High Efficiency Micromachined Sub-THz Channels for Low Cost Interconnect for Planar Integrated Circuits

Bo Yu, Yuhao Liu, Yu Ye, Junyan Ren, Xiaoguang “Leo” Liu, and Qun Jane Gu

Abstract—This paper presents for the first time the design, fabrication, and demonstration of micromachined silicon dielectric waveguide based sub-THz interconnect channel for high efficiency, low cost sub-THz interconnect, aiming to solve the long-standing intra-/inter-chip interconnect problem. Careful studies of the loss mechanisms in the proposed sub-THz interconnect channel are carried out to optimize the design. Both theoretical and experimental results are provided with good agreement. To guide the channel design, a new Figure-of-Merit is also defined. The insertion loss of this first prototype with a 6-mm long interconnect channel is about 8.4 dB at 209.7 GHz, with a 3-dB bandwidth of 12.6 GHz.

Index Terms—Channel, dielectric waveguide, interconnect, micromachined, sub-THz, THz.

I. INTRODUCTION

The I/O bandwidth growth of intra/inter-chip communications doubles every two years over the past decade and the trend is projected to continue in the future [1]. However, the number of I/O pins increases slowly over the time due to physical constraints. To overcome this increasing gap between the I/O bandwidth and pin numbers, the transmitting data bandwidth per I/O, defined as bandwidth density, should keep up with the interconnect bandwidth requirement. In addition, the energy used for data communications may potentially be orders of magnitude higher than the energy used for data processing and storage [2]. There are two scenarios for the interconnect: intra-chip interconnect, which is the communication among CPU cores or among the high speed processing components inside a chip, and inter-chip interconnect, which is the communication between chips. Therefore, to ultimately solve the problem of intra-/inter-chip interconnect, both bandwidth density and energy efficiency should be boosted.

Interconnect research has been active in two areas: optical interconnect [3]–[6] and electrical interconnect [7]–[11]. Optical interconnects have the advantages of low loss and high bandwidth, but it is still very challenging to integrate highly efficient light sources with current CMOS processes [4]. Electrical interconnect schemes are compatible and scalable with silicon processes. However, the transmission media, metal wires, has severe conduction loss with high data rates or operating at high frequencies, thus limiting the supported bandwidth. Therefore, both electrical and optical interconnect face big challenges to fill this gap individually.

Sub-THz interconnect, using the spectrum sandwiched between optical and microwave frequencies, holds high potentials to fill the interconnect gap with wide bandwidth density and high energy efficiency by leveraging advantages of both optical and electrical interconnect approaches: low loss quasi-optical channels as well as advanced high speed semiconductor devices, illustrated in Fig. 1.

The comparison with the state of the art is shown in Table I. Wireless chip-to-chip communication, demonstrated in [7]–[10], suffers from large losses. For example, the path loss is greater than 40.9 dB with 40-mm distance at 260 GHz [7] and 66 dB with 1-m distance at 45 GHz [8]. The challenge of wireless chip-to-chip communications is that the path loss is inversely proportional to $\lambda^2$, which impedes high frequency adoption in wireless scheme. Besides, the interference between channels is a big issue for wireless based schemes. Chang’s group [11] demonstrates a design on wired interconnect based on-chip transmission line, which also faces the challenge of increasingly high losses versus frequencies.

In general, the interconnect can be classified into three types, transmission line (including microstrip line, coplanar waveguide (CPW), grounded CPW, etc.) [12], [13], metallic waveguide [14]–[18], and dielectric waveguide [19], [20]. The dielectric waveguides with low loss dielectric material have much less losses than transmission lines and metallic waveguides since the conduction loss is avoided. For example, a metal based transmission line has almost three orders higher loss than the loss of silicon ribbon at THz frequency [21].
TABLE I
COMPARISON AMONG DIFFERENT TECHNOLOGIES OF CHIP-TO-CHIP COMMUNICATIONS

<table>
<thead>
<tr>
<th>Design</th>
<th>Wired/Wireless</th>
<th>Carrier Freq (GHz)</th>
<th>Distance (mm)</th>
<th>Path Loss (dB)</th>
<th>Normalized Path Loss* (dB) with 10-mm l at 200 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>[8]</td>
<td>Wireless</td>
<td>45</td>
<td>1000</td>
<td>66</td>
<td>38.9</td>
</tr>
<tr>
<td>[9]</td>
<td>Wireless</td>
<td>60</td>
<td>100</td>
<td>58</td>
<td>48.5</td>
</tr>
<tr>
<td>[10]</td>
<td>Wireless</td>
<td>43</td>
<td>100</td>
<td>45</td>
<td>38.4</td>
</tr>
<tr>
<td>[11]</td>
<td>Wired</td>
<td>60</td>
<td>3</td>
<td>6.6</td>
<td>N/A</td>
</tr>
<tr>
<td>This work</td>
<td>Wired</td>
<td>210</td>
<td>6</td>
<td>8.4</td>
<td>8.7</td>
</tr>
</tbody>
</table>

The path loss normalization is calculated based on Friis equation for wireless type.

The loss for the CMOS transmission line is about 1 dB/mm at 100 GHz and 2 dB/mm at 150 GHz and increases fast with frequency [22], [23]. Moreover, from the process-compatible point of view, the dielectric waveguides are easier to fabricate and potentially compatible with integrated silicon circuits compared with metallic waveguides. Metallic waveguide poses a big challenge to integrate with ICs due to the waveguide flange connection. Therefore, to enable sub-THz interconnect, the channel should have wide bandwidth and small size for large bandwidth density, low loss for high energy efficiency, and good compatibility with silicon processes for low cost as well as good isolation among channels.

To satisfy these requirements, dielectric waveguides [24] are employed by taking advantage of quasi-optical channels. Low loss THz channels have been investigated with loss of 0.1 dB/m [21]. However, till now, no investigations have been conducted on planar silicon processes compatible sub-THz channels for intra-/inter-chip interconnects.

In this work, we demonstrate the feasibility of using a micro-machined dielectric waveguide as a low-loss chip-to-chip interconnect channel that is compatible with conventional semiconductor and packaging processes. Fig. 2 illustrates the concept [25]. The signal, transmitting from chip A, propagates through the channel and reaches chip B. The bending structures at two ends of the channel are to establish the link for planar processes.

Compared with the authors’ previous work [25], this paper presents a thorough analysis of the design trades-offs for the silicon dielectric waveguide as well as the analysis of the bending loss, radiation loss, and mode conversion loss. We propose a Figure of Merit (FoM) to quantify channel design by incorporating bandwidth-area efficiency and channel loss. The first demonstration is presented at 210 GHz due to the constraint of the measurement equipment availability. The channel design methodology can be readily applied to higher frequencies in the THz range.

This paper is organized as follows. Section II reviews and presents the fundamental concept and design methods of silicon dielectric waveguide based sub-THz interconnect. Section III discusses the design and considerations of the coupling structure. Section IV presents the fabrication, the measurement, and discussions of the sub-THz interconnect.

II. SILICON DIELECTRIC WAVEGUIDES AS SUB-THZ INTERCONNECT MEDIUM

To enable high data transmission rate, the proposed sub-THz dielectric interconnect channel must be optimized for both bandwidth and loss. The bandwidth of a dielectric waveguide is primarily determined by the dispersion characteristics of the chosen mode of the propagating wave and the orthogonality and/or separation from other modes. In this demonstration of the proposed concept, we choose to operate our waveguide in the lowest-order mode ($E_{311}$) to simplify the design and implementation. The loss of the waveguide is determined by many factors, including the dielectric loss of the material, the geometry of the waveguides, such as bending and discontinuity structures, and possibilities of mode conversions. The following sub-sections provide detail discussions of these factors for design guidance.

A. Material Loss

Material loss can be a critical contribution to the total channel loss. It is therefore desirable to use low-loss materials as the dielectric medium. Several materials have been studied at sub-THz frequencies, such as silicon [21], [26]–[30], quartz [21], and plastic [31], [32]. These research provides evidence that dielectric sub-THz interconnect channels can be designed with low loss. In particular, the loss of high resistivity (HR) silicon has been reported to be as low as 0.1 dB/m at 200 GHz [21], which is one of the reasons that HR silicon is used in this work. A second reason to choose HR silicon is due to its relatively high dielectric constant ($\epsilon_r = 11.9$), which helps to confine the electromagnetic (EM) wave inside the waveguide. A higher level of confinement can reduce cross-talk between adjacent channels, lower packaging parasitics, and facilitates wave guidance in non-straight channels, such as bending structures.

B. Waveguide Geometry

Rectangular silicon waveguides with air surrounding are used in this work. Compared to the other geometries, such as circular waveguides, rectangular waveguides are easier to fabricate using microfabrication technologies, such as the deep reactive-ion etching (DRIE). Also, rectangular waveguides are not prone to have polarization mode dispersion issues [33], which minimize polarization mismatches and losses.

Fig. 3(a)-(c) show the effective index $n_{eff}$, the wave confinement factor $\Gamma$, and the attenuation constant $\alpha$ of rectangular
waveguides of various dimensions \( w \) of 100–500 \( \mu \text{m} \), and \( h \) of 100–700 \( \mu \text{m} \) based on full-wave simulation in ANSYS high frequency structure simulator (HFSS). Due to the unshielded characteristic, the size of the wave port in simulation is set significantly larger than the cross section. The channel is enclosed by an air box with radiation boundary. \( \Gamma \) is given by

\[
\Gamma = \frac{P_{\text{in}}}{P_t}
\]

where \( P_{\text{in}} \) is the power inside the waveguide, and \( P_t \) is the total cross section power. \( \Gamma \) represents how much EM wave energy is propagating inside the dielectric channel. It can be seen that at small \( w \) and \( h \) (compared to the propagation wavelength \( \lambda_p \), which is about 1.3 mm at 200 GHz, the wave is weakly confined and has a large portion of wave propagating outside the channel. At large \( w \) and \( h \) (> 300 \( \mu \text{m} \)), the wave is mostly confined inside the silicon channel, with \( n_{eff} \), \( \Gamma \), \( \alpha \) values saturating with larger dimensions. The cross section views of electric field distribution with various \( h \) are plotted in Fig. 4. It is observed that larger portion of the electric field is confined within the waveguide as \( h \) increases. The attenuation constant \( \alpha \) is a weak function of \( w \) due to the mode polarization along \( h \) direction.

The dispersion is another important aspect to optimize this channel. The phase and group velocity are utilized to check the dispersive characteristics as shown in Fig. 5. When the frequency is higher than 150 GHz, the dispersion decreases. Therefore, from a bandwidth point of view, it is desirable to operate at either the weakly or strongly confined states. However, from a loss point of view, we prefer to avoid the weakly confined region even though its straight channel loss is lower. This is due to the excessive radiation loss caused by the bending structures and large cross-talk among channels when waves are not confined. It is noted that the attenuation constant is still low (\(< 0.09 \text{ dB/mm from simulation} \)) for the highly confined case.

Isolation is also important with multiple channels, which is dependent on the channel space. The simulated isolation
C. Radiation Loss

To implement the intra-/inter- chip interconnect for planar processes, the bending structure is the most intuitive and convenient approach. However, the bending structures may introduce additional loss due to radiation and mode conversion as shown in Fig. 7. Bending structure has been studied as early as 1920s [33]. Researchers have reported very low bending loss designs [27], [28]. However, these works are for electrically large bending structures, such as [27] with about 113 wavelengths and [28] with about 6 wavelengths. Because the practical constraints of integrated circuit fabrication and packaging, a large bending structure is not feasible to integrate and the investigation of a small bending structure is needed.

Radiation loss is caused because the portion of EM waves leaking into the air cannot preserve the phase front after the bending. As shown in Fig. 7(a), the portion of the waves propagating in air does not follow the curvature of the bend and results in the power loss. A method proposed in [34] can be used to analyze the radiation loss. The bending structure can be divided into infinitesimal sections as shown in Fig. 8.

Considering each section as an array of point sources [35] and assuming that the power beyond the first null of the beam, that is, the power in minor lobes, will be lost after bending, the attenuation constant is defined as

$$\alpha_{rad} \approx \frac{1}{Z_c} \frac{P_1}{P_1},$$

(2)

where $P_1$ is the power loss through radiation, $P_l$ is the total cross section power, and $Z_c$ is the field propagation distance with a unit power loss [34]. Based on diffraction theory, $Z_c$ is derived as

$$Z_c = \frac{h/2}{\tan \varphi} = \frac{h/2}{\tan[\sin^{-1}(\frac{\lambda_0}{h})]} \approx \frac{h^2}{2\lambda_g},$$

(3)

where $h$ is the channel height, $\varphi$ is the half beam angle, and the $\lambda_0$ is the guided wavelength in the channel. $P_l$ and $P_i$ can be obtained from

$$P_i = \int_{X_r}^{\infty} H^2(x),$$

(4)

and

$$P_i = \int_{-\infty}^{X_r} H^2(x) dx,$$

(5)

where $H(x)$ is the magnetic field distribution, and $X_r$ is the distance from the waveguide center to the position of the first null. Beyond the radius ($r + X_r$), the waves phase velocity would have to exceed the velocity of unguided waves, which results in the loss of the corresponding wave power. To preserve the phase front [34], $X_r$ has the relationship of

$$(r + X_r) \frac{d\theta}{dt} = \frac{\omega}{\beta_0}.$$  

(6)

In addition, $H(x)$ is given by

$$H(x) = \begin{cases} 
H_0 \cos(mx), & -\frac{h}{2} \leq x \leq \frac{h}{2} \\
H_0 \cos(m \frac{h}{2}) e^{-C_0 (|x| - \frac{h}{2})}, & \text{else} 
\end{cases}$$

(7)

where $H_0$ is the amplitude of $H(x)$, $C_0$ is the decay rate along the $h$ direction, and $m$ is the mode number. Substituting (3), (4), and (5) into (2) yields

$$\alpha_{rad} = \frac{1}{2} \frac{C_0 \cos^2(m \frac{h}{2})e^{-2C_0 (|x| - \frac{h}{2})}}{[\frac{\lambda}{2} + \frac{1}{m} \sin(mh) + \frac{1}{C_0 \cos^2(m \frac{h}{2})}]h^2},$$

(8)
where \( r \) is the channel radius, \( \beta_z \) is the guided propagation constant, and \( \beta_0 \) is the propagation constant in free space. By multiplying the EM waves propagation distance \( l \), the arc length of a quarter circle, the radiation loss for a bend is given by

\[
\text{Loss} = \alpha_{\text{rad}} = \alpha_{\text{rad}} \frac{\pi r}{2}.
\]

Note that although (9) shows a linear relationship between \( \text{Loss} \) and \( r \), the \( \alpha_{\text{rad}} \) term is also dependent on \( r \) with an \( e^{-r} \) relationship, which decays much faster. Therefore, the radiation loss is lower for a larger \( r \), which is verified in Fig. 9 with both analytical and full-wave simulation results. The theoretical radiation loss analysis at the bend is based on Fig. 8 by assuming the bending structure has infinite width in the \( w \) direction. In real case, the width is finite. Hence, the field distributions are not exactly same, which is responsible for the major discrepancy between theoretical and simulated results in Fig. 9.

**D. Mode conversion Loss**

The discussions so far seem leading to a conclusion that a waveguide with large \( w \), \( h \), and \( r \) has smaller bending loss. However, the large waveguide dimensions may introduce a competing loss mechanism by introducing mode conversion as shown in Fig. 7(b).

Mode conversion can lead to additional loss when \( w \) and \( h \) of the waveguide are large enough to allow multiple modes to exist at the operating frequency. The issue is exacerbated by the bending structure of the proposed interconnect scheme. Fig. 10 shows \( n_{\text{eff}} \) of several possible modes for a channel of \( w = 300 \mu m \) at various \( h \) values. For example, the \( E_{y11} \) mode may be converted to \( E_{y12} \) mode when \( h \) is larger \( 260 \mu m \) after bending.

Fig. 11 shows the bending loss, consisting of radiation loss and mode conversion loss, versus \( h \) with a fixed \( r \) of \( 300 \mu m \) and a fixed \( w \) of \( 300 \mu m \) at 200 GHz. Multi-mode wave ports are used in HFSS to extract the power conversion among the lowest 3 modes. When \( h \) is less than \( 500 \mu m \), radiation loss dominates. Smaller height leads to a larger portion of the waves leaking into the air and causes larger radiation loss as shown by the curve with up triangles. When \( h \) is larger than \( 500 \mu m \), higher order modes may be excited and propagate, causing increasing mode conversion loss as shown by the curve with down triangles. The total loss is plotted as the curve with squares. The minimum loss of 0.3 dB occurs around \( 500 \mu m \).

A contour map of the bending loss versus \( w \) and \( h \) is plotted in Fig. 12. The minimum bending loss occurs for \( h = 500 \mu m \) and \( w = 400 \mu m \).

**E. Figure-of-Merit**

As discussed in Section I, the performance of chip-to-chip interconnect is determined by the bandwidth density and energy efficiency. The bandwidth density is defined as bandwidth per cross section area,

\[
\rho_{BW} = \frac{BW}{\text{Area}}.
\]

A larger bandwidth density means a higher spatial utilization efficiency to enable a higher data rate transmission per unit space. To evaluate the performance of sub-THz interconnect channel, we propose a Figure-of-Merit, defined as

\[
\text{FoM} = \frac{\rho_{BW}}{\text{Channel Loss}} = \frac{\text{BW/Area}}{\text{Channel Loss}(r, h, w)},
\]

Fig. 9. Calculated and simulated bending loss with respect to \( r \) at 200 GHz with 200-\( \mu m \) \( w \) and 500-\( \mu m \) \( h \).

Fig. 10. Calculated \( n_{\text{eff}} \) versus frequency for the first three modes at 200 GHz with 300-\( \mu m \) \( w \).

Fig. 11. Simulated total bending loss versus \( h \) at 200 GHz with 300-\( \mu m \) \( w \) and 300-\( \mu m \) \( r \).
where channel loss depends on the $r$, $h$, and $w$ of the channels. Note that the dominant loss mechanism depends on the communication distance. The total channel loss is dominated by bending loss for short distance communications ($l < 10 \text{ mm}$) while material loss is the dominant factor for the longer ones. Assuming that the data bandwidth is 10 percent of the carrier frequency, the contour map of FoM versus $h$ and $w$ is plotted in Fig. 13. The highest FoM occurs around $h = 400 \mu\text{m}$ and $w = 400 \mu\text{m}$.

Higher operating frequency leads to better FoM. This is because at higher operating frequencies, the optimal waveguide dimensions, $h$, $w$, and $s$, are all inversely proportional to the operating frequency. Assuming a constant fractional bandwidth, higher operating frequency leads to the significant increasing of FoM as shown in Fig. 14.

### III. Coupling Structure

To transmit signals between IC chips and channels, coupling structures are needed with the requirements of high directivity, high radiation efficiency, and with the broadside radiation pattern. High directivity and high coupling efficiency aim to maximize the power transfer to the receiver through the interconnect channel. The reasons to choose broadside radiation patterns compared to end-fire radiation patterns are: a) better isolation due to non-direct interference from the reflected waves to the circuits in surrounding places; b) flexible channel location because the coupling structure can be located in the whole chip instead of just on the chip peripherals otherwise in end-fire radiation pattern cases.

In this work, a patch antenna based coupling structure is chosen due to the mature design method [33]. Fig. 15(a) illustrates the coupling structure. The signal is excited through a coplanar waveguide (CPW), and transitioned to a microstrip line before feeding the coupling structure. Rogers 3850, with 25.4-µm thickness and dielectric constant of 2.9, is chosen as the coupling structure substrate. To simplify the fabrication, a via-less CPW to microstrip line transition is adopted [36]. In order to prevent the energy leakage through the substrate, choosing the ground plane width smaller than half wavelength of the signal avoids the generation of parallel plate modes and high order modes [37]–[39]. The dimensions of the coupling structure are labeled with EM simulation results in Fig. 15. Also, both near field and far field patterns are presented. From these patterns, it can be seen that the signal effectively propagates upward to the perpendicular direction from the antenna.

Noted that the coupling structure exhibits a limited bandwidth whereas the bandwidth of the dielectric waveguide is very large as shown in Fig. 17. Future demonstrations of the sub-THz interconnect concept will focus on improving the bandwidth of the coupling structures.

### IV. Experimental Demonstration

#### A. Design of the Demonstration Setup

The simulation results of the complete sub-THz interconnect channel with a pair of patch antenna coupling structures and feeding structures are shown in Fig. 16. The waves are radiated by the coupling structure A, and then coupled to the waveguide. Propagating through the waveguide, the waves are collected by the coupling structure B. Fig. 16(a) also shows partial portions of EM waves leaking into air due to the bending structure causes coupling loss. Besides, a larger
beam width compared with waveguides cross section size also results in the finite power collection capability. The simulated $S_{21}$ as shown in Fig. 16(b) indicates the minimum insertion loss of 5.9 dB of the complete interconnect structure. The band-limited behavior $S_{21}$ is due to the limited bandwidth of the patch antenna; the silicon channel waveguide itself is very wideband as shown in Fig. 17.

B. Fabrication

The fabrication processes of the sub-THz interconnect channel are summarized in Fig. 18. A 500-µm thick HR silicon wafer (resistivity of 10000 Ω · cm) is first patterned with a thick (∼17 µm) photoresist (AZ9260) to define the waveguide geometries: $h$ and $r$. Then, the HR silicon wafer is attached to a carrier substrate and etched through in a DRIE process. The individual channels are isolated after etching. Fig. 19(a) shows the photographs of the channel from different perspectives.

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Fig. 15. Simulated $S_{11}$ for the patch antenna based channel coupling structure with insets of schematic, and field distribution at 210 GHz.

Fig. 16. (a) Simulated magnitude of E-field distribution of a complete sub-THz interconnect channel with a pair of channel feeding structures at 210 GHz, and (b) simulated $S_{21}$ with 300-µm $w$, 500-µm $h$, and 300-µm $r$.

Fig. 17. Simulated S-parameters versus frequency for the straight silicon channel waveguides without or with bending structure with 6-mm $l$, 300-µm $w$, 500-µm $h$, and 300-µm $r$.

Fig. 18. Fabrication procedure of (a) the silicon channel, and (b) the patch antenna.

Fig. 19. (a) Photographs of the channel from different perspective, and (b) SEM photograph of the patch antenna.
Note that the silicon waveguide is etching from side view to simplify the fabrication complexity as shown in Fig. 18(a)(3).

The coupling structure is fabricated on a Rogers 3850 substrate. Photolithographic thin film patterning is used to achieve a fine feature definition. One side of the copper laminate is completely removed first. The antenna structure is then patterned with a Ti/Au thin film of thickness 50/300 nm by a lift-off process. Fig. 19(b) shows the fabricated scanning electron microscope (SEM) photograph of the antenna coupling structure.

A 3D printed holder with a low dielectric constant material (Acrylate based polymer, $\varepsilon_r \approx 2.7$) is used to support the channel. According to the full-wave simulations, the holder introduces negligible effects on the signal propagation. The alignment of the channel and two coupling structures is very critical. To ensure good alignment, the channel holder and alignment marks are introduced. The channel holder is first attached to the two alignment marks, which should be put exactly between alignment marks; the holder’s slot, where the channel is to insert, is then in the center of alignment marks and holds the channel.

C. Measurement Results and Discussions

Fig. 20 and 21 illustrate the measurement setup, which is based on an Agilent network analyzer (PNA-X N5247A). A pair of Virginia Diodes frequency extension modules (VDI WR5.1-VNX) up-converts the signal frequency to G band (140–220 GHz). WR-5 waveguides are used to guide the wave toward the tip of the probes. SOLT (Short, Open, Load, Thru) calibration method is employed to set the reference plane at the edge of the patch antenna for each side. The measured interconnect path includes two patch antenna based coupling structures, and the sub-THz interconnect channel.

Fig. 22 shows the comparison of measurement results between the cases with and without the sub-THz interconnect channel, which indicates the insertion loss is significantly improved. Fig. 23 also shows the measured S-parameters comparing with the simulation results after using extracted material parameters, specifically the substrate loss tangent and metal effective conductivity. The minimum insertion loss is 8.4 dB, which is about 2.5 dB higher than the simulation result in Fig. 16(b) due to two major reasons. First, the real loss tangent of the substrate for the coupling structure is larger than their typical values at such high frequencies, with the extraction method described in the next paragraph. Second, titanium is used as an adhesion layer under the gold thin film. Because the skin depth of gold and titanium at 210 GHz are 172 nm and 807 nm, separately, the underlying titanium layer with 50 nm thickness can be penetrated completely. The effective conductivity of the Ti/Au thin-film is extracted to be $2.92 \times 10^7$ S/m.

The substrate loss tangent is extracted through fabricated transmission lines as shown in Fig. 23(a). After comparing the S$_{21}$ between measurement and simulation results, the
V. Conclusion

This paper for the first time presents the design, analysis and demonstration of a micromachined silicon dielectric waveguide based the sub-THz interconnect channel for silicon planar integrated circuits. A detailed analysis of channel losses, including radiation loss and mode conversion loss, channel size optimization for the bending, fabrication procedure, as well as the measurement setup have been conducted. To quantify the design optimization, an FoM is also defined. The analytical, simulated, and measured results agree well, demonstrating much lower loss than other electrical interconnect methods while maintaining better process compatibility than optical interconnect. In addition, this technique can be readily scaled up to THz frequencies due to a better FoM at higher frequencies. Therefore, the authors envision that THz interconnect has the potential to eventually solve the long-standing interconnect problems of intra-/inter-chip communications.

REFERENCES


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Qun Jane Gu is a recipient of NSF CAREER award and 2015 COE Outstanding Junior Faculty Award. She is a co-author of several best paper awards, including the Best Student Paper Award of 2010 IEEE Asia-Pacific Microwave Conference (APMC), the Best Paper Award of 2011 IEEE RFIT (RF Integrated Circuit Technology Conference), the Best Student Papers the Third Place of 2012 IEEE MTT-S International Microwave Symposium, and the Best Conference Paper Award of 2014 IEEE Wireless and Microwave Technology Conference (2014 WAMICON).