

Bidirectional Microwave-Optical Conversion with an Integrated Soft-Ferroelectric Barium Titanate Transducer

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Efficient, low-noise, and high-bandwidth transduction between optical and microwave photons is key to long-range quantum communication between distant superconducting quantum processors. Recent demonstrations of microwave-optical transduction using the broadband direct electro-optic (Pockels) effect in optical thin films made of AlN or LiNbO₃ have shown promise. To improve efficiency and added noise, materials with larger Pockels coefficients, such as the soft ferroelectrics BaTiO₃ or SrTiO₃, are needed. However, these materials require adapted designs and fabrication approaches due to their nonlinear and, in some cases, hysteretic electro-optic response. Here, we engineer an on-chip, triply resonant transducer comprising low-loss BaTiO₃-on-SiO₂ waveguides monolithically integrated with a superconducting microwave resonator made of Nb. We demonstrate bidirectional microwave-optical transduction and reach total off-chip efficiencies of 1×10^{-6} using pulsed pumping. The device design permits *in situ* poling of the ferroelectric material without introducing excess microwave loss, exploiting superconducting air bridges fabricated using only subtractive patterning steps. In addition, we investigate optically induced heating, revealing fast thermalization and quasiparticle resilience of the microwave resonator. Our transducer concept and fabrication process are applicable to other materials with a large bias-induced Pockels effect like SrTiO₃ and pave the way for efficient, low-power quantum interconnects.

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I. INTRODUCTION

Quantum computing promises to revolutionize many areas across academia and industry [1]. Because of their long coherence times, fast gates, and scalability, on-chip superconducting qubits represent a front runner for universal quantum computation [2], and superconducting quantum computers are rapidly approaching practical quantum advantage. Eventually, additional functionality may be provided by the ability to send and receive quantum information between distant quantum computing units, allowing them to be interconnected in a quantum network. However, operating at frequencies of approximately 5 GHz, superconducting qubits must be cooled to temperatures of

approximately 10 mK in order to suppress thermal decoherence and ensure high-fidelity computation. Despite efforts to create cryogenically cooled, low-loss microwave quantum links over distances as long as 30 m [3], an entire system maintained at millikelvin temperatures is not a viable solution for long-range interconnects between quantum processing units.

Optical photons, on the other hand, offer negligible thermal noise, high bandwidth, and orders-of-magnitude lower propagation losses at room temperature, using either optical fibers [4,5] or free-space transmission [6,7]. Optical quantum systems have thus been employed to pioneer entanglement distribution over hundreds of kilometers [4,5,8–10], paving the way toward a quantum internet [11,12]. Recently, demonstrations of efficient, low-noise, microwave-optical photon interconversion [13,14] have sparked the idea of networking superconducting quantum systems by means of optical quantum communication [15]. Central to reaching this goal is a quantum transducer capable of entangling microwave and optical photons with high fidelity and large bandwidth.

Currently, the most advanced approaches to an integrated quantum transducer are piezo-optomechanical [16–21] and direct electro-optic [22–25] systems. Whereas the former relies on an intermediary mechanical resonance serving as

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an interface to both microwave and optical modes, the latter directly couples microwave and optical photons via the Pockels effect. Classically, direct electro-optic coupling can be described as a modulation of the optical refractive index proportional to the Pockels coefficient r and an applied microwave field. State-of-the-art thin-film electro-optic platforms offer only comparatively low vacuum coupling rates g_0 on the order of approximately 1 kHz as opposed to approximately 1 MHz for piezo-optomechanical systems. Nevertheless, the direct electro-optic approach is appealing because it avoids the complexity and potential added noise of an intermediary mode and offers the possibility of larger bandwidth. In bulk devices, electro-optic transduction has been demonstrated with efficiency close to unity and

added noise below one photon [26]. First proof-of-concept demonstrations have shown microwave-optical entanglement [27] as well as optical control [25] and readout of a superconducting qubit [28]. In addition to scalability and higher efficiency for a given pump power, integrated devices offer the prospect of gigahertz bandwidths [29].

Following initial demonstrations using AlN ($r = 1$ pm/V) [22,30], most transducer devices today employ optical waveguides made of the ferroelectric LiNbO₃ ($r = 30$ pm/V) [23–25,31,32] [Fig. 1(a)], where peak efficiency and noise performance are limited by heating from optical pump dissipation. The primary route to reducing heating from optical pumping has so far been through enhancement of the optical quality factor, with intrinsic quality factors $Q_{a,0}$

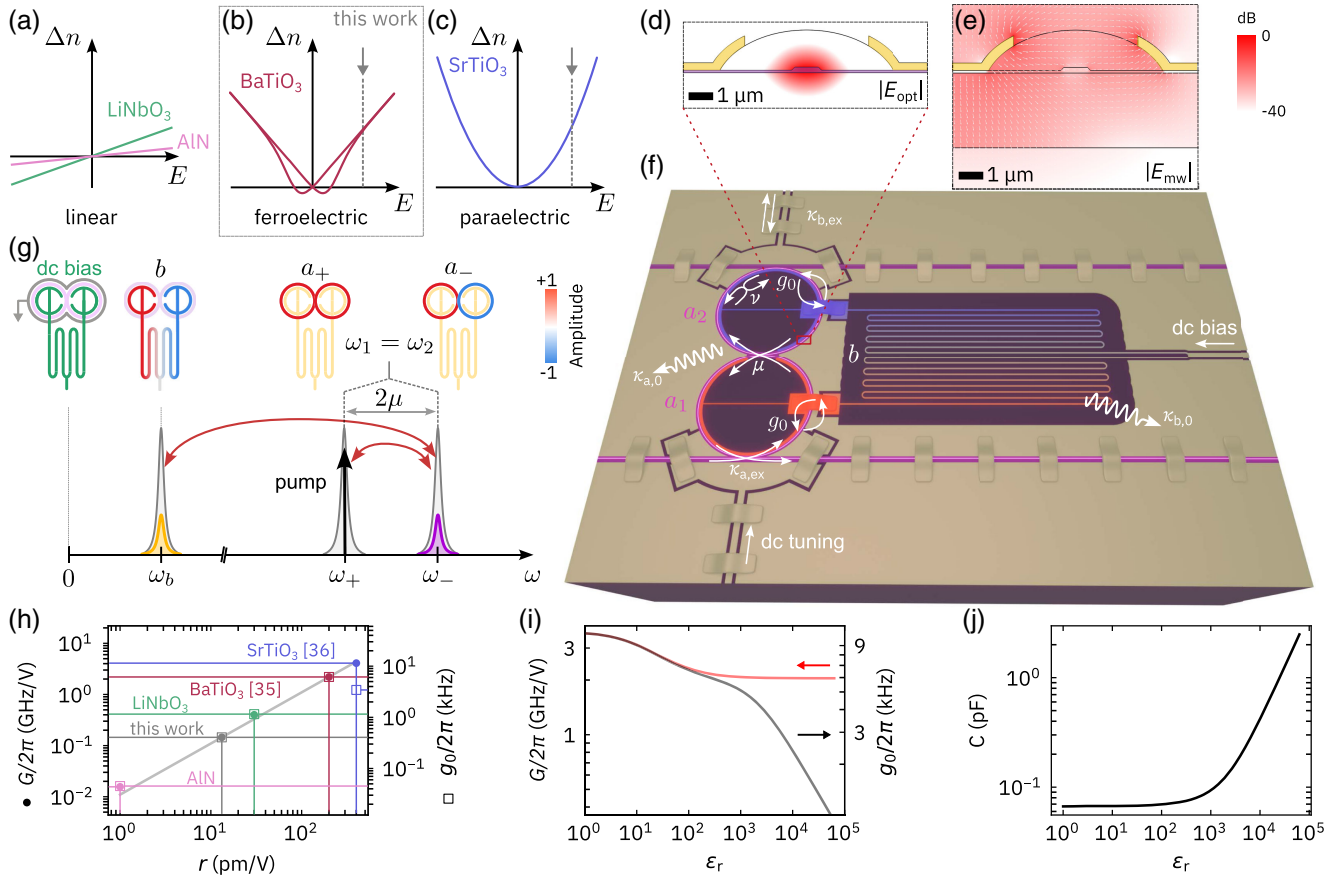


FIG. 1. Electro-optic transducer design for materials with nonlinear Pockels effect. (a)–(c) Diagrams of refractive index change Δn vs applied dc electric field E for (a) linear electro-optic materials AlN, LiNbO₃, and nonlinear electro-optic materials: ferroelectric BaTiO₃ (b) and SrTiO₃ in the quantum paraelectric phase (c). Gray arrows and dashed lines in (b) and (c) indicate working points at nonzero bias fields. (d), (e) Simulated optical and microwave field amplitudes, respectively, in a cross section of BaTiO₃ ridge waveguide and superconducting electrodes. (f) Schematic of the integrated, triply resonant transducer consisting of a photonic molecule formed by two evanescently coupled optical ring resonators (a_1 , a_2) and a $\lambda/2$ -type microwave resonator (b). A voltage applied at the dc-bias port poles the electro-optic material, and a voltage at the dc-tuning port tunes the two ring resonators into resonance. (g) Scattering picture of triply resonant microwave-optical transduction. The hybridized photonic-molecule modes $a_{\pm}$ produce an antisymmetric beating that is phase matched to the microwave field distribution b . Pumping on the lower optical resonance coherently scatters photons between b and $a_{\pm}$ when the microwave resonance frequency ω_b matches the optical splitting 2μ . Shown is also the symmetric dc-bias field poling the material. (h) Simulation of electro-optic frequency pulling factor $G = \partial\omega_a/\partial V$ (circles) and single-photon coupling strength g_0 (squares) as a function of the Pockels coefficient r for different materials. (i) Simulated G (red) and g_0 (black) for $r = 200$ pm/V as function of ϵ_r . (j) Total microwave resonator capacitance C as function of ϵ_r . For (i) and (j), $\omega_b/2\pi = 6.8$ GHz is kept constant.

recently reaching 2.9×10^7 (1.93×10^8) in optimized integrated [33] (bulk [34]) devices. Although the dissipated optical pump power at constant efficiency and critical coupling, i.e., $P_{\text{diss}} \propto 1/(Q_a^2 g_0^2)$, drops quadratically with increasing loaded optical quality factor Q_a [35], this comes at the cost of reduced optical bandwidth [36]. If a transduction bandwidth of approximately 10 MHz is to be maintained, the optical resonator must be increasingly overcoupled for $Q_{a,0} > 4 \times 10^7$ at an optical vacuum wavelength of 1550 nm. In the overcoupled regime, $P_{\text{diss}} \propto 1/Q_{a,0}$ and further improvements of $Q_{a,0}$ permit, at best, a linear suppression of P_{diss} . Additionally, reducing intrinsic optical losses in integrated waveguides becomes increasingly difficult and is accompanied by trade-offs, i.e., between optical quality factor, device footprint, and waveguide geometry [33]. Hence, it will be challenging to achieve further advances using established materials in integrated devices. Increasing the vacuum electro-optic coupling rate g_0 , on the other hand, always enables a quadratic reduction in dissipated optical pump power for a given efficiency and is key to achieving scalable and low-noise transduction.

There is therefore interest in integrating alternative materials with larger Pockels coefficients r , which would enable higher vacuum coupling rates ($g_0 \propto r$ [35]) and thus more efficient and broadband transducers operating at lower optical pump powers. Promising candidates are the soft ferroelectric perovskites BaTiO_3 and SrTiO_3 . For BaTiO_3 , effective Pockels coefficients of $r = 500$ and 200 pm/V have been reported in thin films at 300 and 4 K, respectively [37], while for SrTiO_3 , values of up to $r = 400$ pm/V were demonstrated in a bulk device at 4 K [38]. Whereas single-crystalline LiNbO_3 , a hard ferroelectric with a Curie temperature over 1300 K [39], features a permanent polarization, a stable ferroelectric phase, and a linear electro-optic response in the temperature range of interest for quantum applications (10 mK–300 K) [Fig. 1(a)], the large electro-optic response of BaTiO_3 and SrTiO_3 stems from softening of the ferroelectricity in the vicinity of phase transitions, usually marked by a divergence in permittivity. Indeed, the Curie temperature of BaTiO_3 is 420 K [40], and a phase transition occurs at around 240 K, accompanied by a peak in the electro-optic response ($r \approx 700$ pm/V) [37]. In BaTiO_3 , the Pockels effect is hysteretic due to ferroelectric domain poling [Fig. 1(b)], and maximizing the electro-optic response of BaTiO_3 requires a static electrical bias field to orient the domains [41]. In contrast, the Curie temperature of SrTiO_3 is close to 0 K, giving rise to a quantum paraelectric phase near this temperature [42] and, in turn, a large quadratic electro-optic response [38,43] [Fig. 1(c)]. Thus, both BaTiO_3 in its ferroelectric phase [Fig. 1(b)] and SrTiO_3 in its quantum-paraelectric phase [Fig. 1(c)] require a dc-bias field to control and harness their nonlinear electro-optic response for applications such as quantum transduction.

Despite the promising electro-optic properties of soft ferroelectric materials, their use in integrated devices faces several challenges: (i) There are often no commercially available (single-)crystalline thin films with the necessary optical quality. (ii) The material properties depend on growth parameters, film thickness, strain, temperature, and environmental conditions, such as the presence of oxygen or water vapor [41,44]. (iii) An electrical microwave circuit designed for dc biasing of the active material is required to enable control of the nonlinear electro-optic response.

Here, we take a first step to address these challenges by developing the design and fabrication processes for an integrated triply resonant electro-optic transducer for soft ferroelectric thin films. The transducer enables *in situ* ferroelectric domain poling without incurring microwave losses, as well as independent tuning of the photonic resonances without sacrificing electro-optic mode overlap. Using a device fabricated with BaTiO_3 , we show bidirectional continuous-wave (cw) and pulsed transduction with a peak efficiency of 1×10^{-6} and linear behavior over a wide range of pump powers. Further, we investigate optically induced heating of the microwave circuit. Because of their different dynamics, we can distinguish fast in-cavity dielectric heating and slower substrate heating, as well as quasiparticle effects. We find that in-cavity dielectric heating dominates over the other mechanisms at optical pump powers relevant for quantum transduction.

II. TRANSDUCER DESIGN

Our on-chip transducer consists of a photonic molecule integrated with a quasi-lumped-element superconducting microwave resonator [Fig. 1(f)]. Two optical ring resonators comprising SiO_2 -clad ridge waveguides [Fig. 1(d)] are coupled evanescently at rate μ to form the photonic molecule [23–25,31,32]. Using a voltage applied at the dc-tuning port [Fig. 1(f)], the detuning $\Delta_\pi = \omega_1 - \omega_2$ between the ring modes a_1 and a_2 with frequencies ω_1 and ω_2 is controlled to form hybrid symmetric and anti-symmetric normal modes $a_\pm = (a_1 \pm a_2)/\sqrt{2}$ with frequencies $\omega_\pm$, respectively [Fig. 1(g)]. The microwave resonator is composed of two capacitors enclosing the optical waveguides [Fig. 1(e)] and a connecting meandered inductor. It is dimensioned to have an open-ended half-wavelength ($\lambda/2$) microwave field distribution (mode b) matching the antisymmetric optical beating $a_+^\dagger a_- + a_-^\dagger a_+$ (Appendix A 4). To achieve triply resonant frequency conversion, the photonic-molecule splitting $\omega_- - \omega_+ = \sqrt{(2\mu)^2 + (\Delta_\pi)^2}$ should match the microwave resonance frequency ω_b . The dc-bias port [Fig. 1(f)] is connected to the microwave resonator at the midpoint where the voltage amplitude vanishes. Designing the microwave mode to have mirror symmetry with respect to a line through the bias port enables orientation of the ferroelectric domains

without incurring excess microwave loss, as confirmed by finite-element simulations of the full device [see Appendix A 4, Fig. 7(a)]. The symmetry of the layout is further maintained by having two optical bus waveguides pass along the upper- and lower-ring resonators. Whereas the lower bus waveguide is designed for critical optical coupling, the upper one is displaced a few microns from the upper ring resonator to avoid optical coupling. Instead, it couples to another device on the chip. For microwave input and output, a coplanar waveguide capacitively couples to the microwave resonator at the top. In contrast to all previous integrated designs [23–25,31,32], the dc-tuning electrode is capacitively coupled to the transducer electrode inside the rings instead of tuning with respect to ground. This means that the tuning section is also part of the microwave resonator and contributes to transduction, resulting in a record-high usable electrode coverage of 91%. The tuning electrode as well as the microwave port electrode on the opposite side of the transducer are purposely wide to ensure both strong optical dc tuning and a large external coupling rate $\kappa_{b,ex}$ to the microwave bus.

The electro-optic interaction in our device is described by the Hamiltonian $H_{\text{int}} = \hbar g_0 (a_+^\dagger a_- + a_-^\dagger a_+) (b + b^\dagger)$ with vacuum electro-optic coupling constant $g_0 = GV_{\text{ZPF}}$. Here, $G = \partial\omega_a/\partial V$ is the frequency pulling factor and $V_{\text{ZPF}} = \sqrt{\hbar\omega_b/(8C)}$ is the microwave resonator's zero-point voltage across each ring capacitor. Note that the factor of 1/2 compared to the usual definition of V_{ZPF} originates from the two ring capacitances to ground acting like a voltage divider [see Appendix A 4, Fig. 8(a)]. Critical parameters determining the strength of g_0 are the Pockels coefficient r and the relative permittivity ϵ_r of the electro-optic material, the microwave-optical mode overlap, and the effective transducer capacitance C . We can estimate g_0 via numerical simulations. First, an equivalent circuit is derived for the microwave resonator that is motivated by the $\lambda/2$ field distribution obtained from a fully three-dimensional finite-element simulation [Appendix A 4, Fig. 7(a)]. This breaks down the problem into simulations of the electro-optic mode overlap and individual capacitances and inductances. Exploiting angular symmetry, the overlap integral is calculated over a two-dimensional cross section of the optical waveguide and transduction electrodes [Figs. 1(d) and 1(e)].

We thus compute g_0 and G as a function of r for various electro-optic materials [Fig. 1(h)]. Note that, roughly speaking, $g_0 \propto r/\sqrt{C}$, whereas $G \propto r$, and both scale approximately linearly with microwave-optical mode overlap. (See Appendix A for the full expressions.) The simulation predicts $g_0/2\pi = 0.05$ kHz for AlN ($r = 1$ pm/V, $\epsilon_r = 9$) [45], $g_0/2\pi = 1.2$ kHz for LiNbO₃ ($r = 30$ pm/V, $\epsilon_r = 25$) [45], and $g_0/2\pi = 6.1$ kHz for BaTiO₃ ($r = 200$ pm/V, $\epsilon_r = 200$ [37]). For quantum paraelectric SrTiO₃, a value of $g_0/2\pi \approx 3.4$ kHz is obtained

from bulk values ($r = 400$ pm/V, $\epsilon_r = 2 \times 10^4$) [38], though the significantly larger biasing fields possible in integrated devices may enable larger values of r and g_0 . The value of g_0 obtained for LiNbO₃ is comparable to that of other published devices [23–25,31]. As expected, there is an approximately linear relationship between g_0 or G and r [solid gray line in Fig. 1(h) computed with $\epsilon_r = 200$]. Deviations from the linear relation are caused by the vastly different material permittivities, in particular, that of SrTiO₃. Moreover, ϵ_r can be tuned significantly in soft ferroelectrics via doping, strain, temperature, and bias voltage [46–48]. A recent report on the correlation between the large Pockels response and the permittivity of soft ferroelectrics [47] highlights the importance of considering r and ϵ_r together, as a large permittivity may, depending on geometry, reduce the electro-optic mode overlap by suppressing the microwave field penetration into the optical waveguide or increase the transducer capacitance.

To elucidate the dependence of our device design on ϵ_r , we simulate g_0 and G as a function of ϵ_r between 1 and 10^5 for the expected value of $r = 200$ pm/V for BaTiO₃ [Fig. 1(i)]. The device geometry is kept constant except for the inductor length, which is adapted to compensate the change in capacitance and maintain the same microwave resonance frequency. At low permittivity ($\epsilon_r \approx 1$), the field penetrates the waveguide core both directly through the slab in contact with the electrodes and through the optical cladding ($\epsilon_{r,\text{SiO}_2} \approx 4$). With increasing permittivity, the field penetration through the cladding becomes insignificant and plays virtually no role for $\epsilon_r > 10^2$, where the field strength in the waveguide core saturates to half of that in the slab, roughly corresponding to the ratio of the slab and total waveguide thicknesses. As a consequence, G and g_0 reduce gradually by the same amount but begin to plateau around $\epsilon_r \approx 10^2$ [Fig. 1(i)], a fact that should be considered in waveguide optimization. For values of ϵ_r exceeding 2×10^3 , we see another drop in g_0 , but not G , caused by the increase of transducer capacitance [Fig. 1(j)]. In this regime, the capacitance of the transduction electrodes dominates over the stray capacitance and lowers V_{ZPF} . This effect can be mitigated to some extent by making the device smaller. Note that, for the device studied in this work, the zero-point voltage $V_{\text{ZPF}} = 2.8$ μ V is still dominated by the stray capacitance of the meander inductor.

Our simulations show that this novel transducer design can leverage the large electro-optic effect offered by soft ferroelectrics to enable devices with significantly larger vacuum coupling strengths than achievable with established materials; that is, up to $5\times$ larger than with LiNbO₃, assuming reported values. Despite the large permittivities of these materials, strong microwave-optical overlap is maintained by the use of a ridge waveguide with low ridge-slab contrast and electrodes in direct contact with the slab.

III. DEVICE FABRICATION AND CHARACTERIZATION

Our transducer device [Fig. 2(a)] is fabricated using only subtractive patterning steps (see Appendix C 1). Optical ridge waveguides are defined by dry etching of a 225-nm BaTiO₃ film bonded to an oxidized Si substrate and are later cladded with SiO₂. The microwave circuit, including the air bridges, is made of sputtered Nb. The fabrication process differs from that of other transducer realizations [22–25,30–32] in that lithographically defined superconducting air bridges are employed to enable optical waveguide-superconductor and superconductor-superconductor crossings for complex layouts. The subtractive two-step metallization process with dry etching preserves submicron critical dimensions, e.g., for nanowire inductors in the main metallization layer. The air bridges are compatible with both oxide- [Fig. 2(c)] and air-clad optical waveguides [Fig. 2(d)]. Note that air bridges are included in the current-carrying portion of the microwave resonator connecting the

inductor wire (2 μm wide, 300 nm thick) to the transduction electrodes, not just as patches between regions of the ground plane, through which no significant current flows. The waveguide cross section in Fig. 2(b) showcases the dome-shaped cladding that enables clean subtractive electrode fabrication in contrast to a lift-off process, allowing thicker superconducting films. Without the dome shape, the sputter-deposited Nb layer would not be cleared cleanly from any vertical sidewall during dry etching, risking the creation of a short circuit. Overall, our process flow provides a robust, high-yield method for incorporating multiple low-loss air bridges made of niobium into a superconducting coplanar-waveguide circuit, as opposed to the ubiquitous aluminum air bridges. This includes crossings of photonic structures such as waveguides, where absorption of optical fields must be avoided.

The BaTiO₃ optical bus waveguides lead to the chip facets, which are defined by dry etching deep trenches into the Si substrate. Approaching the facet, the ridge

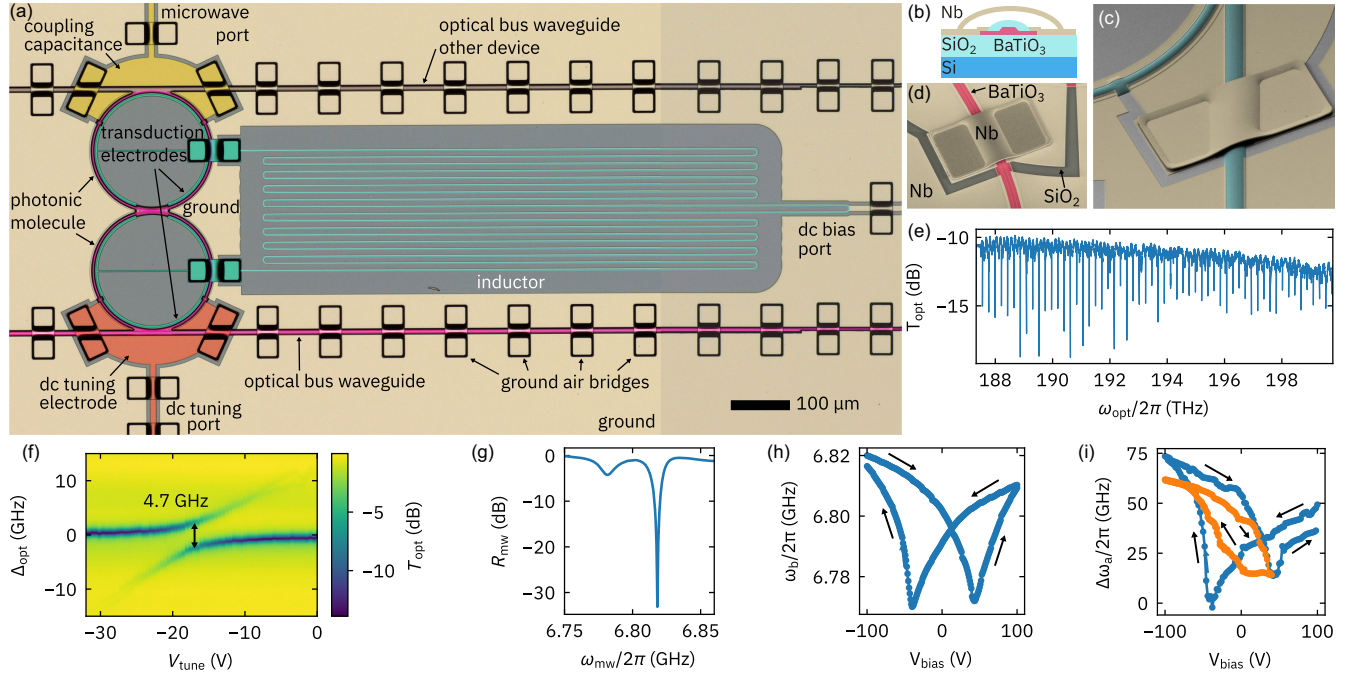


FIG. 2. Triply resonant photonic-molecule transducer with dc bias. (a) False-colored optical micrograph of the photonic-molecule transducer. Optical signals are evanescently coupled from the optical bus waveguide to the lower ring of the photonic molecule, which is evanescently coupled to the upper ring at its opposite side. Symmetric electrodes for dc tuning (orange) of the photonic-molecule and microwave bus coupling (yellow) are placed on each ring. (b) Schematic cross section of the measured device, showing BaTiO₃ ridge waveguides on SiO₂ on Si and SiO₂ cladding. Nb electrodes are in contact with the BaTiO₃ slab, with Nb air bridges across, where required. (c) False-colored SEM micrograph of a representative Nb air bridge used throughout the design. (d) For illustration, an air bridge across an air-clad waveguide on another fabricated chip is shown. (e) Optical spectrum with a free-spectral range of 218 GHz measured in transmission at room temperature after optical packaging. (f) Optical avoided crossing as a function of dc-tuning voltage showing photonic-molecule hybridization with a ring-ring coupling of $2\mu = 2\pi \times 4.7$ GHz. (g) Microwave reflection measured at 10 mK at $V_{\text{bias}} = -100$ V revealing the microwave resonance at 6.82 GHz. (h) Tuning curve of the microwave resonance frequency as a function of dc-bias voltage showing ferroelectric hysteresis (arrows indicate tuning direction). (i) Tuning curve of the optical resonance measured in reflection at 75 mK. The blue data points show the initial hysteresis curve after cooldown, which includes a drift due to material conditioning (open curve at 100 V). The orange data points show hysteresis when ramping V_{bias} between -100 V and $+40$ V, with the latter voltage corresponding to the coercive field. The tuning voltage in (e),(g)–(i) is 0 V.

waveguide is tapered to achieve good mode matching with an ultrahigh-numerical-aperture (UHNA) optical fiber (Appendix C 2). Broadband optical fiber-to-chip coupling with 200-nm bandwidth is achieved, as evidenced by the envelope of the optical transmission spectrum in Fig. 2(e), in contrast to the comparatively narrow-band grating couplers used in most transducer demonstrations to date [23–25].

We initially characterize the optical transmission spectrum [Fig. 2(e)] at room temperature. Without bias or tuning voltages applied, a series of resonances of the lower ring resonator coupled to the optical bus waveguide are observed with a free-spectral range of 218 GHz and an average intrinsic optical linewidth of $\bar{\kappa}_{a,0}/2\pi = 776$ MHz ($\bar{Q}_{a,0} = 2.5 \times 10^5$) (see Appendix E for further optical characterization). Monitoring the spectrum of a selected resonance at 1603.5 nm as a function of tuning voltage [Fig. 2(f)], we observe an avoided crossing between modes of the two ring resonators. The resonant photonic-molecule splitting is thus determined to be $2\mu = 2\pi \times 4.7$ GHz at vanishing ring-ring detuning $\Delta_{rr} = 0$. Because this value of μ is the largest observed and the closest to the microwave resonance frequency, this pair of coupled optical resonances is selected for transduction measurements.

The chip is packaged for cryogenic measurements by aligning UHNA fibers and gluing them to the chip (Appendix C 2). Wire bonding the chip to a microwave printed circuit board and enclosing the assembly inside a copper cavity completes the packaging. During sample handling, before mounting in the dilution refrigerator, the gluing point of one of the optical fibers was damaged, resulting in misalignment and no measurable coupling to the bus waveguide. Consequently, the packaged transducer could only be probed optically in reflection. The reflected signal originates from moderate coherent backscattering in the ring resonators.

The packaged transducer is cooled in a dilution refrigerator to a base temperature of 8 mK without any voltage applied to the device. The microwave resonance is then probed in reflection [Fig. 2(g)], at a bias voltage of -100 V, revealing the critically coupled transducer resonance at $\omega_b = 2\pi \times 6.82$ GHz, close to the simulated value. A spurious mode at 6.78 GHz is also detected, the frequency of which does not tune with temperature and might originate from the package (see Appendix F for more detailed microwave characterization). At single-photon microwave probe power levels (i.e., -120 dB m), we find an intrinsic microwave linewidth of $\kappa_{b,0} = 2\pi \times 10$ MHz ($Q_{b,0} = 7.0 \times 10^2$), limited by dielectric loss. The external coupling rate was extracted to be $\kappa_{b,ex} = 2\pi \times 7.2$ MHz ($Q_{b,ex} = 9.5 \times 10^2$) for coupling to the microwave feedline.

Ramping the bias voltage from -100 to 100 V and back [Fig. 2(h)], we observe tuning of the microwave resonance due to the change of permittivity with applied field, with pronounced hysteresis due to the ferroelectricity of

BaTiO₃. We observe similar hysteretic tuning for the optical resonance [Fig. 2(i)], albeit with more noise due to the measurement being made in reflection, which precludes an unambiguous assignment of the resonance frequency because of interference effects (Appendix F 3). The initially recorded optical tuning curve cycling between 100 and -100 V also exhibits drift, as this measurement was taken immediately after cooldown. The partial tuning curve cycling between -100 and $+40$ V, the latter voltage corresponding to the coercive field at which ferroelectric domains are reoriented, shows a weaker hysteresis. For transduction measurements, the device was biased at -100 V, where the ferroelectric domains are maximally oriented, achieving the strongest Pockels effect [41]. The optical frequency tuning rates observed in Figs. 2(f) and 2(i) can be used to infer an effective Pockels coefficient of 273 pm/V at room temperature and -18 V, and 13 pm/V at 75 mK and -100 V (see Appendix A 3 and E 2).

IV. BIDIRECTIONAL CONTINUOUS-WAVE TRANSDUCTION

Having characterized the optical and microwave resonances individually, we set out to measure microwave-optical transduction. For these experiments, the device is poled at a bias voltage of -100 V. Because the resonant photonic-molecule splitting [$2\mu = 2\pi \times 4.7$ GHz at room temperature, Fig. 2(f)] is smaller than the microwave resonance frequency [$\omega_b = 2\pi \times 6.82$ GHz, Fig. 2(g)], the ring-ring detuning Δ_{rr} , controlled by the dc-tuning port, must be adjusted to achieve frequency matching, albeit at the cost of mode overlap. Unfortunately, without the possibility of measuring the optical spectrum in transmission, we cannot easily determine the photonic-molecule splitting at millikelvin temperatures. We instead adjust the dc-tuning voltage to maximize the observed transduction efficiency (Appendix G).

For microwave-to-optical transduction [Fig. 3(a)], we apply a microwave modulation field b_{in} at the microwave coupling port with frequency ω_{mw} around ω_b . The photonic molecule is pumped with a continuous-wave laser (frequency ω_p , power -11 dBm in fiber) near the resonance pair at 1603.5 nm. The electro-optic interaction drives both Stokes scattering $b \rightarrow a_{\pm}^{\dagger}$ to a photon with lower frequency $\omega_p - \omega_{mw}$, as well as anti-Stokes scattering $b \rightarrow a_{\pm}$ to a photon with higher frequency $\omega_p + \omega_{mw}$ [Fig. 3(b)]. Using optical heterodyne detection allows us to distinguish both processes. Because of the misalignment of the second optical fiber, we detect the transduced photons emitted in the backward direction (a_{out}), i.e., in reflection, relying on residual optical backscattering in our device.

We first position the microwave tone on resonance $\omega_{mw} = \omega_b$ and measure the transduced optical signal as a function of optical pump detuning $\Delta_p = \omega_p - \omega_+$ [Fig. 3(c)]. Normalizing the detected signals by the independently

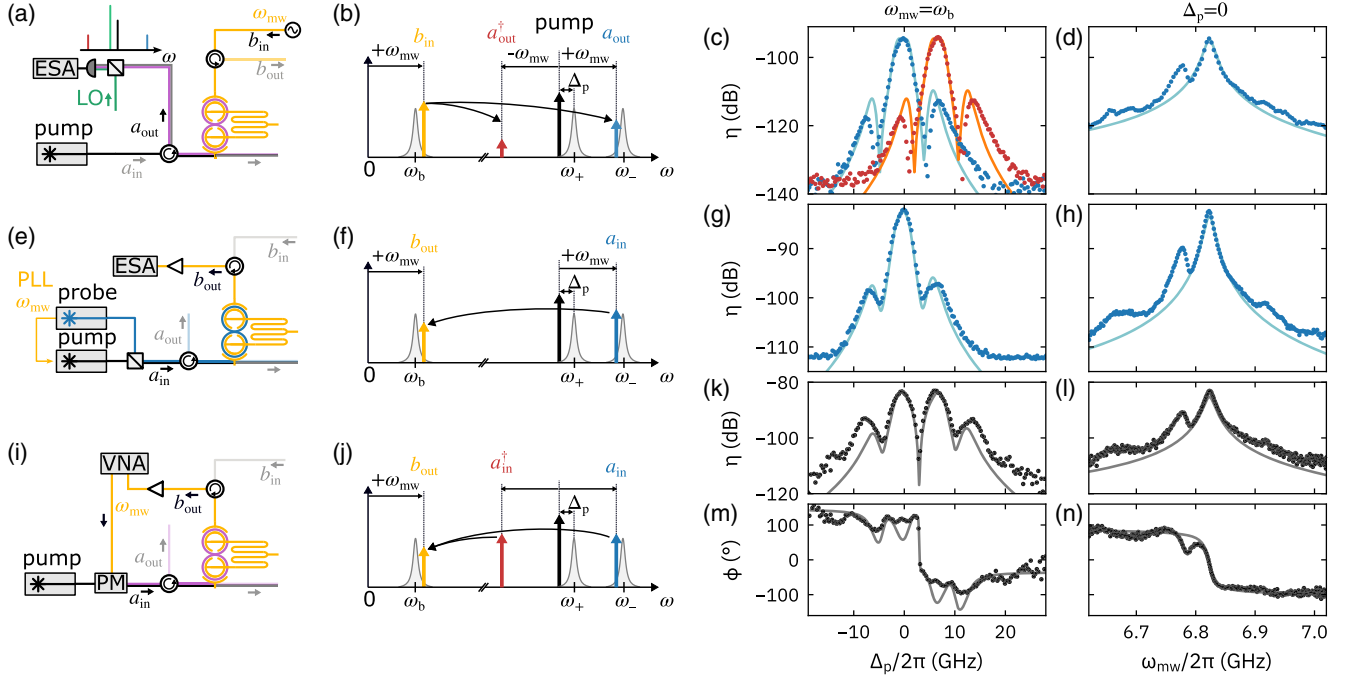


FIG. 3. Bidirectional coherent triply resonant microwave-optical transduction. (a)–(d) Microwave-to-optical transduction. (a) Setup in which the transducer is driven by a microwave tone b_{in} and an optical pump. The optical signal a_{out} in backward direction is combined with a local oscillator (LO) and detected using heterodyne detection on an electrical spectrum analyzer (ESA). (b) Scattering diagram showing the microwave and optical resonances, the microwave tone (b_{in} , yellow), and the transduced optical Stokes ($a_{out}^\dagger$, red) and anti-Stokes (a_{out} , blue) sidebands. The optical pump is detuned by Δ_p from the lower photonic resonance ω_+ . (c) Transduction efficiency of the Stokes (red) and anti-Stokes (blue) processes on microwave resonance ($\omega_{mw} = \omega_b$) as a function of Δ_p . (d) Anti-Stokes transduction efficiency for optical pumping at $\Delta_p = 0$ as a function of ω_{mw} . (e)–(h) Single-sideband optical-to-microwave transduction. (e) Setup in which a probe laser is locked to the pump laser at a blue-detuned frequency offset ω_{mw} and both are driving the device at a_{in} . The transduced signal b_{out} is amplified and detected on an ESA. (f) Scattering diagram showing probe laser photons a_{in} being scattered into the microwave mode at ω_{mw} . (g), (h) Transduction efficiency as a function of Δ_p and ω_{mw} , respectively. (i)–(n) Coherent optical-to-microwave transduction: (i) Setup in which a VNA drives a phase modulator (PM) at frequency ω_{mw} , creating coherent sidebands on the pump laser driving the transducer at a_{in} . The transduced signal b_{out} is amplified and detected at the VNA. (j) The blue and red phase modulation sidebands are transduced simultaneously to the microwave mode and interfere because of their fixed phase relation. (k), (l) Efficiency and (m), (n) phase of the transduced signal as a function of Δ_p and ω_{mw} , respectively. For all measurements, the optical pump power P_p is -11 dB m in the fiber. The solid lines in (c), (d), (g), (h), (k), (l), (m), (n) are theoretical curves determined by the S -matrix coefficients of the linear transduction model.

calibrated optical and microwave powers, as well as transmission losses, we extract the off-chip transduction efficiency η (Appendix D 2), defined as the photon number efficiency between the optical fiber and the microwave coax line of the cryostat. We note that η includes fiber-to-chip loss, estimated at -5 dB from room-temperature characterization. The recorded efficiency curves for the Stokes and anti-Stokes processes are displaced in frequency with respect to each other by the photonic-molecule splitting and reach a maximum around $\omega_p = \omega_-$ and $\omega_p = \omega_+$, respectively. The traces display the characteristic response of a triply resonant device: At the central peak, both optical pump and signal are resonant with respective modes of the photonic molecule. Two side peaks with lower amplitude occur when either the optical pump or the signal are resonant with an optical mode. These doubly resonant

peaks are displaced from the triply resonant peak by the photonic-molecule splitting.

Setting the pump laser to the lower photonic-molecule resonance ($\Delta_p = 0$), we next measure transduction efficiency of the anti-Stokes sideband as a function of microwave frequency ω_{mw} [Fig. 3(d)]. As expected, the data reveal the microwave resonance, including the previously observed spurious mode (Fig. 25). The transduction bandwidth is determined from the linewidth to be $\kappa_b = 2\pi \times 14$ MHz.

To measure the optical-to-microwave response of the system, we combine the optical pump with light from a blueshifted probe laser that is phase locked to the pump laser with a programmable frequency offset denoted ω_{mw} [Fig. 3(f)]. The light from both lasers is injected into the device in the forward direction [a_{in} , Fig. 3(e)]. The transduced microwave signal is then amplified using a

high-electron-mobility transistor (HEMT) amplifier and detected using an electrical spectrum analyzer. As before, we measure the transduction efficiency as a function of optical pump detuning, keeping the pump-probe offset fixed at $\omega_{\text{mw}} = \omega_b$ [Fig. 3(g)]. The data show a triply resonant response similar to that of the microwave-to-optical experiment. However, the overall efficiency is more than 10 dB higher. We attribute this difference to the low backscattering efficiency in our device reducing the observed microwave-to-optical signal; the backscattering rate has no effect on the optical-to-microwave data. Fixing the optical pump signal at $\Delta_p = 0$, a sweep of the pump-probe frequency offset ω_{mw} again traces out the microwave resonance [Fig. 3(h)].

To understand the electro-optic response of the device, we treat the transducer as a linear system and compute its S matrix (Appendix B 3). We model the data using theoretical curves determined by the S -matrix coefficients (Fig. 3), adjusting the photonic-molecule coupling strength μ , the ring-resonator detuning Δ_r , and the optical loss rates, as well as the coherent optical forward-backward coupling rates $\nu_{1,2}$, arising from scattering at imperfections in the optical resonator. The microwave loss rates are determined from independent spectroscopy [Fig. 2(g)] and the vacuum electro-optic coupling strength g_0 from electro-optic dc tuning (Appendix E 2). All system parameters are summarized in Table V of Appendix G.

We find very good agreement between the theory curves and the data; the triply resonant response as well as the dips in between the doubly and triply resonant peaks are reproduced by the model. We find that the reduced microwave-to-optical transduction efficiency of the signal emitted in the backward direction is well explained by a low forward-backward coupling strength of $\nu_{1,2} \approx 0.1\kappa_a$. However, since neither the peaks at triple or double resonance are split [Figs. 3(c) and 3(g)], intraring forward-backward coupling cannot clearly be distinguished from residual backscattering that may occur, e.g., at the end facet of the bus waveguide.

The measured bidirectional efficiencies are consistent with the vacuum electro-optic coupling rate of $g_0 = 2\pi \times 406$ Hz extracted independently from dc tuning of the optical resonance frequency (see Appendix E 2). This corresponds to an effective material Pockels coefficient of $r = 13$ pm/V, which is more than an order of magnitude lower than the value $r = 200$ pm/V previously reported for BaTiO₃ thin films at cryogenic temperatures [37] [Fig. 1(h)]. Possible explanations include variations in material quality and processing, but also a difference in strain in the BaTiO₃ layer. Indeed, it has been reported that the electro-optic response in perovskite titanate films, and strong electro-optic materials, in general, depends on strain [49]. This is confirmed by first-principles calculations for BaTiO₃ thin films [44] and it has also been shown to affect the phase diagram of SrTiO₃ [50]. Since those

materials typically also feature large thermal expansion coefficients [49], strain engineering is likely going to be crucial for optimal performance at cryogenic temperatures.

Finally, we measure phase-coherent optical-to-microwave transduction using a vector network analyzer (VNA). An external phase modulator driven by the VNA creates blue and red sidebands at $\pm\omega_{\text{mw}}$ on the pump light injected into the device [Fig. 3(i)]. As the two optical sidebands have a fixed phase relation, i.e., $a_{\text{in}} \propto i(e^{-i\omega_{\text{mw}}t} + e^{i\omega_{\text{mw}}t})$, their simultaneous transduction leads to interference, which becomes apparent when sweeping Δ_p at fixed $\omega_{\text{mw}} = \omega_b$ [Fig. 3(k)]. Now, two triply resonant peaks arise, with destructive interference occurring at a pump detuning symmetric with respect to the two photonic-molecule modes due to the π phase shift between the two phase-modulation sidebands, i.e., $a_{\text{in}} = -a_{\text{in}}^\dagger$. In the phase response [Fig. 3(m)], a π phase shift at the point of symmetric detuning is also observed and well captured by the model. The smaller features, however, are exaggerated by the simulation, suggesting a slight deviation of the parameters from those of the actual device. When the pump is tuned to ω_+ ($\Delta_p = 0$), the microwave frequency dependence of the transduced signal [Fig. 3(l)] is dominated by the blue sideband, as the red sideband is far off resonance and, therefore, looks virtually identical to the previous measurement [Fig. 3(h)]. Additionally, the phase of the measured signal [Fig. 3(n)] agrees well with the predicted oscillator response, confirming that the process is coherent.

V. OPTICALLY INDUCED HEATING

We now investigate optical-pump-induced heating of the superconducting circuit, which is known to cause microwave decoherence, add noise, and limit the transduction efficiency [51]. Tolerance to optical pumping is thus an important aspect of a quantum transducer.

There are two main pathways through which optical photons can contribute to heating of the microwave circuit [Fig. 4(a)]: (i) Stray light created by scattering at primarily the fiber-chip interface irradiates the circuit, or (ii) resonant light coupled to the photonic molecule is absorbed in the dielectric or scattered to the superconductor. Whereas stray light causes a near homogeneous irradiation of both the capacitors and the inductor, in-cavity light is confined to the optical rings and predominantly affects the capacitors. Because of these distinct intensity distributions, stray light and in-cavity light result in different heating dynamics.

We probe the possible heating mechanisms using pulsed optical pumping. Laser pulses with a duration of 1 μ s and a repetition rate of 1 kHz are generated by an acousto-optic modulator (AOM), amplified by an erbium-doped fiber amplifier (EDFA) and sent to the device [Fig. 4(a)]. To investigate stray light effects, we deliver off-resonant light through the second misaligned optical fiber such that no light is coupled to the on-chip waveguide [Fig. 4(a)].

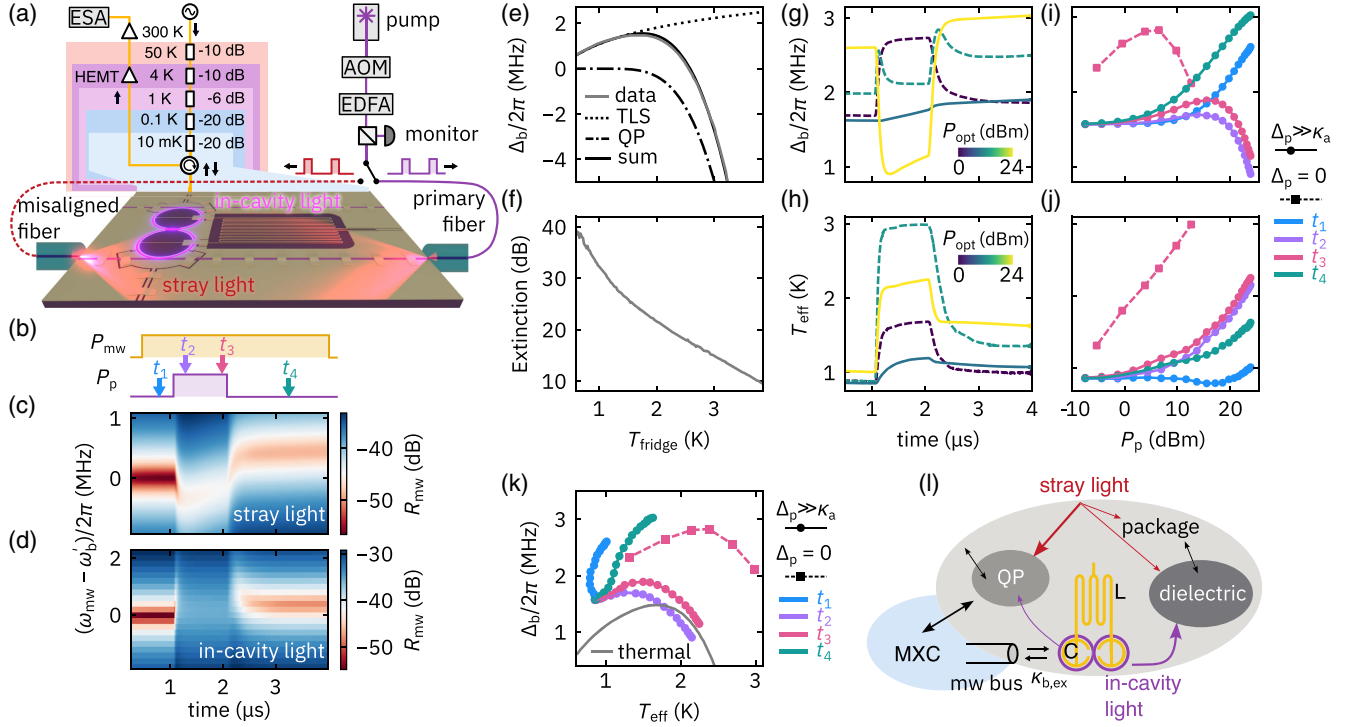


FIG. 4. Microwave response under optical pump. (a) Experimental setup: A pump laser is pulsed using an AOM and amplified with an EDFA. The pulse power is monitored using a fast photodetector and sent to the device either through the primary or misaligned optical fiber. Time-resolved microwave spectroscopy is performed using a pulsed high-power microwave source and an ESA. (b) Pulse sequence with the microwave pulse starting before and ending after the optical pulse. Arrows mark times at which microwave resonator properties were evaluated for (i)–(k). Exemplary time-resolved microwave spectra with (c) off-resonant stray light pulses and (d) resonant, in-cavity light pulses, using in-fiber optical powers of 18 and 19 dBm, respectively. During the optical pulse, a shift and reduced extinction of the microwave dip are clearly observed that partially recovers after the pulse. (e) Microwave resonator frequency shift Δ_b as a function of cryostat temperature T_{fridge} without optical light. A theoretical model for Δ_b that is the sum of QP and TLS contributions is fit to the data. (f) Microwave resonator extinction ratio as a function of cryostat temperature used to calibrate effective temperature T_{eff} in optical heating experiments. (g) Frequency shift Δ_b and (h) effective temperature T_{eff} determined as a function of time during the pulse sequence. Different traces correspond to selected peak optical powers (color bar) for off-resonant stray light (solid lines, circles) or resonant in-cavity light (dashed lines, squares). (i) Frequency shift Δ_b and (j) effective temperature T_{eff} as a function of peak optical power at various times in the pulse sequence [arrows in (b)] for off-resonant stray light (solid lines, circles) and resonant in-cavity light (dashed lines, squares). (k) Frequency shift vs effective temperature for optical and thermal experiments at various times in the pulse sequence [arrows in (b)] for off-resonant stray light (solid lines, circles) and resonant in-cavity light (dashed lines, squares). (l) Diagram of heating mechanisms. The microwave mode is coupled to three baths: QPs, the dielectric, and the MXC through the microwave bus. Optical stray light primarily heats the QP bath, while in-cavity light predominantly causes fast heating of the dielectric at the capacitor gap. Both QP and dielectric baths are thermalized to the chip package, which itself is cooled by the MXC.

We note that the observed behavior induced by off-resonance stray light is qualitatively independent of the fiber used, but the fiber-chip coupling point of the misaligned fiber is located closer to the device than that of the primary fiber with good coupling efficiency. To study heating from in-cavity light, we couple light through the primary fiber and step the optical detuning Δ_p across the resonances.

The microwave resonator's response before, during, and after the optical pulses is probed using microwave pulses synchronized to the optical pulses [Fig. 4(b)]. We record reflected microwave power as a function of time and probe frequency ω_{mw} . Microwave spectra for stray light [Fig. 4(c)] and in-cavity light [Fig. 4(d)] both show a

reduction in the resonant extinction during the optical pulse, but a redshift vs a blueshift of the resonance, respectively. By fitting the spectra at each time bin, we determine the time evolution of both the frequency shift and the extinction of the resonance dip.

In order to analyze the dynamics of our system in thermal nonequilibrium, we separately measure the microwave resonance shift [Fig. 4(e)] and extinction [Fig. 4(f)] as a function of cryostat temperature T_{fridge} . The thermal data show an initial resonance blueshift with increasing temperature that turns into a redshift for temperatures above 2 K [Fig. 4(e), see Appendix F 3 for details]. While a redshift is expected from Bardeen-Cooper-Schrieffer (BCS) theory of quasiparticles in a superconductor [52], the blueshift is not.

Instead, a blueshift has been reported to originate from a temperature-dependent contribution to the dielectric constant from two-level systems (TLSs) in the dielectric [53]. Here, it most likely stems from TLSs in BaTiO₃ and the SiO₂ optical cladding in the transducer capacitor gap. We fit a theory curve to our data that is the sum of the dielectric and the BCS models [Fig. 4(e) and Appendix F3], showing very good agreement, and we also plot the individual contributions. A determination of the microwave quality factor was complicated by the hybridization with the spurious mode and the frequency dependence of the external coupling rate to the microwave bus (Appendix F1). Instead, we use the relationship between the measured extinction of the microwave resonance and the cryostat temperature T_{fridge} [Fig. 4(f)] as a calibration for the effective temperature T_{eff} during optical pumping. Since the measured decrease in extinction with temperature is due to a resonance broadening, it is consistent with quasiparticle-induced losses, which also tend to increase with temperature, whereas TLS-induced losses tend to decrease [54]. This points to T_{eff} being dominated by the quasiparticle bath, while the temperature of the dielectric may be different. In the following, we use the thermal model to identify dielectric and quasiparticle heating with blue and red frequency shifts, respectively.

First, we look at the dynamics of the frequency shift [Fig. 4(g)] and effective temperature [Fig. 4(h)]. Traces for resonant light (primary fiber) and off-resonant light (misaligned fiber) at various optical powers are shown. Off-resonant light at low power causes a blueshift and a weak temperature increase with a time constant approximately 1 μs . At high power, the temperature increase becomes much faster, following essentially the rise of the optical power (approximately 0.1 μs). The frequency now shifts strongly to the red starting from a blueshifted background level. Resonant light causes a stronger temperature increase during the optical pulse with fast rise time, quasi instantaneously following the optical pulse shape. At low power, the frequency is strongly blueshifted during the pulse. At high power, a redshift with slightly lower rise time superimposes on the blueshift, as seen by the peaks at the beginning and end of the optical pulse. All traces show a strong rise of the background temperature after the optical pulse due to in-pulse heating [Fig. 4(h)], proportional to optical power, which is slowly decaying to a steady state (i.e., before pulse) at a rate of approximately 10 kHz (Appendix H1) by thermalization with the substrate. The background frequency shift also increases with optical power [Fig. 4(g)], but the difference between levels after and before the pulse is less pronounced. Part of the background level is caused by continuous output noise of the EDFA (average power -22 dBm).

We next analyze the dependence of Δ_b [Fig. 4(i)] and T_{eff} [Fig. 4(j)] on optical power P_p between -7.5 and $+24$ dBm at selected times before (t_1), during (t_2 , t_3),

and after (t_4) the optical pulse [Fig. 4(b)]. For off-resonant light, we observe an increasing blueshift up to power levels of approximately 15 dBm at all times during the pulse sequence. Increasing power further turns this into a redshift when the pulse is on (t_3) and enhances the blueshift when the pulse is off. Resonant light at t_3 produces a strong initial blueshift at low powers with a peak at approximately 3 dBm, after which a redshift starts to set in. Traces for resonant light at other times can be found in Appendix H2. Note that a power correction is applied to the resonant data (Appendix H3) to allow direct comparison with the transduction measurements (Sec. VI). Compared to stray light, in-cavity light creates a 50% larger peak blueshift at about 10 dB less power. The temperature increase for both in-cavity light and sufficiently strong off-resonant light appears to scale with the logarithm of the optical power [Fig. 4(j)]. Resonant light causes a more than tenfold stronger temperature increase from the background level, which seems to be consistent with the peak blueshift being reached at $1/10$ the optical power.

From the dynamics and power dependence, we conclude that there are three distinct heating processes in the device [Fig. 4(l)]: (i) Slow dielectric heating (blueshift) of the substrate due to stray light occurs at low power and causes a raised temperature level after the optical pulse, when quasiparticles have decayed. (ii) Stray light photons hitting the inductor produce quasiparticle heating (redshift) with a fast response time due to the high quasiparticle recombination rate in Nb. (iii) In-cavity light results in dielectric heating (blueshift) with fast rise and fall times, which may be explained by its strong confinement to the ring resonators inside the capacitor gaps. Because of this confinement, in-cavity light contributes stronger dielectric heating and has a lower probability of generating quasiparticles in the relatively distant inductor where they influence microwave coherence the most. The density of quasiparticles is further diminished by the fast recombination rate in Nb.

These data highlight a transition from dielectric heating (blueshift) dominating at low optical powers to quasiparticle-dominated heating (redshift) at resonant optical powers above 5–10 dBm. At such high optical powers, fast heating to temperatures well above 1 K occurs already during the pulse [Fig. 4(j)]. These findings are relevant for quantum applications, for which the number of thermal microwave photons must be less than 1 ($k_B T_{\text{eff}} \ll \hbar\omega_b$) [55]. To attain this low level of thermal excitation, the optical power P_{opt} in our system must be much less than 0 dBm, for which in-cavity dielectric heating dominates over quasiparticle heating of the superconductor. This result differs from that of a previous study on a device with a 50-nm NbN film, where off-resonant stray light at the chip interface was concluded to be the dominant heating source [30], suggesting that the device studied here is more resilient against stray light, likely due to the use of a thicker

(300-nm Nb) superconducting film. Note that due to the fast submicrosecond response of the local dielectric heating, it cannot be mitigated by reducing the duty cycle of the pump, emphasizing the importance of reducing optical absorption loss as much as possible.

The different relative weights of dielectric and quasiparticle effects for in-cavity light and stray light are illustrated by plotting Δ_b vs T_{eff} [Fig. 4(k)]. For stray light, the traces for times before and after the pulse (t_1 and t_4) show a large frequency blueshift and small temperature increase. The stray light data traces for times during the pulse (t_2 and t_3) tend to move toward the curve described by the thermal data. For resonant in-cavity light, however, the trace is shifted to both larger blueshift and higher effective temperature. This can be interpreted in the sense that stray light causes global heating of the chip that corresponds most closely to an equilibrated thermal state. In-cavity light, however, deviates strongly from thermal equilibrium because it generates a localized heat source with a stronger dielectric component. An additional reason for the large temperature offset between dielectric and quasiparticles may be the fact that the quasiparticle bath is cooled by the circulator and isolator at the mixing chamber plate (MXC) via the microwave bus line, acting as a heat sink [56].

The suggested heating pathways are summarized in Fig. 4(l), showing the quasiparticle (QP) bath, which mostly affects the inductive part of the resonator (L) and the dielectric bath, which primarily affects the capacitor (C). Both baths are driven with different amplitudes by stray light and in-cavity light and decay to the surrounding chip substrate and package. Finally, the microwave bus couples a cold bath at the cryostat base temperature to the resonator. A detailed analysis of the bath temperatures, coupling rates, and resulting added noise of our transducer goes beyond the scope of this work and will be the subject of future studies [51,57].

VI. PULSED OPERATION AND POWER DEPENDENCE

Finally, we investigate the linearity of the transduction efficiency with optical pump power in both pulsed and continuous-wave operation. For pulsed optical-to-microwave transduction, we combine amplified pump pulses with pulses of a weaker probe laser that is frequency locked to the blue sideband of the pump laser [Fig. 5(a)]. A fraction of the pump and probe pulses are measured using a calibrated heterodyne detector to determine their respective power levels [Fig. 5(b) and Appendix D 2]. In this experiment, with increasing pump power, the pulse length is decreased from 1 ms to 40 μs to both maintain good signal-to-noise ratio at low pump power and decrease the duty cycle to keep the average power constant. The repetition rate is fixed at 0.8 kHz. The transduced microwave pulses are amplified and recorded on an electrical spectrum

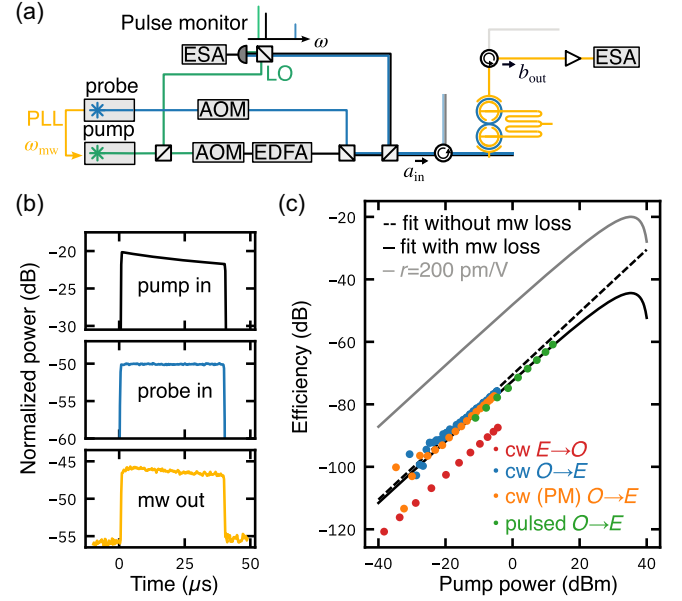


FIG. 5. Pulsed transduction and power dependence. (a) For pulsed optical-to-microwave transduction, a blue-detuned probe laser is locked to the pump laser at a frequency offset $\omega_{\text{mw}} = \omega_b$. Each laser is pulsed using an AOM, and the pump laser is additionally amplified with an EDFA. Both are combined and sent to the device via a_{in} . The transduced signal b_{out} is amplified and detected with an ESA. A fraction of the optical input signals is combined with a LO and detected via a photodetector and an ESA using heterodyne detection for pulse monitoring. (b) Pulse monitor time traces for the pump (black), probe (blue), and transduced microwave signal pulse (yellow). (c) Continuous-wave microwave-to-optical (red dots) and optical-to-microwave (blue and orange dots) off-chip efficiency vs optical pump power, using heterodyne detection, two offset-locked lasers and a phase modulator, respectively. (Green dots) Pulsed optical-to-microwave off-chip efficiency vs optical pump power using two phase-locked lasers. The black line is a model prediction of the efficiency vs optical pump power taking into account the degradation of the microwave quality factor determined in Appendix F 3, Fig. 28(c).

analyzer [Fig. 5(b)]. Note that the shapes of the pump and probe pulses are imprinted onto the transduced signal.

Figure 5(c) shows off-chip conversion efficiency as a function of off-chip peak optical pump power for cw microwave-to-optical and optical-to-microwave transduction, as well as pulsed optical-to-microwave transduction. Clearly, all optical-to-microwave data agree within 3 dB and follow a linear trend from pump powers of -30 to $+12$ dBm. As discussed in Sec. IV, cw microwave-to-optical transduction has a lower off-chip efficiency due to inefficient signal extraction in the backward direction.

The pulsed optical-to-microwave transduction data show clean, linear behavior up to the largest pump power of $+12.5$ dBm, reaching an off-chip efficiency of 1×10^{-6} . Higher pump powers could not be delivered to the device, as the fiber-chip efficiency degraded suddenly

and irreversibly, likely due to thermal misalignment of the fiber where it was glued to the chip. Damage to the transducer device itself can be ruled out since optical and microwave quality factors did not degrade.

To put the reported efficiencies into perspective, we plot a linear trend [Fig. 5(c)] for total efficiency versus pump power corresponding to $\eta/P_{\text{pump}} = -73$ dB/mW, which is about 5 orders of magnitude lower than the recently reported value for an optimized LiNbO₃ transducer (-23 dB/mW) [25]. This difference can be largely attributed to the maturity of the LiNbO₃ platform and its highly optimized optical quality factor, which enters the efficiency quadratically. Given the relatively low Pockels coefficient found for our device, an improved BaTiO₃ transducer achieving the previously observed $r = 200$ pm/V [37] would result in a more than 20 dB increase in transduction efficiency [Fig. 5(c)]. We use the measured decrease of microwave quality factor with optical power [Fig. 4(f) and Appendix F 3] to estimate when optically induced microwave decoherence limits transduction efficiency and estimate that this happens only above 30 dBm, an optical power that is far beyond the range where low-noise transduction is possible.

VII. CONCLUSION

We have presented an integrated, triply resonant electro-optic transducer using the soft ferroelectric BaTiO₃, exploiting its bias-dependent Pockels effect. In bidirectional microwave-optical conversion measurements, the efficiencies for both continuous-wave and pulsed pumping showed high linearity up to pump powers of -2 and $+12$ dB m, respectively. Heating of the transducer under pulsed optical pumping was studied using time-resolved microwave spectroscopy, allowing us to distinguish quasiparticle and dielectric heating by the induced red or blue microwave frequency shift, respectively. Overall, optical heating was found to be dominated by in-cavity light, with stray light contributing less than 10% of the temperature increase. Interestingly, for in-cavity light, in contrast to stray light, we observed stronger dielectric heating and weaker quasiparticle heating, which we attribute to the different spatial intensity distribution and the use of a thick superconducting film.

Although the maximum observed off-chip pulsed transduction efficiency of -60 dB (-73 dB/mW pump power) is low in this first demonstration, a number of improvements in device fabrication and design may significantly increase the efficiency of this novel platform. Through process optimization or strain engineering [49,58], it should be possible to increase the low-temperature effective Pockels coefficient from the current value of 13 pm/V to the level reported previously in the literature of 200 pm/V [37], enhancing efficiency by approximately 20 dB. The intrinsic optical quality factor is limited by sidewall scattering and could potentially be improved with better fabrication processes and use of wider waveguides [59]

from the present value of 1×10^5 to the absorption-limited value of 3.8×10^6 [60], further increasing the efficiency by approximately 30 dB. In addition, adjusting the device geometry should eliminate the mismatch between photonic-molecule resonant splitting and the microwave resonance frequency, providing another 4 dB improvement. A more efficient usage of the angle-dependent Pockels effect can be achieved in a racetrack resonator geometry instead of circular rings, potentially increasing the efficiency by 6 dB [41]. Finally, off-chip efficiency can be improved by reducing fiber-to-chip coupling losses from -5 to -1 dB through optimized waveguide tapers and chip facets. Achieving all improvements simultaneously would boost microwave-optical cooperativity by up to 60 dB and external efficiency by 4 dB, yielding, in principle, an off-chip efficiency approaching unity.

In addition to improving efficiency, it is crucial for high-fidelity quantum transduction that input-referred added noise be maintained well below one photon [15,21,26,55]. This study reveals submicrosecond in-cavity dielectric heating as a significant noise source, which we distinguish from quasiparticle heating using the observed microwave frequency shift, as these noise sources cannot readily be distinguished from their response times alone. Our analysis of optically induced broadening of the microwave resonance has shown that the effective temperature of the microwave resonator rapidly increases to over 1 K at 0 dBm peak optical power, corresponding to three photons of added noise. To validate this estimate, we plan to carry out a more detailed investigation of added noise in our transducer [26,51,57].

By providing an integrated means to pole the electro-optic material with a dc-bias field, our approach to cavity electro-optics enables the use of soft ferroelectrics with nonlinear Pockels behavior. However, even for hard ferroelectrics, e.g., LiNbO₃, the dc-bias port is an important advancement over previous designs. It provides a means to adjust the optical resonance frequencies of the device, as a voltage applied to this port shifts the entire optical spectrum of the photonic molecule. This may be useful for remote entanglement [15,61], where the optical photons emitted by two or more transducers must be indistinguishable, or for wavelength-division multiplexing, where multiple devices need to be aligned with different communication channels at separate frequencies.

Our numerical simulations show that the large Pockels coefficients of perovskite titanates, despite their high permittivities, can be harnessed to yield vacuum electro-optic coupling rates significantly exceeding those achievable with LiNbO₃. We emphasize that both the design and the fabrication process can be transferred with minor modifications to other material platforms, e.g., LiNbO₃ or SrTiO₃, and can be used with electrode layouts comprising crossings of either air- or oxide-clad optical structures, or both. To fully exploit the potential of soft ferroelectrics for electro-optic quantum transduction, future

studies should investigate methods to both reduce optical loss and maximize the Pockels response at low temperature [58]. This first demonstration opens up a route to microwave-optical quantum transducers with stronger electro-optic nonlinearity, a prerequisite for high-fidelity optical quantum links between superconducting qubits.

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C. M. designed the transducer with support from A. R.; C. M. and U. D. developed the fabrication process with support from A. R. and A. O.; D. C. and C. M. fabricated the samples with support from U. D., A. O., and A. R.; M. G. and C. M. set up and carried out coupling rate calculations with guidance from T. K.; C. M. performed the measurements with guidance from T. K. and D. I.; C. M. analyzed the data supervised by T. K. and D. I.; T. K. developed the theoretical model of the transducer. T. K. and D. I. designed and built the experimental setup with support from C. M., A. R., and M. G.; C. M. and T. K. wrote the paper with input from all authors. P. S. conceived and supervised the project.

The authors declare no conflicts of interest.

DATA AVAILABILITY

The data that support the findings of this article are openly available [62].

APPENDIX A: ELECTRO-OPTIC COUPLING STRENGTH

Here, we derive expressions for the electro-optic coupling in our device and provide details on the numerical simulations.

1. Field quantization

First, we recall that the classical electromagnetic energy density ρ_{em} is defined as the absolute value of

the Poynting vector [63],

$$\rho_{\text{em}} = \frac{1}{2}(\mathbf{E} \cdot \mathbf{D} + \mathbf{H} \cdot \mathbf{B}), \quad (\text{A1})$$

with electric field $\mathbf{E}$, electric displacement field $\mathbf{D} = \epsilon_0 \epsilon_r \mathbf{E}$, vacuum permittivity ϵ_0 , relative permittivity ϵ_r , magnetic field $\mathbf{H}$, magnetic flux density $\mathbf{B} = \mu_0 \mu_r \mathbf{H}$, vacuum permeability μ_0 , and relative permeability μ_r . The vacuum speed of light is $c_0 = 1/\sqrt{\epsilon_0 \mu_0}$. Assuming, for simplicity, an isotropic, nonmagnetic ($\mu_r = 1$) dielectric medium with scalar ϵ_r , we can simplify the energy density to

$$\rho_{\text{em}} = \frac{1}{2} \epsilon_0 (\epsilon_r |\mathbf{E}|^2 + c_0^2 |\mathbf{B}|^2), \quad (\text{A2})$$

which is the sum $\rho_{\text{em}} = \rho_e + \rho_m$ of the electrical and magnetic field contributions, i.e.,

$$\rho_e = \frac{1}{2} \epsilon_0 \epsilon_r |\mathbf{E}|^2, \quad (\text{A3})$$

$$\rho_m = \frac{1}{2} c_0^2 |\mathbf{B}|^2. \quad (\text{A4})$$

The total electromagnetic energy can then be obtained by integration, i.e.,

$$W_{\text{em}} = \int \rho_{\text{em}} dx dy dz, \quad (\text{A5})$$

$$W_e = \int \rho_e dx dy dz, \quad (\text{A6})$$

$$W_m = \int \rho_m dx dy dz. \quad (\text{A7})$$

For lumped-element capacitors and inductors, we have

$$W_e = \frac{1}{2} C V^2, \quad (\text{A8})$$

$$W_m = \frac{1}{2} L I^2, \quad (\text{A9})$$

where C is the capacitance, V is the voltage across the capacitor, L is the inductance, and I is the current flowing through the inductor. In the following, we consider time-averaged energy only, for which holds $\langle W_{\text{em}} \rangle = 2 \langle W_e \rangle$, such that we can focus on the electric energy.

To quantize the cavity electro-optic system, we consider the microwave and optical eigenmodes as harmonic oscillators with ladder operators a and b , and resonance frequencies ω_a and ω_b , respectively. Their respective uncoupled Hamiltonians are

$$H_a = \hbar\omega_a \left(a^\dagger a + \frac{1}{2} \right), \quad (\text{A10})$$

$$H_b = \hbar\omega_b \left(b^\dagger b + \frac{1}{2} \right). \quad (\text{A11})$$

The quantized electric field operators, in a rotating frame at the optical frequency ω_a , can be defined as [64]

$$\mathbf{E}_a = \sqrt{\frac{\hbar\omega_a}{2\epsilon_0 U_a}} (a \mathbf{u}_a e^{-i\omega_a t} + a^\dagger \mathbf{u}_a^* e^{+i\omega_a t}), \quad (\text{A12})$$

$$\mathbf{E}_b = \sqrt{\frac{\hbar\omega_b}{2\epsilon_0 U_b}} (b \mathbf{u}_b + b^\dagger \mathbf{u}_b^*), \quad (\text{A13})$$

with the spatial mode functions $\mathbf{u}_i$ and the effective volumes $U_i = \int \epsilon_r |\mathbf{u}_i|^2 dx dy dz$ for $i \in [a, b]$. Here, it is useful to define the field normalization factors

$$N_i = \sqrt{\frac{\hbar\omega_i}{2\epsilon_0 U_i}}. \quad (\text{A14})$$

Inserting the expressions (A12) and (A13) for the field operators into the time-averaged energy $W_{\text{em}} = 2W_e$ yields the Hamiltonian Eqs. (A10) and (A11), respectively.

2. Electro-optic interaction Hamiltonian

Based on Refs. [22–24], we first consider the optical electric energy density and express it in terms of the displacement field $D^i = \epsilon_r^{ij} E^j$,

$$\rho_{\text{el,opt}} = \frac{1}{2} \epsilon_0 E_{\text{opt}}^i \epsilon_{r,\text{opt}}^{ij} E_{\text{opt}}^j \quad (\text{A15})$$

$$= \frac{1}{2} \epsilon_0 D_{\text{opt}}^i (\epsilon_{r,\text{opt}}^{-1})^{ij} D_{\text{opt}}^j. \quad (\text{A16})$$

Here and in what follows, it is implicit that indices appearing twice are summed over.

We can now introduce the electro-optic interaction as a perturbation of the optical energy density from a microwave field $\mathbf{E}_{\text{mw}}$ that couples to the optical impermeability tensor $(\epsilon_{r,\text{opt}}^{-1})^{ij} \rightarrow (\epsilon_{r,\text{opt}}^{-1})^{ij} + r^{ijk} E_{\text{mw}}^k$. While the first term is just the linear energy density, the second one describes the interaction energy. We can thus define the electro-optic interaction Hamiltonian

$$H_{\text{EO}} = \frac{1}{2} \epsilon_0 \int r^{ijk} D_{\text{opt}}^i D_{\text{opt}}^j E_{\text{mw}}^k dx dy dz. \quad (\text{A17})$$

Moving back to electric fields, we have

$$H_{\text{EO}} = \frac{1}{2} \epsilon_0 n_{\text{opt}}^4 \int_{\text{EO}} r^{ijk} E_{\text{opt}}^i E_{\text{opt}}^j E_{\text{mw}}^k dx dy dz, \quad (\text{A18})$$

where $\int_{\text{EO}}$ denotes the integral over the volume of the electro-optic material, where $r \neq 0$. Inserting the field operators $\mathbf{E}_a$ and $\mathbf{E}_b$ from Eqs. (A12) and (A13) for the optical and microwave modes, respectively, we obtain

$$\begin{aligned} H_{\text{EO}} &= \frac{1}{2} \epsilon_0 n_{\text{opt}}^4 \int_{\text{EO}} r^{ijk} E_a^i E_a^j E_b^k dx dy dz \\ &= \epsilon_0 n_{\text{opt}}^4 N_a^2 N_b \\ &\quad \times \int_{\text{EO}} r^{ijk} (u_a^i)^* u_a^j a^\dagger a (u_b^k b + (u_b^k)^* b^\dagger) dx dy dz. \end{aligned} \quad (\text{A19})$$

In the last equation, we kept only terms conserving the optical photon number. We finally arrive at an expression for the vacuum coupling strength g_0 . Since the microwave spatial mode function u_b is real valued, we have $H_{\text{EO}} = \hbar g_0 a^\dagger a (b + b^\dagger)$ for a single optical ring resonator, with [23]

$$\hbar g_0 = \epsilon_0 n_{\text{opt}}^4 N_a^2 N_b \int_{\text{EO}} r^{ijk} (u_a^i)^* u_a^j u_b^k dx dy dz. \quad (\text{A20})$$

Considering only a single in-plane field component, as is approximately the case in our device, we get

$$\hbar g_0 = \epsilon_0 n_{\text{opt}}^4 N_a^2 N_b \int_{\text{EO}} r |u_a|^2 u_b dx dy dz. \quad (\text{A21})$$

We note that this describes the coupling between the microwave mode and one of the optical ring resonators. The full electro-optic interaction Hamiltonian contains the contributions of both ring resonators and takes into account that the $\lambda/2$ -type microwave mode function u_b [Fig. 7(a)] is equal and opposite at both rings, i.e.,

$$H_{\text{EO}} = \hbar g_0 (a_1^\dagger a_1 - a_2^\dagger a_2) (b + b^\dagger). \quad (\text{A22})$$

3. Separating longitudinal and cross-sectional integration

Since the overlap only happens in the active material of the optical rings, we can separate the integration into a longitudinal (dz) and cross-sectional part ($dx dy$). For the longitudinal integral along one optical ring resonator with radius R , we get (neglecting curvature)

$$\int r dz = r \alpha_{\text{overlap}} L_{\text{opt}}, \quad (\text{A23})$$

with total optical length of $L_{\text{opt}} = 2\pi R$ being the ring circumference. The factor α_{overlap} describes the effective overlap fraction of the microwave and optical fields, which takes into account the angle dependence of the Pockels effect. Motivated by the results reported in Refs. [37,41], we model the angular dependence as a sinusoidal function

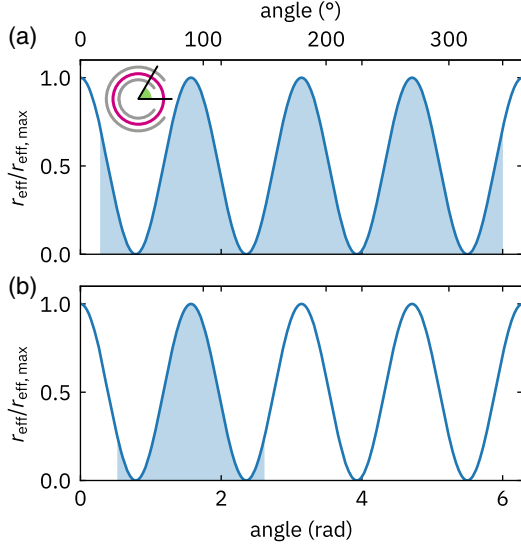


FIG. 6. Angular dependence of Pockels effect. Model for the angular dependence of the Pockels effect (blue solid line) and visualization of the overlap factor α_{overlap} as the blue shaded area under the curve for (a) the transducer electrode and (b) the tuning electrode, corresponding to 42.0% and 13.2%, respectively.

with a maximum every 90° (blue solid line in Fig. 6). The blue areas under the curves mark the angular interval of the rings, which is covered by the transducer electrodes [Fig. 6(a)] and tuning electrodes [Fig. 6(b)]. The transducer electrodes cover an interval of $2\pi\alpha_{\text{coverage}} = 2\pi \times 0.91$, but due to the angle dependence, the effective overlap is only $\alpha_{\text{overlap}} = 0.42$. We thus obtain

$$\hbar g_0 = L_{\text{opt}} \alpha_{\text{overlap}} \epsilon_0 r n_{\text{opt}}^4 N_a^2 N_b \int_{\text{EO}} |u_a|^2 |u_b| dx dy. \quad (\text{A24})$$

Note that we replaced the complex amplitude of the microwave field by its absolute value to ensure that no arbitrary phase in the numerical simulation leads to issues with the integration. Furthermore, the EO subscript for the integral denotes that we integrate only over the electro-optic material cross-sectional surface, as the electro-optic coefficient vanishes everywhere else.

The frequency pulling factor $G = \partial\omega_{\text{opt}}/\partial V = g_0/V_{\text{ZPF}}$ can be calculated analogously. We note that $N_b u_b$ is the zero-point microwave electric field. Noting that

$$\frac{N_b u_b}{V_{\text{ZPF}}} = \frac{E_b}{V_{\text{sim}}}, \quad (\text{A25})$$

where for E_b we use the simulated microwave field for the applied voltage V_{sim} , we get

$$G = \frac{L_{\text{opt}} \alpha_{\text{overlap}} \epsilon_0 r n_{\text{opt}}^4 N_a^2}{\hbar V_{\text{sim}}} \int_{\text{EO}} |u_a|^2 |E_b| dx dy. \quad (\text{A26})$$

The only quantities left to compute are the field normalization factors. For the optics, this is straightforward by integrating the cross-sectional field distribution,

$$N_a = \sqrt{\frac{\hbar \omega_a}{2\epsilon_0 L_{\text{opt}} \int n_{\text{opt}}^2 |\mathbf{u}_a|^2 dx dy}}. \quad (\text{A27})$$

4. Microwave equivalent circuit

An analogous normalization factor equation like Eq. (A27) holds for the microwave field as well, but requires simulating the whole microwave mode [Fig. 7(a)], which is computationally expensive. Instead, we simulate only the fraction of the microwave mode in the transducer electrode and normalize it with the fraction of the energy that is contained in that transducer electrode, i.e., the ring. We do this by introducing the participation ratio [65]

$$\rho_c = \frac{W_{e,\text{ring}}}{W_{e,\text{total}}}. \quad (\text{A28})$$

As illustrated by the transducer equivalent circuit [Fig. 8(a)], this corresponds to the ratio of the capacitive energies and, hence, the capacitances, i.e.,

$$\rho_c = \frac{C_{\text{ring}}}{2(C_{\text{ring}} + C_{L,\text{mw}})}, \quad (\text{A29})$$

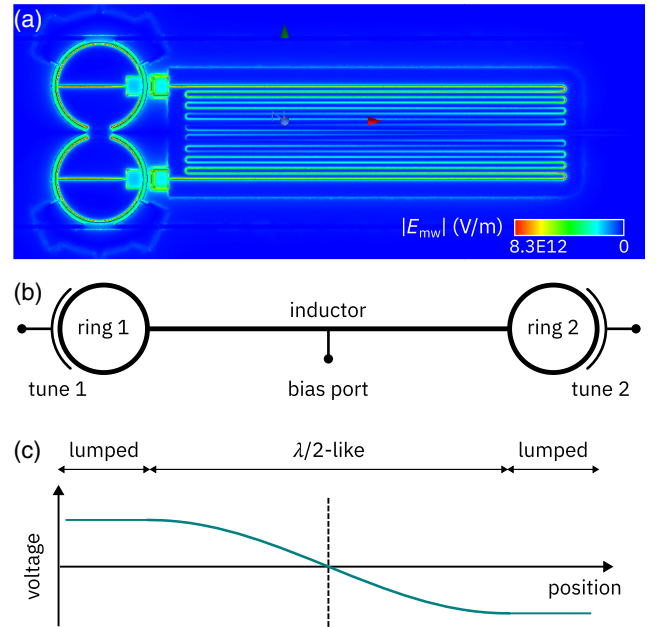


FIG. 7. Microwave mode. (a) Three-dimensional finite-element eigenmode simulation of the microwave mode field distribution performed with the ANSYS high-frequency structure simulator (HFSS). (b) Schematic of the electrical circuit with unfolded meander inductor and with lumped (ring) capacitances at each end. (c) Sketch of the $\lambda/2$ -type field distribution in the inductor according to (a) that vanishes at the bias port.

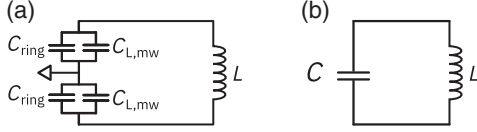


FIG. 8. Participation ratio. (a) Standard equivalent circuit and (b) extended circuit representation of the transducer device. The ring capacitances C_{ring} go from one inductor end to ground, and the stray capacitance from the inductor is modeled as parallel capacitances $C_{L,\text{mw}}$, also to ground.

where C_{ring} is the electrode capacitance at each ring, and $C_{L,\text{mw}}$ is the parasitic capacitance of each inductor arm between ring electrode and bias port. Additionally, we need to take into account that the transducer electrode overlaps only with the fraction α_{coverage} of the optical ring resonator. This gives the corrected normalization

$$N_b = \sqrt{\frac{\rho_c \hbar \omega_b}{2\epsilon_0 L_{\text{opt}} \alpha_{\text{coverage}} \int \epsilon_r |\mathbf{u}_b|^2 dx dy}}. \quad (\text{A30})$$

To obtain the participation ratio ρ_c , we first determine the capacitance of each ring electrode C_{ring} . On the one hand, we can do this in an electrostatic simulation of the whole device using the finite-element solver ANSYS MAXWELL. The resulting dc capacitances of the different circuit elements are listed in Table I. The ring capacitance is about 57.3 fF. On the other hand, we can directly compute the capacitance from the field distribution $\mathbf{u}_b$ of a Comsol finite-element simulation by equating the capacitive energy with the applied simulation voltage to the energy of the simulated field,

$$W_{e,\text{ring}} = \frac{1}{2} C_{\text{ring}} V_{\text{sim}}^2 \quad (\text{A31})$$

$$= \frac{1}{2} \epsilon_0 \alpha_{\text{coverage}} 2\pi R \int \epsilon_r |\mathbf{u}_b|^2 dx dy. \quad (\text{A32})$$

TABLE I. Simulated dc capacitances. Three-dimensional finite-element simulation of dc-capacitance matrix of the electro-optic transducer using ANSYS MAXWELL. Conductor labels correspond to those in Fig. 7(b). The active material permittivity is set to $\epsilon_r = 200$ in this case.

fF	ring 2	inductor	ring 1	ground	tune 1	tune 2
ring 2	57.3	-2.1	-1.6	-36.8	-0.4	-16.4
inductor	-2.1	177.4	-2.1	-170.5	-1.4	-1.4
ring 1	-1.6	-2.1	57.3	-36.8	-16.5	-0.4
ground	-36.8	-170.5	-36.8	343.2	-49.5	-49.5
tune 1	-0.4	-1.4	-16.5	-49.5	67.8	-0.1
tune 2	-16.4	-1.4	-0.4	-49.5	-0.1	67.7

Equation (A32) can be solved for the ring capacitance,

$$C_{\text{ring}} = \frac{\epsilon_0 \alpha_{\text{coverage}} 2\pi R}{V_{\text{sim}}^2} \int \epsilon_r |\mathbf{u}_b|^2 dx dy. \quad (\text{A33})$$

Using a Comsol cross-sectional simulation, we obtain $C_{\text{ring}} = 61.1$ fF for the ring electrode, which is fairly close to the value obtained from the three-dimensional dc finite-element simulation (Table I).

We are now missing only the value of the microwave stray capacitance of the inductor arm $C_{L,\text{mw}}$ which can, in principle, be estimated from the dc-capacitance matrix listed in Table I. Instead, we determine $C_{L,\text{mw}}$ from the equivalent circuit capacitance C [Fig. 8(b)]. An accurate value for the inductance L and the equivalent circuit capacitance C can be obtained by performing a three-dimensional finite-element eigenmode simulation of the full circuit, but with an added parallel capacitance C_p . The simulation setup is shown in Fig. 9(a), where the coupling

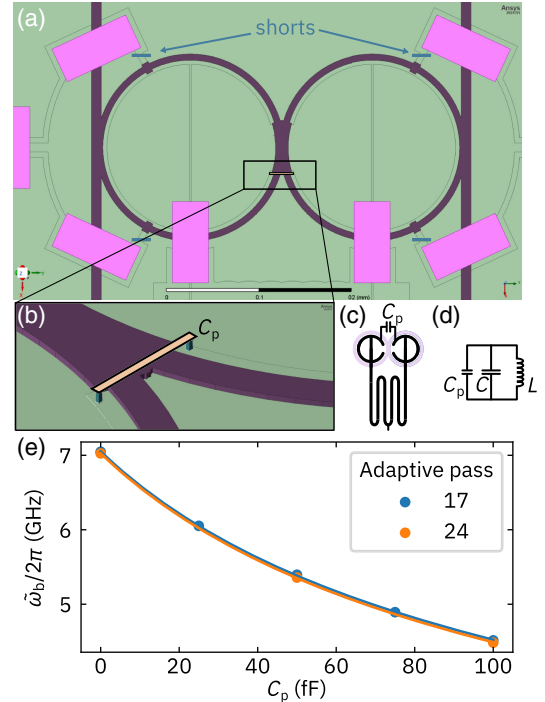


FIG. 9. Extraction of geometric inductance and capacitance using a parallel capacitance. (a) ANSYS HFSS simulation geometry of the capacitive part of the circuit with patches (blue) to short the coupling electrodes to ground. (b) Out-of-plane pins (blue) connected with a two-dimensional sheet (orange) which is defined as a lumped-element boundary condition with capacitance C_p . (c) Schematic representation of the modified device circuit with the added parallel capacitance C_p between the ring electrodes. (d) Equivalent circuit representation with the parallel capacitance in addition to the equivalent capacitance C and inductance of the device L . (e) Resonance frequency of the device as a function of parallel capacitance C_p using the mesh after 17 (blue dots) and 24 adaptive passes (orange dots) and fit to Eq. (A34).

capacitances were set to ground by shorting with small patches (blue). The parallel capacitance is then implemented as a bridge between the two ring electrodes at the shortest possible distance [Fig. 9(b)]. The bridge is implemented using two blocks (blue) that connect the ring electrodes to a two-dimensional sheet, on which a lumped-element capacitance boundary condition with value C_p is defined. A schematic of the modified device circuit is shown in Fig. 9(c). Effectively, this represents an additional parallel capacitance C_p [Fig. 9(d)] to the equivalent circuit capacitance C [Fig. 8(b)].

The resonance frequency of this modified circuit is then simulated as a function of the parallel capacitance C_p , as shown in Fig. 9(e). The advantage of using a lumped element is that the mesh for the simulation needs only to be calculated once and can then be used to simulate for all values of C_p . The observed dependence of the resonance frequency on the parallel capacitance is modeled as

$$\tilde{\omega}_b = \sqrt{\frac{1}{L(C + C_p)}}, \quad (\text{A34})$$

with the inductance L and the equivalent capacitance C of the device as fit parameters. The results are listed in Table II. We can then use the fact that the circuit is symmetric, and together with the rules for series and parallel capacitances, we have

$$C = \frac{C_{\text{ring}} + C_{L,\text{mw}}}{2}, \quad (\text{A35})$$

which we can solve for the inductor's capacitance

$$C_{L,\text{mw}} = 2C - C_{\text{ring}}, \quad (\text{A36})$$

which equates to $C_{L,\text{mw}} = 80.7$ fF using the value of C_{ring} from Table I. With this, we arrive at a participation ratio of $\rho_c = 0.43$.

Finally, we evaluate Eq. (A30) replacing u_b with the simulated field E_b and using Eqs. (A33) and (A29), which gives

TABLE II. Simulated microwave equivalent circuit parameters. Results of fitting Eq. (A34) to the data obtained from three-dimensional finite-element eigenmode simulations of the electro-optic transducer with a parallel capacitance using ANSYS MAXWELL [Fig. 9(a)].

C (fF)	L (nH)
69.0	7.4

$$\begin{aligned} N_b &= \sqrt{\frac{\rho_c \hbar \omega_b}{2C_{\text{ring}} V_{\text{sim}}^2}} = \sqrt{\frac{\hbar \omega_b}{4(C_{\text{ring}} + C_{L,\text{mw}})} \frac{1}{|V_{\text{sim}}|}} \\ &= \frac{V_{\text{ZPF}}}{|V_{\text{sim}}|}. \end{aligned} \quad (\text{A37})$$

We thus obtain the zero-point voltage

$$V_{\text{ZPF}} = \sqrt{\frac{\hbar \omega_b}{8C}}, \quad (\text{A38})$$

dropping across one ring electrode. The effective microwave voltage operator [35] can be written as

$$V_b = V_{\text{ZPF}}(b + b^\dagger). \quad (\text{A39})$$

5. Incorporating scaling with material permittivity

In order to fairly compare the vacuum coupling strength of the same device design and cross-sectional geometry, but for different active materials with similar optical properties but sometimes vastly different permittivities ($\epsilon_{r,\text{LiNbO}_3} = 30$, $\epsilon_{r,\text{BaTiO}_3} = 2 \times 10^2$, $\epsilon_{r,\text{SrTiO}_3} \approx 2 \times 10^4$), we incorporate the scaling of the inductor, which affects both the inductance and stray capacitance of the circuit, and thereby also the total capacitance and participation ratio. We do this by requiring a constant resonance frequency ω_b of the transducer

$$\omega_b = \frac{1}{\sqrt{LC}}. \quad (\text{A40})$$

We then assume that, by varying the length of the meander inductor wire by the factor α_{ind} , its inductance and capacitance vary proportionally, such that the resonance frequency scales as

$$\omega_b = \frac{1}{\sqrt{\alpha_{\text{ind}} L (C_{\text{ring}} + \alpha_{\text{ind}} C_{L,\text{mw}})/2}}. \quad (\text{A41})$$

We can solve this quadratic equation for α_{ind} of the form $0 = p_1 x^2 + p_2 x + p_3$ with the coefficients

$$p_1 = LC_{L,\text{mw}}, \quad (\text{A42})$$

$$p_2 = LC_{\text{ring}}, \quad (\text{A43})$$

$$p_3 = -\frac{2}{\omega_b^2}. \quad (\text{A44})$$

The inductor scaling parameter is then determined via

$$\alpha_{\text{ind}} = \frac{-p_1 \pm \sqrt{p_2^2 - 4p_1 p_3}}{2p_1}, \quad (\text{A45})$$

where we use the positive solution.

APPENDIX B: LINEAR SYSTEM RESPONSE

1. General picture

In this appendix, we give an introduction to the input-output formalism and coupled-mode theory to describe the linearized response of the cavity electro-optic system studied in this work. We describe the system as a collection of resonant modes with annihilation operators q_i ($q_i \in a_1, a_2, b, \dots$), resonance frequencies ω_i , and intrinsic loss rates $\kappa_{i,0}$. Some of the modes are coupled to external bus modes $q_{i,\text{in}}$ ($q_{i,\text{out}}$) for input (output) at the external coupling rates $\kappa_{i,\text{ex}}$. Total linewidths are denoted $\kappa_i = \kappa_{i,0} + \kappa_{i,\text{ex}}$. Additionally, there are pairwise coherent couplings characterized by (possibly complex-valued) couplings g_{ij} (f_{ij}) for counterrotating (corotating) terms. The generic Hamiltonian reads

$$H = \sum_i \hbar \omega_i q_i^\dagger q_i + \sum_{i < j} \hbar \left(g_{ij} q_i^\dagger q_j + g_{ij}^* q_i q_j^\dagger \right) + \sum_{i < j} \hbar \left(f_{ij} q_i^\dagger q_j^\dagger + f_{ij}^* q_i q_j \right). \quad (\text{B1})$$

The Heisenberg equations of motion including damping and external inputs read

$$\dot{q}_i = -i\omega_i q_i - \frac{\kappa_i}{2} q_i - \sqrt{\kappa_{i,\text{ex}}} q_{i,\text{in}} - i \sum_{j \neq i} (g_{ij} q_j + f_{ij} q_j^\dagger). \quad (\text{B2})$$

Using the vector notation

$$\mathbf{q} = [q_1, q_1^\dagger, q_2, q_2^\dagger, \dots]^T, \quad (\text{B3})$$

we can write Eq. (B2) in matrix form, i.e.,

$$\dot{\mathbf{q}} = M' \mathbf{q} - K \mathbf{q}_{\text{in}}, \quad (\text{B4})$$

where M' is the time-domain dynamical matrix and K is the input-output matrix defined by Eq. (B2). The input-output relations $q_{i,\text{out}} = q_{i,\text{in}} + \sqrt{\kappa_{i,\text{ex}}} q_i$ have the matrix form

$$\mathbf{q}_{\text{out}} = \mathbf{q}_{\text{in}} + K^T \mathbf{q}. \quad (\text{B5})$$

In the frequency domain (with Fourier frequency ω), the equations of motion become

$$(-i\omega - M') \mathbf{q}(\omega) = -K \mathbf{q}_{\text{in}}(\omega), \quad (\text{B6})$$

where

$$\mathbf{q}(\omega) = [q_1(\omega), q_1^\dagger(-\omega), q_2(\omega), q_2^\dagger(-\omega), \dots]^T. \quad (\text{B7})$$

The solution to Eq. (B6) can be found by matrix inversion, i.e.,

$$\mathbf{q}(\omega) = -M(\omega)^{-1} K \mathbf{q}_{\text{in}}(\omega), \quad (\text{B8})$$

where $M(\omega) = -i\omega - M'$. Inserting into the input-output relation, Eq. (B5) yields

$$\mathbf{q}_{\text{out}}(\omega) = \mathbf{q}_{\text{in}}(\omega) - K^T M(\omega)^{-1} K \mathbf{q}_{\text{in}}(\omega) \quad (\text{B9})$$

$$= S(\omega) \mathbf{q}_{\text{in}}(\omega), \quad (\text{B10})$$

where we defined the S matrix

$$S(\omega) = 1 - K^T M(\omega)^{-1} K. \quad (\text{B11})$$

2. Photonic molecule with forward-backward scattering

First, we apply the formalism derived above to describe the coupled optical resonators in the transducer. As shown in Fig. 10(b), the two ring resonators are coupled evanescently at rate μ , and resonator 1 is coupled to the optical bus waveguide at rate $\kappa_{1,\text{ex}}$. Moreover, the fabricated waveguides exhibit imperfections, such as particles on the surface, surface roughness, or other scattering sites on the inside (black dots) that cause scattering to both unguided and guided modes [Fig. 10(a)]. In particular, these imperfections can couple the clockwise (c) and counterclockwise (cc) propagating optical modes at rates ν_i [66]. We note that ν_i can, in principle, be complex valued to account for incoherent, non-Hamiltonian scattering, which is manifested as asymmetric linewidths of forward-backward split optical resonances. The system thus consists of the four modes $a_{1,\text{c}}, a_{1,\text{cc}}, a_{2,\text{c}}, a_{2,\text{cc}}$ described by the Hamiltonian

$$H_a = - \sum_{i=1,2} \sum_{\sigma=c,\text{cc}} \hbar \Delta_{p,i} a_{i,\sigma}^\dagger a_{i,\sigma} - \hbar \mu \left(a_{1,\text{c}}^\dagger a_{2,\text{cc}} + a_{1,\text{cc}}^\dagger a_{2,\text{c}} \right) + \text{H.c.} - \sum_{i=1,2} \hbar \text{Re}(\nu_i) a_{i,\text{c}}^\dagger a_{i,\text{cc}} + \text{H.c.} \quad (\text{B12})$$

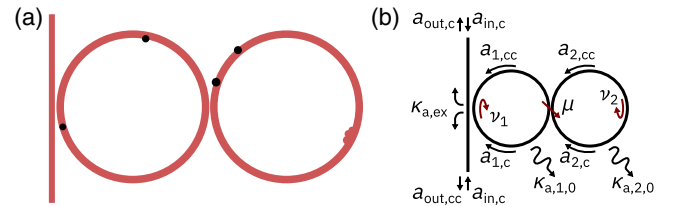


FIG. 10. Photonic molecule with backscattering. (a) Physical representation of photonic-molecule circuit, consisting of two coupled optical rings. Imperfections leading to backscattering such as scattering sites on or in the waveguide (black dots) or waveguide cross section irregularities (bottom right) are illustrated. (b) Coupling and loss rates between the involved clockwise and counterclockwise propagating resonator modes $a_{i,\text{c}}, a_{i,\text{cc}}$ ($i \in \{1, 2\}$) and in- and output modes a_{in} and a_{out} .

in the rotating frame of the pump laser, where $\Delta_{p,i} = \omega_p - \omega_i$ are the detunings of the pump laser from the optical resonances. Since Eq. (B12) does not couple $a_{i,\sigma}$ and $a_{i',\sigma'}^\dagger$ in the equations of motion, we define the state vectors

$$\mathbf{q}_a = [a_{1,c} \ a_{1,cc} \ a_{2,c} \ a_{2,cc}]^T, \quad (\text{B13})$$

$$\mathbf{q}_{a,\text{in}} = [a_{\text{in},c} \ a_{\text{in},cc}]^T, \quad (\text{B14})$$

$$\mathbf{q}_{a,\text{out}} = [a_{\text{out},c} \ a_{\text{out},cc}]^T. \quad (\text{B15})$$

To set up the equations of motion as in Eq. (B4), we define the optical input-output matrix

$$K_a = \begin{bmatrix} \sqrt{\kappa_{a,\text{ex}}} & 0 \\ 0 & \sqrt{\kappa_{a,\text{ex}}} \\ 0 & 0 \\ 0 & 0 \end{bmatrix}, \quad (\text{B16})$$

and the optical time-domain dynamic matrix

$$M'_a = \begin{bmatrix} +i\Delta_{p,1} & +i\nu_1 & 0 & +i\mu \\ +i\nu_1 & +i\Delta_{p,1} & +i\mu & 0 \\ 0 & +i\mu & +i\Delta_{p,2} & +i\nu_2 \\ +i\mu & 0 & +i\nu_2 & +i\Delta_{p,2} \end{bmatrix} + \begin{bmatrix} -\frac{\kappa_1}{2} & 0 & 0 & 0 \\ 0 & -\frac{\kappa_1}{2} & 0 & 0 \\ 0 & 0 & -\frac{\kappa_2}{2} & 0 \\ 0 & 0 & 0 & -\frac{\kappa_2}{2} \end{bmatrix}, \quad (\text{B17})$$

where the imaginary part of ν_i is included [66].

The frequency-domain dynamical matrix can then be written as

$$M_a(\omega) = \begin{bmatrix} \frac{1}{\chi_1(\omega)} & -i\nu_1 & 0 & -i\mu \\ -i\nu_1 & \frac{1}{\chi_1(\omega)} & -i\mu & 0 \\ 0 & -i\mu & \frac{1}{\chi_2(\omega)} & -i\nu_2 \\ -i\mu & 0 & -i\nu_2 & \frac{1}{\chi_2(\omega)} \end{bmatrix}, \quad (\text{B18})$$

with the optical response functions

$$\chi_i(\omega) = \frac{1}{\kappa_i/2 - i(\omega + \Delta_{p,i})}. \quad (\text{B19})$$

For optical spectroscopy in transmission or reflection, we set $\omega = 0$, and ω_p corresponds to the optical probe

frequency. With Eq. (B18) we can calculate the S -matrix elements using Eq. (B11). To fit optical resonances in transmission, we use S_{11} . In principle, S_{21} corresponds to the reflected pump amplitude. However, real spectra are dominated by other, frequency-dependent optical reflections from various interfaces, e.g., at the fiber-chip coupler, that make it difficult to reliably fit optical reflection spectra.

In the case when $\nu_i = 0$, solving for the eigenvectors of Eq. (B18) yields the symmetric and antisymmetric eigenmodes a_+ and a_- of the photonic molecule mentioned in the main text, identical for clockwise and counterclockwise propagation. When $\nu_i \neq 0$, these modes are perturbed and further split by about $\nu_1 + \nu_2$. Instead of working with the eigenmodes, we choose to treat the problem in the uncoupled basis $a_{i,\sigma}$ because it simplifies the external couplings.

3. Transduction model

We now use the formalism from Appendix B 1 to describe the full transducer by including microwave resonator and the electro-optic interaction (Fig. 11). For clarity, we divide the system Hamiltonian into its optical, microwave, and electro-optic parts, i.e.,

$$H = H_a + H_b + H_{\text{EO}}, \quad (\text{B20})$$

with microwave Hamiltonian

$$H_b = \hbar\omega_b b^\dagger b \quad (\text{B21})$$

and the electro-optic interaction Hamiltonian

$$H_{\text{EO}} = -\hbar g_0 \sum_{\sigma=c,cc} a_{1,\sigma}^\dagger a_{1,\sigma} (b + b^\dagger) + \hbar g_0 \sum_{\sigma=c,cc} a_{2,\sigma}^\dagger a_{2,\sigma} (b + b^\dagger). \quad (\text{B22})$$

Note that the sign difference in the electro-optic couplings between the two optical rings originates from the anti-symmetric microwave mode shape. Next, we linearize the electro-optic interaction assuming that the optical fields are driven to large amplitude by the pump laser [67],

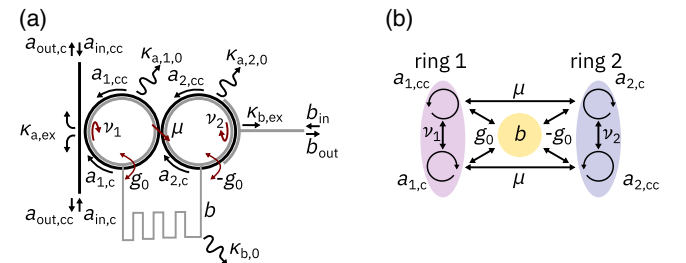


FIG. 11. Full transduction model. (a) Schematic of the device. (b) Effective mode coupling diagram.

i.e., $a_{i,\sigma} \rightarrow \alpha_{i,\sigma} + a_{i,\sigma}$ with $\alpha_{i,\sigma} = \langle a_{i,\sigma} \rangle$. This results in the linearized electro-optic interaction

$$H_{\text{EO}} = -\hbar g_0 \sum_{\sigma=c, \text{cc}} (\alpha_{1,\sigma}^* a_{1,\sigma} + \alpha_{1,\sigma} a_{1,\sigma}^\dagger)(b + b^\dagger) + \hbar g_0 \sum_{\sigma=c, \text{cc}} (\alpha_{2,\sigma}^* a_{2,\sigma} + \alpha_{2,\sigma} a_{2,\sigma}^\dagger)(b + b^\dagger). \quad (\text{B23})$$

With the full system Hamiltonian derived, we can now apply the formalism of Appendix B 1 to obtain the scattering matrix. We thus define the vectors

$$\mathbf{q} = \left[a_{1,c} \ a_{1,c}^\dagger \ a_{1,\text{cc}} \ a_{1,\text{cc}}^\dagger \ a_{2,c} \ a_{2,c}^\dagger \ a_{2,\text{cc}} \ a_{2,\text{cc}}^\dagger \ b \ b^\dagger \right]^T, \quad (\text{B24})$$

$$\mathbf{q}_{\text{in}} = \left[a_{\text{in},c} \ a_{\text{in},c}^\dagger \ a_{\text{in},\text{cc}} \ a_{\text{in},\text{cc}}^\dagger \ b_{\text{in}} \ b_{\text{in}}^\dagger \right]^T, \quad (\text{B25})$$

$$\mathbf{q}_{\text{out}} = \left[a_{\text{out},c} \ a_{\text{out},c}^\dagger \ a_{\text{out},\text{cc}} \ a_{\text{out},\text{cc}}^\dagger \ b_{\text{out}} \ b_{\text{out}}^\dagger \right]^T. \quad (\text{B26})$$

The input-output matrix for the full system is given by

$$K = \begin{bmatrix} \sqrt{\kappa_{a,\text{ex}}} & 0 & 0 & 0 & 0 & 0 \\ 0 & \sqrt{\kappa_{a,\text{ex}}} & 0 & 0 & 0 & 0 \\ 0 & 0 & \sqrt{\kappa_{a,\text{ex}}} & 0 & 0 & 0 \\ 0 & 0 & 0 & \sqrt{\kappa_{a,\text{ex}}} & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & \sqrt{\kappa_{b,\text{ex}}} & 0 \\ 0 & 0 & 0 & 0 & 0 & \sqrt{\kappa_{b,\text{ex}}} \end{bmatrix}. \quad (\text{B27})$$

The time-domain dynamical matrix is

$$M' = M'_a + M'_b + M'_{\text{EO}}, \quad (\text{B28})$$

with optical part

$$M'_a = \begin{bmatrix} +i\Delta_{p,1} & 0 & +i\nu_1 & 0 & 0 & 0 & +i\mu & 0 & 0 & 0 \\ 0 & -i\Delta_{p,1} & 0 & -i\nu_1^* & 0 & 0 & 0 & -i\mu & 0 & 0 \\ +i\nu_1 & 0 & +i\Delta_{p,1} & 0 & +i\mu & 0 & 0 & 0 & 0 & 0 \\ 0 & -i\nu_1^* & 0 & -i\Delta_{p,1} & 0 & -i\mu & 0 & 0 & 0 & 0 \\ 0 & 0 & +i\mu & 0 & +i\Delta_{p,2} & 0 & +i\nu_1 & 0 & 0 & 0 \\ 0 & 0 & 0 & -i\mu & 0 & -i\Delta_{p,2} & 0 & -i\nu_2^* & 0 & 0 \\ +i\mu & 0 & 0 & 0 & +i\nu_2 & 0 & +i\Delta_{p,2} & 0 & 0 & 0 \\ 0 & -i\mu & 0 & 0 & 0 & -i\nu_2^* & 0 & -i\Delta_{p,2} & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix} + \begin{bmatrix} -\frac{\kappa_1}{2} & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & -\frac{\kappa_1}{2} & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & -\frac{\kappa_1}{2} & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & -\frac{\kappa_1}{2} & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & -\frac{\kappa_2}{2} & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & -\frac{\kappa_2}{2} & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & -\frac{\kappa_2}{2} & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & -\frac{\kappa_2}{2} & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}, \quad (\text{B29})$$

microwave part

$$M'_b = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & -\frac{\kappa_b}{2} - i\omega_b & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & -\frac{\kappa_b}{2} + i\omega_b & 0 \end{bmatrix}, \quad (\text{B30})$$

and the electro-optic interaction part

$$M'_{\text{EO}} = g_0 \cdot \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & +i\alpha_{1,c} & +i\alpha_{1,c} \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & -i\alpha_{1,c}^* & -i\alpha_{1,c}^* \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & +i\alpha_{1,cc} & +i\alpha_{1,cc} \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & -i\alpha_{1,cc}^* & -i\alpha_{1,cc}^* \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & -i\alpha_{2,c} & -i\alpha_{2,c} \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & +i\alpha_{2,c}^* & +i\alpha_{2,c}^* \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & -i\alpha_{2,cc} & -i\alpha_{2,cc} \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & +i\alpha_{2,cc}^* & +i\alpha_{2,cc}^* \\ +i\alpha_{1,c}^* & +i\alpha_{1,c} & +i\alpha_{1,cc}^* & +i\alpha_{1,cc} & -i\alpha_{2,c}^* & -i\alpha_{2,c} & -i\alpha_{2,cc}^* & -i\alpha_{2,cc} & 0 & 0 \\ -i\alpha_{1,c}^* & -i\alpha_{1,c} & -i\alpha_{1,cc}^* & -i\alpha_{1,cc} & +i\alpha_{2,c}^* & +i\alpha_{2,c} & +i\alpha_{2,cc}^* & +i\alpha_{2,cc} & 0 & 0 \end{bmatrix}. \quad (\text{B31})$$

The frequency-domain dynamical matrix is given by

$$M(\omega) = M_a(\omega) + M_b(\omega) + M_{\text{EO}}, \quad (\text{B32})$$

with optical part

$$M_a = \begin{bmatrix} \frac{1}{\chi_1(\omega)} & 0 & -i\nu_1 & 0 & 0 & 0 & -i\mu & 0 & 0 & 0 \\ 0 & \frac{1}{\chi_1^*(-\omega)} & 0 & +i\nu_1^* & 0 & 0 & 0 & +i\mu & 0 & 0 \\ -i\nu_1 & 0 & \frac{1}{\chi_1(\omega)} & 0 & -i\mu & 0 & 0 & 0 & 0 & 0 \\ 0 & +i\nu_1^* & 0 & \frac{1}{\chi_1^*(-\omega)} & 0 & +i\mu & 0 & 0 & 0 & 0 \\ 0 & 0 & -i\mu & 0 & \frac{1}{\chi_2(\omega)} & 0 & -i\nu_2 & 0 & 0 & 0 \\ 0 & 0 & 0 & +i\mu & 0 & \frac{1}{\chi_2^*(-\omega)} & 0 & +i\nu_2^* & 0 & 0 \\ -i\mu & 0 & 0 & 0 & -i\nu_2 & 0 & \frac{1}{\chi_2(\omega)} & 0 & 0 & 0 \\ 0 & +i\mu & 0 & 0 & 0 & +i\nu_2^* & 0 & \frac{1}{\chi_2^*(-\omega)} & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}, \quad (\text{B33})$$

microwave part

$$M_b = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & \frac{1}{\chi_b(\omega)} & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & \frac{1}{\chi_b^*(-\omega)} \end{bmatrix}, \quad (\text{B34})$$

and $M_{\text{EO}} = -M'_{\text{EO}}$. Here, we also defined the microwave response function

$$\chi_b(\omega) = \frac{1}{\kappa_b/2 - i(\omega - \omega_b)}. \quad (\text{B35})$$

We can now solve the equations of motion for this linearized system. In the first step, we calculate the optical resonator pump amplitudes using Eq. (B18) by

$$\mathbf{q}_p = -M_a(0)^{-1} K_a \mathbf{q}_{\text{in},p}, \quad (\text{B36})$$

where

$$\mathbf{q}_p = [\alpha_{1,c}, \alpha_{1,cc}, \alpha_{2,c}, \alpha_{2,cc}]^T. \quad (\text{B37})$$

With the pump field injected in forward (clockwise) direction only, the pump input vector is

$$\mathbf{q}_{\text{in},p} = \sqrt{\frac{P_{p,\text{on}}}{\hbar\omega_p}} [1 \ 0 \ 0 \ 0]^T, \quad (\text{B38})$$

with the on-chip pump power $P_{p,\text{on}} = \eta_{f2c} P_p$, where P_p is the pump power in the fiber and η_{f2c} is the fiber-chip coupling efficiency.

In the second step, we insert the pump resonator amplitudes $\alpha_{i,\sigma}$ into the full matrix and solve the full problem using Eqs. (B32) and (B11). The optical sideband transmission and the microwave reflection can then be calculated as

$$T_{\text{opt}} = \left| \frac{\langle a_{\text{out},c} \rangle}{\langle a_{\text{in},c} \rangle} \right|^2 = |S_{11}(\omega)|^2, \quad (\text{B39})$$

$$R_{\text{mw}} = \left| \frac{\langle b_{\text{out}} \rangle}{\langle b_{\text{in}} \rangle} \right|^2 = |S_{99}(\omega)|^2, \quad (\text{B40})$$

respectively. The transduction efficiencies for the anti-Stokes and Stokes sidebands, respectively, are calculated using

$$\eta_{\text{AS}} = \left| \frac{\langle a_{\text{out},c} \rangle}{\langle b_{\text{in}} \rangle} \right|^2 = |S_{19}(\omega)|^2, \quad (\text{B41})$$

$$\eta_{\text{S}} = \left| \frac{\langle a_{\text{out},c}^\dagger \rangle}{\langle b_{\text{in}} \rangle} \right|^2 = |S_{29}(\omega)|^2. \quad (\text{B42})$$

4. Microwave spurious mode coupling

We now derive the linear response for the coupling between the microwave resonator and the spurious mode detailed in Appendix F. The system comprises two microwave resonators b and b_2 with resonance frequencies ω_b and $\omega_{b,2}$, and intrinsic loss rates $\kappa_{b,0}$ and $\kappa_{b,2,0}$, respectively. The main microwave mode b is coupled to the bus waveguide at a rate $\kappa_{b,\text{ex}}$. The two resonators are coupled to each other at a rate γ .

The system Hamiltonian reads

$$H = +\hbar\omega_b b^\dagger b + \hbar\omega_{b,2} b_2^\dagger b_2 + \hbar\gamma (b^\dagger b_2 + b_2^\dagger b). \quad (\text{B43})$$

We define the input and output vectors of the MW field

$$\mathbf{q}_{\text{in}} = [b_{\text{in}} \ b_{\text{in}}^\dagger]^T, \quad (\text{B44})$$

and

$$\mathbf{q}_{\text{out}} = [b_{\text{out}} \ b_{\text{out}}^\dagger]^T. \quad (\text{B45})$$

We infer the scattering matrix elements S using Eq. (B11) with the input-output matrix

$$K_b = \begin{bmatrix} \sqrt{\kappa_{b,\text{ex}}} & 0 \\ 0 & \sqrt{\kappa_{b,\text{ex}}} \\ 0 & 0 \\ 0 & 0 \end{bmatrix}, \quad (\text{B46})$$

and the time-domain dynamical matrix

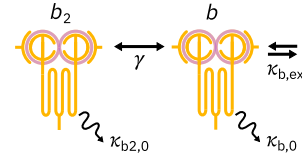


FIG. 12. Coupled microwave mode schematic. The main microwave mode b_1 is coupled to the bus waveguide at a rate $\kappa_{b,\text{ex}}$ and is additionally coupled to a spurious mode b_2 at a rate γ . The intrinsic loss rates of the two modes are $\kappa_{b,0}$ and $\kappa_{b,2,0}$, respectively.

$$M'_b = \begin{bmatrix} -i\omega_b & 0 & +i\gamma & 0 \\ 0 & +i\omega_b & 0 & -i\gamma \\ +i\gamma & 0 & -i\omega_{b2} & 0 \\ 0 & -i\gamma & 0 & +i\omega_{b2} \end{bmatrix} + \begin{bmatrix} -\frac{\kappa_b}{2} & 0 & 0 & 0 \\ 0 & -\frac{\kappa_b}{2} & 0 & 0 \\ 0 & 0 & -\frac{\kappa_{b2,0}}{2} & 0 \\ 0 & 0 & 0 & -\frac{\kappa_{b2,0}}{2} \end{bmatrix}, \quad (\text{B47})$$

from which the frequency-domain dynamical matrix follows, i.e.,

$$M_b(\omega) = \begin{bmatrix} \frac{1}{\chi_b(\omega)} & 0 & -i\gamma & 0 \\ 0 & \frac{1}{\chi_b^*(-\omega)} & 0 & +i\gamma \\ -i\gamma & 0 & \frac{1}{\chi_{b2}(\omega)} & 0 \\ 0 & +i\gamma & 0 & \frac{1}{\chi_{b2}^*(-\omega)} \end{bmatrix}. \quad (\text{B48})$$

From this, we determine the S matrix S_b . In Appendix F, we fit the microwave spectra using

$$R_{\text{mw}}(\omega) = \left| \frac{\langle b_{\text{out}} \rangle}{\langle b_{\text{in}} \rangle} \right|^2 = |S_{b,11}(\omega)|^2. \quad (\text{B49})$$

As fit function we use $R_0 R_{\text{mw}}(\omega)$ where R_0 is the off-resonant reflection level.

APPENDIX C: SAMPLE PREPARATION

1. Sample fabrication

We have developed a process to fabricate low-loss, high-confinement photonic waveguides and resonators in BaTiO₃ and integrate them with Nb superconducting circuits. We start with a 225-nm-thick BaTiO₃ thin film bonded on 3 μm of SiO₂ on silicon [Fig. 13(a)]. BaTiO₃ ridge waveguides are patterned using electron beam lithography and dry etched to produce a 100-nm-thick slab and a 125-nm-thick ridge [Fig. 13(b)]. The waveguides are protected with a 1.5-μm cladding of SiO₂ [Fig. 13(c)]. Electrodes are subsequently fabricated by sputtering Nb and patterning using electron beam lithography [Fig. 13(d)]. For waveguide-superconductor and superconductor-superconductor crossings, Nb air bridges are fabricated [Fig. 13(e)] by sputtering Nb and subsequent optical lithography [Fig. 13(e)]. In the last step, optical facets are etched such that waveguides reach to the edge of the chip for low-loss fiber coupling [Fig. 13(f) shows a side view of the chip before optical facet fabrication]. Optical facets are fabricated using optical lithography, SiO₂ dry etching, and deep reactive-ion etching into the Si substrate

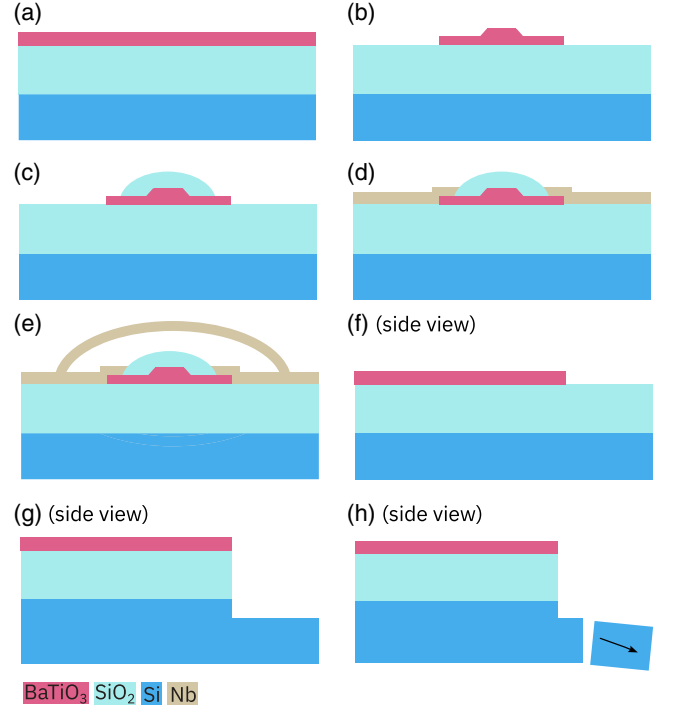


FIG. 13. Fabrication flow. (a) Initial material stack. (b) Optical ridge waveguide with etched slab. (c) Oxide-clad ridge waveguide. (d) Added main electrode layer. (e) Added superconducting air bridges. (f) Chip edge prior to facet fabrication. (g) Dry-etched facet. (h) Diced chip edge.

[Fig. 13(g)]. The chip is diced to facilitate access to the facet [Fig. 13(h)].

Using this process, a variety of elements are fabricated to produce the optical and electronic circuits constituting the transducer. Figure 14(a) shows a scanning electron micrograph (SEM) of a cross section of an optical ridge waveguide, Fig. 14(b) a close-up view of the waveguide facet, Fig. 14(c) the optical waveguide coupling region seen from the top, and Fig. 14(d) a Nb air bridge of an air-clad device. Figure 14 depicts the device under test measured in this study, with oxide-clad waveguides.

2. Packaging

The high-numerical-aperture input and output fibers (UHNA7, mode-field diameter 3.2 μm) are first aligned to the chip by optimizing the optical transmission through the bus waveguide using motorized stages. The fibers are then glued to the chip facets using an optical epoxy (Norland NOA61) and subsequently stabilized with additional gluing points 1–2 mm from the edge of the chip, as shown in Fig. 15. For mode matching with the optical fiber mode, the optical bus waveguide is first converted from a ridge waveguide with a ridge width of 1.2 μm to a rib waveguide by tapering the slab width from 6.0 to 1.2 μm over 77 μm. In a second step, the single-mode rib waveguide with a width of 1.2 μm is tapered down over 118 μm

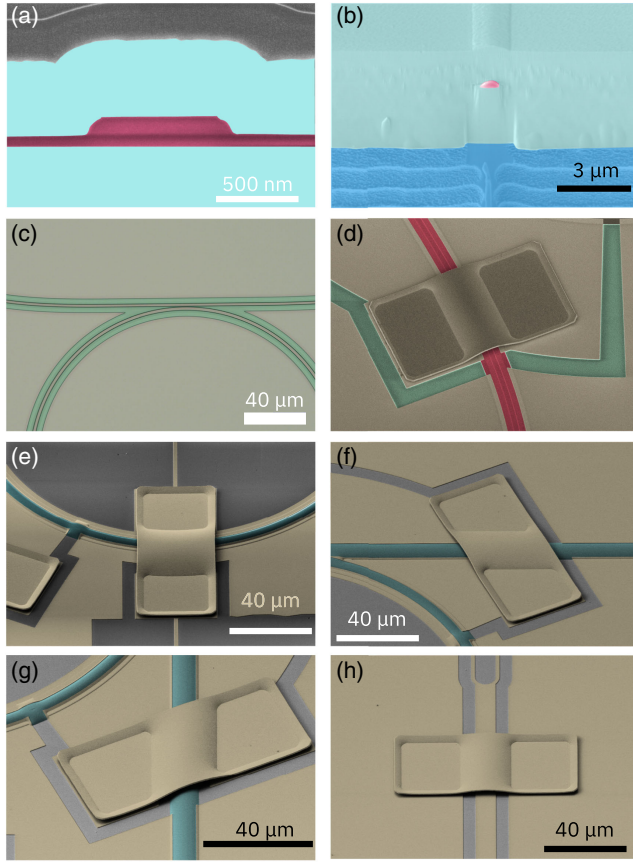


FIG. 14. Fabrication results. False-colored scanning electron micrographs: (a) Cross section of an oxide-clad BaTiO₃ ridge waveguide. (b) Optical chip facet with BaTiO₃ taper end. False-colored optical micrographs: (c) Directional coupler section between uncladded ring and bus waveguide. (d) Nb air bridge over uncladded optical waveguide. (e) Air bridge between transducer electrode and inductor over cladded waveguide. (f),(g) Air bridge at coupling capacitor from different angles. (h) Air bridge over dc bias line.

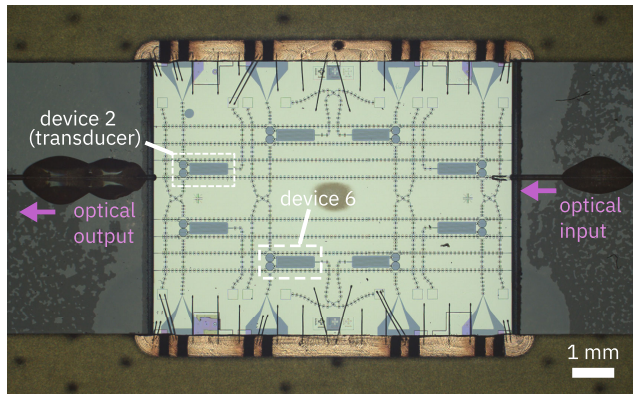


FIG. 15. Cryogenic microwave-optical copackaged chip. (a) Sample chip with optical edge-coupling gluing points. (b) Electrical on-chip circuit and wire bonds to printed circuit board. Lines to the inductor are highlighted in red and to ring electrodes in yellow.

to a width of 550 nm at the edge of the chip. The fiber-to-chip coupling efficiency is inferred to be -5 dB by dividing the total fiber-to-fiber transmission after packaging by 2 [see Fig. 2(e)]. When the packaging is successful, the coupling loss can be stable over multiple cooldown cycles.

Next, the on-chip electrical circuit is connected to the printed circuit board with wire bonds (Fig. 15). In this case, the transducer device (device 2) is fully packaged with three leads for microwave reflection, photonic-molecule tuning, and domain biasing. A second device (device 6) is packaged with only the ports for microwave reflection and domain biasing connected for microwave characterization.

APPENDIX D: EXPERIMENTAL SETUPS

1. Cryostat wiring

The wiring inside the cryostat is shown in Fig. 16. Our cryostat (BlueFors LD400) features a fast sample exchange bottom probe mechanism. While the electrical signals are delivered from the top of the cryostat at the cryostat breakout connectors (fridge BO), the optical fiber is fed through the bottom probe to the device.

For microwave drive signal delivery, SCuNi-CuNi semi-rigid coaxial cables lead from the breakout connectors at

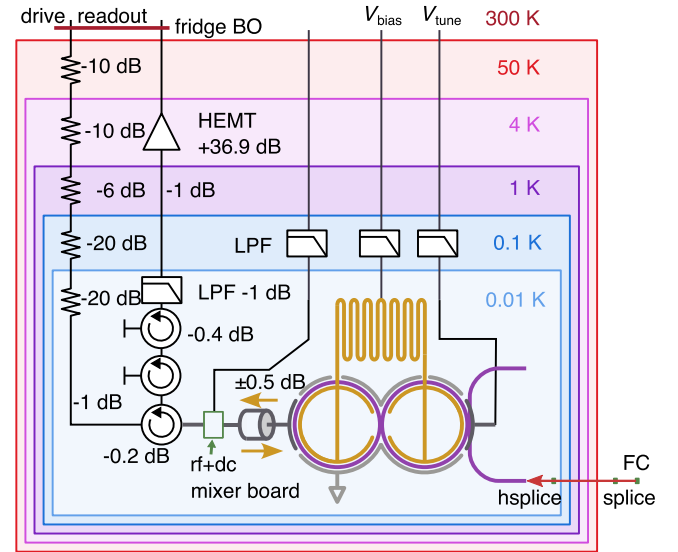


FIG. 16. Cryostat wiring with device under test. Microwave signals are delivered to the device through the drive input at the cryostat breakout (fridge BO) via coaxial cables and a series of attenuators, a circulator, and a rf + dc mixer board. The reflected microwave signal is passed through two isolators, a low-pass filter (LPF) and a HEMT before being delivered to the readout channel at the dilution refrigerator breakout. Bias and tuning voltages V_{bias} and V_{tune} , respectively, are delivered to the device via twisted wire pairs through LPFs. Optical signals are delivered through a fiber going from room temperature into the bottom probe of the cryostat, which is spliced to the packaged device fiber before sample mounting.

the top down to the mixing chamber plate of the cryostat. They are attenuated at the different stages according to the temperature gradient to the previous stage (i.e., from 4 to 1 K) only -6 dB, while from 1 to 100 mK by -20 dB. At the mixing chamber plate, a circulator is connected which, in turn, passes the signals through a custom mixer board for the ability to combine microwave and direct current electrical signals up to ± 30 V. The reflected microwave signal is sent through two isolators via the circulator, which, importantly, are all thermalized to the mixing chamber plate. The signal is then filtered with a low-pass filter and delivered to a HEMT at the 4 K stage via NbTi superconducting coaxial cables, which are thermalized to each stage with 0-dB attenuators. To avoid bulky fiber connectors, the high-numerical-aperture fibers (UHNA7) of the packaged sample are spliced to the standard single-mode fibers (SMF-28) creating a smooth transistion (hsplice in Fig. 16), with typical losses on the order of 1 dB. A second standard splice between two SMF-28 fibers connects the cryostat fiber to the fiber connector (FC).

2. Calibration of transduction efficiency

The optical input power arriving at the fiber inside the probe $P_{\text{opt,in}}$ is calibrated by comparing cw optical heating of the cryostat to that using the resistive probe heater, from which the factor $+11.3$ dB between the TLPM_dev_in power meter reading and power in the fiber is determined (Fig. 18). This also characterizes the transmission of the dilution refrigerator A fiber with its standard (splice) and hybrid (hsplice) splices (Fig. 16) to be -0.7 dB using additional loss characterization of the circulator and FPC_dev_in. For pulsed measurements (Fig. 19), the voltage of the monitoring photodiode with 150 MHz bandwidth (150-MHz PD) is calibrated by measuring the total power output from FPC_dev_in. The ratio between optical signal and pump amplitude is measured via the ratio of their respective heterodyne sidebands on the ESA. Depending on where the optical power is measured, the optical gain (G_{opt}) is defined.

Next, the detected optical power is calibrated by determining the factor between the power injected into FPC_dev_in and the TLPM_dev_out power meter reading, which is $+15.7$ dB. The measured heterodyne signal amplitude on the ESA is then calibrated relative to this power meter, giving a factor of -14.7 dB between them at constant LO power, where the ratio of the two gives the heterodyne gain (G_{het}).

The microwave source power $P_{\text{mw,in}}$ (rf source 1+ power amp, Fig. 18) is calibrated with respect to the ESA, as well as the room-temperature readout and amplification line (RT read amp line). The loss through the cryostat wiring (in Fig. 16 from drive to readout) is determined and compared to the expected losses and gains of all components, and the

difference is added equally to the drive (G_{drive}) and readout (G_{read}) branch.

The photon number conversion efficiencies are then calculated by multiplying the detected powers with all determined gain factors in the signal chain, i.e.,

$$\eta_{\text{opt} \leftarrow \text{mw}} = \frac{P_{\text{ESA}}}{G_{\text{het}} G_{\text{opt} \leftarrow \text{mw}} G_{\text{drive}} P_{\text{mw,in}}} \quad (\text{D1})$$

$$\eta_{\text{mw} \leftarrow \text{opt}} = \frac{P_{\text{ESA}}}{G_{\text{read}} G_{\text{mw} \leftarrow \text{opt}} G_{\text{opt}} P_{\text{opt,in}}}, \quad (\text{D2})$$

where P_{ESA} is the detected output power on the ESA, and $P_{\text{mw,in}}$ and $P_{\text{opt,in}}$ are the microwave and optical input powers, respectively. The number conversion gains are defined by $G_{\text{opt} \leftarrow \text{mw}} = 1/G_{\text{mw} \leftarrow \text{opt}} = \omega_a/\omega_b$.

3. Pulsed microwave characterization

To characterize the microwave reflection under pulsed optical pumping, we use a setup very similar to the pulsed transduction setup (Fig. 17). The main difference is that only the pump laser is sent to the device, and instead of measuring the transduced signal, we measure the high-power microwave reflection from the device. This means that only the pulse power sum measurement with the monitoring diode is necessary. Because of the high microwave probe power levels, we remove the RT read amp line. The measurement and pulse sequence is virtually identical to the pulsed transduction measurement described in Appendix D 5. Time-resolved microwave spectra are recorded by running the full sequence for each frequency step. For time efficiency, only a narrow frequency span of a few megahertz around the resonance was recorded because the main parameters of interest are the extinction and frequency shift. To calculate the extinction, a larger span measurement is taken once using the same routine and setup to obtain the required reference level. Note that the setup is readily converted to the pulsed transduction configuration, which can be used to monitor the fiber-to-chip coupling efficiency between experiments.

4. Continuous-wave transduction

The setup for continuous-wave and bidirectional transduction measurements is shown in Fig. 18. Two continuously tunable external-cavity diode lasers (lasers 1 and 2) are used for optical pumping and probing. The main and pump laser in the cw experiment is laser 2. A portion of the light from laser 2 is sent to the frequency referencing path, where a coarse Mach-Zehnder interferometer (MZI) with a free-spectral range of about 2.5 GHz is used to monitor laser detuning sweeps during transduction measurements. Another fraction of the light is sent as LO to the heterodyne detection setup. Next, a fraction of the light is combined with that of the auxiliary laser (laser 1) to generate the

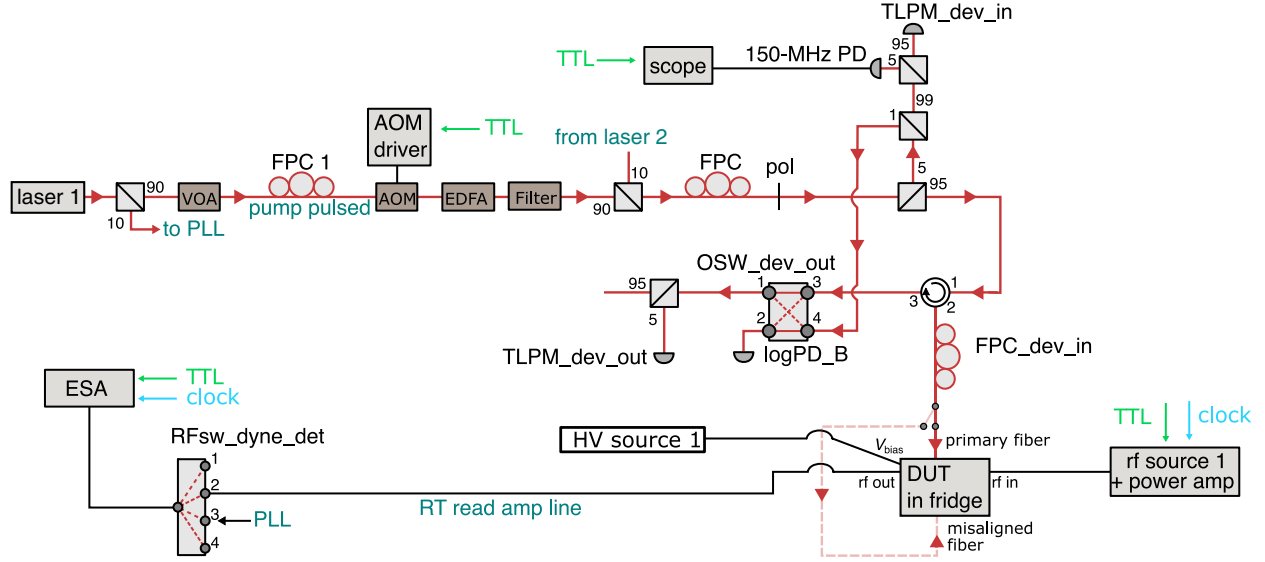


FIG. 17. Pulsed microwave characterization setup. HV, high voltage; DUT, device under test.

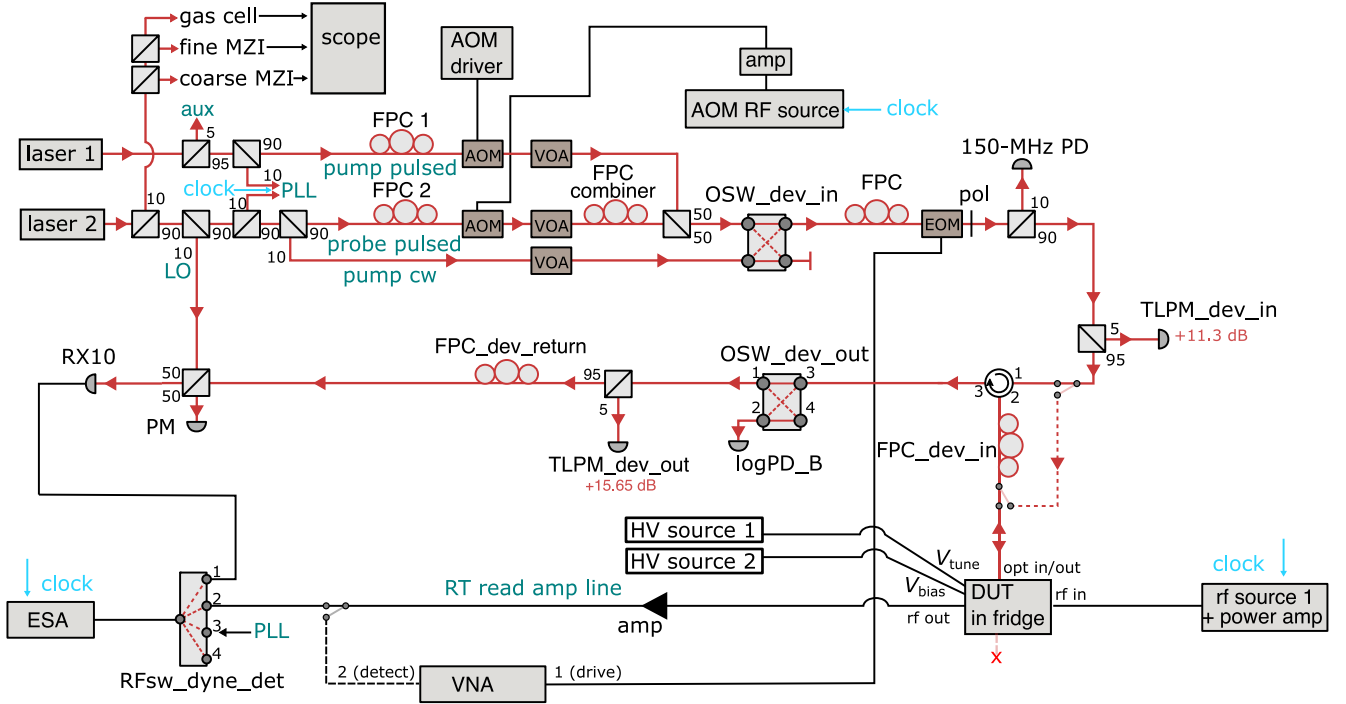


FIG. 18. Continuous-wave transduction setup.

optical beating to offset-lock laser 1 with a defined frequency offset to laser 2 using a phase-locked loop (PLL). Here, only the pulsed laser paths were used. Both lasers are polarization-controlled using fiber polarization controllers (FPCs) to maximize the transmitted power through the AOMs, while minimizing laser-frequency-dependent power variations, and then sent to variable optical attenuators (VOAs) to adjust their power levels. The light from the two lasers is combined on a

50/50 splitter, before which laser 2's polarization is again optimized using an FPC to align the polarization of the two lasers. After passing through an optical switch (OSW_dev_in), the polarization of the light is optimized to maximize the transmission through the electro-optic modulator (EOM). A polarization filter (pol) behind the EOM ensures equal polarization for both lasers. A fraction of the light is sent to a power meter (TLPM_dev_in) to monitor the power sent to the cryostat. The light passes

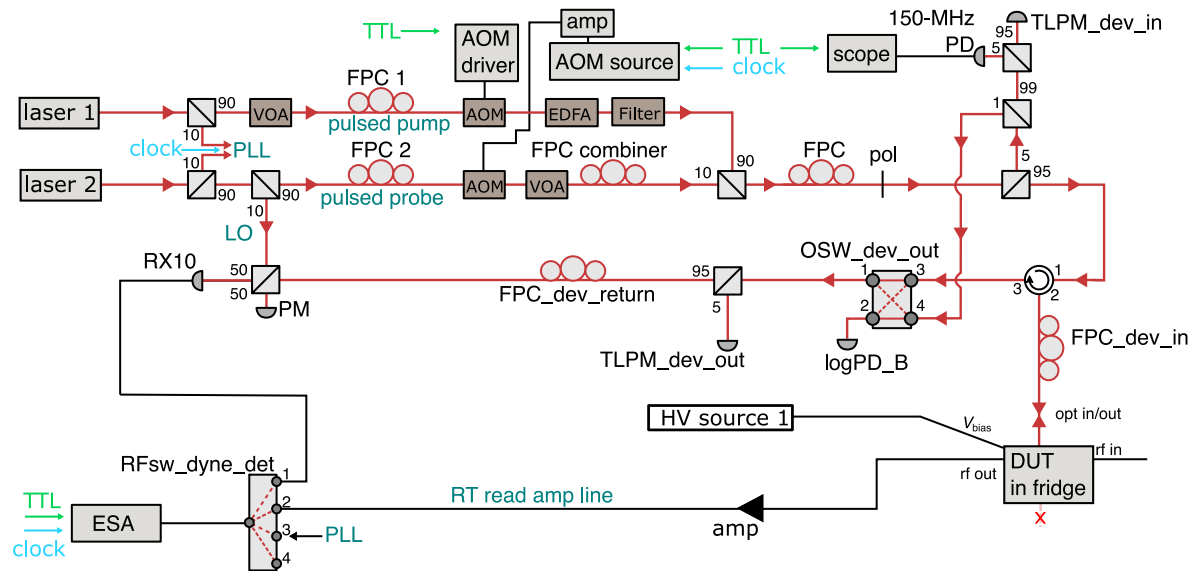


FIG. 19. Pulsed transduction setup.

through a circulator and is sent to the device via a final FPC. Light reflected from the cryostat fiber is sent to switch OSW_dev_out to either monitor the device reflection on the logarithmic photodiode (logPD_B) or to the heterodyne detection path. For calibration purposes, the laser signal can be sent into FPC_dev_in to directly be sent to the detection path. In the heterodyne detection path, the reflected power is monitored with power meter TLPM_dev_out and the polarization optimized using FPC_dev_return to optimize the overlap with the local oscillator. The light is then combined with the local oscillator and detected on the gigahertz photodiode RX10.

Electrical input signals are generated using a radio frequency source (rf source 1) and the heterodyne transduction signal is detected on an ESA. Optical input signals are generated either as a single sideband from offset-locking laser 2 to laser 1 (PLL) or by modulating the EOM using the VNA. The electrical signals from the cryostat are sent through the RT read amp line and detected either on the ESA or the VNA.

Note that the PLL, laser 2's AOM source, rf source 1, and the ESA are phase locked to a common 10-MHz reference clock. This allows the lowest possible noise during electrical detection with the ESA. Since we had only one referenced AOM source, the fixed-frequency AOM source for laser 1 was kept free running at a systematic offset of a few kilohertz from the nominal 110 MHz of both AOMs. This slight offset is beneficial for the electrical detection of the optical-to-microwave signal on the ESA, as the PLL creates crosstalk, which would otherwise be exactly at the detected microwave frequency. This noise issue does not exist for the microwave-to-optical direction, as the detected heterodyne signal is shifted by the AOM frequency (110 MHz).

5. Pulsed transduction

For studying the system in pulsed operation, we only perform optical-to-microwave transduction using two lasers with an offset lock, as the signal is less sensitive to variations in the optical reflection from the device compared to the microwave-to-optical signal. The setup is shown in Fig. 19 and is a variation of the continuous-wave setup discussed previously. It differs, however, in certain important points, which are mainly related to the generation and monitoring of high-power pump pulses.

In this configuration, laser 2 is the signal laser whose AOM is now pulsed using a transistor-transistor-logic (TTL) signal from a pulse generator. To achieve large peak powers, the pulsed optical signal of laser 1, which now acts as the pump, is amplified using an EDFA whose amplified spontaneous emission noise is filtered using a tunable filter. The light from the two lasers is then combined, but now using a 90/10 splitter to maximize the pump pulse power. The combined laser light is again sent through a polarization controller and a polarization filter.

Because of the large power range that needs to be covered, the power monitoring had to be modified. To that end, a portion of the light is sent to a pulse-monitoring branch, where the standard power meter measures the average power (TLPM_dev_in), while the 150-MHz PD is used to monitor the optical pulse shape of the combined lasers on an oscilloscope. Another fraction of the light is sent to the heterodyne detection arm, which we now use to monitor the power ratio of the optical pulses. Because of the frequency offset between the two pulses, this detection method also enables recording of the shape of each individual pulse with a signal-to-noise ratio superior to that of the photodiode monitoring the sum of the lasers.

Using an electrical switch (RFsw_dyne_det) either the optical pulse shapes and ratios are monitored using the heterodyne signal, or the transduction signal from the RT read amp line is measured using an ESA.

APPENDIX E: OPTICAL AND DC CHARACTERIZATION

1. Room-temperature optical transmission spectroscopy before cooldown

In this appendix, the optical response of the device is studied in transmission at room temperature before cooldown, as shown in Fig. 20. The transmission coefficient in this case is defined as $T = |a_{\text{out,cc}}/a_{\text{in,cc}}|^2$. Note that, for all measurements shown in this section, no voltage was applied to the device.

We apply the photonic molecule with backscattering fit model to each resonance in the spectrum and plot the resulting detuning and rates as a function of optical frequency (Fig. 21). Apart from a single outlier at high optical frequencies, the detuning between the rings Δ_π seems to be consistently around -2.2 GHz [Fig. 21(a)]. Note that no bias or tuning voltage has been applied to the device at that point. The forward-backward coupling amplitudes [Fig. 21(b)] scatter quite a lot with no discernible systematic trend, consistent with randomly distributed scattering defects along the ring. The resonant Rabi splitting $\hbar\Omega = 2\mu$ [Fig. 21(c)] shows the expected reduction with increasing optical frequency and reaches almost 5 GHz at the longest wavelengths (lowest frequencies). The external bus coupling rate [Fig. 21(d)] also shows the expected dispersion with more than 800 MHz at the lowest frequencies. The intrinsic loss rates [Fig. 21(e)] are scattered around 776 MHz, where, as previously mentioned, the average value of the two rings (green trace) seems to be more reliable than the individual ring loss rates. The distribution of intrinsic loss rates [Fig. 21(f)] is peaked around the mean value, indicating a reliable fit with the exception of the outlier at 3 GHz, which corresponds to the previously mentioned outlier in ring detunings at large frequencies.

2. Optical tuning at cryogenic temperatures

Next, we extract the resonance frequency ω_a of the photonic molecule as a function of bias voltage in the linear

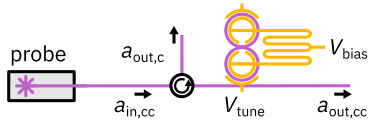


FIG. 20. Optical characterization measurement schematic. Light is injected via the counterclockwise input mode $a_{\text{in,cc}}$ and the transmission to $a_{\text{out,cc}}$ or reflection to $a_{\text{out,c}}$ is measured.

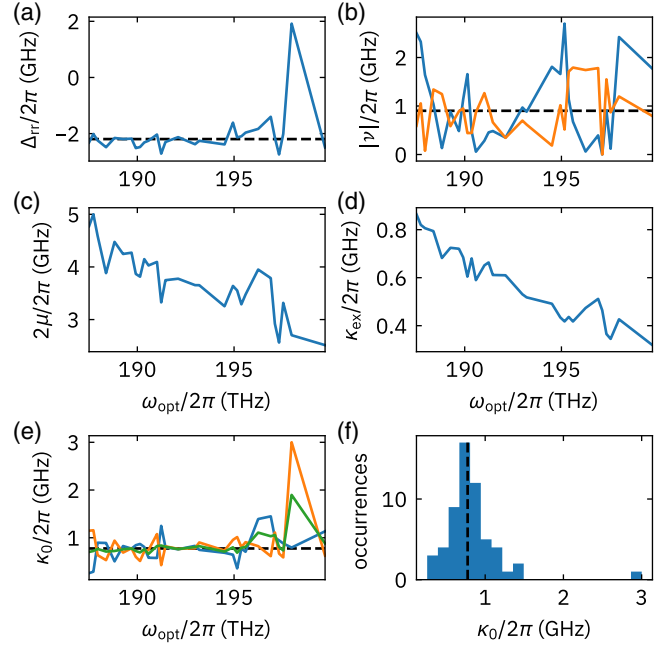


FIG. 21. Detuning and loss and coupling rates vs optical frequency. (a) Ring-ring detuning and (b) backscattering rates as a function of optical frequency. (c) Resonant ring-ring and (d) external bus coupling rate as a function of optical frequency. (e) Intrinsic loss rates as a function of optical frequency (e) and their distribution (f). Fit results of transducer (device 2) optical spectrum. When more than one trace is shown, blue represents the result for the first ring, orange for the second, and green the average of the two. The dashed lines show average ring-ring detuning $\bar{\Delta}_\pi/2\pi = -2.2$ GHz, intrinsic loss rate $\bar{\kappa}_{a,0}/2\pi = 776$ MHz, and backscattering amplitude $|\nu^*|/2\pi = 902$ MHz.

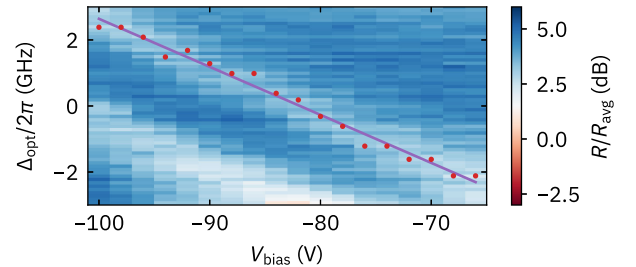


FIG. 22. Cryogenic electro-optic tuning efficiency. Linear fit to the V_{bias} tuning dependence of the optical resonance frequency around the standard working point $V_{\text{bias}} = -100$ V. The extracted slope is $\partial\omega_a/\partial V = 2\pi \times 145$ MHz V^{-1} .

range close to the standard working point of $V_{\text{bias}} = -100$ V (Fig. 22). We then fit a linear function to the trend and extract an electro-optic tuning efficiency of $\partial\omega_a/\partial V = 2\pi \times 145$ MHz V^{-1} . This value can later be used to estimate the vacuum coupling strength, as the dc tunability is typically an upper bound for that in the microwave domain.

3. Room-temperature optical transmission spectroscopy after warm-up

Since no voltages were applied to the transducer device prior to cooling down the sample, we characterize the sample postcooldown at room temperature. Interestingly, the response (Fig. 23) looks vastly different to the expected ferroelectric response of BaTiO₃ at room temperature [41]. While we do not have experimental characterization of this exact device prior to cooldown, other devices fabricated in this platform have shown the expected hysteretic behavior at room temperature. Therefore, we suspect that the material was permanently altered compared to before cooldown. Given that the optical resonance continues to shift in the same direction when the tuning voltage changes polarity, one may conclude that the ferroelectric domains have a permanent preferential orientation after cooldown. This may be due to the fact that the device was biased at $V_{\text{bias}} = -100$ V during most cryogenic measurements. Although the voltage is brought back to ground before warm-up, the coercive field is not reached, and the domains likely remain oriented and keep their orientation during warm-up.

When looking at the optical resonance tuning in the negative voltage branch only (between 0 and -100 V), we note that the response is nearly free of hysteresis [Fig. 23(c)]. The dc-tuning rate is $\partial\omega_a/\partial V = 2\pi \times 395$ MHz/V, corresponding to an effective Pockels coefficient of 115 pm/V. Interestingly, the average intrinsic loss rate (averaged over optical frequency) has increased by about a factor of 2 compared to the initial room-temperature value prior to application of dc voltages and before cooldown (Fig. 21). Furthermore, there seems to be an increase in optical loss rate as a function of tuning voltage, as shown in Fig. 23(d).

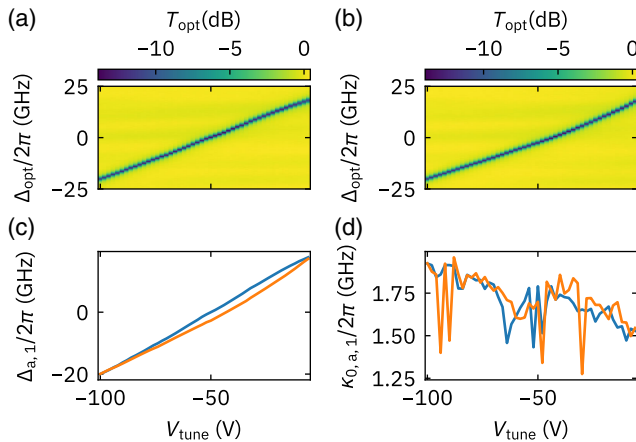


FIG. 23. Postcooldown room-temperature tuning hysteresis and loss. Optical transmission for (a) down and (b) up sweep of the tuning voltage. (c) Optical resonance frequency tuning and (d) intrinsic optical loss rate as a function of tuning voltage for up (orange) and down (blue) sweep.

APPENDIX F: MICROWAVE CHARACTERIZATION

In this appendix, the continuous-wave microwave response of the device is studied in reflection, as shown in Fig. 24. The reflection coefficient in this case is defined as $R = |b_{\text{out}}/b_{\text{in}}|^2$.

1. Bias dependence

Analogous to the characterization of the optical response as a function of bias voltage, we now study the microwave reflection with applied bias. While only one device (device 2) is fully packaged, we wired up a reference device (device 6) for microwave characterization. The microwave reflection as a function of bias voltage of both devices is shown in Fig. 25. The first thing to note is that the response shows a significant bias dependence. Unfortunately, the main transducer device [device 2, Fig. 25(b)] exhibits an anticrossing to an unknown spurious mode at around 6.79 GHz. The observed resonant interaction splitting is $2\gamma = 2\pi \times 31.9$ MHz.

For characterization of general microwave parameters, we initially focus on device 6 [Fig. 25(a)] for a simplified model description with less error sources and fit parameters than the actual transducer device with spurious mode coupling requires. To that end, we take a full hysteresis curve and fit the data with the single resonance model NLQFIT7

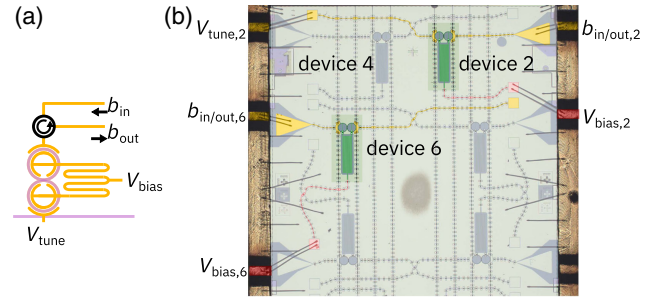


FIG. 24. Microwave characterization measurement schematic. (a) A microwave tone is injected in b_{in} , reflected at the device, and is sent to b_{out} via a circulator. (b) Optical micrograph of the two electrically packaged devices (device 2, device 6).

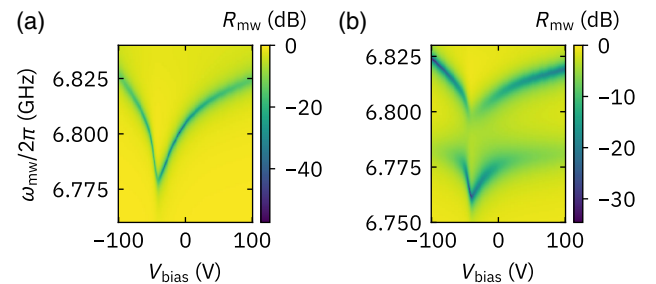


FIG. 25. Microwave reflection vs bias voltage. (a) Device 6 and (b) device 2 microwave reflection vs bias voltage V_{bias} .

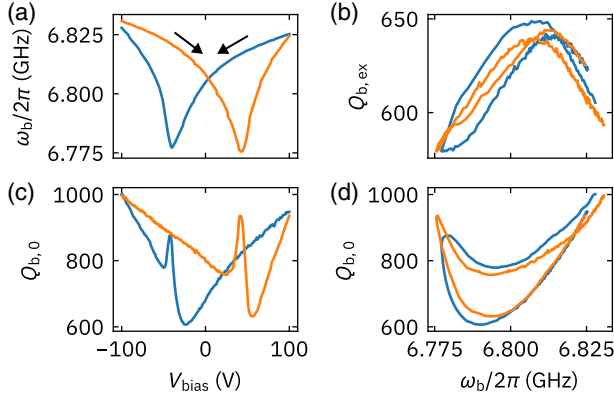


FIG. 26. Bias dependence of microwave properties: device 6. (a) Microwave resonance frequency ω_b and (b) intrinsic quality factor $Q_{b,0}$ as a function of bias voltage V_{bias} . (c) External quality factor $Q_{b,\text{ex}}$ and (d) intrinsic quality factor $Q_{b,0}$ as a function of microwave resonance frequency ω_b .

in the PYTHON package SCIKIT-RF [68]. Figure 26(a) shows the bias dependence of the microwave resonance frequency, revealing an almost perfectly symmetric hysteresis curve as expected for a ferroelectric material. Note that, apart from the asymmetry and noise, the microwave hysteresis of device 6 shows a qualitatively similar shape to that of device 2 [Fig. 2(h)]. Importantly, the coercive field values align well. The bias dependence can most likely be attributed to a bias-dependent microwave permittivity of the material, making it a nonlinear capacitor. Interestingly, the intrinsic quality factor also depends on the applied bias [Fig. 26(b)]. Generally, bias voltages beyond the coercive field seem to lead to higher quality factors, with the minimal quality factor reached for values close to the coercive field. Right at the coercive field, however, a local peak in the quality factor is observed. While this behavior may be explained by a bias-dependent material absorption of microwave radiation, it may also stem from piezoelectric actuation, since it is known that most electro-optic materials also feature a nonzero piezoelectric effect. Right at the coercive field, one would expect the ferroelectric domains to have a net vanishing polarization, which would explain the local peaks of the microwave quality factor. The fact that the largest microwave loss is observed close to the net zero polarization point suggests that the piezoelectric coupling may be largest there. For completeness, the external microwave coupling quality factor is plotted vs frequency [Fig. 26(c)] and shows a slight frequency dependence, which may stem from reflections in the bus line.

In order to extract the bias dependence of the bare transducer resonance frequency of device 2, we use the coupled microwave mode model derived in Appendix B 4. Since a fit of the complex S -parameter response is involved, we implement a fit of the magnitude $|S|$. To improve the fit stability, we fix certain parameters as listed in Table III.

TABLE III. Effective dc-tuning model parameters. The parameters listed here were fixed when fitting the reflection of device 2. We fixed $\kappa_{b,\text{ex}}$ only for experiments where the resonance frequency did not shift significantly, namely, for microwave- and optical-power-dependence measurements. For the temperature and bias dependence, $\kappa_{b,\text{ex}}$ is a free parameter in the fit.

$\omega_{b,2}/2\pi$	6.792 GHz
$\kappa_{b,2,0}/2\pi$	11.4 MHz
$\gamma/2\pi$	15.95 MHz
Reference level R_0	$ S ^2 @ 6.835 \text{ GHz}, -100 \text{ V}_{\text{bias}}$
$\kappa_{b,\text{ex}}/2\pi$	7 MHz

2. Microwave power dependence

So far, we have been probing the microwave system in the medium-to-high-power regime. Typically, however, the signal levels in transducers for quantum transduction are just a few or even single photons. In this low-power regime, absorption from two-level systems may become a significant loss channel, as the circulating power is no longer sufficient to saturate the two-level systems in the dielectric, for example. For this reason, we probe the microwave reflection of the device as a function of incident microwave power. By fitting the microwave response of devices 2 and 6, again using the single resonance model NLQFIT7 in the SCIKIT-RF PYTHON package [68], we can extract the external and intrinsic quality factors $Q_{b,\text{ex}}$ and $Q_{b,0}$, respectively. The external coupling quality factor is independent of microwave power, as expected, but, interestingly, different for the two devices; namely, $Q_{b,\text{ex}} = 960$ for device 2 and $Q_{b,\text{ex}} = 605$ for device 6. This can potentially be attributed to the fact that the transmission lines from the wire bonding pads to the devices have different lengths [see Fig. 24(b)]. While device 6 has a very short line without optical crossings, the transmission line of device 2 has two optical crossings and is longer. The low-power quality factors are 740 and 650 for device 2 and 6, respectively.

3. Temperature dependence

In this section, we sweep the temperature of the cryostat (dilution refrigerator) by either evaporating the helium cooling mixture (warm up) or recondensing it (cool down), while simultaneously measuring the microwave reflection at high probe power. Note that this means that both the bottom probe and mixing chamber temperature are swept simultaneously, which we call T_{fridge} .

In a first analysis step, we extract the observed center frequency of the reflection dip and its extinction as a function of temperature for both devices. While for device 6 the observed reflection dip frequency corresponds to the actual bare resonator frequency ($\omega'_b = \omega_b$), the observed frequency of device 2 is shifted due to coupling to the spurious mode. We therefore need to fit the response with the two-mode model to extract the bare frequency.

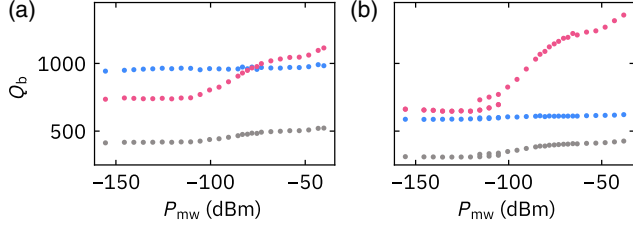


FIG. 27. Microwave quality factor vs microwave power. External (blue), loaded (gray), and intrinsic (red) microwave quality factor for devices (a) 2 and (b) 6 as a function of microwave power P_{mw} . The bias voltage for both measurements is $V_{bias} = -100$ V.

To further quantify the observed frequency shift and loss rate, we fit the spectra to extract the resonance frequency shift and quality factor change as a function of temperature. For device 6, we again use the simple resonance fit NLQFIT7 in the SCIKIT-RF PYTHON package [68]. For device 2, we use the more complex model that takes into account the coupling to the spurious mode (see Appendix B 4). The fixed fit parameters are listed in Table III. The resulting relative frequency shift and intrinsic quality factor as a function of temperature are plotted in Fig. 28 in blue and magenta for device 2 and 6, respectively. As shown in Fig. 28(b), the relative frequency shift is almost identical for both devices. For both devices, the intrinsic quality factors decrease monotonously as a function of temperature for $T_{fridge} > 0.6$ K. Note that the quality factor for device 6 depends only on the loaded linewidth and observed extinction and should, therefore, be accurate. The additional spurious mode for device 2 introduces additional parameters such as the loss and coupling rate of the spurious mode, which cannot be unambiguously separated, meaning that the extracted intrinsic quality factor shown here for device 2 may have a large error.

While the redshift and decrease of intrinsic quality factor are a typical signature of QP generation in the superconductor, the initial blueshift has been attributed to TLSs often present in dielectrics with defects or amorphous composition [53,54]. The observed relative frequency shift

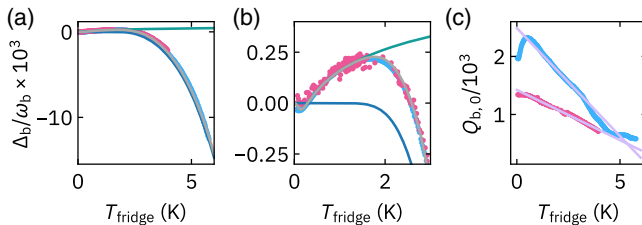


FIG. 28. Microwave resonance shift and quality factor vs temperature. (a) Full and (b) reduced temperature range plot of relative resonance frequency shift Δ_b/ω_b and (c) intrinsic microwave quality factor $Q_{b,0}$ as a function of cryostat temperature T_{fridge} for device 2 (blue) and device 6 (magenta).

TABLE IV. Thermal microwave frequency shift fit parameters. Fit results for free parameters from Eqs. (F1)–(F3) fitted to the data shown in Figs. 28(a) and 28(b).

$Q_{b,0,TLS}$	T_C (K)	α (%)
2038	8.7	7.8

is then described as a sum of both effects [54],

$$\frac{\Delta\omega_b}{\omega_b} = \left(\frac{\Delta\omega_b}{\omega_b}\right)_{TLS} + \left(\frac{\Delta\omega_b}{\omega_b}\right)_{QP}. \quad (F1)$$

The individual contributions can be modeled as [69]

$$\left(\frac{\Delta\omega_b}{\omega_b}\right)_{TLS} = \frac{1}{\pi Q_{b,0,TLS}} \times \text{Re} \left\{ \Psi \left(\frac{1}{2} + i \frac{\hbar\omega_b}{2\pi k_B T} \right) - \ln \left(\frac{\hbar\omega_b}{2\pi k_B T} \right) \right\}, \quad (F2)$$

with the complex digamma function Ψ , and

$$\left(\frac{\Delta\omega_b}{\omega_b}\right)_{QP} = -\frac{\alpha}{2} \left(\frac{\sin[\gamma\phi(T)]}{\sin[\gamma\pi]} \left| \frac{\sigma(T)/2}{\sigma(0)} \right|^\gamma - 1 \right), \quad (F3)$$

with the kinetic inductance fraction $\alpha = L_k/(L_g + L_k)$ and the kinetic and geometric inductance of the circuit L_k and L_g , respectively. The kinetic inductance of a thin superconducting film of length l is related to the complex conductivity $\sigma = \sigma_1 - i\sigma_2$ via $L_k \approx l/(w\omega_b\sigma_2)$ for $R \rightarrow 0$ with the conductor width w and thickness t . For the conductivity, we used the expressions defined in Ref. [54].

We fit the model to the data of device 2 and plot the TLS, QP, and sum of both contributions in green, dark blue, and gray, respectively [Figs. 28(a) and 28(b)]. The fit parameters and results are listed in Table IV. We note that the zero-temperature TLS intrinsic quality factor aligns well with the observed quality factor drop in Fig. 28(c) toward low temperatures, and the critical temperature is also in line with previous experimental studies on the material [70].

Additionally, linear fits of the extracted intrinsic quality factors between 0.6 and 6 K are shown in Fig. 28(c) in light purple.

APPENDIX G: TRANSDUCTION

While it was not possible to directly measure the optical response of the system during cooldown in this study, we can make use of the transduced signal to characterize the optical system. To that end, we perform an optical-to-microwave measurement as a function of tuning voltage V_{tune} , which detunes the two rings with respect to each other. As shown in Fig. 29(a), the response changes

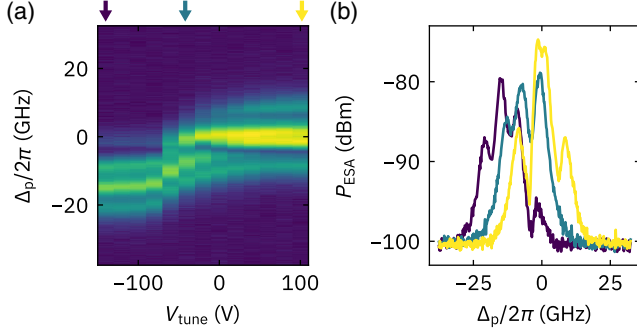


FIG. 29. Optical-to-microwave transduction vs tuning voltage. (a) Detected microwave-optical signal power as a function of tuning voltage V_{tune} and pump detuning Δ_p . (b) Transduction spectra for selected tuning voltages as marked with the colored arrows in (a).

significantly as a function of tuning voltage. For $V_{\text{tune}} = 100$ V, the maximum detected signal is highest [Fig. 29(c)] and the doubly resonant satellite peaks are of similar height [Fig. 29(b), yellow trace]. This suggests that this is the case where the rings are resonant with each other. The slight splitting of the triply resonant central peak can either arise from forward-backward scattering or from a mismatch between microwave frequency and the optical splitting. The frequency mismatch is the more likely explanation, as otherwise the doubly resonant peaks should also show a splitting. In the detuned case [Fig. 29(b), purple trace], the maximum signal is lower as expected, but now a total of four peaks are clearly visible, indicating strong forward-backward coupling. Note that the forward-backward coupling strength is strongly frequency dependent and may be lower in the resonant case.

The model parameters assumed for the theory curves plotted in Fig. 3 are given in Table V.

TABLE V. System parameters for transduction model.

Parameter	Value
$\Delta_{\pi}/2\pi$	0.4 GHz
$\omega_b/2\pi$	6.82 GHz
$2\mu/2\pi$	4.7 GHz
$ \nu_1 /2\pi$	0.35 GHz
$\arg \nu_1$	0
$ \nu_2 /2\pi$	0.15 GHz
$\arg \nu_2$	0
$\kappa_{a,0}/2\pi$	2 GHz
$\kappa_{a,\text{ex}}/2\pi$	0.85 GHz
$\kappa_{b,0}/2\pi$	10 MHz
$\kappa_{b,\text{ex}}/2\pi$	7 MHz
$g_0/2\pi$	406 Hz
P_p	-10 dBm
η_{f2c}	-5 dB

APPENDIX H: OPTICALLY INDUCED MICROWAVE HEATING

1. Slow optical heating

To characterize the slow thermalization rate in the system, we send a 10- μs pulse of off-resonant light with a low repetition rate to the device (Fig. 30). In addition to the fast-respondering redshift, which we attribute to quasi-particle generation in the case of off-resonant light, we observe a slow blueshift and decrease in extinction, indicating heating of the dielectric. After the pulse, the system recovers to its prepulse state after several tens to hundreds of microseconds [Fig. 30(a)], which is significantly slower than the on-resonant dielectric and quasiparticle heating. We attribute this thermalization rate to the thermalization between the sample package and probe with the cryostat mixing chamber stage.

2. On-resonant heating

Comparing Fig. 31 and Figs. 4(i) and 4(j) in the main text, we note that the dynamics before and after the pulse, i.e., at times t_1 and t_4 , respectively, are very similar for both on- and off-resonant light via the primary and misaligned fiber, respectively. Furthermore, we note that the effective temperature and frequency shift are virtually identical at the beginning and end of the pulse for on-resonant light, i.e., at times t_2 and t_3 , respectively (see Fig. 31). For the sake of clarity, we thus chose to plot only the on-resonant data at time t_2 in Figs. 4(i) and 4(j) of the main text.

3. Power correction

The on-resonant heating dynamics via the primary fiber were measured after the transduction measurements (Secs. IV and VI), after which the fiber-chip coupling

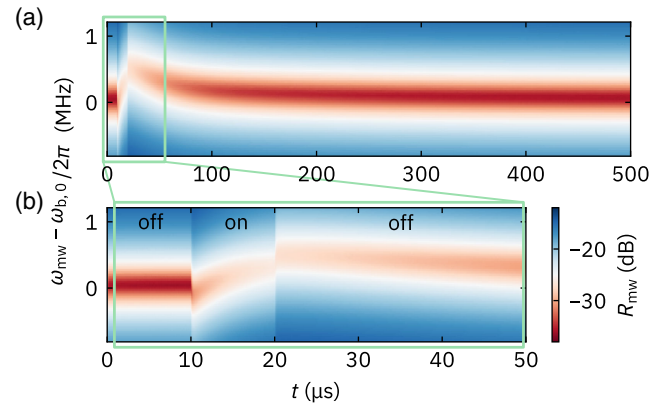


FIG. 30. Slow microwave heating under pulsed operation. (a) Microwave reflection as a function of time and microwave probe frequency ω_{mw} around the prepulse dip frequency $\omega_{b,0}$. (b) Close-up of the first 50 μs of (a).

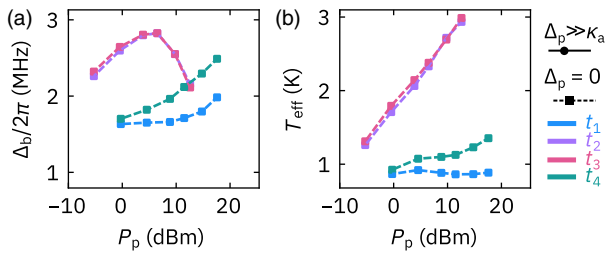


FIG. 31. Extended on-resonant heating data. (a) Microwave resonance frequency shift Δ_b and (b) effective temperature T_{eff} as a function of optical pump power P_p for resonant optical light pulses at different times t_i with $i \in [1, 2, 3, 4]$ correspond to the ones shown in Fig. 4(b). Here, the light was delivered via the primary fiber.

efficiency had degraded by 5 dB. The fiber-chip coupling efficiency was monitored by measuring the transduction efficiency per pump power. To allow a better comparison with the transduction experiments, we subtracted the additional insertion loss of 5 dB from the pump power values for resonant data points during the pulse (t_2 and t_3) in Fig. 31 and Figs. 4(i) and 4(j). This correction accounts for the lower intracavity circulating power at a given in-fiber pump power and ensures that data points with equal pump power in Figs. 4(i) and 4(j) (on resonance) and in Fig. 5(c) correspond to the same on-chip power. Since the heating is dominated by resonant in-cavity light, despite the lower coupling efficiency compared to during the transduction measurements, this correction should result in a good estimate of the effective temperature T_{eff} at a given pump power P_p during the transduction measurements of Secs. IV and VI.

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