 70 GHz, associated with a 5 ps rise time (10% - 90%) [86]. Therefore, these electronic sampling oscilloscopes provide insufficient bandwidth for accurate measurements of sub-10ps pulses.

In this chapter, I introduce a direct time-domain measurement method for characterizing silicon integrated picosecond impulse radiators using a femtosecond laser-gated optoelectronic sampling technique. This measurement method overcomes the bandwidth limitation in the conventional oscilloscope-based methods. Historically, femtosecond laser-gated photoconductive antennas (PCAs) are used as fast sampling tools in the field of THz Time-Domain Spectroscopy (THz-TDS) [87, 88, 89, 90]. High spectral resolution and fast data acquisition speed can be achieved in THz-TDS [87, 81]. THz-TDS is a common tool for transmission or reflection spectroscopy analysis of various passive samples. Femtosecond laser-gated optoelectronic sampling is an attractive solution to accurately characterizing ultrafast electromagnetic signals generated and radiated by fully-electronic silicon chips in the THz regime. Because femtosecond laser-gated optoelectronic sampling has a superior rise time, typically on the order of 100 fs, which is associated with a bandwidth of 4.5 THz [82, 83], it is a powerful characterization tool that can help circuit designers investigate silicon integrated picosecond impulse radiators in the millimeter wave (mm-wave) and THz regimes. This proposed technique fills in a literature gap on developing a femtosecond laser-gated optoelectronic sampling system for direct time-domain characterization of

\(^5\) The fastest commercial real-time oscilloscope cannot be used for measuring picosecond pulses because its sample timing interval is 4.2 ps, which is not sufficiently small to record picosecond pulses with enough number of samples [80].
This paper demonstrates a custom time-domain measurement setup based on femtosecond laser-gated optoelectronic sampling to characterize picosecond pulses radiated by a silicon chip. In this work, I focus on the technical challenge of system-level integration between the custom silicon chips under test and the optoelectronic sampling unit. A technical solution to synchronizing the silicon chips with the optoelectronic sampling system is provided. The reported measurement technique has successfully captured a picosecond pulse with an FWHM of 4.8 ps radiated by a custom silicon chip. This is 46% improvement compared with an FWHM of 7 ps obtained using a conventional method based on a fast commercial electronic sampling oscilloscope. Compared with silicon-based impulse detectors [91, 92, 93, 94, 95], the proposed optoelectronic-based solution provides much higher sampling resolution and significantly larger bandwidth.

The reminder of this chapter is organized as follows: Section 5.2 reviews the fundamentals of femtosecond laser-gated optoelectronic sampling technique, focusing on the physical mechanisms and measurement principles. Section 5.3 reports the details of the proposed time-domain optoelectronic sampling system for silicon integrated picosecond impulse radiators. Section 5.4 discusses the measurement results and compares them with the results obtained using the conventional methods based on a fast commercial electronic sampling oscilloscope. Finally, section 5.5 concludes this chapter.
5.2 Femtosecond Laser-Gated Optoelectronic Sampling Technique

In a femtosecond laser-gated optoelectronic sampling system, PCAs are commonly used as sampling probes. Compared with electro-optical (EO) sampling methods, PCAs do not require additional optical devices to measure laser polarization changes. In this section, first, the physical mechanism of the optoelectronic sampling operation in PCAs is discussed. Second, technical details of a femtosecond laser-gated optoelectronic sampling system are described.

5.2.1 Photoconductive Antennas (PCAs)

PCAs are widely used to measure transient electric field of a THz pulse in the time domain, recording both amplitude and phase information of a THz pulse [64, 18]. Fig. 5.1 shows a simple PCA, consisting of two metal contacts evaporated onto a semiconductor substrate. When a PCA is used as a sampling probe, two electrodes are connected to the input of detection and data-acquisition electronics. A femtosecond laser pulse shines at the gap between the electrodes. In absence of a THz wave, only photo-carriers are produced and no current is generated. In presence of a THz wave, the electric field of the THz wave drives the photo-carriers and results in a non-zero current. This current is amplified and digitized by a data acquisition unit following the PCA. A silicon lens is usually mounted onto the backside of a PCA to increase antennas gain.

According to Ohms law, in a semiconductor material, the generated transient current density $\mathbf{J}(t)$ is the multiplication of transient conductivity response $\sigma_{\text{sub.}}(t)$

---

5.3 Electro-optical (EO) sampling methods exploit the Pockels effect in particular crystalline materials, which become birefringent in the presence of the E-field of THz pulse, causing a change in the optical polarization of the probe pulse. The polarization change is proportional to the E-field strength of THz pulse [96].
Figure 5.1: Schematic of a photoconductive antenna (PCA).

and transient driving electric field $\overrightarrow{E_{THz}}(t)$. The relation can be expressed as

$$\overrightarrow{J}(t) = \sigma_{sub.}(t) \overrightarrow{E_{THz}}(t)$$  \hspace{1cm} (5.3)

The conductivity response of the semiconductor, $\sigma_{sub.}(t)$, depends on the concentration of photo-carriers generated by the femtosecond laser pulse. As illustrated in Fig. 5.2, when the timing delay between the received THz pulse and the femtosecond laser pulse is $\tau$, the generated transient current, $I(t)$, will be expressed as

$$\overrightarrow{I}(t) = A\sigma_{sub.}(t - \tau) \overrightarrow{E_{THz}}(t)$$  \hspace{1cm} (5.4)
where $A$ is the cross-section area of the conducting region in the semiconductor substrate.

Figure 5.2: Optoelectronic sampling mechanism of a PCA detector with (a) slow substrate conductivity response and (b) fast substrate conductivity response.

If the semiconductor substrate is grown at low temperature or is ion-damaged, exhibiting very short photocarrier lifetimes [7], the detection circuit will operate as long as the PCA detector is illuminated by the femtosecond laser pulse. As shown in Fig. 5.2(b), because the duration of a femtosecond laser pulse, e.g. 50 fs, is usually much shorter than that of a THz pulse, Eq. 5.4 can be further simplified using a Dirac delta function,

\[ \vec{I}(t) = K \delta_{\text{sub.}}(t - \tau) \overrightarrow{E_{\text{THz}}}(t) \]  

(5.5)

where $K$ is a constant number related with the conductivity response strength and the cross-section area $A$. 

\[ \tau \]
Therefore, when the timing delay between the received THz pulse and the femtosecond laser pulse is $\tau$, at each sampling, the collected electrical charge $Q(\tau)$ is

$$Q(\tau) = \int_{\tau}^{\tau^+} I(t) \, dt = \int_{\tau}^{\tau^+} K\delta_{\text{sub.}}(t - \tau) E_{THz}(t) \, dt = KE_{THz}(\tau) \quad (5.6)$$

The sampled output, $V_{out}(\tau)$, is proportional to the collected electrical charge, $Q(\tau)$, which can be expressed as

$$V_{out}(\tau) \propto Q(\tau) \propto E_{THz}(\tau) \quad (5.7)$$

Therefore, the time-domain waveform of received THz pulse can be obtained by directly plotting the sampled outputs with different timing delays $\tau$. In this work, this condition is valid in measuring picosecond pulses. Although the above condition is not valid in measuring <100 fs THz pulses, the actual time-domain waveform of THz pulse can still be recovered if substrate conductivity response $\sigma_{\text{sub.}}(t)$ is characterized [97].

As shown in Eq. 5.3, $\sigma_{\text{sub.}}(t)$, the conductivity response of a PCAs semiconductor substrate, determines the bandwidth of the PCA detector. With a femtosecond laser pulse excitation, which can be considered as an instantaneous excitation, $\sigma_{\text{sub.}}(t)$ rises exponentially towards a steady-state value with a time constant given by the carrier scattering time [83]. A typical value of semiconductor carrier scattering time is less than 100 fs [82], which is much shorter than the rise time of state-of-the-art off-the-shelf electronic sampling oscilloscopes (5 ps) [86]. According to Eq. 5.1, the typical bandwidth of a PCA detector (probe) can be estimated to be around 4.5 THz.

Apart from the optoelectronic sampling mechanism, signal-to-noise ratio (SNR) is another concern for a PCA detector. The intrinsic noise of a PCA mainly comes from the Johnson-Nyquist noise, which is related to the average resistance of the PCA.
Substrate materials with shorter photocarrier lifetime result in PCAs with a lower noise current level. However, substrate materials with shorter lifetime have lower photocurrent responsivity as an expense. As a result, to maximize PCAs SNR, it is important to optimize the substrate material [98].

5.2.2 A Femtosecond Laser-Gated Optoelectronic Sampling System

Fig. 5.3(a) shows a femtosecond laser-gated optoelectronic sampling system designed for THz-TDS applications. In such a system, two femtosecond laser sources generate two beams. One is called the pump beam, the other is called the probe beam. These two laser beams have slightly different repetition frequencies, $f_{r1}$ and $f_{r2}$, respectively. The frequency detuning is denoted as $\Delta f_r$. The pump beam excites a PCA emitter, which produces synchronized THz pulse radiations. The probe pulse travels to gate a PCA detector in order to measure the THz pulse radiation. The output of the PCA detector is connected to data-acquisition electronics. The trigger signal required for data acquisition can be extracted from the pump and probe beams.

As shown in Fig. 5.3(b), the repetition frequency detuning between the pump and probe beam enables the PCA detector to sample the whole measurement window, which is defined by the pulse-to-pulse timing spacing of THz pulse radiations. The scan rate is determined solely by the repetition frequency detuning $\Delta f_r$ under the condition that $\Delta f_r$ is much smaller than $f_{r1}$ and $f_{r2}$. The upper limit of $\Delta f_r$ is determined by the bandwidth of detection and date acquisition circuits, as well as the required timing resolution [81]. This configuration has a superior scan rate and does not require a mechanical translation stage for scanning a large range in the time domain. This approach eliminates the noise induced by mechanical movements.

The sampling mechanism can also be explained in the frequency domain, as shown
in Fig. 5.3(c). The THz impulse train under test has a frequency spectrum consisting of discrete repetitive frequency tones with a frequency spacing of \( f_{r1} \). Similarly, the probe femtosecond laser pulse train has a broader frequency spectrum due to its shorter timing duration. The frequency spacing between the discrete frequency tones in the probe laser is \( f_{r2} \). Sampling operation can be explained in the frequency domain as convolution operation between the frequency spectrums of the THz impulse...
train and the probe femtosecond laser. Filters are used to ensure that only 1st-order harmonic mixing outputs are collected. As a result, all the high-frequency components in the THz impulse train are mapped to low-frequency components less than several GHz, which significantly relaxes the requirements on the speed of acquisition electronics.

In a time-resolved sampling measurement, the signal of interest and the trigger (femtosecond laser) should be synchronized. For example, in the femtosecond laser-gated optoelectronic sampling system shown in Fig. 5.3(a), the signal of interest is the radiated THz pulse triggered by the pump beam, and the detection trigger is the probe beam. These two beams are generated by two individual femtosecond laser sources. Repetition frequency fluctuations of the two laser sources generate a deviation from the chosen scan rate, $\Delta f_r$, resulting in timing jitters and distortions during sampling. A smaller effective sample timing interval ($\Delta T$) can be obtained by making $\Delta f_r$ smaller. However, minimum $\Delta T$ is limited. Because of the frequency fluctuations, each THz impulse and femtosecond laser pulse have timing jitters ($1$ and $2$). Therefore, $\Delta T$ should be significantly larger than $1$ and $2$ so that $\Delta T$ is stable and the sampling results are accurate. Timing jitters can be reduced by both averaging operations and frequency synthesizers, like Phase-Locked-Loops (PLLs).

As shown in Fig. 5.3(a), PLLs are used to stabilize the repetition frequencies of two femtosecond laser sources. In the proposed time-domain characterization system based on femtosecond laser-gated optoelectronic sampling, silicon integrated picosecond impulse radiators are synchronized with the femtosecond laser gating the photoconductive detector.

It is noteworthy to compare this femtosecond laser-gated optoelectronic sampling method with the conventional RF direct down-conversion sampling method. In the conventional RF mixer-based sampling method, the THz impulse signal is down-
converted with a tunable single-frequency RF source so that all the discrete frequency components (THz Pulse) shown in the top subfigure of Fig. 5.3(c) can be sampled. Because the THz impulse signal has a broad frequency spectrum with SNR>1 bandwidth ranging from hundreds of GHz to several THz, the tunable single-frequency RF source must have a large tuning range, which makes it very challenging and costly to build. Meanwhile, the front-end mixer in this method needs to operate in a large bandwidth covering several microwave bands, and consequently, multiple mixers operating in different bands must be used, introducing more system complexity, system cost, and calibration burden.

5.3 A Femtosecond Laser-Gated Optoelectronic Sampling System for Silicon Integrated Picosecond Impulse Radiators

This section presents the first successful demonstration of a time-domain femtosecond laser-gated optoelectronic sampling system for silicon integrated picosecond impulse radiators. In this section, first, the measurement setup is described. Second, the technical solution of synchronizing the silicon radiators and the optoelectronic sampling system is discussed.

5.3.1 Configuration of the Proposed Optoelectronic Sampling System

Fig. 5.4 presents the schematic of the proposed characterization system. A PCA detector (Advantest TAS1230) is used to capture the radiated picosecond impulses by the silicon integrated chips under test. The gating femtosecond laser (probe laser) for
the PCA detector is provided by the measurement unit of an Advantest TAS7500TS system. In addition, the measurement unit has another femtosecond laser source (pump laser) that is used to synchronize the silicon chips under test and the optoelectronic sampling system. The synchronization method will be discussed in section 5.3.2. The measurement unit transfers the sampled waveform obtained by the PCA detector to an analysis unit, which performs post-processing. These results can be displayed on the system monitor. Since the analysis unit has a limited memory, the displayed time-domain waveforms have a limited timing duration. To resolve this problem, a digital oscilloscope (Tektronix 4104B-L) is connected to the measurement unit to read the real-time sampled output of the received picosecond pulses. The PCA detector is placed in the far-field region of the radiating elements on the silicon chips. In this measurement setup, no parabolic mirror is used.

In addition to the enhanced measurement bandwidth achieved by the superior rise time of the PCA detector, the femtosecond laser-gated optoelectronic time-domain characterization tool offers a much simpler calibration compared to conventional solutions that are based on high-speed electronic oscilloscopes. In the reported method, the received picosecond pulses are directly sampled at the PCA detector. Therefore, the remaining circuitry between the PCA and the acquisition unit only transfers low-frequency signals (sampled data) and does not require high-frequency calibration. In contrast, in the conventional methods based on high-speed electronic oscilloscopes, because the received ultra-short pulse needs to be delivered to the sampler at the end of the receiving path, broadband calibrations are demanded for multiple passive components that connect the antenna to the sampler. Such a procedure is time-consuming and prone to errors.

In this proposed measurement method, linearity of PCA detectors is an important issue for detection accuracy. According to the working mechanism of PCA detection,
Figure 5.4: Schematic of the proposed femtosecond laser-gated optoelectronic sampling system for broadband characterization of silicon integrated picosecond impulse radiators.

saturation will happen when the input THz signal reaches a power larger than the saturation power ($P_{sat}$). The saturation could be caused by the photo-conductive antenna or the electronics that amplify the detected photocurrent. The saturation results in distortions on the measured pulse waveform. Therefore, saturation of the PCA detector is detrimental and should be avoided. The PCA detector (Advantest TAS1230) used in this work has a peak dynamic range of more than 60 dB [99].

It is verified that the received power by the PCA is below the $P_{sat}$ of the PCA. If the device under test produces power reaching $P_{sat}$ of the PCA detector, wave...
attenuators will be required to make sure that the PCA detector works in the linear regime. Calibration process can be used to compensate the attenuator loss.

![Diagram of synchronization technique](image)

Figure 5.5: (a) Block diagram of the proposed synchronization technique. (b) Actual circuit blocks for synchronization signal generation. (c) Converted electrical pulses from the pump femtosecond laser.

### 5.3.2 Details of the Synchronization Technique

In the reported measurement method, one of the technical challenges is to synchronize the silicon chips under test with the probe femtosecond laser source in the optoelectronic sampling system. To explain the synchronization method, it is necessary to briefly review the working principle of the silicon-based picosecond impulse radiators.
under test [39, 75]. The pulse radiating chips convert an electrical input trigger signal to a radiated picosecond pulse that is synchronized with the trigger signal. In the proposed system, the pump and probe femtosecond laser sources are well synchronized in the measurement unit. Therefore, if the input trigger signal is synchronized with the pump femtosecond laser source, the radiated picosecond pulses by the chips will be synchronized with the optoelectronic sampling system, which meets the requirement for time-resolved sampling measurements (discussed in section 5.2.2).

Fig. 5.5(a) shows the block diagram of the synchronization technique. The output port of the pump laser source is attenuated with an optical attenuator (Thorlabs OVA50-APC) before being converted to an electrical trigger by a photodetector (Thorlabs DET01CFC). An optical isolator protects the laser source by minimizing the reflected power and a bias-tee (Picosecond Pulse Labs) is used to provide a proper dc bias voltage for the photodetector output. The generated electrical trigger is a 50 MHz pulse train with a broadened pulse-width due to the limited bandwidth of the photodetector, as shown in Fig. 5.5(c). A 50 MHz filter (Mini-Circuits SIF-50+) extracts a 50 MHz signal and feeds it to a broadband divide-by-5 frequency divider (Analog Devices 438MS8G). The undesired harmonic spurs are eliminated by a 10 MHz narrowband filter (Mini-Circuits SBP-10.7+). After applying a low-noise amplifier (RF Bay LNA-250) and another 10 MHz narrowband filter, a clean 10 MHz signal is obtained with enough power to synchronize a RF signal generator (Keysight E8257D), which provides the input trigger for the silicon picosecond impulse radiators under test.

In the proposed measurement method, the PCA detector samples the received picosecond pulse radiation with a sampling rate, \( f_s \), which is given by that of the probe femtosecond laser. Here, \( f_s \) is 50 MHz + 5 Hz and the repetition rate of the THz pulse radiation, \( f_r \), is 50 MHz. This means that the effective timing interval of
optoelectronic samples is \((1/50 \text{ MHz} - 1/50.000005 \text{ MHz}) = 2 \text{ fs}\), which is validated with averaging operations and fully characterized. It should be noted that, although the repetition rate of the pump laser is 50 MHz, the proposed optoelectronic sampling system allows us to trigger the silicon radiators with any repetition rate that is a harmonic of 50 MHz. For example, as shown in Fig. 5.6, if the repetition rate of the radiated pulse train, \(f_r\), is a \(N_{th}\) harmonic of 50 MHz, the PCA detector will sample one point for every \(N\) received pulses and the actual sample timing interval of the output waveform is still 2 fs. If the repetition rate is not a harmonic of 50 MHz, the sample timing interval of the output waveform needs to be scaled accordingly. In fact, in our measurements, the repetition rates of the radiated picosecond pulses are larger than 1 GHz.

![Diagram showing sampling mechanism](image)

Figure 5.6: Sampling mechanism of the proposed optoelectronic characterization method when the repetition rate of the signal of interest is a \(N_{th}\) harmonic of 50 MHz.
5.4 Measurement Results

Fig. 5.7 presents the measurement setup used for characterizing the custom silicon picosecond impulse radiators. In this setup, the PCA detector is placed in the far-field region of the chips under test. No parabolic mirror is used in this measurement. A 10 MHz synchronization signal is generated using the circuit shown in Fig. 5.5(a) and its time-domain waveform is shown in Fig. 5.8(a). This synchronization signal has a power of 7.7 dBm at 10 MHz (Fig. 5.8(b)). To investigate the stability of the generated 10 MHz synchronization signal, we have compared its phase noise with that of a state-of-the-art signal generator (Keysight 8257D) that uses an internal reference. This comparison is shown in Fig. 5.8(c).

![RF Signal Generator (10MHz Locked)](image)

PCA Detector

![Chip](image)

Figure 5.7: Measurement setup for characterizing the custom-designed silicon integrated picosecond impulse radiators.

In this work, a custom silicon integrated picosecond impulse radiators [39], is characterized using the measurement setup shown in Fig. 5.7. The RF signal generator (Keysight E8257D) provides sinusoidal trigger signals to the silicon radiators. The
chip radiate picosecond pulses with the same repetition rate as the input trigger. Design details of the silicon radiators are discussed in [39]. In this work, the radiator in [39] is triggered using an 1 GHz sinusoidal signal.

In a prior work [26], the linear regime of the PCA detector (Advantest TAS1230) has been tested. The measured waveforms were not distorted when the tunable source power was increased to its maximum level. Because the two custom fully-electronic

Figure 5.8: (a) Measured time-domain waveform of the 10 MHz synchronization signal. (b) Measured frequency spectrum of the 10 MHz synchronization signal. (c) Measured phase noise of the 10 MHz synchronization signal.
chips tested by using the proposed measurement system produce less power than the photonic chip in [26], the PCA detector remains in the linear regime in the following measurements.

Fig. 5.9 demonstrates the characterization results of the chip reported in [39]. The time-domain waveform of the captured pulse is shown in Fig. 5.9(a). This chip radiates pulses with an FWHM of 4.8 ps. In order to increase SNR, an averaging of 512 is used. The frequency spectrum of the picosecond pulse radiation is obtained by performing DFT on the recorded time-domain waveform (Fig. 5.9(b)). The 4.8 ps pulse radiation has a peak frequency component at 54 GHz. Its 10 dB bandwidth is 53 GHz, and its SNR>1 bandwidth is 144 GHz.

Figure 5.9: Measurement results of the picosecond pulse radiated by the chip using the proposed femtosecond laser-gated optoelectronic sampling system. (a) Measured time-domain waveform and (b) DFT power spectrum.

The chip was also characterized by using the conventional method based on a sampling oscilloscope (Keysight 86100D) with a sampling module of Keysight 86118A. The measurement setup of this method is shown in Fig. 5.10. A custom wideband PCB receiving antenna is used to capture the radiated picosecond pulses [15]. The received signal is delivered to the sampling module (Keysight 86118A) via high
frequency coaxial cables and adapters (1.85 mm). The sampling module (Keysight 86118A), which is the fastest off-the-shelf sampling module available for electrical characterization, has a 3 dB bandwidth of 70 GHz. It is important to note that even if a higher-speed sampling oscilloscope existed, it would still require time-consuming procedures to assembly low loss waveguides and adapters to connect antennas and the sampler. Meanwhile, these mm-wave waveguides and adapters increase the cost of the measurement system. Fig. 5.11 shows the characterization results using this conventional method. Because the sampling oscilloscope and the PCB receiving antenna have limited bandwidth, the captured picosecond pulse has an FWHM of 7 ps, which shows that the proposed optoelectronic sampling method has 46% improvement on the measured pulse FWHM compared with the conventional method.\textsuperscript{5,5} In addition, the DFT power spectrum has a peak frequency component at 40 GHz, which is smaller than that obtained by using the proposed optoelectronic sampling technique. This comparison demonstrates that the bandwidth of the coaxial cables, adapters, and electronic oscilloscopes is crucial for accurate characterization of picosecond pulses.

![Figure 5.10: Measurement setup of using the conventional method based on a high-speed sampling oscilloscope.](image)

\textsuperscript{5,5} The improvement is calculated as \((7\text{ps} - 4.8\text{ps})/4.8\text{ps} = 45.8\%\).
Figure 5.11: Measurement results of the picosecond pulse radiated by the chip I using the conventional method. (a) Measured time-domain waveform and (b) DFT power spectrum.

The measurement scenario, shown in Fig. 5.7, was also simulated by using a FDTD-based EM simulator, CST Microwave Studio. In the simulation, a far-field E-field probe is placed at a distance of 4 cm away from the custom silicon radiator. Fig. 5.12 presents the simulation results. The time-domain waveform of the simulated picosecond pulse has an FWHM of 3.6 ps. Its DFT power spectrum has a peak frequency component at 60 GHz and a 10 dB bandwidth of 66 GHz. The differences between the measured (Fig. 5.9) and simulated (Fig. 5.12) results are due to the fact that an ideal E-field probe rather than the actual PCA detector is modeled in the simulation. Therefore, the simulated results are free from the effects induced by the PCA detector. The electronics following the PCA, such as the internal amplifiers and filters, operate at low frequencies (smaller than 50 MHz) and can be easily calibrated.

5.6 Because there is no noise added in the simulation, SNR>1 bandwidth is not shown in the simulated results.
Figure 5.12: Simulation results of the picosecond pulse radiated by the chip I in the measurement setup shown in Fig. 5.7. (a) Simulated time-domain waveform and (b) DFT power spectrum.

5.5 Conclusions

This chapter presents a femtosecond laser-gated optoelectronic sampling technique for direct time-domain characterization of silicon integrated picosecond impulse radiators. The typical measurement rise time of the custom characterization system is less than 100 fs, which is more than 50 times smaller than that of the fastest commercial sampling oscilloscope (5 ps). This proposed time-domain characterization system can help circuit designers investigate performance of silicon radiators in the mm-wave and THz regimes. The proposed optoelectronic sampling system was used to measure picosecond pulses radiated by a custom-designed silicon radiator. A pulse FWHM of 4.8 ps has been recorded by using the proposed system. A 46% improvement on the measured pulse FWHM is achieved compared with the conventional method based on high-speed electronic sampling oscilloscopes. Finally, a comparison between the proposed femtosecond laser-gated optoelectronic sampling technique and the conven-
Comparison between the Proposed Femtosecond Laser-Gated Optoelectronic Sampling Technique and the Conventional Method based on a State-of-the-Art Electronic Sampling Oscilloscope

<table>
<thead>
<tr>
<th>Measurement Domain</th>
<th>Femtosecond Laser-Gated Optoelectronic Sampling Technique</th>
<th>Conventional Oscilloscope-based Method</th>
</tr>
</thead>
<tbody>
<tr>
<td>Time Domain</td>
<td>Time Domain</td>
<td></td>
</tr>
<tr>
<td>Sampling Method</td>
<td>Femtosecond Laser-Gated Optoelectronic Sampling</td>
<td>Electronic equivalent-time sampling</td>
</tr>
<tr>
<td>Measurement Bandwidth</td>
<td>4.5 THz *</td>
<td>70 GHz b</td>
</tr>
<tr>
<td>Rise Time (10% - 90%)</td>
<td>100 fs c</td>
<td>5 ps b</td>
</tr>
<tr>
<td>Measured FWHM of Radiated Pulse by the Chip [39]</td>
<td>4.8 ps</td>
<td>7 ps b</td>
</tr>
<tr>
<td>Sampling Trigger Frequency</td>
<td>50 MHz + 5 Hz d</td>
<td>DC - 3.2 GHz</td>
</tr>
<tr>
<td>Effective Sample Timing Interval</td>
<td>2 fs e</td>
<td>62.5 fs b</td>
</tr>
</tbody>
</table>

* This value is estimated based on the typical rise time of a PCA detector (100 fs), as discussed in section 5.2.1.

b These values are obtained based on the fastest commercial electronic sampling oscilloscope used in this work, Keysight Technologies Infiniium DCA-X 86118D wide-bandwidth oscilloscope with an 86118A sampling module (70 GHz) [86].

c This is a typical value of a PCA detector, as discussed in section 5.2.1.

d This value is limited by the probe femtosecond laser source in the optoelectronic sampling system used in this work.

e This value is estimated by calculating the difference between $1/50$ MHz and $1/50.000005$ MHz. In practice, this number is limited by the jitter of the optoelectronic characterization system and the averaging number.

Table 5.1: Comparison between the proposed femtosecond laser-gated optoelectronic sampling technique and the conventional method based on a state-of-the-art electronic sampling oscilloscope.

The conventional method based on a 70 GHz electronic sampling oscilloscope is summarized in Table 5.1.
This dissertation describes three design techniques for broadband mm-wave and THz impulse radiators using silicon technologies and one measurement method for characterizing silicon-based impulse radiators in the time domain directly. The work presented in this dissertation resolves the shortcomings and challenges in the existing techniques.

The first design technique focuses on using silicon technologies to achieve the functionalities of the conventional photonics-based mm-wave and THz impulse radiators, which usually demand non-silicon process technologies. As described in the Section 2, a Germanium-based waveguide-coupled PCA was implemented using a CMOS-compatible silicon photonics process technology, which enables potentials of integrating both photonics-based mm-wave and THz devices and electronics control units on a single chip. Additionally, by using silicon-based technologies and under mass production, the cost of the prototype chip can be smaller than those of conventional photonics counterparts. Thanks to the photoconductive properties of Germanium, the prototype chip can operate with 1550 nm wavelength sources, opening an access to a broader collection of well-developed and low-cost laser sources in the fiber-optics communication fields. Furthermore, a waveguide-coupled optical excitation scheme
for the PCA switch is introduced to allow on-chip manipulations on the optical excitation signals, i.e. femtosecond laser. The prototype chip can radiate 1.14 ps impulses with an SNR>1 bandwidth of 1.5 THz. The average radiated power is 0.337 $\mu$W and its DC-to-RF efficiency is $3.6 \times 10^{-5}$. The future research focus is to improve the radiation bandwidth, radiated power, and energy efficiency. This work presents a research direction of using silicon technologies to implement photonics-based mm-wave and THz impulse radiators that can be integrated with electronics to form powerful mm-wave and THz SOCs.

The second design technique (Section 3) improves the performance of electronic oscillator-based mm-wave and THz impulse radiators using silicon technologies. In the second design technique, a 30GHz impulse radiator is implemented for mm-wave 3D radar imaging using a 130 nm SiGe BiCMOS process technology. An asymmetric cross-coupled pulsed VCO topology is proposed to ensure the coherency of the generated impulse with respect to the input trigger that switches the pulsed VCO. The timing jitter of the generated short impulse can be minimized by optimizing the size ratio between the two transistors in the cross-coupled pair in the pulsed VCO, without the needs of PLLs or DLLs as in the conventional oscillator-based solutions. As a result, the prototype chip has less system complexity, less power consumption, and less die area. With a 106 mW average dc power consumption, the prototype chip can radiate impulses with an FWHM pulse-width of 60 ps and a RMS jitter of 178 fs when the input trigger signal has a 150 fs RMS jitter. The chip achieves an average EIRP of 0.5 dBm and a peak EIRP of 6.5 dBm, when producing impulses with an FWHM pulse-width of 74 ps and a center frequency of 30 GHz. This work, for the first time, presents a proof-of-concept of using a synthetic array of a fully integrated silicon-based impulse radiator to perform high resolution lens-free 3D imaging. 3D radar images of metallic objects and dielectric objects have been successfully demon-
strated. In this work, a depth resolution of 9 mm and a lateral resolution of 8 mm at a range of 10 cm in the air have been achieved.

The third design technique (Section 4) improves the performance of silicon-based Digital-to-Impulse (D2I) methods. A nonlinear Q-switching impedance (NLQSI) technique is proposed to have an autonomous mechanism to adjust the quality-factor of a RLC tank according to the transient voltage output so as to reduce late-time ringing in the generated sub-5 ps impulses. The NLQSI technique can also adjust the pulse peak amplitude. In addition, a four-way on-chip impulse combining is introduced as a single-chip solution for radiating picosecond pulses with amplitude modulation capability. A prototype chip is implemented that consists of four NLQSI-based impulse generation channels, an on-chip impulse combiner, and an on-chip antenna. An on-chip impulse-coupling scheme eliminates the undesired ringing caused by the transmission line effects of the supply routings. For the first time, an asynchronous optoelectronic sampling technique is used to perform direct time-domain characterizations on the picosecond pulses directly radiated by a silicon chip. Based on the measurements results, the prototype chip radiates 4 ps pulses with an $SNR > 1$ bandwidth of 161 GHz. The NLQSI technique can produce a 35% change of pulse peak amplitude. By switching on and off the four impulse generation channels, 14 different levels of peak amplitudes of radiated impulses have been measured.

The optoelectronic sampling technique (Section 5) used for characterizing the NLQSI-based prototype chip resolves the challenges of using conventional electronic equipment to characterize ultra-broadband silicon chips in either frequency domain or time domain, as discussed in the Section 1.3. The proposed optoelectronic sampling technique is a direct time-domain characterization method for silicon-based integrated picosecond impulse radiators. In the proposed measurement system, a 1550 nm femtosecond laser source is used to generate an electrical trigger signal fed to a
picosecond impulse radiator, and another synchronized 1550 nm femtosecond laser source is used to gate a PCA detector. The PCA detector serves as a sampling unit with a typical measurement rise time of less than 100 fs, which is more than 50 times smaller than that of the fastest commercial electronic sampling oscilloscope (5 ps). Custom-designed circuitry is used to synchronize the silicon radiators with the PCA detector, which is crucial for time-resolved measurements. The repetition frequency detuning between the trigger signal fed to the silicon radiators and the femtosecond laser gating the PCA detector enables the detector to sample the whole measurement window. The proposed measurement technique presents a 46% improvement on the measured pulse FWHM compared with the conventional method based on high-speed electronic sampling oscilloscopes.

To sum up, the design techniques presented in this dissertation demonstrate that it is promising to use silicon technologies to implement miniaturized mm-wave and THz impulse radiators, exhibiting advantages of low cost and easier integrations with digital processing units. In addition, optoelectronics sampling technique can be used as a direct time-domain characterization method for silicon-based mm-wave and THz impulse radiators with an ultra-broad radiation bandwidth ($\text{SNR}>1 \text{ BW is greater than 100 GHz}$).

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