

On-chip microwave interferometry using phononic integrated circuits

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 (Received 20 February 2026; revised 30 April 2026; accepted 14 May 2026; published 8 June 2026)

Piezoelectric phononic integrated circuits (PnICs), chip-scale platforms in which GHz frequency acoustic fields are routed in μm -scale waveguide geometries, have steadily advanced in performance and complexity, providing new degrees of freedom for manipulating microwave signals. Here, we consider the use of PnICs for implementing on-chip microwave interferometers ($f = 3.4\text{ GHz}$) using a gallium nitride (GaN) on silicon carbide (SiC) PnIC platform. We demonstrate controlled mode amplitude splitting, differential control (local phase shift), and mode recombination in an unbalanced ($\Delta L = 1.6\text{ mm}$) phononic Mach-Zehnder interferometer (MZI), achieving a maximum fringe visibility >0.985 . The high visibility does not imply high sensitivity due to the presence of parasitic acoustic modes in the MZI. We quantify the effect of these parasitic guided modes in the interferometer, and discuss potential design improvements for achieving quasi-single-mode operation in PnICs, with a view toward enabling applications in microwave and acoustic signal processing and sensing.

DOI: [10.1103/96ns-kvtd](https://doi.org/10.1103/96ns-kvtd)

I. INTRODUCTION

Piezoelectric phononic integrated circuits (PnICs) [1], like their more established photonic counterparts [2], aim to manipulate GHz-frequency acoustic fields in wave-length (μm)-scale waveguide geometries. While the ideas around acoustic waveguiding are not new [3], a combination of material advances (in particular, heteroepitaxial growth [4] and wafer bonding techniques [5]) with progress in nanofabrication and numerical simulation methods [6] has made it feasible to implement device geometries that can potentially achieve a performance advantage over existing approaches. Given that the acoustic field in these devices is usually generated from a microwave signal of interest, PnIC platforms naturally provide new degrees of freedom for controlling microwaves on chip. In particular, the combination of strong confinement [7] and low propagation loss [8], with the $\approx 10^5 \times$ reduction in propagation velocity (and wavelength) by working with acoustic fields, enables one to significantly increase the time window over which the signal can be manipulated on chip [9]. For instance, using compact spiral delay lines [8], one can achieve microwave signal delays

exceeding $2.5\text{ }\mu\text{s}$ in a footprint $<0.25\text{ mm}^2$. These long on-chip signal storage times enable one to effectively magnify weak perturbative effects, with resulting implications for both signal processing [9] and sensing.

Here, we demonstrate a GHz-frequency microwave interferometer designed around a gallium nitride (GaN) on silicon carbide (SiC) PnIC platform [8,10], which serves as a prototypical building block for both sensing and signal processing applications. An interferometer splits an input field into two different paths, which are then recombined downstream. The magnitude of the recombined field is exquisitely sensitive to any differential phase accumulated by the field as it traverses the two paths, and it can be made insensitive to common-mode noise sources such as pump intensity and temperature fluctuations [11]. This makes interferometry the standard for implementing phase-sensitive detection [12], best exemplified by LIGO's historic detection of gravitational waves emitted by a binary black hole merger [13].

An alternative way to view an interferometer is to think of it as a signal transducer wherein the signal of interest perturbs the phase of the propagating field and the interferometer maps that phase change into a corresponding output power change. In this view, the LIGO interferometer converts gravitational-wave-induced differential cavity length fluctuation to an output optical power fluctuation [14]. This idea can be extended to other signals of potential interest and underlies the use of interferometers as sensors.

Chip-scale platforms benefit from an inherent phase stability advantage compared with bulk systems because the

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compact footprint effectively renders most external noise sources as *common-mode*, to which the interferometer is insensitive. In addition, they leverage traditional electronics foundry manufacturing methods to facilitate efficient interfaces to signal conditioning and readout electronics at scale. As a case in point, integrated photonic Mach-Zehnder interferometers (MZIs) [2] currently underpin optical approaches to both classical [15,16] and quantum computing [17], as they provide a natural route toward calculating the underlying matrix-vector products.

Implementing microwave interferometers, instead of the standard mixer-based phase detector approach [18], has the potential to improve the overall sensitivity (by reducing the noise floor and increasing conversion gain) of amplitude noise, and particularly phase noise measurements [19,20] of microwave signals. In addition, interferometric approaches have shown a sensitivity advantage for sensing applications, in particular electron spin resonance detection of nanoscale samples [21,22]. While these interferometers are traditionally implemented purely using electromagnetic components, here we consider their implementation using acoustic fields as an intermediary in PnICs. As noted above, the $\approx 10^5 \times$ reduction in the speed of sound makes acoustic implementations the only viable solution wherein long path lengths (signal delays > 50 ns) are needed in a chip-scale platform. The effective phase change ($\Delta\phi$) scales with the path length L as $\Delta\phi \approx \Delta\beta L$ where $\Delta\beta$ is the change in the wave vector of the propagating mode due to a perturbation. Therefore, detecting small frequency and phase fluctuations, as occurs frequently in precision measurements or characterizing stable oscillator sources, comes down to having long path lengths and maximizing the signal observation time on chip. On the sensing front, due to the vastly reduced mode volumes of acoustic fields, the same perturbing fields can induce larger perturbations (larger coupling strength leading to a proportionally larger phase change $\Delta\beta$) in PnIC implementations [23]. The underlying acoustic wave interferometers can also be applied to traditional acoustic wave sensor modalities [24], as they provide a natural way to do differential measurements that are common-mode noise insensitive.

To realize these potential benefits, it is important to understand the fidelity with which microwave interferometry can be implemented using PnICs. The main issue here is that standard acoustic waveguide geometries are inherently multimode, unlike their photonic counterparts, and high visibility interference requires maintaining effectively single-mode operation through the mode generation, injection [25], splitting [26,27], routing [28–32], dynamic control [33–36], recombination, and detection stages. While our ability to control and manipulate GHz-frequency acoustic fields in piezoelectric PnIC platforms has steadily advanced [37–39], understanding how well quasi single-mode operation can be achieved in complex

PnIC circuits remains an open question. This requires a detailed understanding of the nature of the parasitic modes that one can potentially excite in PnIC circuits, the role of transducers and interfaces in mediating this parasitic mode excitation, and their overall impact on the measured interference fidelity. We note that by fidelity, we are referring to an estimate of the purity of single-mode operation in a PnIC interferometer, or equally valid, an estimate of the power in parasitic modes.

Here, we attempt to address these issues by studying microwave interferometry implemented using on-chip tunable unbalanced phononic MZIs. While we focus on the GaN-on-SiC platform here, the general ideas are applicable to other PnIC platforms under consideration, such as lithium niobate on sapphire [29] and scandium aluminum nitride on SiC [30,31,36].

II. DEVICE DESIGN AND CHARACTERIZATION

Figure 1(a) shows a representative microscope image of a fabricated tunable unbalanced acoustic wave MZI.

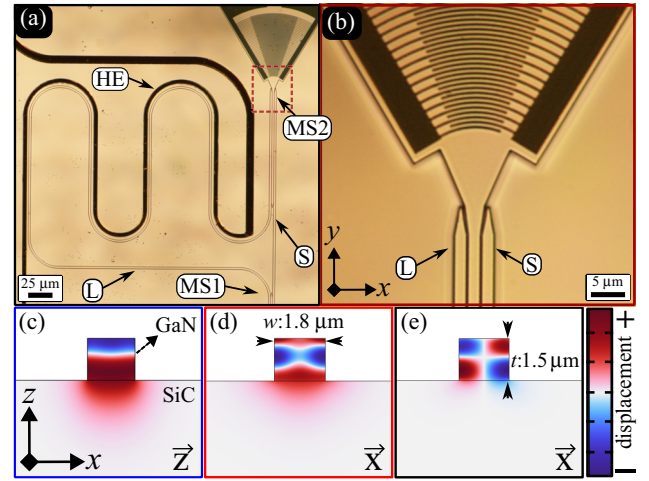


FIG. 1. (a) Microscope image of a representative tunable unbalanced acoustic wave MZI showing the long (L) and short (S) arms. The acoustic field at 3.4 GHz is generated from a microwave signal using a focusing interdigitated transducer (IDT) that also serves to focus the generated sound wave into the waveguide. The field is then split (unequally, 75:25) into the L and S arms using directional couplers (MS1) to account for the differential propagation loss in the two arms. An enlargement of the receive port coupler is shown in (b). At the receiver end, the acoustic waves in the two arms are recombined using a 50:50 directional coupler arrangement (MS2) and the field is electrically read out using a nominally identical receiving IDT. The resistive heating element (HE) for tuning the acoustic velocity along the L arm is shown by the dark trace. (c)–(e) Propagating modes in the waveguide geometry, as calculated by a finite-element method (FEM) solver with the relevant displacement component indicated: (c) Lamb-like mode of interest ($\bar{z}$), (d) Love-like parasitic mode ($\bar{x}$), (e) higher-order Love-like mode ($\bar{x}$).

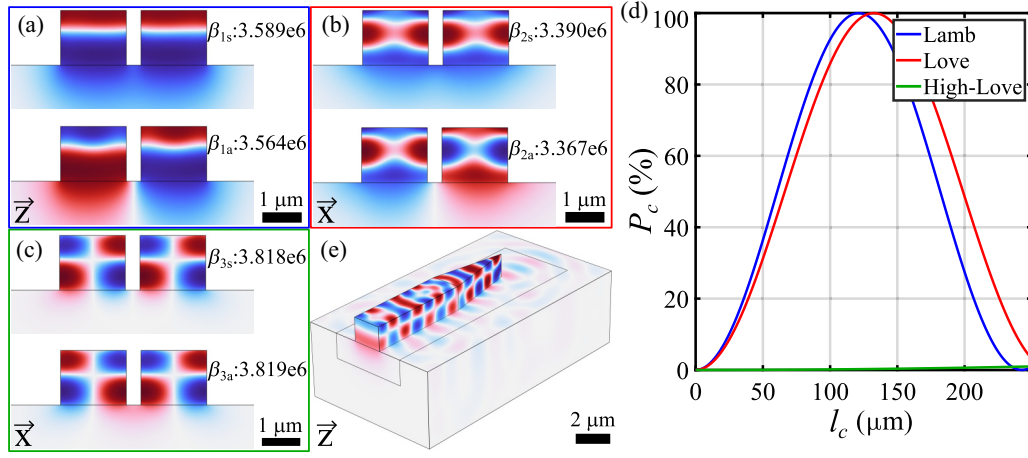


FIG. 2. Efficacy of directional couplers as parasitic mode filters. (a)–(c) Even (symmetric, β_s) and odd (antisymmetric, β_a) supermodes formed in the waveguide for the Lamb-like [Fig. 1(c)] and Love-like [Figs. 1(d) and 1(e)] modes in the GaN-on-SiC waveguide platform. The difference between the modes' propagation constants ($\Delta\beta$) is inversely related to the power coupling length, plotted in (d). Here, P_c indicates the fraction of the power propagating in waveguide 1 in a given mode that couples into waveguide 2 over a coupling length l_c . The directional coupler filters out the higher-order Love-like mode effectively, but the fundamental Love-like mode has almost the same coupling efficiency as the Lamb-like mode of interest. (e) The waveguides are terminated by tapers of length $2\lambda_a$, which leak the mode into the substrate and minimize backreflection into the waveguide.

The fabrication procedure is identical to our previous work [8,10] with the addition of an extra aligned lithography step to pattern the heater electrode should be (indicated by the dark trace in Fig. 1(a)). Acoustic fields at 3.4 GHz are launched from a microwave signal using focusing interdigitated transducers (FIDTs), which also serve to focus the generated field down to the waveguide width ($w = 1.8 \mu\text{m}$). The GaN layer thickness (t) is $1.5 \mu\text{m}$.

The acoustic field in the entrance waveguide is then split 75:25 into long (L) and short (S) arms using a directional coupler, shown in Fig. 1(b). The asymmetric split between the L and S arms is implemented to ensure equal received power from the two arms, accounting for differential propagation loss; see Appendix D. At the receiver end, the acoustic fields from the L and S arms are recombined into a central waveguide using a similar splitter, and then electrically read out by a nominally identical FIDT. Figures 1(c)–1(e) show the three prominent propagating waveguide modes in our geometry. Figure 1(c), which we refer to as a Lamb-like mode on account of its predominant displacement being along $\vec{z}$, is the main mode of interest, as ideally this is the mode that is most efficiently launched by the FIDT. Figures 1(d) and 1(e) show parasitic Love-like modes, which also propagate in this waveguide geometry at the same frequency. We refer to them as Love-like because their predominant displacement is along $\vec{x}$. The fundamental Love-like mode [Fig. 1(d)] is the main parasitic mode of interest here as its propagation characteristics are very similar to the Lamb-like mode [cf. Fig. 2(d)]. We note that the simulations are performed using

waveguide geometries with vertical sidewalls. While we have not measured the sidewall angle on these devices, we expect the sidewalls to be near vertical ($< 10^\circ$) based on the etch chemistry used.

Nominally, the Love-like modes should not be efficiently launched in unbounded geometries using FIDTs. But due to a confluence of reasons, these parasitic modes are launched quite efficiently in our experiments, and corrupt the overall spatial mode profile in the central waveguide, which interfaces to the FIDTs [40]. A key element of our design, therefore, is to ensure that there is no direct acoustic transmission link between the transmit and receive FIDTs, and the L and S arms are both coupled to from the central waveguide via directional couplers. The residual field in the central waveguide is then leaked into the substrate via tapers as shown in Fig. 2(e) and avoids backreflections into the waveguide.

To work out the efficiency of directional couplers as mode filters, we can calculate the beat length (L_b) for the different modes of interest for a fixed coupler geometry as shown in Fig. 2. Figures 2(a)–2(c) show the propagation constants of the even (symmetric, β_s) and odd (antisymmetric, β_a) supermodes whose difference ($\Delta\beta$) is related to L_b as $L_b = \pi/\Delta\beta$. The beat length here refers to the length of the coupling region required to transfer all the power originating in one waveguide to the other. For GaN-on-SiC waveguide geometries with $w = 1.8 \mu\text{m}$, $t = 1.5 \mu\text{m}$, and the gap (g) between waveguides of 400 nm , we estimate the beat lengths for the three modes shown in Figs. 1(c)–1(e) to be $121 \mu\text{m}$, $133 \mu\text{m}$, and $3927 \mu\text{m}$,

respectively. From the beat length data, we can see that the directional coupler serves to effectively filter the higher-order Love-like mode, but it cannot effectively separate the fundamental Love-like mode from the Lamb-like mode. It is important to note that due to acoustic anisotropy, the behavior of these couplers is dependent on the in-plane propagation direction. The data shown in Fig. 2 correspond to a directional coupler oriented along the x - z plane with the mode propagating along y . The axes correspond to the material orientation used while setting up the finite-element modeling (FEM) simulation in COMSOL. The modal behavior, when oriented along the complementary y - z plane with mode propagation along x , is shown in Fig. 10 in Appendix B.

III. CHARACTERIZING PHONONIC MZIS

Figure 3(a) shows representative transmission (S_{21}) spectra, offset for clarity by -30 dB, for phononic MZI with varying path length differences ΔL of $500\ \mu\text{m}$, $1027\ \mu\text{m}$, and $1621\ \mu\text{m}$, respectively. The spectra have been time gated with a temporal bandpass filter (Kaiser window) over $40\ \text{ns}$ - $2\ \mu\text{s}$ [40,41] to remove the direct electromagnetic crosstalk and increase signal-to-noise ratio (SNR) for the received acoustic signal. The overall acoustic signal transmission is limited in these experiments due to a mismatch between the transmit and receive IDT frequencies, see Appendix A Fig. 7. The transmission spectra in Fig. 3(a) clearly show periodic modulation of the transmitted signal power, with the average modulation period (equivalently the free spectral range (FSR) shown in the inset of Fig. 3(b)) scaling $\propto 1/\Delta L$. Unlike traditional photonic and microwave interferometers, the spectral transmission of phononic MZIs is significantly influenced by the dispersive behavior of the transmit and receive IDTs. To first order, we can remove the effect of the IDTs by normalizing the measured gated S_{21} by the raw reflection (S_{11}) spectra of the individual IDTs, and work with a scaled transmission coefficient $S_{21,s} = S_{21}/(1 - S_{11})(1 - S_{22})$.

Figure 3(b) shows the normalized transmission ($S_{21,n} = S_{21,s}/\max(S_{21,s})$), which is used to estimate the fringe visibility. From the figure, we can estimate the maximum observed fringe visibility ($V = (1 - \min(S_{21,n}))/ (1 + \min(S_{21,n}))$) as 0.985 , where $\min(S_{21,n})$ refers to the normalized measured transmission minimum closest to the maxima, indicated by the small red square in Fig. 3(b). As we move away from the central peak, the fringe visibility falls off rapidly. We note that the fringe visibility reported is very close to that estimated from the gated S_{21} spectrum without any scaling or normalization.

For phononic MZIs to be used in precision microwave interferometry, it is critical to understand the effect of parasitic acoustic modes as they place a limit on the achievable fringe visibility and corrupt the overall transmission spectrum. The overall interferometer response can

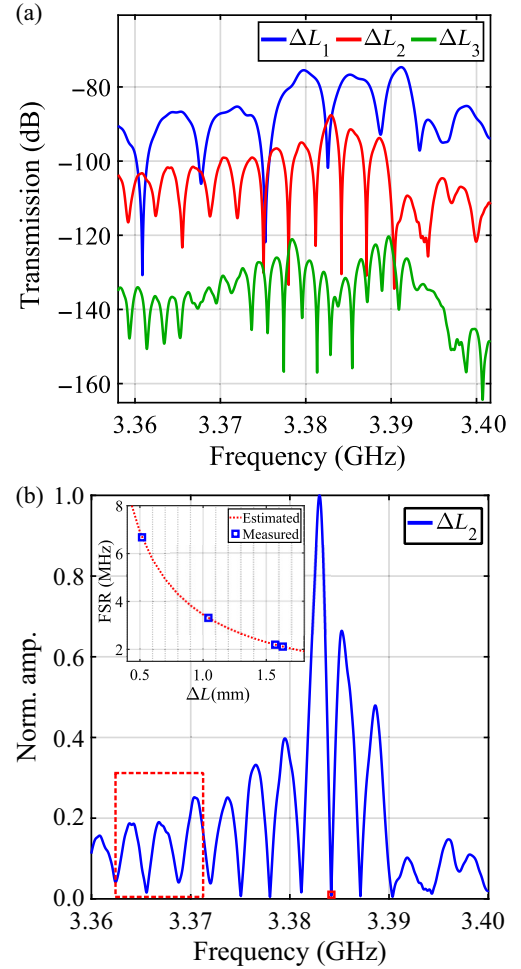


FIG. 3. (a) Time-gated transmission (S_{21}) spectra of unbalanced phononic MZI devices with different path-length differences ($\Delta L_1 = 500\ \mu\text{m}$, $\Delta L_2 = 1027\ \mu\text{m}$, and $\Delta L_3 = 1621\ \mu\text{m}$). For visual clarity, the spectra for ΔL_2 and ΔL_3 are vertically offset by -30 dB and -60 dB, respectively. (b) Normalized S_{21} amplitude spectrum for the ΔL_2 MZI device, plotted to remove the spectral characteristics of the transmit and receive IDTs. The red dashed box shows the effect of the parasitic mode as an unwanted broadening of the MZI transmission. The red square shows the minima used to estimate the maximum observed interference visibility. The inset compares the average FSR measured and theoretically estimated free spectral range (FSR) of the MZIs as function of the path difference ΔL .

be modeled by accounting for the two modes as discussed in Appendix B. We would like to note that while the modes are orthogonal and intermodal acoustic wave interference is negligible, the generated voltage signals are phase coherent and therefore lead to intermodal cross terms (see Appendix B) that distort the MZI response, as can be seen in the MZI transmission. While the average dip spacing [FSR, plotted in the inset of Fig. 3(b)] is consistent with the $1/\Delta L$ scaling, the red dashed boxed regions in Fig. 3(b) highlight the excess broadening of the maxima in the MZI

transmission in certain frequency ranges, clearly indicating the presence of parasitic modes. Because of the presence of these intermodal cross terms and their dependence on the specific MZI design as detailed in Appendix B, the region of maximum fringe visibility becomes hard to predict. Moreover, measurement of high fringe visibility here in and of itself does not indicate high measurement sensitivity as in a traditional single-mode MZI, but shows that these parasitic modes remain phase coherent as they propagate through the MZI.

One can directly visualize the parasitic mode in the time domain by Fourier transforming the spectral data, as shown in Fig. 4(a) and using the vector network analyzer as a time-domain reflectometer [40]. By knowing the geometrical path lengths and the respective mode group velocities, one can delineate the various routes by which signals make their way from the transmit to the receive FIDTs, as shown schematically in Fig. 4(b). The signals propagating on the short arm are designated S_i , whereas those taking the longer arm L_i , where the i identifies the particular mode (Lamb-like $\Rightarrow$ 1 or Love-like $\Rightarrow$ 2). tt^m refer to the different triplet transit signals that take the routes as shown in the schematic Fig. 4(b). A triplet transit, borrowing terminology from surface acoustic waves devices [9], here refers to signals that take longer times in their transit from transmit to receive ports via reflections along the way. For instance, a signal could reflect off the receiving IDT, and then get detected after a secondary reflection as shown in Fig. 4(b). Given the small difference between the propagation constants of the two modes, one needs long path lengths for the two modes to show up as distinct events at the receiver IDT. This is best seen by comparing the long-arm L signal for the shorter MZI (ΔL_1 , blue) with the longest one (ΔL_3 , red) wherein the L signal clearly splits into two distinct pulses (L_1, L_2) corresponding to the two different modes. Based on the simulated group velocities of the two modes, we can identify them as indicated in the figure, where L_1 is the Lamb-like mode and L_2 is the Love-like mode. To get a rough estimate of the length needed to resolve the parasitic mode, we start with a given pulse width (Δt) set by the vector network analyzer (VNA) acquisition parameters (bandwidth and number of frequency points) and the temporal dispersion introduced by propagation and the IDT spectral characteristics. Assuming we need a pulse separation $\approx \Delta t$ to consider two pulses to be considered resolved, for modes with an average velocity v_g and difference Δv_g , one requires path lengths (L) on the order of $L = \Delta t(v_g^2/\Delta v_g)$. For the modes of interest here, with an average group velocity of 3542 m s^{-1} , a temporal pulse width of 30 ns and a velocity difference of $\approx 280 \text{ m s}^{-1}$ between the two modes, this works out to $L \approx 1.3 \text{ mm}$. We would like to reiterate that the identification of modes with the temporal pulses in Fig. 4(a) is done based primarily on the estimated velocity difference from FEM simulations,

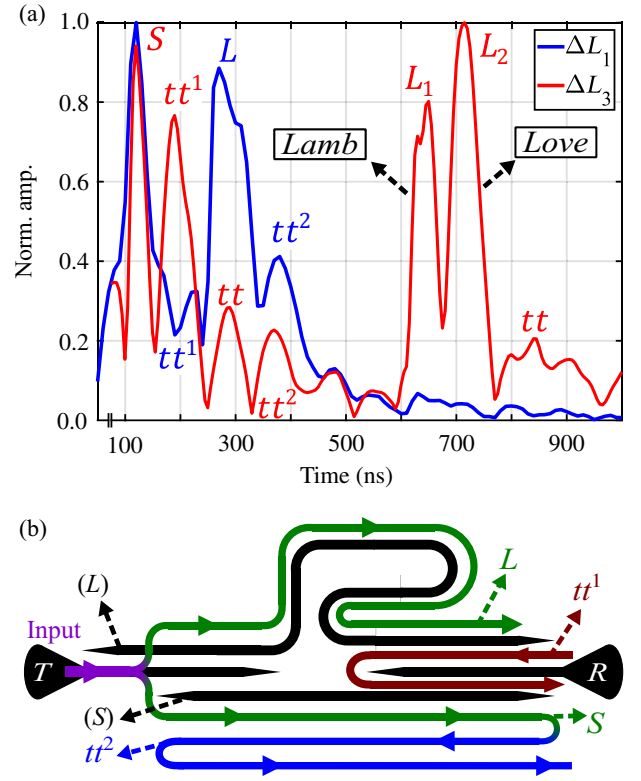


FIG. 4. (a) Time-domain transmission response of two phononic MZIs ($\Delta L_1 = 0.5 \text{ mm}$ (blue) and $\Delta L_3 = 1.6 \text{ mm}$ (red)) obtained by Fourier transforming the spectral data [41]. The different pulse labels correspond to the different paths taken from source to receive IDTs, as shown schematically in (b). The direct paths are labeled by d and the triplet transits by tt . Indices 1, 2 appearing in subscripts refer to modes (Lamb-like or Love-like), and those in superscripts refer to the paths taken by the tt pulses. The ΔL_m notation refers to the path-length difference in the MZI as in Fig. 3(b). The temporal splitting of the d_2 pulse in the long arm into two distinct Lamb-like and Love-like pulses is indicated. The mode designation is based on the relative group velocity inferred from the FEM simulation.

but agrees reasonably well across multiple devices with different long-arm lengths (L_L). Without field mapping, one would ideally need to measure this temporal separation at one additional point along the circuit to make the assignment unambiguous.

Based on the time-domain data, it is clear that the parasitic Love-like mode is excited quite efficiently, or more importantly, filtered insufficiently in these devices; see discussion at the end of Appendix B. While we do not definitively understand its origin, we believe it stems from a combination of factors. First, for fabrication reasons (mainly inertness of gold (Au) to most wet etching and cleaning chemicals), we chose to work with chromium (Cr)/Au electrodes, which results in significant acoustic wave scattering due to the mass loading of the Au electrodes [42]. This is further compounded by the choice

of focusing IDTs, wherein transverse mode excitation via busbars [43] is known to introduce shear displacements that couple efficiently to the Love-like mode. Finally, the waveguide-taper interface [1] is not optimized from a mode matching perspective, which leads to higher-order mode excitation in the waveguide. It is important to note that optimizing mode matching for acoustic fields is significantly more complex than for analogous optical fields due to the presence of additional slow surface modes.

While suppression of the Love-like mode excitation directly in the entrance waveguide is difficult without an extensive redesign of the focusing IDT and waveguide-taper interface, one can instead attempt to filter the parasitic mode out more efficiently using directional couplers that have inherently higher mode selectivity. This can be achieved in principle by exploiting the waveguide geometry. Reducing waveguide widths below 800 nm effectively renders the waveguide single mode, as the Love-like mode is cut-off below this width. The trade-off with moving to narrow-width devices is that the width $\ll \lambda_a$, which reduces the coupling efficiency. In addition, the aspect ratio makes it challenging to fabricate long (mm-scale) devices with high yield. An alternative approach might be to use the narrow-width waveguide in an asymmetric directional coupler as a mode filter, as shown in Fig. 5, wherein both the Lamb-like and Love-like modes from the entrance waveguide couple into a Lamb-like mode in the narrower waveguide, which can then be tapered back to normal width. The main trade-off with this approach is that if the coupling lengths of the input Lamb-like and Love-like waves are very different, then one achieves higher mode purity at the output at the cost of a higher insertion loss (IL) penalty. Implementing such designs would enable quasi-single-mode operation in phononic MZIs, but their operation needs to be verified in practice.

We would like to note that we have kept the GaN layer thickness fixed at $1.5\ \mu\text{m}$ in these discussions in order to be compatible with the standard GaN thicknesses used in current GaN high electron mobility transistor (HEMT) foundry processes with a long-term view toward monolithic integration.

IV. DYNAMIC CONTROL: THERMAL TUNING AND RESIDUAL CROSSTALK

To exert dynamic control on propagating acoustic fields with a view toward tunable microwave interferometry, we add a resistive heating element (HE) in close proximity to the long arm of the interferometer, as shown by the dark HE trace in Fig. 1(a). Figure 6 shows the change in the MZI transmission as a function of power delivered to the heater. To calibrate the frequency shift associated with temperature, the chip is first uniformly heated using a Peltier module, and the resulting frequency shift of the MZI fringe is measured as a function of sample temperature.

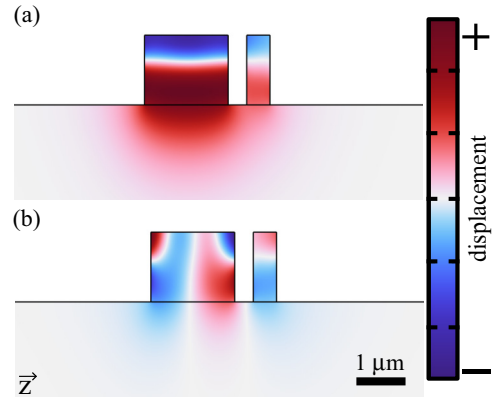


FIG. 5. Using asymmetric directional couplers to improve parasitic mode filtering. Reducing waveguide widths below 800 nm makes the waveguide effectively single mode, but makes it difficult to couple into the mode effectively by requiring diffraction-limited focusing transducers ($w \approx \lambda_a/2$) and affecting device yield due to high aspect ratio. Using the narrow waveguide as part of a directional coupler enables both the Lamb-like and Love-like modes in the entrance waveguide to couple to the Lamb-like mode in the narrow waveguide.

As the chip temperature is increased from 18°C to 100°C , we observe a frequency shift of $22.8\ \text{ppm}/^\circ\text{C}$, which is consistent with previously reported values for GaN devices [44]. Using this calibration, the change in

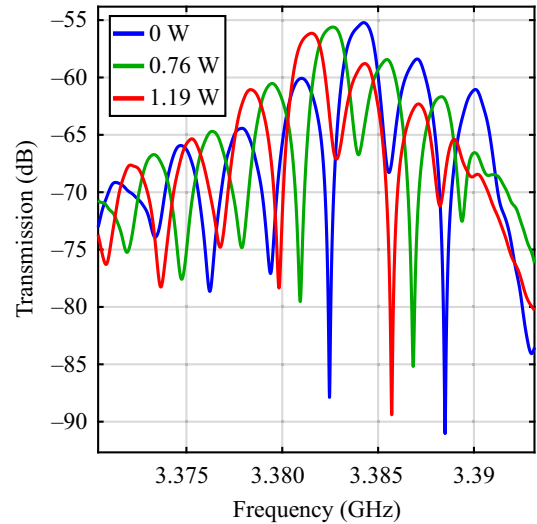


FIG. 6. Dynamic control of phononic MZIs. A heater element meandering along the long (L) arm of the phononic MZI [cf. Fig. 1(a)] shifts the acoustic velocity along the L arm, which results in a shift in the MZI transmission that scales with local temperature, here represented by the power supplied to the heating element. We estimate a frequency shift of $22.8\ \text{ppm}/^\circ\text{C}$. The change in transmission amplitude is attributed to thermal crosstalk, which affects the IDTs; see text for details.

group velocity, Δv_g , is extracted as a function of temperature, yielding the temperature-dependent relation $v_g(T)$; see data in Appendix C. This relation is then used to quantify the group-velocity shift in the heated long arm of the MZI. We assume that the section of the long arm whose temperature is modified via heating corresponds exactly to the heater length. This is not strictly correct, but is a reasonable approximation given the fraction of the long arm that is covered by the heating element in our case. We also assume that the temperature change in the S arm of the MZI is negligible. In the absence of local temperature measurements, this is the only assumption we can make, but as we show in the Supplemental Material (Sec. E), the high thermal conductivity of SiC makes the thermal crosstalk between the MZI arms significant.

From the chip-scale heating data, we obtain a temperature coefficient of the group velocity of $\Delta v_g/\Delta T = -0.067 \text{ ms}^{-1} \text{ }^\circ\text{C}^{-1}$. By fitting the heater-induced frequency shift Δf to the extracted $v_g(T)$ dependence via $\Delta v_g/v_g = (\Delta f/f)(\Delta L/L_h)$, we determine group-velocity changes of -1.92 m s^{-1} and -3.32 m s^{-1} for heater input powers of 0.76 W and 1.19 W, respectively, which correspond to temperature changes of 29.2°C and 50.1°C . Here, ΔL corresponds to the path difference between the MZI arms, and L_h is the length of the long arm that is heated. It is important to remember that the MZI has two modes propagating in the arms that can have very different temperature dependences, so the extracted temperature coefficients by the procedure reported above should be interpreted as representing a weighted average.

While the spectral shift in the MZI response with temperature is clear, we also observe a change in the transmission amplitude. We believe this occurs due to unintended thermal crosstalk between the HE and the IDTs, where the IDT response is spectrally shifted due to the residual temperature change. This residual crosstalk also highlights one of the critical differences between implementing thermal tuning in phononic MZIs and their photonic counterparts. Unlike in photonic implementations, wherein the HE can be placed directly on top of the waveguide (suitably far away to avoid optical absorption and scattering), the only viable approach in phononic MZIs is heating from the side because there is no easy way using dielectric interlayers to prevent the acoustic field from leaking out of the waveguide core. This side heating makes crosstalk control more challenging, especially given the very high thermal conductivity of the fast substrates (sapphire, silicon carbide, and diamond) usually required for implementing PnICs; see the Supplemental Material (Sec. E) for additional details.

The effect of this thermal crosstalk becomes increasingly important as these PnIC circuits start to scale in complexity and the component density increases in parallel. Moving forward, therefore, alternative methods for exerting dynamic control, such as using the piezoelectric

effect to induce geometry deformations [35] or exploiting acoustoelectric interactions in waveguide geometries [1], become attractive even if the induced phase change is weaker, provided that crosstalk can be minimized.

V. CONCLUSIONS

We have shown that phononic MZIs can be used to perform tunable microwave interferometry on chip. Given that acoustic waveguides are inherently multimoded, sensitivity is determined primarily by the ability to suppress parasitic mode excitation in the interferometer. We have shown one route toward mode suppression by exploiting the mode selectivity of directional couplers, but the performance needs to be significantly improved for precision microwave interferometry. One issue we have not considered in this work is the overall IL that such phononic MZIs present in comparison to more traditional microwave implementations. Addressing this excess IL is critical if PnIC-based implementations of microwave circuitry are to displace their traditional counterparts. A key open question [1] remains the ultimate achievable efficiency with which GHz frequency acoustic fields can be focused into and out of wavelength (μm)-scale waveguide geometries, while preserving high mode fidelity.

ACKNOWLEDGMENTS

We would like to acknowledge funding support from the UK's Engineering and Physical Sciences Research Council (EP/V005286/1, EP/N015126/1, EP/Z000688/1). We thank Elnaz Shokati, Laurent Kling, Robert Thomas, James Cockburn, Martin Cryan, Martin Kuball, Bruce Drinkwater, John Haine, and Mark Beach for discussions and suggestions.

DATA AVAILABILITY

The data that support the findings of this article will be made available once the current research plan is completed, expected by August 2026. More information and data are available from the authors upon reasonable request.

APPENDIX A: SPECTRAL MISMATCH BETWEEN TRANSMIT AND RECEIVE FIDTS

As discussed in the main text, the overall transmission through the MZI is limited to $\approx -75 \text{ dB}$, after removal of the electromagnetic crosstalk via time gating. The main source of this increased IL is the mismatch between the spectral response of nominally identical FIDT devices [10]. Figure 7 shows the rf reflection spectra (S_{11} and S_{22}) for the ΔL_1 and ΔL_3 phononic MZI devices reported in Fig. 3. To compensate for the weak k_{eff}^2 of GaN, we designed narrow-bandwidth IDTs that had higher

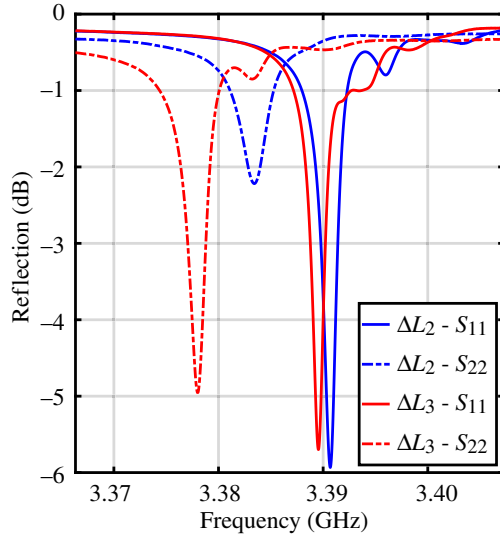


FIG. 7. RF reflection spectra of the transmit (S_{11} , solid) and receive (S_{22} , dashed) FIDTs for the phononic MZI devices reported in the main text Fig. 3: ΔL_2 (Blue) and the ΔL_3 (Red). The mismatch between the spectra is the main cause of the increased insertion loss (IL) observed in our MZI transmission measurements.

microwave-to-acoustic conversion efficiency at a given frequency. However, this design trade-off meant that the devices were highly susceptible to fabrication variations. As can be seen, our S_{11} and S_{22} spectra are offset by > 10 MHz, which is $\approx 5 \times$ larger than the IDT bandwidth.

We believe the mismatch between the IDTs [10] primarily occurs due to a combination of charging effects and a possible mismatch of FIDT orientation with respect to the crystal axes of SiC. While the IL is definitely critical and needs to be addressed from an applications perspective, it does not directly affect any of the main results of the paper.

Due to the random nature of the fabrication variations, there are devices with their transmit and receive IDT spectra better aligned. A reference spectrum is shown in Fig. 8, where now the gated S_{21} peaks at ≈ -45 dB, an improvement of ≈ 30 dB over the devices reported in the main text. The devices reported in the main text were chosen mainly because they formed a complete dataset of ΔL .

APPENDIX B: EFFECT OF PARASITIC MODES ON MZI TRANSMISSION

The presence of a parasitic mode on the MZI transmission can be estimated as follows. We assume that at a given frequency ω , we excite two modes with propagation constants β_1 and β_2 in the MZI with $\Delta\beta \ll (\beta_1 + \beta_2)/2$. Representing the signal traveling on the short and long arms by L and S, and using 1 and 2 to represent the modes, the complex voltage signal at the receiver (V_{out}) can be

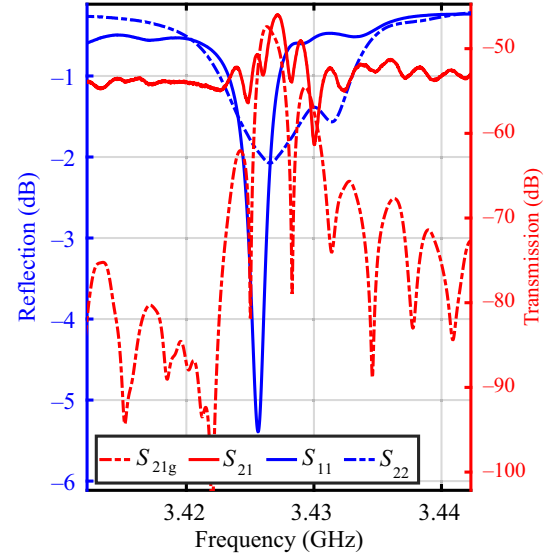


FIG. 8. Raw RF reflection (S_{11} , solid blue, S_{22} , dashed blue), and raw and gated transmission (S_{21} , solid red, S_{21g} , dashed red) spectra of the phononic MZI devices of ΔL_2 of a different set fabricated on the same run. The transmission spectra are gated with a temporal bandpass filter (Kaiser window) applied between 40 ns and 1.5 μ s [41] to remove the electromagnetic crosstalk.

represented as

$$V_{\text{out}} = V_{S1}e^{j\beta_1 L_S} + V_{L1}e^{j\beta_1 L_L} + V_{S2}e^{j\beta_2 L_S} + V_{L2}e^{j\beta_2 L_L}, \quad (\text{B1})$$

where $V_{S,i}$ and $V_{L,i}$ are the voltages launched into the short and long arms in the two modes at the first mode splitter. The detected voltage amplitude $|V_{\text{out}}| = \sqrt{V_{\text{out}}^* V_{\text{out}}}$ and the path-length difference $\Delta L = L_L - L_S$.

It is clear from expanding Eq. (B1) that in addition to the single-mode interference terms ($e^{\pm j\beta_{1,2}\Delta L}$), one gets additional cross terms between the two modes of the form $e^{\pm j\Delta\beta_{1,2}L_L}$ and $e^{\pm j\beta_1 L_L - \beta_2 L_S}$.

Figure 9 shows this parasitic modal interference captured with a simple model. The model employs a Gaussian pulse profile for both waves (in the long and short arms of the MZI) with an equal amplitude and a bandwidth of 10 MHz. The excitation is centered at 3.4 GHz and spans a frequency range of 100 MHz. The propagation constants for the two modes, β_1 and β_2 , are obtained from FEM simulations as a function of frequency, ranging from 3.35 GHz to 3.45 GHz. Figure 9(a) shows the interference pattern generated at the receiver if the two modes were acting alone, which corresponds to the standard single-mode MZI case. Figure 9(b) shows the cumulative effect of the two modes showing the characteristic beating pattern. The simulations correspond here to the case of ΔL_2 devices reported in the main text. As noted above, the presence of these intermodal cross terms and their dependence on the specifics of the MZI design make it challenging to

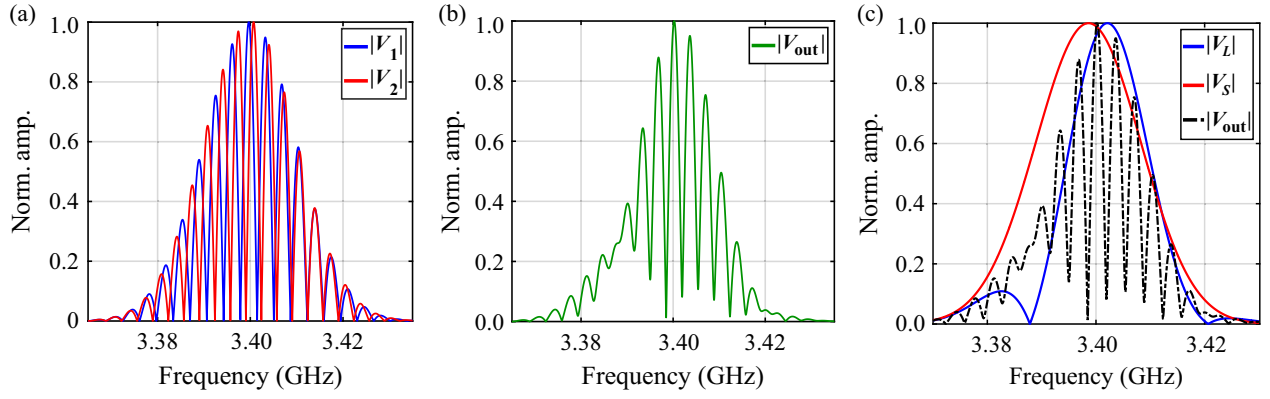


FIG. 9. Effect of parasitic modes on the simplified model MZI transmission analyzed for the ΔL_2 device reported in the main text, [cf. Fig. 2(b)]. (a) If the two modes (Lamb-like and Love-like) act independently, they each produce an interference pattern at the receiver that corresponds to the standard single-mode MZI response. (b) With the two modes present simultaneously, one now sees intermodal interference and a characteristic beat pattern in the MZI transmission. The interference between modes occurs only in the voltage domain as the signals are phase coherent. It is important to see that even with two modes present, one can still observe high visibility interference in a narrow spectral region. (c) To understand the contributions due to the different cross terms, we can plot the normalized output amplitude of the short arm ($|V_S|$, red, $V_{L,i} = 0$) and long arm ($|V_L|$, blue, $V_{S,i} = 0$) contributions separately, overlaid with the model output. The long-arm response exhibits a distorted Gaussian profile, suggesting the presence of two guided modes that accumulate relative phase over the longer propagation length. No such distortion is observed in the short arm, as its length is insufficient to resolve intermodal interference.

predict the overall MZI spectral response without detailed inline measurements of the individual modes as they propagate through the MZI. We would like to stress again that observation of high fringe visibility here does not directly imply high measurement sensitivity as would be the case in a single-mode MZI. On the other hand, high visibility does show that the parasitic modes remain coherent with the mode of interest as they propagate through the MZI.

While this model is simplified, it captures the main features of our experiment, in particular the observation that with two modes with similar β , one can still observe high visibility interference in a narrow spectral region, which depends on the specific implementation of the MZI as detailed above. We note that while the model captures the underlying physics, it cannot be used to fit the experimental data directly because we are attributing all the phase difference between the two modes to the difference in propagation constants and not including any differential phase shifts encountered by the modes as they propagate through the MZI.

As noted in the main text, the β values for the two modes are sensitive to the waveguide orientation with respect to the crystal axes. Figure 10 shows the beat length for the Lamb-like and Love-like modes with the waveguide cross section in the y - z plane, and the mode propagating along x . The beat lengths, and therefore the mode suppression, are quite different. Because we are not certain about the coupler orientation with respect to the GaN (and SiC) in-plane crystal axes, there is a resultant uncertainty in the directional coupler mode suppression.

This orientation dependence of mode suppression can partly explain the time-domain data for the relative heights

of the L_1 and L_2 pulse heights in Fig. 4 in the main text. As noted above, the Lamb wave should, in principle, be better excited via the FIDTs, but in the instance that the directional coupler is misaligned, the Love-like mode transmission can be higher. In any case, moving forward, coupler orientation with a crystal axis can be potentially used as a tool for engineering the desired mode suppression.

APPENDIX C: MZI GLOBAL HEATING CALIBRATION

The calibration curves for acoustic frequency shift (ppm/ $^{\circ}\text{C}$) are shown in Fig. 11. As noted in the main text, the calibration curves are extracted by globally heating

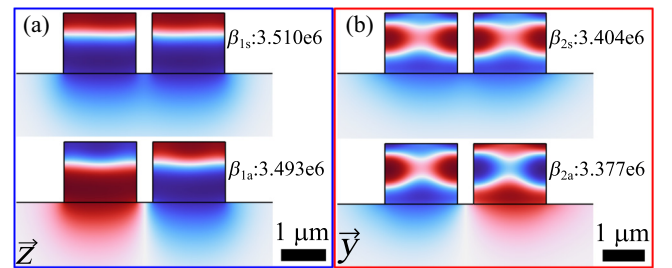


FIG. 10. Mode profiles of Lamb mode (a) and Love mode (b) correspond to a directional coupler oriented along the y - z plane with the mode propagating along x . Symmetric and antisymmetric modes are denoted as β_s and β_a , respectively, and the corresponding beat length for the Love-like mode is now $116\ \mu\text{m}$ and for the Lamb-like mode is $194\ \mu\text{m}$. These results should be contrasted with the values reported in Fig. 2 in the main text.

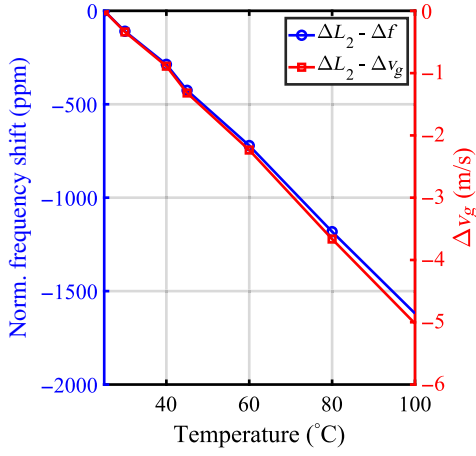


FIG. 11. Normalized frequency shift of the MZI ΔL_2 device reported in the main text Fig. 3 as a function of the overall chip temperature. The data are extracted by tracking the frequency of a particular fringe dip as the sample temperature is varied. Note here that both arms (and the IDTs) are changed simultaneously. This fringe shift is then used to extract an effective group-velocity shift, with the caveat that there are two propagating modes in the MZI.

both MZI and microring resonator devices and noting the shift in mode transmission frequency as a function of temperature. This curve is then used to work out the local temperature in the MZI long arm due to global heating by relating the measured frequency shift to an induced phase shift under the assumption that the fraction of the MZI long arm that is modulated is exactly equal to the heating element. We would like to reiterate that while we map the fringe frequency shift to an effective group velocity shift, there are two distinct modes in the MZI, and this mapping should be interpreted as an effective weighted average of the two modes.

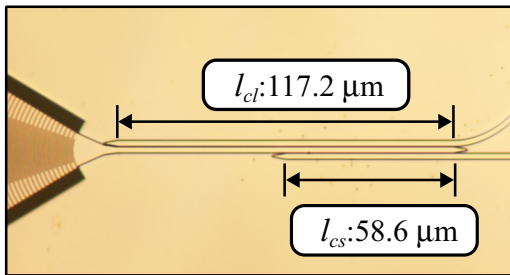


FIG. 12. Microscope image of the PnIC MZI input coupler region (MS1). l_{cl} couples into the long arm and l_{cs} couple into short arm. The length of both is directly correlated to the transferred power ratio (75:25).

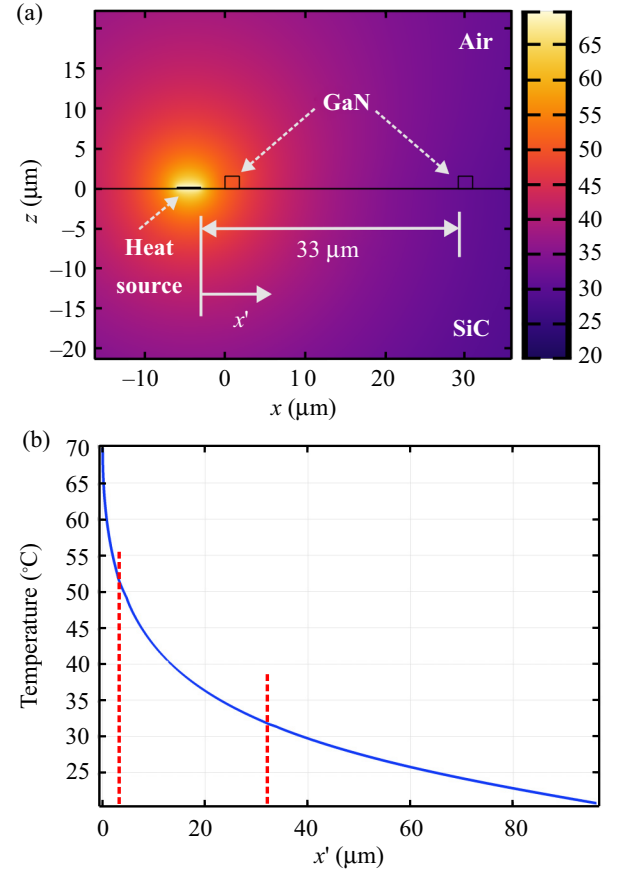


FIG. 13. (a) Steady-state FEM simulation of the 2D temperature distribution originating from localized heating. The heat source adjacent to the waveguide is set to 70 °C. The neighboring waveguide (L arm) reaches a temperature of ≈ 55 °C, while the short-arm waveguide (S) is heated to ≈ 32 °C. (b) Temperature variation along the SiC surface, measured from the edge of the heating pads to the simulation boundary. The location of the two waveguides is shown by red dashed lines, and the coordinate axis (x') starts from the heater edge, as shown in (a).

APPENDIX D: MZI ASYMMETRIC INPUT COUPLER

The input coupler of the PnIC MZI [MS1 in Fig. 1(a)] is designed to split the input acoustic field asymmetrically between the long and short arms. It achieves this via an asymmetric coupling length, as shown in Fig. 12. The 3 : 1 ratio can be inferred by noting that in the first section of the coupler of length $l_{cl}/2$, which is designed to be half the coupling length of the Lamb-like mode, $\approx 1/2$ of the input power is dropped into the long arm, and in the next section $\approx 1/4$ of the remaining power is dropped in the two arms equally, giving us the desired power splitting. We would like to reiterate that the 3 : 1 ratio used in the design is purely based on FEM simulations.

APPENDIX E: FEM SIMULATIONS OF TEMPERATURE DISTRIBUTION DUE TO LOCAL HEATING

Given SiC's high thermal conductivity, it is important to estimate the scale of the temperature fall-off from the localized heating. Figure 13 shows an FEM simulation of the cross-section of the MZI with the heating element located close to the long arm (L), and the short arm located $\approx 33 \mu\text{m}$ away. The simulation shows that with a heater temperature of 70°C , we obtain a temperature of 55°C in the L-arm waveguide, whereas the S-arm temperature reaches 32°C . This simulation clearly shows that for the current device geometries, heater-induced thermal crosstalk between MZI arms needs to be properly accounted for in future experiments via local temperature or Raman measurements in the arms. As noted in the main text, addressing this crosstalk issue is a key challenge moving forward to ensure scaling of device complexity.

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