

# Surpassing millisecond coherence times in on-chip superconducting quantum memories by optimizing materials, processes, and circuit design

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## Abstract

The performance of superconducting quantum circuits for quantum computing has advanced tremendously in recent decades; however, a comprehensive understanding of relaxation mechanisms does not yet exist. In this work, we utilize a multimode approach to characterizing energy losses in superconducting quantum circuits, with the goals of predicting device performance and improving coherence through materials, process, and circuit design optimization. Using this approach, we measure significant reductions in surface and bulk dielectric losses by employing a tantalum-based materials platform and annealed sapphire substrates. With this knowledge we predict and experimentally verify the relaxation times of aluminum- and tantalum-based transmon qubits. We additionally optimize device geometry to maximize coherence within a coaxial tunnel architecture, and realize on-chip quantum memories with single-photon Ramsey times of 2.0–2.7 ms, limited by their energy relaxation times of 1.0–1.4 ms. To our knowledge this is the highest coherence achieved in an on-chip quantum memory, and demonstrates an advancement towards a more modular and compact coaxial circuit architecture for bosonic qubits with reproducibly high coherence.

**Keywords:** Transmon qubit, Coherence, Microwave loss, Quantum memory, Bosonic qubit

## Introduction

The emergence of superconducting qubits as a promising platform for quantum computing has been facilitated by over two decades of steady improvements to coherence and gate fidelity[1]. This has enabled the demonstration of

many milestones, including quantum error correction or mitigation[2–9], quantum algorithms[10, 11], quantum simulations[12–15], and quantum supremacy[16] using large numbers of qubits. However, the realization of a practical quantum computer requires far higher gate fidelities[17, 18],

which necessitate further mitigation of decoherence mechanisms in quantum circuits. Substantial exploration in the past has shown that the sources of decoherence can be traced to intrinsic sources of energy loss from the circuits' constituent materials and has revealed the existence of significant bulk and surface dielectric loss[19–23], two-level-system (TLS) loss[24–28], and residual quasiparticle or vortex loss in superconductors[29–33]. Accordingly, improvements to coherence have been made by employing intrinsically lower-loss materials such as sapphire substrates[34, 35], tantalum thin-films[36, 37], and contamination-minimizing fabrication processes such as acid-based etching[30, 33] and substrate annealing[22, 37, 38]. Additionally, dramatic improvements have also been achieved by modifying circuit geometry to reduce sensitivity to loss, an approach that has given rise to 3D transmon qubits[39] and cavity-based quantum memories with millisecond coherence times[40–42].

Improving coherence requires understanding the underlying loss mechanisms. Determining where the dominant losses originate as well as the extent to which those losses dominate is crucial to maximizing the performance of superconducting qubits. There have been significant efforts to understand and mitigate surface dielectric loss in thin-film resonators[38, 43–46]; however, recent studies have shown that bulk dielectric loss can play a significant role[47, 48]. A systematic approach is therefore desired to characterize intrinsic losses and improve coherence in a predictable way.

Conveniently, superconducting microwave resonators are powerful characterization tools because they can be measured easily with high precision and their quality factors are limited by the same intrinsic sources of loss as transmon qubits[33]. Additionally, their sensitivities to particular sources of loss can be tuned by modifying their geometries, a feature that has been heavily utilized in other studies to investigate various sources of loss in thin-film resonators and bulk superconductors[19, 21, 23, 33, 38, 44–46, 49–51]. In a multimode approach to loss characterization, a single device can have multiple resonance modes that are each sensitive to different sources of loss. This allows for the use of a single device to study multiple sources of loss, eliminating systematic errors due to device-to-device or run-to-run

variation[33]. Furthermore, by measuring multiple multimode devices, the device-to-device variation of intrinsic loss can be determined, allowing for the evaluation of the consistency of a particular materials system or fabrication process and the prediction of the expected energy relaxation rate of a quantum circuit.

In this work, we introduce the tripole stripline, a multimode superconducting microwave resonator whose modes can be used to distinguish between surface losses, bulk dielectric loss, and package losses in thin-film superconducting quantum circuits. We use this loss characterization device to measure and compare the losses associated with thin-film aluminum and tantalum deposited on a variety of sapphire substrates that differ by their growth method and preparation. While previous work has shown improved device coherence using tantalum-based fabrication processes[36, 38] and annealed sapphire substrates[37, 52], we use our technique to show that the aforementioned improvements originate definitively from the reduction of surface loss in tantalum-based devices and of bulk dielectric loss in annealed sapphire substrates.

With the tripole stripline, we gain a comprehensive understanding of how materials and fabrication processes limit the coherence of superconducting quantum circuits. We use this knowledge to predictively model the loss of aluminum- and tantalum-based transmon qubits. We then confirm through transmon coherence measurements that reduction of surface loss yielded by a tantalum-based process results in a  $T_1$  improvement of a factor of two in tantalum-based transmons over aluminum-based transmons. Understanding the loss mechanisms that limit coherence informs optimization and circuit design choices to further improve device coherence. We optimize device geometry to maximize coherence in a particular coaxial architecture, and design a stripline-based quantum memory with coherence times exceeding one millisecond. This far surpasses those of previous implementations of thin-film quantum memories[20, 53], and enables the miniaturization of highly coherent bosonic qubits within larger multiqubit systems for quantum information processing.

## Results

### Characterizing microwave losses in thin-films with tripole striplines

Differentiating between the various sources of loss in superconducting quantum circuits requires an appropriately designed loss characterization system. We implement such a system in the coaxial tunnel architecture[20] using multimode thin-film stripline resonators fabricated on sapphire substrates. The devices are inserted into a cylindrical tunnel waveguide package made of conventionally-machined high-purity (5N5) aluminum (Fig. 1a). End-caps close the tunnel, creating a superconducting enclosure with well-defined package modes that are high (>18 GHz) in frequency (see Methods “[Device packaging](#)”).

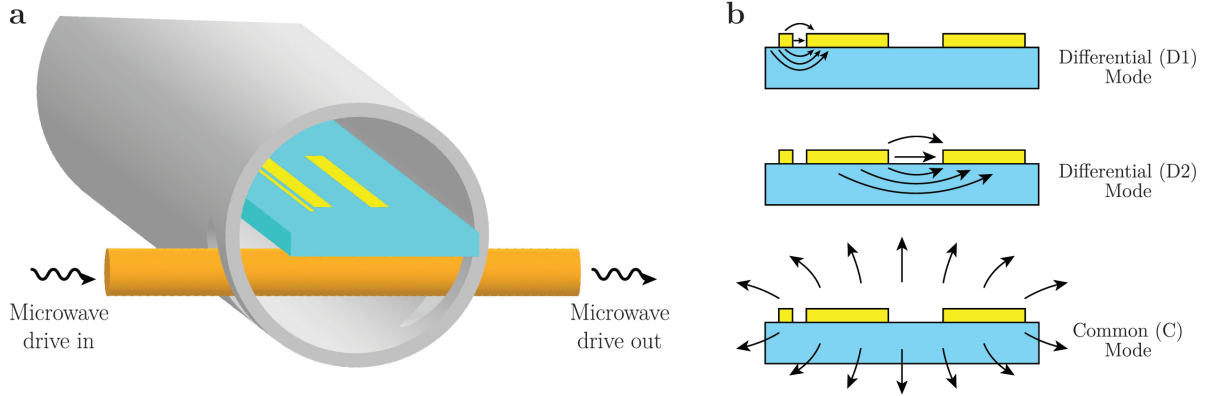
We design the multimode tripole stripline to distinguish between package losses due to induced current flowing in dissipative regions of the cylindrical tunnel package, bulk dielectric loss in the substrate, and surface dielectric losses associated with the various interfaces between substrate, superconductor, and air/vacuum. The tripole stripline is comprised of three superconducting strips placed adjacently to each other on a substrate with different widths and spacings (Fig. 1b). The arrangement of the three strips affect the spatial distributions of the electromagnetic fields of the three fundamental modes, thereby determining their sensitivities to particular sources of loss. The D1 differential mode is highly sensitive to surface losses due to its spatially-localized electromagnetic field in the small 10  $\mu\text{m}$  spacing between the 10  $\mu\text{m}$  narrow strip and the adjacent 400  $\mu\text{m}$  wide strip. On the other hand, the large 1.2 mm spacing between the two wide strips supports the D2 differential mode whose fields are much more dilute, resulting in lower surface loss while still retaining large sensitivity to bulk dielectric loss. Finally, the common (C) mode supports a spatially-diffuse electromagnetic field that induces larger electromagnetic fields on the walls of the package, rendering this mode sensitive to package loss. The differential sensitivity of these modes to different sources of loss allows us to distinguish between them by measuring the mode quality factors.

Losses in the tripole stripline can be described using a generalized energy participation ratio model[25, 43, 54]:

$$\frac{1}{Q_{\text{int}}} = \frac{1}{\omega T_1} = \sum_i \frac{1}{Q_i} = \sum_i p_i \Gamma_i, \quad (1)$$

where  $Q_{\text{int}}$  is the total internal quality factor of the resonator,  $\omega$  is the resonance frequency, and  $T_1$  is the energy decay time. The total loss can be broken down into a sum of losses  $1/Q_i$  from distinct loss channels, where  $\Gamma_i$  is the generalized intrinsic loss factor[44] associated the  $i$ th loss channel, and  $p_i = U_i/U_{\text{tot}}$  is the geometric energy participation ratio calculated by computing the fraction of energy stored in the  $i$ th lossy region when a resonance mode is excited. The participation ratio is therefore determined by the spatial distribution of the electromagnetic field of the resonance mode and as a result can be calculated in finite-element simulation and engineered to alter the mode’s sensitivity to specific loss channels (see Methods “[Calculation of participation ratios](#)”). The loss factors, on the other hand, are intrinsic material- and process-dependent quantities such as loss tangents and surface resistances that must be measured.

We use the participation ratio model in order to quantify the losses in the tripole stripline (see Supplementary Table S5). We define surface loss as  $1/Q_{\text{surf}} = p_{\text{surf}} \Gamma_{\text{surf}}$ , where  $p_{\text{surf}} = p_{\text{SA}} + p_{\text{MS}} + p_{\text{MA}}$  is the sum of surface dielectric participations in thin (3 nm) dielectric (relative permittivity  $\epsilon_r = 10$ ) regions located at the substrate-air (SA), metal-substrate (MS), and metal-air (MA) interfaces.  $\Gamma_{\text{surf}}$  is the corresponding surface loss factor that describes the intrinsic loss in these three interrelated regions. This formulation of surface loss differs from that of other studies[44–46], which aim to independently characterize the surface loss factors  $\Gamma_{\text{SA}}$ ,  $\Gamma_{\text{MS}}$ , and  $\Gamma_{\text{MA}}$ ; here,  $\Gamma_{\text{surf}}$  is a weighted sum of the three surface loss factors and characterizes the overall surface loss due to the presence of oxides, amorphous species, interdiffusion, organic residues, point-like defects, or lattice distortions. Because these physical signatures of loss are heavily influenced by processes such as substrate preparation, metal deposition, and circuit patterning, the three surfaces loss factors are



**Fig. 1: Tripole striplines in the coaxial tunnel architecture.** **a** Superconducting thin-film strips are patterned on a substrate and loaded into a cylindrical tunnel made of high-purity aluminum. Resonator frequencies range from 4 – 7 GHz (see Supplementary Table S5). A transversely oriented coupling pin is used to capacitively drive the resonators in hanger configuration. **b** Cross-sectional view of the tripole stripline, showing the arrangement of the strips and electric field behaviors for each mode. While the electric field of the D1 mode is confined mostly on the surface, the electric field of the D2 mode penetrates far deeper into the bulk, rendering it sensitive to losses over a significant portion of the bulk of the substrate.

interdependent; therefore,  $\Gamma_{\text{surf}}$  is the most relevant descriptor of intrinsic surface loss because it characterizes a particular materials platform and fabrication process in order to predict the total surface loss in a device.

We consider surface loss to be distinct from the bulk loss  $1/Q_{\text{bulk}} = p_{\text{bulk}}\Gamma_{\text{bulk}}$  which is dielectric in nature and may be dependent on the crystalline order of the substrate. Additionally, we define package losses  $1/Q_{\text{pkg}} = p_{\text{pkg}_{\text{cond}}}\Gamma_{\text{pkg}_{\text{cond}}} + p_{\text{pkg}_{\text{MA}}}\Gamma_{\text{pkg}_{\text{MA}}} + p_{\text{seam}}\Gamma_{\text{seam}}$  as a combination of conductor loss due to residual quasiparticles, MA surface dielectric loss due to the metal oxide on the surface of the tunnel package, and seam loss  $p_{\text{seam}}\Gamma_{\text{seam}} = y_{\text{seam}}/g_{\text{seam}}$  due to a contact resistance that manifests when two metals come into contact, which occurs when the tunnel package is closed with the end-caps[33, 55]. The large  $p_{\text{surf}}$  in the tripole stripline’s D1 mode and large  $p_{\text{pkg}_{\text{cond}}}$ ,  $p_{\text{pkg}_{\text{MA}}}$ , and  $y_{\text{seam}}$  in the C mode yields a participation matrix that is well-conditioned to extract the loss factors with minimal error propagation, a crucial requirement for characterizing microwave losses in this way.

## Extracting intrinsic loss factors from resonator measurements

We demonstrate loss characterization by fabricating and measuring tripole stripline resonators. Tripole striplines made of tantalum were fabricated on a HEMEX-grade sapphire substrate. The substrate was annealed at 1200 °C in oxygen before tantalum was deposited via DC magnetron sputtering at 800 °C. The striplines were patterned using a subtractive process (see Methods “[Device fabrication](#)”) and then loaded into multiplexed coaxial tunnel packages (see Methods “[Device packaging](#)”) and measured in hanger configuration in a dilution refrigerator at a base temperature of 20 mK. The frequency response of each mode was measured using a vector network analyzer, and the internal quality factor as a function of average circulating photon number  $\bar{n}$  was extracted by fitting the resonance circle in the complex plane (see Methods “[Measurement of resonator quality factor](#)”)[56].

The differences in power dependence of  $Q_{\text{int}}$  of the tripole stripline’s modes reflect the modes’ sensitivities to surface loss (Fig. 2a). The D1 mode has the largest power dependence, with  $Q_{\text{int}}$  changing by over an order of magnitude from one to one million photons circulating in the resonator.

We attribute this power dependence to the presence of anomalous two-level systems (TLSs) that couple to the electric field of the mode and provide additional pathways for energy relaxation to occur. Beyond a critical photon number, the TLSs become saturated and effectively decouple from the mode, causing  $Q_{\text{int}}$  to increase. The power dependence of each mode is fit to the following TLS model:

$$\frac{1}{Q_{\text{int}}} = \frac{1}{Q_0} + \frac{p_{\text{surf}} \tan \delta_{\text{TLS}}}{\sqrt{1 + (\bar{n}/n_c)^\beta}}, \quad (2)$$

where  $1/Q_0$  is the power-independent contribution to the total internal loss,  $\tan \delta_{\text{TLS}}$  is the ensemble TLS loss tangent,  $n_c$  is the critical photon number beyond which the TLSs begin to saturate, and  $\beta$  is an empirical parameter that describes TLS interaction[24–26, 38, 57, 58]. While the D2 and C modes are also power-dependent, they are far less so, with  $Q_{\text{int}}$  changing by less than a factor of two over the same range of  $\bar{n}$ . This is consistent with these modes having nearly two orders of magnitude smaller surface participation, which allows the D2 and C modes to attain single-photon  $Q_{\text{int}}$  that are over an order of magnitude higher than that of the D1 mode. The D2 mode, being relatively insensitive to both surface and package losses has a single-photon  $Q_{\text{int}}$  of around  $3 \times 10^7$ , which to our knowledge far exceeds the highest single-photon  $Q_{\text{int}}$  measured in a lithographically-patterned thin-film resonator to date.

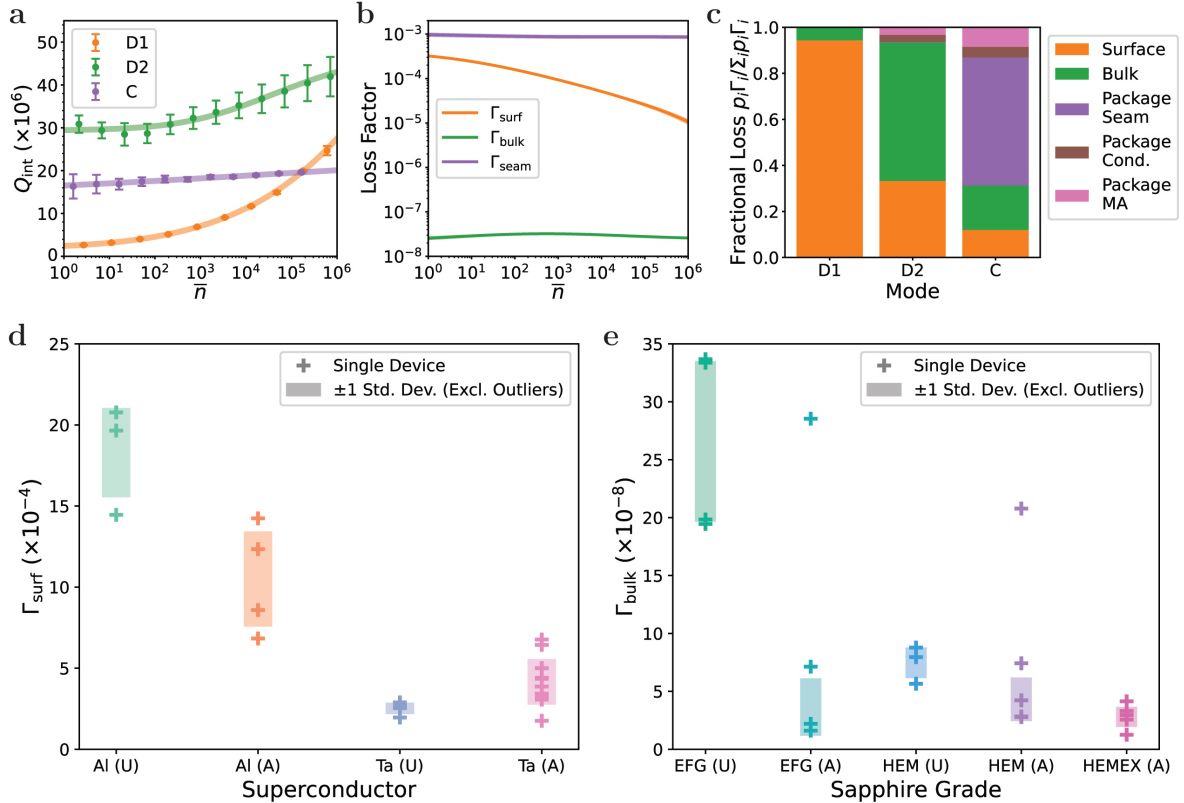
To extract the intrinsic loss factors and distinguish them from the geometric contribution to the total loss, we use the participation ratio model to define a linear system of equations  $\kappa_j = \sum_i P_{ji} \Gamma_i$ , where  $\kappa_j = 1/Q_j$  for the  $j$ th mode of the tripole stripline, and  $P_{ji}$  is the participation matrix (see Supplementary Table S5) of the loss characterization system. The problem reduces to solving a matrix equation using a least-squares algorithm with solution  $\vec{\Gamma} = \mathbf{C} \tilde{\mathbf{P}}^T \vec{\tilde{\kappa}}$ [59], where  $\tilde{P}_{ji} = P_{ji}/\sigma_{\kappa_j}$  and  $\tilde{\kappa}_j = \kappa_j/\sigma_{\kappa_j}$  are the measurement-error-weighted participation matrix and internal loss, respectively, and  $\mathbf{C} = (\tilde{\mathbf{P}}^T \tilde{\mathbf{P}})^{-1}$  is the covariance matrix. The measurement uncertainty  $\sigma_{\kappa_j}$  of the internal loss  $\kappa_j$  propagates onto the uncertainty of the extracted loss factor as  $\sigma_{\Gamma_i} = \sqrt{C_{ii}}$  (see Methods “[Extraction of loss factors using least-squares minimization](#)”). We use

the TLS model from Eq. (2) as an interpolating function to determine  $Q_{\text{int}}$  at all values of  $\bar{n}$  and extract the intrinsic surface, bulk, and package-seam loss factors as a function of  $\bar{n}$  using the least-squares algorithm (Fig. 2b). The contributions of conductor and MA surface losses from the package were calculated using previously measured loss factors for 5N5 aluminum (see Methods “[Subtraction of package conductor and dielectric losses](#)”).

Mapping the mode quality factors to geometry-independent loss factors in this way allows us to observe general trends in different sources of loss. We see that the surface-dependent D1 mode is power-dependent while the others are significantly less so. This implies that the surface loss factor is power-dependent while the other loss factors are not, and that the small power dependence of the D2 and C modes stem from their small but nonzero surface participation. Indeed, this is confirmed in Fig. 2b, where we observe the extracted surface loss factor is heavily power-dependent in sharp contrast with the bulk and seam loss factors. The relative power independence of the bulk and package loss factors also implies that the TLSs that dominantly couple to superconducting microwave resonators are localized in surface dielectric regions[25]. The distinction between surface and bulk dielectric loss is also apparent in the several orders of magnitude difference between the corresponding loss factors. We extract a single-photon bulk loss factor of  $(2.6 \pm 0.2) \times 10^{-8}$ , while the extracted single-photon surface loss factor is nearly 4 orders of magnitude higher at  $(3.4 \pm 0.3) \times 10^{-4}$ , which is qualitatively similar to what is observed in other studies[38, 48, 51].

To quantify the extent to which each mode is limited to a particular source of loss, we calculate a single-photon loss budget by plotting the fractional loss contribution  $p_i \Gamma_i / \sum_i p_i \Gamma_i$  of each source of loss for each mode in Fig. 2c. The loss budget for the three modes shows that the tripole stripline fulfills the ideal conditions for a loss characterization system: each mode’s  $Q_{\text{int}}$  is dominated by a different source of loss.

To measure how the choice of sapphire grade, wafer annealing treatment, and superconducting thin-film process affects the bulk and surface loss factors (Figs. 2d, e), we repeat the multimode approach for a variety of materials and process



**Fig. 2: Extraction of intrinsic losses with the tripole stripline.** **a** Power dependence of internal quality factor of the modes of a particular tripole stripline device, made using tantalum patterned on an annealed HEMEX sapphire substrate. Circles are measured  $Q_{\text{int}}$ ; lines are TLS fits using Eq. (2). Error bars represent the propagated fit error on  $Q_{\text{int}}$  obtained from least-squares minimization of Eq. (9) and for some points are small enough to not be visible. **b** Power dependence of extracted loss factors (solid lines). Propagated errors (shaded regions) are small ( $\sim 10\%$ ) and are hidden within the width of the solid lines. Seam loss here has been normalized to be dimensionless,  $\Gamma_{\text{seam}} = \omega\epsilon_0/g_{\text{seam}}$ . **c** Single-photon loss budget for the modes of the tripole stripline. While the D1 mode is clearly dominated by surface loss, the D2 mode is dominated by bulk dielectric loss, and the C mode is dominated by seam loss. **d, e** Comparison of surface (**d**) and bulk (**e**) loss factors from multiple tripole stripline devices made using either aluminum- or tantalum-based fabrication processes on annealed (A) or unannealed (U) sapphire substrates.

combinations. Multiple devices were measured for each set of materials and fabrication processes to capture the device-to-device variation of loss factors. We remark that while some outliers exist, the majority of the data points for each materials and process combination are well clustered; average and standard deviation of the loss factors are calculated excluding outliers with median relative deviation greater than 3 (see Supplementary Table S3).

We find that surface loss factors can be highly dependent on initial substrate treatment,

type of superconductor, and lithography process. Aluminum-based fabrication processes on unannealed substrates yield the largest surface loss factors, while annealing the substrate improves the surface losses by a factor of two. However, the tantalum-based fabrication process yields over a factor of 2 reduction in surface loss when compared to the best aluminum-based process regardless of whether the substrate was annealed. Cross-sectional transmission electron microscopy (TEM) of aluminum- and tantalum-based devices revealed marked differences in the MS interface.

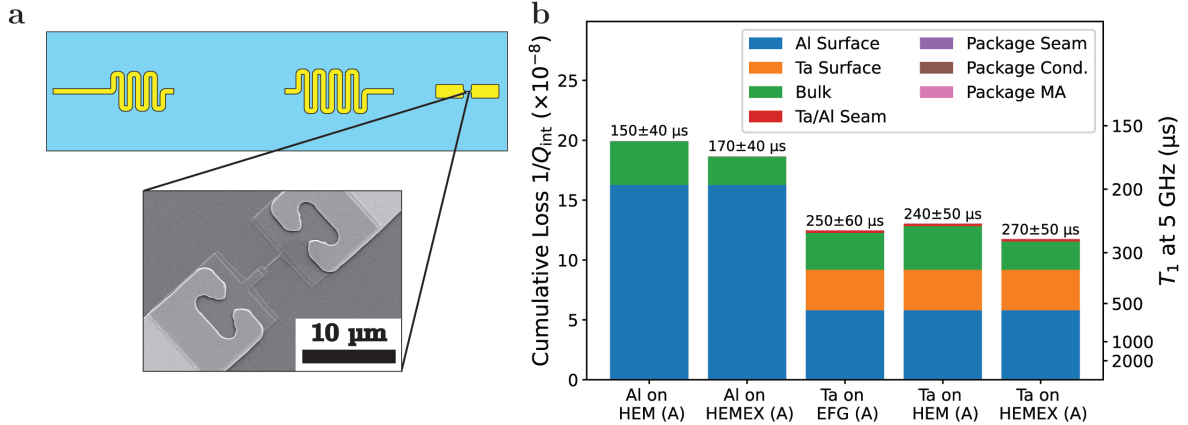
Whereas the aluminum films had a thin,  $\approx 2$  nm-thick amorphous region between the metal and the substrate, the tantalum films had a clean interface with nearly epitaxial film growth and no observable sign of an amorphous region (see Supplementary Note 8: “TEM film characterization”). It should be noted that the aluminum-based devices were deposited using electron-beam evaporation and patterned using a liftoff process while the tantalum-based devices were deposited via high-temperature sputtering and patterned using a subtractive process (see Methods “[Device fabrication](#)”). Therefore, the effects of these processes on surface quality must be considered as a convolution of the materials used and the fabrication processes employed. The differences in surface quality of aluminum- and tantalum-based thin-films may be due to differences in deposition conditions, lithographic patterning, or materials compatibility with the substrate, all of which can influence how the film grows on the substrate[60, 61].

Extracted bulk loss factors also vary based on choice of sapphire grade and annealing treatment. We find that annealing EFG- and HEM-grade sapphire results in reductions in bulk dielectric loss by factors of approximately 8 and 2, respectively. Additionally, annealing HEMEX-grade sapphire yields the lowest bulk loss with the smallest amount of device-to-device variation as measured over six devices. The improvement through annealing is correlated with improved surface morphology observed through atomic force microscopy (AFM), which revealed atomically-flat surfaces with a monatomically-stepped terrace structure after annealing (see Supplementary Note 6: “Sapphire annealing”). It should be noted that while the difference between unannealed EFG and HEM is in qualitative agreement with other studies[48, 62], the absolute bulk loss tangents differ significantly. This discrepancy can be due to the effects of the substrate undergoing the fabrication process. The samples in Read et al. [48] were cleaned, cleaved and measured with no further processing. Our measurements were taken after the substrate had been through the entire fabrication process; most notably, the wafer was diced, which is a more violent process that causes chipping of the sapphire at the edges and may cause more subsurface damage that could affect the bulk loss factor.

Finally, while we observe moderate device-to-device variation in surface and bulk loss factors, we observe the extracted seam losses to vary by over two orders of magnitude over multiple nominally identically-prepared cylindrical tunnel packages (Supplementary Table S2). Device-to-device variation in interface quality due to residual contamination, interface roughness, and clamping force can result in large variations in seam conductance. This highlights the significance of package losses in the coaxial architecture as a potential source of large device-to-device variation in  $Q_{\text{int}}$ . However, tripole striplines are capable of characterizing this variation due to the seam-loss-sensitivity of the common mode. Moreover, the high-Q modes in this section and in future sections are designed to be insensitive to seam loss, rendering it a relatively insignificant contributor to the total internal loss. We can nevertheless calculate an expected seam conductance per unit length  $g_{\text{seam}} = (2.1 \pm 2.0) \times 10^2 (\Omega\text{m})^{-1}$  by excluding outliers with a large relative deviation from the median (see Supplementary Table S2); the large uncertainty on this value is a reflection on the intrinsic variation we should expect in a device made using this particular architecture.

## Validating the loss model with qubit measurements

Microwave loss characterization is useful insofar as it can be applied to understand the losses of a candidate device of desired geometry. We demonstrate utility of our loss characterization technique by using the extracted loss factors from the previous section to predict the internal quality factors of transmon qubits. We subsequently verify our predictions by comparing them with measured transmon coherence. Transmon qubits of a particular design (Fig. 3a) were co-fabricated with the tripole striplines to ensure that the transmon validation devices were subjected to the same processing as the loss characterization devices. Tantalum-based transmons were fabricated by subtractively patterning the capacitor pads using tantalum, and Dolan bridge-style Al/AIO<sub>x</sub>/Al Josephson junctions were additively patterned using double-angle shadow evaporation followed by liftoff (see Methods “[Device fabrication](#)”). Aluminum-based transmons were fabricated in the same way as the junctions in the tantalum-based process, except



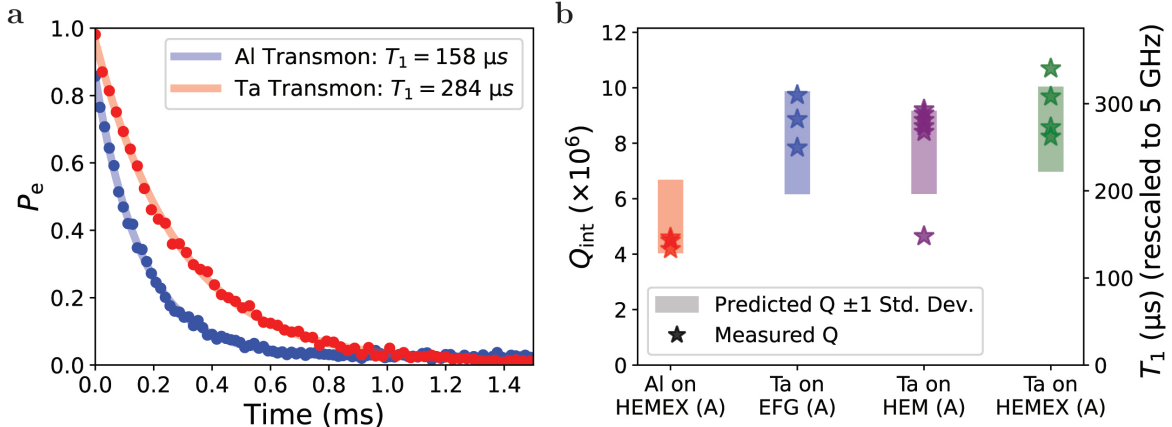
**Fig. 3: Prediction of transmon loss.** **a** 3D transmon qubit design, from which the participation ratios were calculated. Inset: SEM of Josephson junction and near-junction region on a tantalum-based transmon. Ta leads connect to the Al junction through an overlapping Ta/Al contact region. **b** Predicted loss and expected  $T_1$  for transmons made using different materials and processes (Al vs. Ta capacitor pads). Loss budget is also computed, showing the dominant sources of loss in Al- and Ta-based transmons.

the capacitor and the Josephson junction were patterned in a single electron-beam lithography step.

To describe the losses in aluminum and tantalum-based transmon qubits, we once again invoke the participation ratio model (see Supplementary Table S8). The aluminum-based transmon is limited by the same sources of loss as aluminum tripole striplines: surface losses associated with the aluminum thin-film growth and patterning, bulk dielectric losses associated with the substrate, and package losses. The tantalum-based transmon has both tantalum and aluminum regions and is susceptible to surface loss associated with both the tantalum capacitor pads and the aluminum region near the junction. Additionally, due to the use of two separate metals deposited in different deposition steps, the contact between the tantalum and aluminum may also manifest loss in analogy with seam loss. Tantalum oxide or other contaminants located in the Ta/Al interface may contribute to an effective resistance in series with the Josephson junction. We characterized this loss in the microwave regime using a segmented stripline that is highly sensitive to Ta/Al contact loss and extract a seam resistance of  $260 \pm 47 \text{ n}\Omega$  (see Supplementary Note 2: “Extracting Ta/Al contact loss”), which would limit the quality factor of the transmon to over  $5 \times 10^8$ , indicating that Ta/Al contact loss is negligible.

By combining the transmon participations with the extracted loss factors from the tripole and segmented striplines, we can compute the expected device coherence of aluminum- and tantalum-based transmon qubits on different types of annealed sapphire (Fig. 3b). Aluminum-based transmons are expected to achieve  $T_1$  of  $150 - 170 \mu\text{s}$  at 5 GHz, limited primarily by surface loss due to the aluminum-based process. By replacing the capacitor pads with tantalum using its respective process, the reduced surface loss is expected to yield dramatically improved  $T_1$ 's that exceed  $240 \mu\text{s}$ , regardless of sapphire grade. However, nearly half of the tantalum-based transmon's loss is from the near-junction aluminum region which is now the dominant factor that limits transmon relaxation. We attribute this to the small capacitance of the junction electrodes, which induces large electric fields that are localized in a small area, leading to high surface participation in the aluminum region (see Supplementary Table S8). Additionally, as new materials systems are developed that result in reduced surface loss, bulk dielectric loss begins to play a significant role. Already, bulk loss accounts for 15-20% of the tantalum-based transmon's loss; as a result, the microwave quality of the substrate must be considered as coherence continues to improve[48]. Finally, losses associated with the Ta/Al contact





**Fig. 4: Predicted vs. measured transmon quality factors.** **a** Representative Al- and Ta-based transmon  $T_1$  curves showing an almost factor of 2 improvement by adopting a tantalum-based process. **b** Measured transmon  $Q_{\text{int}}$  compared with predictions. Stars represent the 90th percentile transmon  $Q_{\text{int}}$  of a distribution formed from repeated coherence measurements over a 10-hour period. Shaded regions represent a predicted range spanning one standard deviation away from predicted transmon  $Q_{\text{int}}$ . Measured qubit frequencies ranged from 4.5 GHz to 6.7 GHz.

region and the package are predicted to be negligible; the first being due to the low Ta/Al contact resistance, the second being due to the compact electromagnetic field profile of the transmon.

To verify the predicted transmon losses and validate our understanding of decoherence mechanisms and their roles in determining coherence, several aluminum- and tantalum-based transmons were fabricated on different grades of annealed sapphire, and their measured quality factors were compared with the ranges predicted using the transmon loss model. Consistent with the predicted transmon loss, representative  $T_1$  measurements show an almost factor of two improvement in a tantalum-based transmon over an aluminum-based transmon (Fig. 4a). Each transmon’s coherence was also measured over a period of at least 10 hours to capture temporal fluctuations, and the predictive loss model showed remarkable consistency with the 90th percentile of transmon  $T_1$ , with the vast majority of measured  $Q_{\text{int}}$  falling inside one standard deviation of the predicted  $Q_{\text{int}}$  (Fig. 4b). These measured  $Q_{\text{int}}$ ’s are also similar to those measured in other studies[36, 37]. The choice of comparing 90th percentile  $T_1$  measurements with the loss predictions was done to discount the effects of fluctuations of TLSs interacting in the region of the Josephson junction. Despite the statistical expectation of zero TLSs present in

the junction[24, 63–65], this region’s small area and high energy density renders the transmon highly sensitive to deviations from that expectation due to stochastically fluctuating TLSs both in space and frequency[66, 67] over long periods of time. As a result, the transmon  $T_1$  can fluctuate tremendously over hours-long timescales (see Supplementary Note 5: “Temporal fluctuations of coherence in transmons, quantum memories, and resonators”). In contrast, this behavior is not seen in our resonators; resonator  $Q_{\text{int}}$ ’s measured over long timescales fluctuate by approximately  $\pm 10\%$ . We attribute this to the resonator’s much larger area and more uniformly distributed electric field; single TLS fluctuations are not expected to dramatically affect resonator  $Q_{\text{int}}$ . As a result, loss factors extracted from resonator measurements can be used to predict the upper (90th percentile) range of  $T_1$ ’s achievable by a transmon as its coherence fluctuates over long timescales.

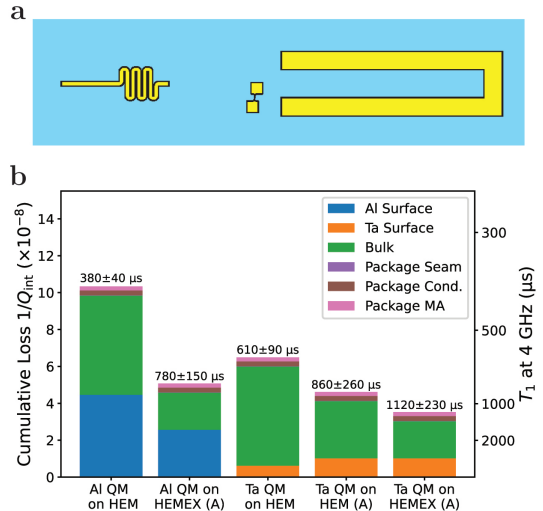
### Optimized geometry to maximize coherence in a quantum memory

Our loss analysis has thus far shown that tantalum-based transmons can achieve high  $T_1$ ’s but are significantly limited by surface participation near the junction. This motivates a more

optimized design choice where we use a linear resonator to encode quantum information[40]. Linear resonators tend to have their electromagnetic fields distributed over a larger area, which leads to reduced surface participation and therefore a higher  $Q_{\text{int}}$ , regardless of what materials or fabrication processes are employed. Furthermore, the lack of a Josephson junction renders the resonator much less sensitive to TLS fluctuations, leading to more temporally-stable coherence and dramatically suppressed pure dephasing. This has been demonstrated to great success using 3D cavity resonators as quantum memories[40–42], where logical qubits are encoded using the bosonic states of the resonator. While thin-film resonators have also been shown to be a viable candidate to be used as quantum memories[20, 53], their coherence has thus far been far below their 3D counterparts. However, with the advancements in materials and fabrication processes demonstrated in this work, thin-film resonator  $Q_{\text{int}}$ 's exceeding  $3 \times 10^7$  have been achieved at single-photon powers (Fig. 2a). It is therefore possible to optimize the design of a resonator to support a highly coherent on-chip quantum memory within the coaxial architecture. Such a device would have all the advantages of a planar device due to its more compact design and ability to be patterned with lithographic precision.

To implement a highly coherent on-chip quantum memory, we have developed the hairpin stripline, a multimode device whose fundamental mode is optimized to balance package and surface loss to maximize its  $Q_{\text{int}}$  (Fig. 5a). The electromagnetic fields of this memory mode are localized primarily between the two arms of the hairpin, rendering it insensitive to package losses, while the large spacing between the two arms dilutes the electric field at the surfaces, thereby reducing surface loss. An ancilla transmon couples dispersively both to the memory mode to enable its fast control, and to the second-order mode of the hairpin stripline, which acts as a readout mode without needing to introduce additional hardware complexity (see Supplementary Table S11).

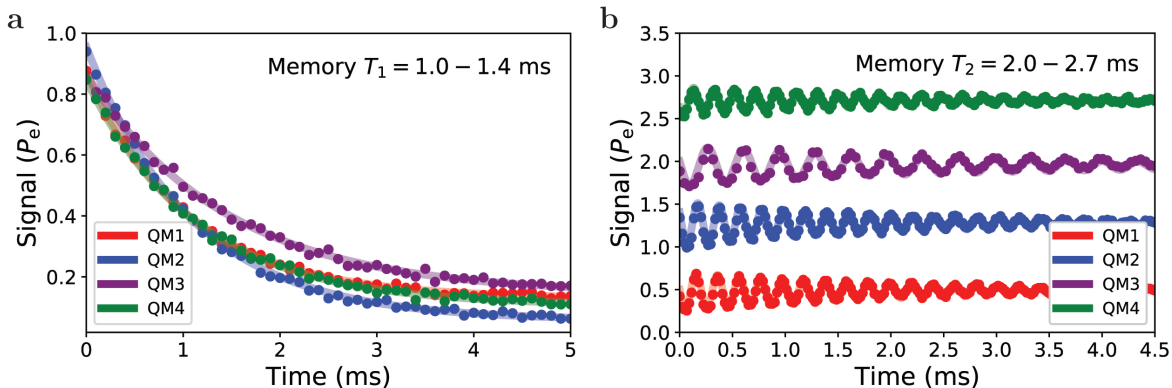
To demonstrate the improvements in coherence achievable by optimizing materials and process choices, we apply the predictive loss model to the hairpin stripline and show that an aluminum-based process employed on unannealed HEM sapphire is not expected to produce remarkable



**Fig. 5: Hairpin stripline quantum memory.** **a** Hairpin stripline quantum memory design. The ancilla transmon couples to the fundamental mode that acts as a storage resonator, and to the higher order mode that acts as a readout resonator. A Purcell filter is used to enhance the external coupling of the readout mode. **b** Predicted loss and expected  $T_1$  for hairpin striplines made using different substrate preparations and different superconducting thin-films.

coherence (Fig. 5b), and replacing the aluminum with a tantalum-based process leads to a modest expected improvement. Additional modest improvements are expected when annealed sapphire substrates are used; however, when both high temperature substrate annealing and tantalum processes are employed, the hairpin stripline is expected to achieve a  $T_1$  of  $(1.1 \pm 0.2)$  ms, which rivals the coherence of commonly used quantum memories realized in 3D coaxial  $\lambda/4$  post-cavities[40]. This dramatic improvement is only achieved when both materials and geometry are optimized to minimize both surface and package participation, resulting in a predominantly bulk loss-limited device.

Four hairpin stripline-based quantum memories were fabricated using a tantalum process on annealed HEMEX-grade sapphire substrates. The devices were measured in the same cylindrical tunnel packages used to measure the triple striplines and transmon qubits. Memory  $T_1$  and  $T_2$  in the Fock ( $|0\rangle, |1\rangle$ ) manifold were



**Fig. 6: Hairpin stripline quantum memory coherence.** **a** Fock state  $T_1$  measurement of four on-chip quantum memory devices. The Fock  $|1\rangle$  state was prepared using Selective Number-dependent Arbitrary Phase (SNAP) gates and the memory state was inferred after a variable delay by selectively flipping the ancilla qubit conditioned on the memory being in the Fock  $|1\rangle$  state, and measuring the ancilla state. Memory  $T_1$ 's for QM1-4 extracted by fitting an exponential to the ancilla state as a function of time were 1.05, 1.09, 1.44, and 1.14 ms. **b** Memory  $T_2$  in the Fock  $(|0\rangle, |1\rangle)$  manifold for the four devices measured in **a**. The Fock state  $\frac{1}{\sqrt{2}}(|0\rangle + |1\rangle)$  was prepared using SNAP gates and after a variable delay a small displacement was applied to interfere with the memory state, followed by measurement in the same way as in **a**[40]. Ancilla state as a function of time for QM2-4 were offset vertically by 0.75, 1.5, and 2.25, respectively, for visibility, and were fit to an exponentially decaying sinusoid. Extracted memory  $T_2$ 's for QM1-4 were 2.02, 2.00, 2.68, and 2.14 ms.

measured using the same pulse sequences as in Reagor et al. [40]. Quantum memory coherence was remarkably consistent with predictions (Fig. 6); Fock state decay times were measured to be 1-1.4 ms. Additionally, measured Fock  $T_2$  times approached  $2T_1$ , which bounds  $T_\phi > 24$  ms, similar to 3D cavity-based quantum memories[42, 68]. Additionally, continuous coherence measurements over 20 hours showed minimal temporal fluctuations in  $T_1$  and  $T_2$ ; coherence fluctuated by no more than  $\pm 10\%$  over hours-long timescales, a markedly different behavior from transmon qubits and consistent with a much reduced sensitivity to TLS fluctuations (Supplementary Fig. S9c).

## Discussion

We have introduced a technique for characterizing microwave losses in thin-film resonators. We have shown that depending on resonator geometry, the surface, bulk and package losses can be significant contributors to the total internal loss of a microwave resonator. We have also observed that our tantalum-based fabrication processes tend to yield higher internal quality factors due to

improvements in surface quality, and that annealing sapphire substrates results in dramatically reduced bulk dielectric loss tangents. Additionally, we have shown that by understanding sources of loss in resonators, we can make and experimentally verify predictions of losses in co-fabricated transmon qubits. By analyzing the sources of loss that limit state-of-the-art devices, we have utilized a powerful tool that revealed comprehensively what limits transmon coherence, and motivated the design of an optimized stripline-based quantum memory using thin-film superconductors patterned on a substrate. While our loss characterization results are specific to our materials and fabrication processes, the participation ratio model provides a versatile approach to loss characterization that can be adapted for the co-planar waveguide, flip-chip, or cavity-based cQED architectures; additional materials and loss channels can be straightforwardly studied by designing the appropriate participation matrix and introducing new devices or modes to characterize them (see Supplementary Note 2: “Extracting Ta/Al contact loss”).

The implementation of a quantum memory in a stripline enables a coaxial architecture that is more scalable, more modular, and more compact than the more traditional cavity approach[20]. Ancilla-memory couplings can be lithographically defined, enabling far greater precision in device design. By employing a well-controlled fabrication process, consistently high device coherence can be achieved. Multiqubit systems can be more straightforwardly and compactly designed, as the striplines themselves are more compact than their 3D counterparts. Multiple devices can be fabricated on a single wafer and easily redesigned without modifying the package, allowing increased modularity. Additionally, the low pure dephasing observed in these devices allows for the implementation of noise-biased qubits, which can enable lower error correction thresholds towards the implementation of surface codes of dual-rail qubits[9, 69]. Stripline-based quantum memories therefore provide a promising building block for realizing large-scale quantum computing with bosonic modes.

Finally, the loss characterization studies presented in this work have shown clear paths forward for improving coherence in superconducting qubits. Transmons are significantly limited by surface participation near the Josephson junction; as a result, developing better processes or using intrinsically lower-loss materials in these regions may be critical towards improving transmon coherence to one millisecond and beyond. Additionally, improvements in surface loss must also accompany improvements in bulk dielectric loss; this is especially important for stripline-based quantum memories, which are dominated by bulk loss. This work demonstrates important techniques that help to provide understanding of coherence-limiting mechanisms and inform optimization and design choices for superconducting quantum circuits.

## Methods

### Device fabrication

All devices were fabricated on c-plane sapphire substrates grown using either the edge-film growth (EFG) method or heat-exchange method (HEM). HEMEX wafers were additionally graded HEM wafers based on superior optical

properties[48, 70]. All substrates were initially cleaned in a piranha solution (2:1  $\text{H}_2\text{SO}_4:\text{H}_2\text{O}_2$ ) for 20 minutes[36], followed by a thorough rinse in DI water for 20 minutes. Substrates were then optionally annealed in a FirstNano EasyTube 6000 furnace at 1200 °C in an oxygen-rich environment. The furnace was preheated to 200 °C and purged with nitrogen prior to wafer loading. The furnace was then purged with pure oxygen, followed by a gradual heating to 1200 °C at a controlled ramp of 400 °C/hr while continuously flowing oxygen. Once the furnace reached 1200 °C, the gas flows were shut off and the wafers were allowed to anneal for one hour in the oxygen-rich ambient conditions. Finally, the wafers were passively cooled over approximately 6 hours by turning off the furnace heaters and flowing a 4:1 mixture of  $\text{N}_2:\text{O}_2$  gas.

For tantalum-based devices, tantalum was deposited after the cleaning and optional annealing by DC magnetron sputtering while maintaining a substrate temperature of 800 °C. Tantalum was sputtered using an Ar pressure of 6 mTorr and a deposition rate of 2.5 Å/s. After deposition, the substrate was cooled at a controlled rate of 10 °C/min to prevent substrate damage due to the differential contraction of the Ta film and the sapphire surface. Tantalum films deposited this way were consistently in the (110) or (111)-oriented  $\alpha$ -phase as shown by X-ray diffractometry (XRD) (Supplementary Fig. S11b) and have  $T_c > 4.1$  K with  $\text{RRR} > 15$  (our best film has a  $T_c = 4.3$  K and  $\text{RRR} = 55.8$ , see Supplementary Fig. S11a). To pattern the tantalum, S1827 photoresist was spun on the wafer after deposition and patterned using a glass photomask and a Suss MJB4 contact aligner. After developing in Microposit MF319 developer for 1 minute, the wafer was hard-baked for 1 minute at 120 °C and treated with oxygen plasma using an AutoGlow 200 at 150 W and 300 mTorr  $\text{O}_2$  for 2 minutes to remove resist residue. The tantalum was etched at a rate of 100 nm/min in an Oxford 80+ Reactive Ion Etcher using  $\text{SF}_6$  with a flow rate of 20 sccm, a pressure of 10 mTorr, and an RF power of 50 W. After etching, photoresist was removed by sonicating for 2 min each in N-Methylpyrrolidone (NMP), acetone, isopropanol, and DI water. To remove any remaining organic residue, an additional 20 minute piranha cleaning step was performed, and to remove excess tantalum oxide that may have

grown due to the strong oxidizing nature of the piranha solution, an oxide strip was performed by dipping the wafer in Transene 10:1 BOE for 20 minutes[38, 71], followed by a 20 minute rinse in DI water.

Josephson junctions and aluminum devices were patterned using electron-beam lithography. The wafer was first dehydrated by baking at 180 °C for 5 minutes. Then, a bilayer of 700 nm MMA (8.5) MAA EL13 and 200 nm of 950K PMMA A4 was spun, with a 5 minute bake at 180 °C following the spinning of each layer. To eliminate charging effects during electron-beam writing, a 15 nm aluminum anti-charging layer was deposited by electron-beam evaporation. Electron-beam lithography was then performed using a Raith EBPG 5200+ to define the Dolan-bridge shadowmask. The anticharging layer was then removed by immersing the wafer in Microposit MF312 developer for 80 seconds, and the pattern was developed in 3:1 IPA:H<sub>2</sub>O at 6 °C for 2 minutes. The wafer was then loaded into the load-lock of a Plassys UMS300 electron-beam evaporator, where an Ar ion beam clean was performed at 400 V to remove the tantalum oxide and other surface residues prior to aluminum deposition. The wafer was tilted by  $\pm 45$  degrees and the ion beam cleaning was performed for 34 seconds at each angle in order to remove the oxide on the tantalum sidewall and to clean the region underneath the Dolan bridge. The same cleaning process was employed prior to deposition of the aluminum-based devices. Following the ion beam clean, the wafer was transferred to the evaporation chamber where a double-angle evaporation of aluminum was performed at  $\pm 25$  degrees (20 nm followed by 30 nm) with an interleaved static oxidation step using an 85:15 Ar:O<sub>2</sub> mixture at 30 Torr for 10 minutes. After the second aluminum deposition, a second static oxidation step was performed using the same Ar:O<sub>2</sub> mixture at 100 Torr for 5 minutes in order to cap the surface of the bare aluminum with pure aluminum oxide. Liftoff was then performed by immersing the wafer in NMP at 90 °C for 1 hour, followed by sonication for 2 minutes each in NMP, acetone, isopropanol, and DI water. The wafer was then coated with protective resist before dicing into individual chips with an ADT ProVectus 7100 dicer, after which the chips were cleaned by sonicating in NMP, acetone, isopropanol, and DI water.

## Device packaging

All striplines, transmons, and quantum memories were measured in cylindrical tunnel packages made out of conventionally-machined high-purity (5N5) aluminum (Supplementary Fig. S2). The packages underwent a chemical etching treatment using a mixture of phosphoric and nitric acid (Transene Aluminum Etchant Type A) heated to 50 °C for two hours[30]. The tunnels were approximately 34 mm long and 5 mm in diameter. Coupling was accomplished by a transverse feedline, allowing for multiple tunnels to be arranged side-by-side and measured in a multiplexed hanger configuration[20]; the same feedline is used for qubit, storage mode, and readout drives. The 40 mm  $\times$  4 mm chips on which the devices are fabricated are inserted into the tunnel package and clamped on either end by beryllium-copper leafsprings. The clamps on either end of the tunnel also serve as end-caps for the tunnels themselves, thereby defining the locations of the seams and completing the enclosure.

## Measurement setup

A fridge wiring diagram can be found in Supplementary Fig. S1. Device packages are mounted to the mixing chamber stage of a dilution refrigerator operating at 20 mK. The packages are enclosed in multiple layers of shielding. First, a light-tight gold-plated copper shield internally coated with Berkeley black acts as an IR photon absorber[72]. A superconducting shield made of 1/64" thick lead foil is wrapped around the copper shield. Finally, a mu-metal can serves as the outermost shield to attenuate the ambient magnetic fields at the package. Input lines are attenuated at both the 4 K stage (20 dB) and mixing chamber stage (50-60 dB depending on the line; 20 dB of reflective attenuation is achieved through the use of a directional coupler) and are filtered at multiple locations using 12 GHz K&L low-pass filters and custom-made eccosorb CR-110 IR filters. Output lines are also low-pass filtered and isolated from the devices using circulators and isolators. A SNAIL parametric amplifier (SPA) is used on the qubit output line to provide quantum-limited amplification for qubit readout. HEMT amplifiers at the 4 K stage provide additional low-noise amplification for the output signals.

Resonators are measured in frequency domain using a vector network analyzer (Agilent E5071C). Qubits and quantum memories are measured in time domain using an FPGA-based quantum controller (Innovative Integration X6-1000M) which can output arbitrary waveforms at  $\approx 50$  MHz that are then up-converted to GHz frequencies using an LO tone generated by an Agilent N5183A (Readout drive uses a Vaunix LMS-103 for the LO) and a Marki IQ-0307-LXP mixer. Qubit, readout, and storage mode drives are all generated the same way and are combined and amplified using a Mini-Circuits ZVA-183-S+. The signals are finally attenuated by a room-temperature 3 dB attenuator to reduce the thermal noise temperature before being fed into the fridge. Readout responses from the fridge are amplified with a room-temperature amplifier (MITEQ LNA-40-04001200-15-10P) and isolated before being down-converted using a Marki IR-0618-LXP mixer (the same LO is used for both the up-conversion and down-conversion of the readout signals). Down-converted signals are then amplified using a Mini-Circuits ZFL 500 before being fed into the ADC of the FPGA. All signal generator sources and VNA are clocked to a 10MHz Rb frequency standard (SRS FS725).

## Calculation of participation ratios

Energy participation in various lossy regions are calculated using the commercial finite-element electromagnetic solver Ansys HFSS and the two-step meshing method detailed in [19]. Thin-film conductors are approximated in a 3D electromagnetic simulation as perfectly conducting 2D sheets. Field behavior at the edges of the thin-films are approximated using a heavily meshed 2D cross-sectional electrostatic simulation with explicitly-defined surface dielectric regions of assumed thickness  $t_{\text{surf}} = 3$  nm and relative permittivity  $\epsilon_r = 10$  to maintain consistency with other works[38, 43, 44]. The true thickness and relative permittivity of these regions are unknown; while nanometer-scale microscopy of these interfaces can yield qualitative information about these interfaces, it cannot definitively reveal the dielectric properties or the presence or absence of physical signatures of loss. We therefore treat the true surface region thickness and relative permittivity as material/process parameters that re-scale the surface loss tangents and thereby define the intrinsic

loss factor that corresponds to  $p_{\text{surf}}$  as  $\Gamma_{\text{surf}} = \sum_{k=\text{SA,MS,MA}} \frac{p_k}{p_{\text{surf}}} \frac{t_{k_0}}{t_{\text{surf}}} \frac{\epsilon_{r_0}}{\epsilon_r} \tan \delta_k$ , where  $\tan \delta_k$ ,  $t_{k_0}$ , and  $\epsilon_{r_0}$  are the true dielectric loss tangent, thickness of the surface regions, and true dielectric constant, respectively;  $t_{\text{surf}} = 3$  nm and  $\epsilon_r = 10$  are the initially assumed thickness and dielectric constant, respectively[33].

We define a combined surface participation term,  $p_{\text{surf}} = p_{\text{SA}} + p_{\text{MS}} + p_{\text{MA}}$  and define the corresponding surface loss factor as a weighted sum of the SA, MS, and MA loss factors (Supplementary Fig. S3). This construction of surface participation prevents us from distinguishing between the different surface losses, but because the relative scaling of these participations is roughly the same for all resonator geometries in this architecture, the geometric ratio  $p_k/p_{\text{surf}}$  is geometry-independent; therefore,  $\Gamma_{\text{surf}}$  still carries predictive power to estimate the loss of a desired resonator geometry. This formulation could also be modified to consider conductor loss in the thin-films, whose participation scales similarly to the surface dielectric participations. In such a case,  $\Gamma_{\text{surf}}$  is a surface loss factor that contains contributions from dielectric and conductor loss. Here, we assume conductor loss to be negligible, as aluminum thin-films have been shown to have residual quasiparticle fractions as low as  $x_{\text{qp}} = 5.6 \times 10^{-10}$ [72], where  $x_{\text{qp}} \sim \Gamma_{\text{cond}}$ . Assuming our tantalum films also have similarly low  $x_{\text{qp}}$ , we estimate the thin-film conductor loss to limit the tripole stripline modes to  $Q_{\text{int}} > 10^{10}$ .

We use the following integral equations to calculate the various on-chip and package participations in the coaxial tunnel architecture:

$$p_{\text{SA,MS}} = \frac{t_{\text{surf}} \int_{\text{SA,MS}} \epsilon |\vec{E}|^2 d\sigma}{\int_{\text{all}} \epsilon |\vec{E}|^2 dv} \quad (3)$$

$$p_{\text{MA}}, p_{\text{pkgMA}} = \frac{t_{\text{surf}} \int_{\text{MA}} \epsilon_0 |\vec{E}_{\text{vac}}|^2 d\sigma}{\epsilon_{r,\text{MA}} \int_{\text{all}} \epsilon |\vec{E}|^2 dv} \quad (4)$$

$$p_{\text{bulk}} = \frac{\int_{\text{bulk}} \epsilon |\vec{E}|^2 dv}{\int_{\text{all}} \epsilon |\vec{E}|^2 dv} \quad (5)$$

$$p_{\text{pkgcond}} = \frac{\lambda \int_{\text{surf}} \mu_0 |\vec{H}_{||}|^2 d\sigma}{\int_{\text{all}} \mu_0 |\vec{H}|^2 dv} \quad (6)$$

$$y_{\text{seam}} = \frac{\int_{\text{seam}} |\vec{J}_S \times \hat{l}|^2 dl}{\omega \int_{\text{all}} \mu_0 |\vec{H}|^2 dv}. \quad (7)$$

For  $p_{\text{SA,MS}}$ , integration was done over a surface located 3 nm below the 2D sheet. For  $p_{\text{MA}}$  and  $p_{\text{pkg}_{\text{MA}}}$ , integration was done over a surface located 3 nm above the 2D sheet. Because the MA surface dielectric region is not explicitly defined in the 3D simulation, the vacuum electric field was re-scaled to that of the MA field by invoking the continuity of the displacement field,  $\epsilon_r E_{\text{MA},\perp} = \epsilon_0 E_{\text{vac},\perp}$ .  $p_{\text{pkg}_{\text{cond}}}$  was calculated by integrating the magnetic field energy density over the surface of the package wall and multiplying it by the effective penetration depth  $\lambda$  of high-purity aluminum, which was previously measured to be 50 nm[30]. Finally,  $y_{\text{seam}}$  was calculated by integrating the current flow across the seam; both  $y_{\text{seam}}$  and  $g_{\text{seam}}$  have units  $(\Omega\text{m})^{-1}$ [55].

For transmons, a significant portion of the total magnetic energy is stored in the kinetic inductance of the Josephson junction; therefore, the total magnetic energy is calculated to include the energy stored in the junction,  $U_{H_{\text{tot}}} = \int_{\text{all}} \mu_0 |\vec{H}|^2 dv + \frac{1}{2} L_J I_J^2$ . Near-junction ( $<5 \mu\text{m}$  away) surface participations are calculated using an additional local 3D electrostatic simulation, and we invoke a similar argument as in Wang et al. [19] and exclude the participation contribution from a region within 100 nm of the junction itself. This exclusion follows from the assumption that surface dielectric loss is dominated by a TLS density of  $\sim 1 \mu\text{m}^{-2}\text{GHz}^{-1}$  and therefore the small region that is the junction itself should likely include zero TLSs and be lossless[24, 63–65]. This assertion that the junction be lossless is further supported by earlier studies that have bounded the loss tangent of the junction oxide to below  $4 \times 10^{-8}$ [73], and by recent quasiparticle tunneling experiments that have shown parity lifetimes on the order of hundreds of milliseconds if the appropriate radiation shielding and microwave filtering are used[72], which has been replicated in this work (see Methods “Measurement setup”).

### Extraction of loss factors using least-squares minimization

Starting with the matrix equation  $\kappa_j = \sum_i P_{ji} \Gamma_i$ , we use the least-squares fitting algorithm to extract the loss factors  $\Gamma_i$  and propagate the measurement error  $\sigma_{\kappa_j}$  onto the fit error  $\sigma_{\Gamma_i}$ [59]. If the

rank of  $\mathbf{P}$  is equal to or greater than the number of loss channels (i.e  $N_{\text{rows}} \geq N_{\text{columns}}$ ), the least-squares sum can be written down as:

$$S = \sum_j \left( \sum_i \tilde{P}_{ji} \Gamma_i - \tilde{\kappa}_j \right)^2, \quad (8)$$

where  $\tilde{P}_{ji} = P_{ji}/\sigma_{\kappa_j}$  and  $\tilde{\kappa}_j = \kappa_j/\sigma_{\kappa_j}$  are the measurement-error-weighted participation matrix and internal loss, respectively. We can then express the least-squares sum in matrix form as  $S = (\tilde{\mathbf{P}}\vec{\Gamma} - \vec{\tilde{\kappa}})^T (\tilde{\mathbf{P}}\vec{\Gamma} - \vec{\tilde{\kappa}})$  and solve for  $\vec{\Gamma}$  by setting  $\partial S/\partial \vec{\Gamma} = 0$  to obtain  $\vec{\Gamma} = \mathbf{C}\tilde{\mathbf{P}}^T \vec{\tilde{\kappa}}$ , where  $\mathbf{C} = (\tilde{\mathbf{P}}^T \tilde{\mathbf{P}})^{-1}$  is defined as the covariance matrix. We calculate the propagated error as  $\vec{\sigma}_{\vec{\Gamma}}^2 = \langle \delta \vec{\Gamma} \delta \vec{\Gamma}^T \rangle = \mathbf{C} \tilde{\mathbf{P}}^T \langle \delta \vec{\kappa} \delta \vec{\kappa}^T \rangle \tilde{\mathbf{P}} \mathbf{C}^T$ , and  $\langle \delta \tilde{\kappa}_i \delta \tilde{\kappa}_j \rangle = \langle \frac{1}{\sigma_i} \frac{1}{\sigma_j} \delta \kappa_i \delta \kappa_j \rangle = \delta_{ij}$ , so  $\vec{\sigma}_{\vec{\Gamma}}^2 = \mathbf{C} \tilde{\mathbf{P}}^T \tilde{\mathbf{P}} \mathbf{C}^T = \mathbf{C} \mathbf{C}^{-1} \mathbf{C}^T = \mathbf{C}$ . Therefore, the propagated error on the extracted loss factors are given by  $\sigma_{\Gamma_i} = \sqrt{C_{ii}}$ .

### Subtraction of package conductor and dielectric losses

Package losses are comprised of conductor, surface dielectric (MA), and seam losses. To quantify the conductor and MA losses, we use previously obtained loss factors for conventionally-machined 5N5 aluminum, measured using a multimode resonator made entirely of 5N5 aluminum called the forky whispering-gallery-mode resonator[33]. From the extracted losses of the two measured devices (F1(e) and F2(e)) we obtain  $R_s = (0.61 \pm 0.28) \mu\Omega$  and  $\tan \delta_{\text{pkg}_{\text{MA}}} = (4.1 \pm 1.8) \times 10^{-2}$ , where  $\Gamma_{\text{pkg}_{\text{cond}}} = R_s/(\mu_0 \omega \lambda)$  and  $\Gamma_{\text{pkg}_{\text{MA}}} = \tan \delta_{\text{pkg}_{\text{MA}}}$ , where  $R_s$  is the surface resistance of the superconductor,  $\lambda$  is the effective penetration depth of the superconductor,  $\omega$  is the frequency of the resonant mode, and  $\tan \delta$  is the surface dielectric loss tangent of the MA interface. Applying these loss factors to the tripole striplines measured in Fig. 2a, we obtain a package loss limit due to conductor and MA dielectric loss to be  $1/Q_{\text{D1}} = 1/17 \times 10^9$ ,  $1/Q_{\text{D2}} = 1/4.7 \times 10^8$ , and  $1/Q_{\text{C}} = 1/1.3 \times 10^8$  for the D1, D2, and C modes respectively. These package loss contributions indicate that they can be treated as residual losses, as they account for no more than 10-15% of the total loss of the common mode, with seam losses being the dominant source of package loss.

## Measurement of resonator quality factor

Microwave resonators were measured in the frequency domain using a vector network analyzer (VNA). The scattering parameter  $S_{21}$  describes the response to driving the resonator as a function of frequency and is given by

$$S_{21}(\omega) = ae^{i\alpha}e^{-i\omega\tau} \left[ 1 - \frac{(Q_L/|Q_c|)e^{i\phi}}{1 + i2Q_L(\omega/\omega_r - 1)} \right], \quad (9)$$

where  $a$  is the global attenuation of the input line,  $\alpha$  is the global spurious phase shift,  $\tau$  is the electrical delay,  $\omega_r$  is the resonance frequency, and  $Q_c = |Q_c|e^{-i\phi}$  is a complex coupling quality factor where  $\phi$  describes the asymmetry in the hanger response[56]. The real-valued loaded quality factor  $Q_L$  is the total quality factor due to both internal and external (coupling) loss,  $1/Q_L = 1/Q_{\text{int}} + \cos\phi/|Q_c|$ , where  $Q_{\text{int}}$  is the internal quality factor due to intrinsic material and process-based losses. The fitting methods used in Probst et al. [56] are robust in that fitting resonators that are overcoupled or undercoupled by as much as a factor of 10 is readily possible. Resonators measured in this work had  $|Q_c| = 2 - 10 \times 10^6$  and were therefore never too overcoupled or undercoupled. The excitation field of the resonator is determined by the input power  $P_{\text{in}}$  and can be expressed in terms of an average photon number in the resonator as  $\bar{n} = \frac{2}{\hbar\omega_r^2} \frac{Q_L^2}{Q_c} P_{\text{in}}$  (see Supplementary Note 10: ‘‘Derivation of resonator average photon number’’).

## Transmon coupling quality factor

Measured transmon  $T_1$  is proportional to the loaded quality factor of the mode,  $(\omega T_1)^{-1} = Q_L^{-1} = Q_{\text{int}}^{-1} + Q_c^{-1}$ . The quality factor predictions made in Fig. 3b are based on internal losses only; therefore, the coupling quality factor must be measured for transmons in order to properly compare predicted  $Q_{\text{int}}$  with measured  $Q_{\text{int}}$ . While this can be done using a finite element electromagnetics solver, the true  $Q_c$  is dependent on the transmon chip’s placement within the tunnel package and can vary by as much as 50% if the chip’s position varies by as little as 0.5 mm from the nominal.

We therefore determined the  $Q_c$  in situ by calibrating the qubit Rabi rate in the  $g$ - $e$  manifold as a function of drive power. The bare transmon Hamiltonian in the presence of a drive can be expressed as

$$H = \hbar\omega_q \hat{a}^\dagger \hat{a} - E_J \left[ \cos \hat{\Phi}_q + \left( 1 - \frac{1}{2} \hat{\Phi}_q^2 \right) \right] + \hbar\Omega_{\text{Rabi}} \cos \omega_d t (\hat{a}^\dagger + \hat{a}), \quad (10)$$

where  $\hat{a}$ ,  $\omega_q$ ,  $E_J$ , and  $\hat{\Phi}_q$  represent the transmon ladder operator, qubit transition frequency, Josephson energy, and flux operator, respectively. The term in the square brackets describes the non-linearity of the transmon, and is assumed to be small enough such that it can be applied perturbatively towards a simple harmonic oscillator Hamiltonian. The drive can be parameterized by a drive strength or Rabi rate  $\Omega_{\text{Rabi}}$  and a drive frequency  $\omega_d$ . We move into the rotating frame of the drive followed by the rotating frame of the transmon and the displaced frame of the drive to arrive at the following transformed Hamiltonian  $\tilde{H}$ :

$$\tilde{H} = -E_J \cos \hat{\Phi}_q - E_J \left( 1 - \frac{1}{2} \hat{\Phi}_q^2 \right), \quad (11)$$

where  $\hat{\Phi}_q = \phi_q(\tilde{a}^\dagger + \tilde{a} - \xi^* - \xi)$ , and  $\xi(t) = -\frac{i\Omega_{\text{Rabi}}e^{-i\omega_d t}}{\omega/Q_L + i2\Delta}$ , where  $\Delta = \omega_q - \omega_d$  and  $|\xi|^2$  is the photon number. Since the transmons are driven in a hanger configuration and can be approximated as a harmonic oscillator as long as leakage to higher computational states is negligible, we can relate the photon number to  $Q_c$  by  $\bar{n} = \frac{2}{\hbar\omega_r^2} \frac{Q_L^2}{Q_c} P_{\text{in}}$ . We can therefore derive the relation between the coupling  $Q$  and the qubit Rabi rate to be  $Q_c = \frac{2P_{\text{in}}}{\hbar\Omega_{\text{Rabi}}^2}$ . From this relation, we measure transmon  $Q_c$  to vary between  $30 - 70 \times 10^6$  due to variations in chip positioning within the tunnel, where the nominal positioning was simulated to yield  $Q_c \approx 40 \times 10^6$ . For our highest  $Q$  transmons, the external loss accounts for as much as 25% of the total loss. A solution to this extra loss is to simply undercouple the transmons even more from the drive line.

The  $Q_c$  for the hairpin striplines, on the other hand, were simulated to be approximately  $10^9$ .



Imprecision in chip positioning can also lead to significant variations in  $Q_c$  for this device; the resulting  $Q_c$  can vary between  $5 - 15 \times 10^8$ . However, the hairpin striplines have measured  $Q_L = 25 - 35 \times 10^6$ ; we therefore estimate that coupling loss accounts for less than 5% of the total loss of the hairpin stripline quantum memories.

## Data availability

Data available upon request.

## Code availability

Codes available upon request.

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## Author contributions

S.G. designed and simulated the tripole stripline, transmon qubits, and hairpin stripline quantum memories and fabricated the Ta-based devices, while Y.L. fabricated the Al-based devices and

implemented the measurement room-temperature electronics apparatus. S.G. measured the transmons and quantum memories, while Y.W. measured the tripole stripline resonators. S.G. and Y.W. performed loss analysis and loss budgeting. A.B. assisted with device fabrication and measurement. C.U.L and L.K assisted with device design and error analysis. K.K. performed TEM. C.Z., R.L., Y.J., and M.L. performed XRD and DC resistivity measurements. S.G., Y.W., L.F., and R.J.S. wrote the manuscript with feedback from all co-authors.

## **Competing interests**

L.F. and R.J.S. are founders and shareholders of Quantum Circuits Inc. (QCI)

**Supplementary information: Surpassing  
millisecond coherence times in on-chip  
superconducting quantum memories by  
optimizing materials, processes, and circuit  
design**



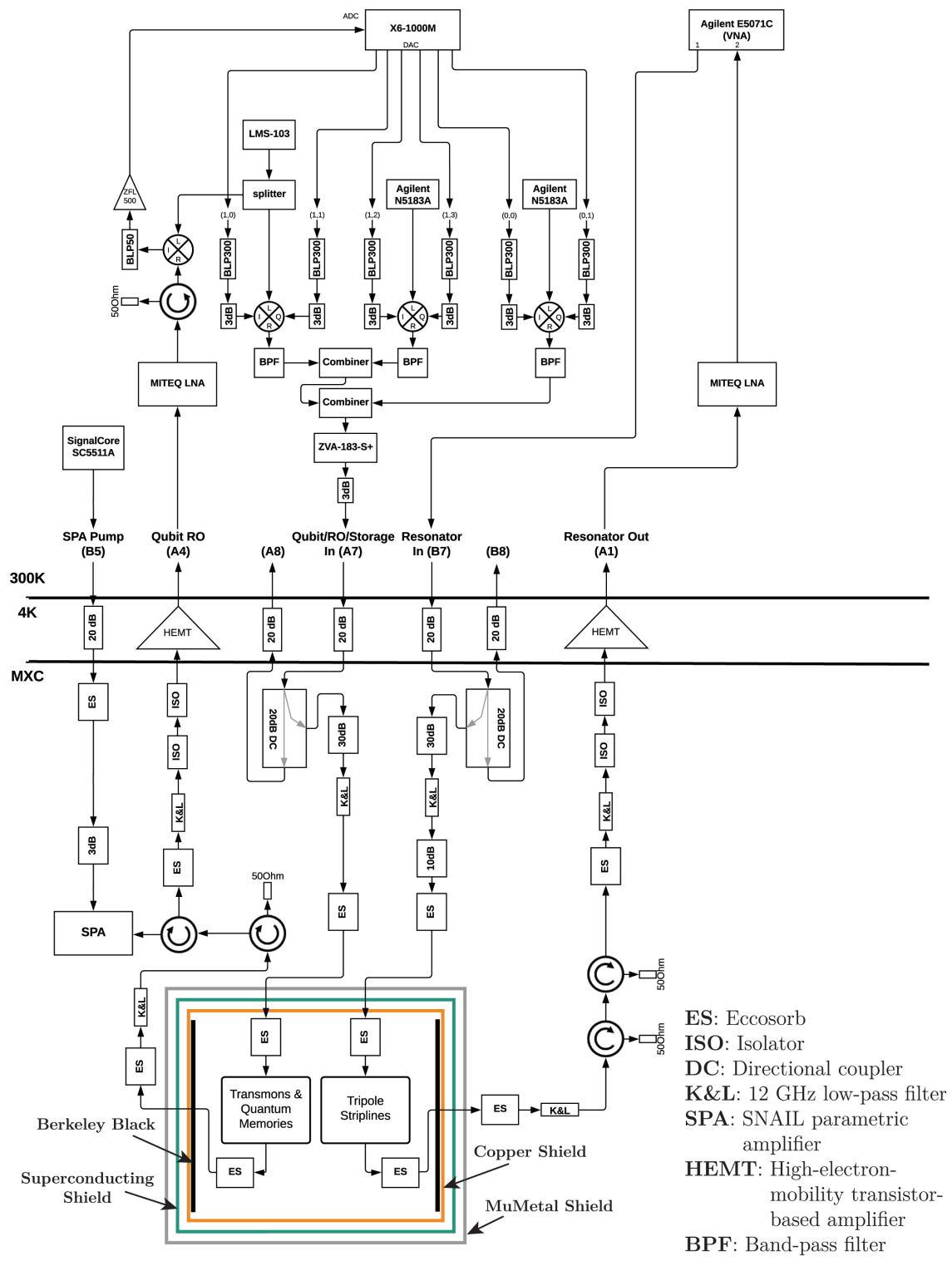


Fig. S7: Measurement setup

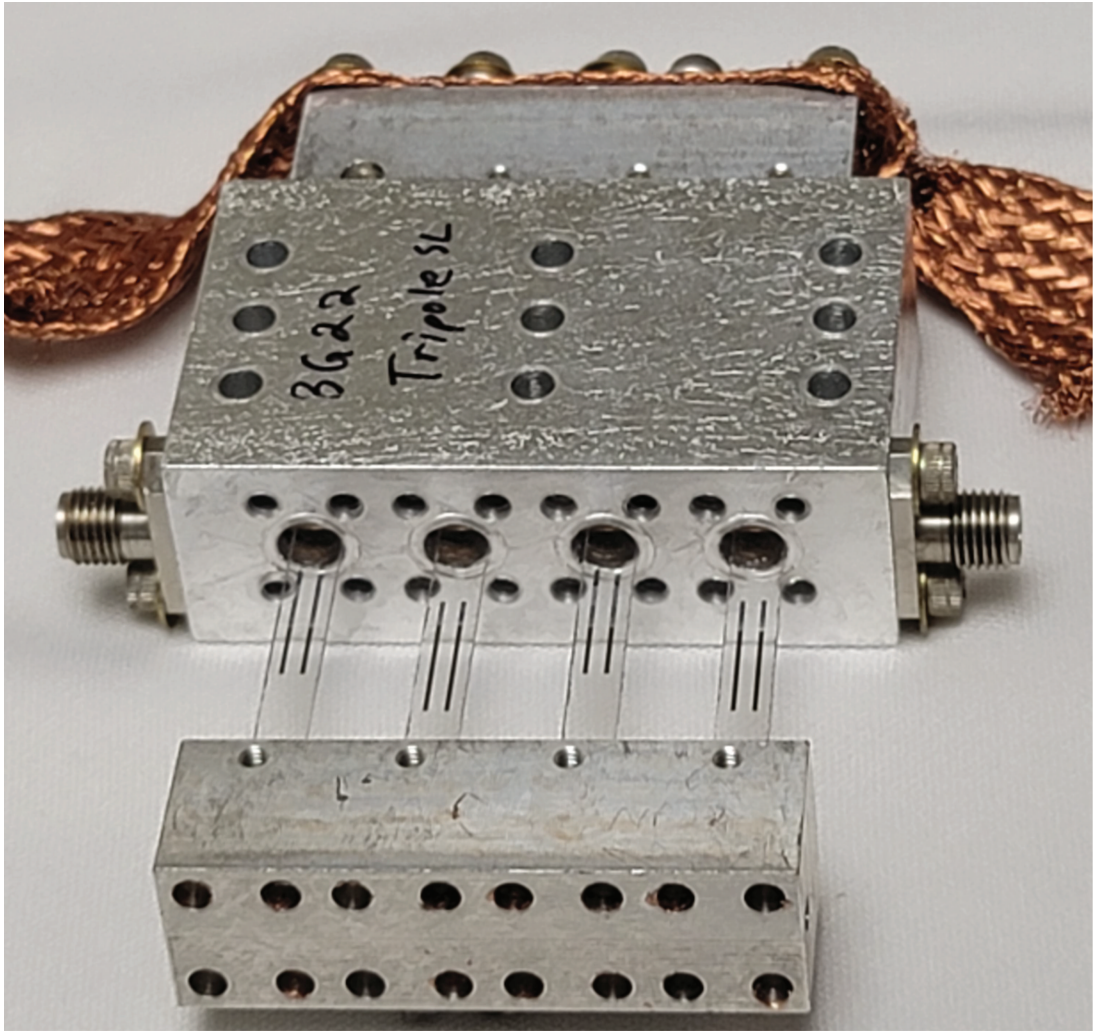


Fig. S8: Device packaging

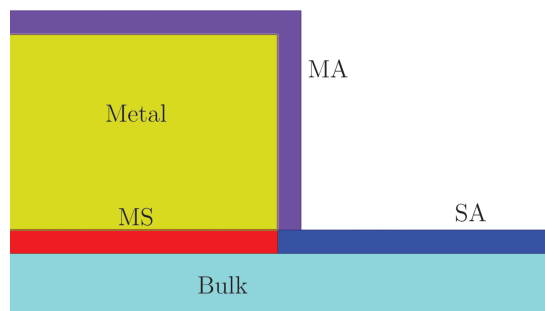


Fig. S9: Surface dielectric interfaces

**Table S1:** Summary of loss characterization devices

Device ID	Sapphire Growth Method	Anneal	Thin-Film Superconductor	Mode $Q_{\text{int}}(\bar{n} = 1)$ ( $\times 10^6$ )			Loss Factors ( $\bar{n} = 1$ )		
				D1	D2	C	$\Gamma_{\text{surf}}$ ( $\times 10^{-4}$ )	$\Gamma_{\text{bulk}}$ ( $\times 10^{-8}$ )	$1/g_{\text{seam}}$ ( $\Omega\text{m}$ ) ( $\times 10^{-3}$ )
AM22 TSL1	HEM	No	Al	0.55	8.95	9.00	$14.5 \pm 1.0$	$7.96 \pm 0.5$	$13.4 \pm 0.9$
AM22 TSL2				0.41	8.79	4.30	$19.7 \pm 3.2$	$5.65 \pm 1.5$	$51.6 \pm 2.4$
AM22 TSL3				0.39	7.13	6.41	$20.8 \pm 6.3$	$8.78 \pm 3.0$	$22.4 \pm 1.2$
DZ22 TSL3	HEM	Yes	Al	0.56	9.25	3.13	$14.2 \pm 3.5$	$7.42 \pm 1.7$	$81.2 \pm 1.2$
DZ22 TSL4				0.82	5.41	6.81	$8.58 \pm 0.5$	$20.8 \pm 0.4$	$13.1 \pm 1.1$
A23A1 TSL2	HEMEX	Yes	Al	1.53	10.33	12.21	$6.83 \pm 0.4$	$4.14 \pm 1.8$	$1.95 \pm 1.6$
A23A1 TSL3				0.88	7.56	10.17	$12.3 \pm 0.5$	$2.57 \pm 0.7$	$1.56 \pm 0.3$
EF21 TSL1	EFG	No	Ta	2.41	6.24	10.21	$1.96 \pm 0.2$	$19.8 \pm 0.7$	$0.87 \pm 0.7$
EF21 TSL2				1.91	6.22	9.93	$2.89 \pm 0.1$	$19.5 \pm 1.2$	$1.26 \pm 1.2$
EF21 TSL3				1.58	3.66	0.81	$2.72 \pm 0.2$	$33.7 \pm 0.2$	$186 \pm 1.0$
EF21 TSL4				1.63	3.65	0.48	$2.54 \pm 0.3$	$33.4 \pm 2.0$	$331 \pm 1.5$
EC21 ASL1	EFG	Yes	Ta	$2.67^1$	-	5.04	$1.96 \pm 0.2$	$19.8 \pm 0.7$	-
EC21 ASL2				$3.41^1$	-	10.99	$2.89 \pm 0.1$	$19.5 \pm 1.2$	-
EC21 ASL3				$3.19^1$	-	10.16	$2.72 \pm 0.2$	$33.7 \pm 0.2$	-
EC21 ASL4				$2.14^1$	-	15.38	$2.54 \pm 0.3$	$33.4 \pm 2.0$	-
R22 TSL1	HEM	Yes	Ta	1.59	24.04	24.20	$5.00 \pm 1.0$	$2.78 \pm 0.5$	$1.08 \pm 0.4$
R22 TSL3				2.04	26.01	7.06	$3.87 \pm 0.2$	$2.83 \pm 0.2$	$11.9 \pm 1.4$
R22 TSL4				1.78	19.66	13.69	$4.35 \pm 0.6$	$4.23 \pm 0.3$	$3.92 \pm 0.3$
BF22 TSL1	HEMEX	Yes	Ta	1.24	19.57	14.25	$6.44 \pm 0.7$	$3.30 \pm 0.5$	$2.38 \pm 0.2$
BF22 TSL2				2.53	27.40	14.96	$3.07 \pm 0.1$	$2.95 \pm 0.1$	$2.81 \pm 0.2$
BF22 TSL3				2.38	41.23	10.80	$3.40 \pm 0.3$	$1.25 \pm 0.2$	$5.18 \pm 0.2$
BF22 TSL4				2.44	29.47	16.54	$3.22 \pm 0.1$	$2.56 \pm 0.2$	$2.45 \pm 0.3$

<sup>1</sup>ASL differential (D) mode Q

TSL: Tripole stripline

ASL: Adjacent stripline

**Table S2:** Summary of seam loss

Device ID	$1/g_{\text{seam}}$ ( $\Omega\text{m}$ ) ( $\times 10^{-3}$ )	Median Relative Deviation (MRD)
AM22 TSL1	$13.4 \pm 0.9$	1.9
AM22 TSL2	$51.6 \pm 2.4$	10.3
AM22 TSL3	$22.4 \pm 1.2$	3.9
DZ22 TSL3	$81.2 \pm 1.2$	16.8
DZ22 TSL4	$13.1 \pm 1.1$	1.9
A23A1 TSL2	$1.95 \pm 1.6$	0.6
A23A1 TSL3	$1.56 \pm 0.3$	0.7
EF21 TSL1	$0.87 \pm 0.7$	0.8
EF21 TSL2	$1.26 \pm 1.2$	0.7
EF21 TSL3	$186 \pm 1.0$	39.9
EF21 TSL4	$331 \pm 1.5$	71.8
R22 TSL1	$1.08 \pm 0.4$	0.8
R22 TSL3	$11.9 \pm 1.4$	1.6
R22 TSL4	$3.92 \pm 0.3$	0.1
BF22 TSL1	$2.38 \pm 0.2$	0.5
BF22 TSL2	$2.81 \pm 0.2$	0.4
BF22 TSL3	$5.18 \pm 0.2$	0.1
BF22 TSL4	$2.45 \pm 0.3$	0.5
Average (excl. MRD > 3)	$4.75 \pm 4.5$	-

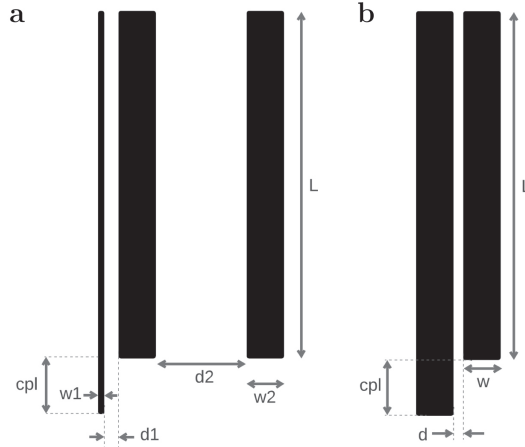
Note:  $\text{MRD} = |X_i - \tilde{X}|/\tilde{X}$ , where  $\tilde{X}$  is the median.

**Table S3:** Average surface & bulk loss factors

Material/Process System	$\Gamma_{\text{surf}}$ ( $\times 10^{-4}$ )
Al (Unannealed)	$18.3 \pm 2.7$
Al (Annealed)	$10.5 \pm 2.9$
Ta (Unannealed)	$2.53 \pm 0.4$
Ta (Annealed)	$4.15 \pm 1.4$
Material/Process System	$\Gamma_{\text{bulk}}$ ( $\times 10^{-8}$ )
EFG (Unannealed)	$26.6 \pm 6.9$
EFG (Annealed)	$3.64 \pm 2.5$
HEM (Unannealed)	$7.46 \pm 1.3$
HEM (Annealed)	$4.31 \pm 1.9$
HEMEX (Annealed)	$2.80 \pm 0.9$

Note: Outliers: MRD > 3.

# 1 Design of loss characterization devices



**Fig. S10: Multimode stripline device design.** **a** Tripole stripline (TSL). **b** Adjacent stripline (ASL).

Several different device designs were used in this work (Fig. S10). Tripole striplines (TSL) had 3 different designs that differed in stripline length ( $L$ ), wide conductor width ( $w_2$ ), and narrow conductor spacing ( $d_1$ ). The dimensions of the TSL are given in Table S4. TSLv1 and TSLv2 are nearly identical in design except for the stripline length. TSLv2 is shorter to increase the mode frequency, allowing us to benefit from slightly higher gain from our HEMT amplifier at those frequencies. TSLv3, on the other hand, had different values of  $d_1$  and  $w_2$ , in order to increase consistency and reproducibility during patterning with electron-beam lithography. In all TSL devices, the narrow stripline was longer than the wide striplines by length  $cpl$ ; this introduces a field perturbation in the D1 mode that increases its coupling to the drive line, allowing the mode to be excited during measurement.

Devices EC21 ASL1-4 were measured using a different device called the adjacent stripline (ASL). Rather than having three striplines placed next to each other, the ASL contains two striplines that are spaced by a distance  $d = 10 \mu\text{m}$  apart. This device manifests two modes; a surface-sensitive differential (D) mode and a bulk and package-sensitive common (C) mode. While this device cannot distinguish between package seam loss and bulk dielectric loss, we measured these devices in the same tunnel package as another set of devices, BF22, from which we had extracted the device-specific seam losses. By using these extracted seam losses, we subtracted the seam loss contribution to the modes of the ASL and contracted the participation matrix to solve for the surface and bulk loss factors. The dimensions of the ASL are given in Table S4. It should be noted that the ASL was meandered in order to confine the eigenfield to slightly mitigate seam loss.

Participation matrices are given for the different types of tripole striplines and the adjacent stripline in Table S5. Because the substrate thickness differed for some devices, the participations slightly changed even among identical device designs.

The participation matrix of a loss characterization device determines its measurement sensitivity; that is, the lowest loss factor that can be resolved with a fractional error  $\sigma_i/\Gamma_i < 1$ [33]. This sensitivity is dependent on the loss factors themselves. We calculate the measurement sensitivity (Fig. S11) for TSLv1, TSLv3, and ASLv1 by fixing the package losses using previously measured package conductor and MA loss factors[33] and the average measured package  $g_{\text{seam}}$  excluding outliers (Fig. S2). We note that the sensitivities for TSLv1 and TSLv2 do not differ significantly, due to their participations being very similar to each other. From the sensitivity plots, we see that both the TSLv1 and TSLv3 designs can resolve bulk loss factors as low as  $2 \times 10^{-9}$  and surface loss factors as low as  $1 \times 10^{-6}$ . However, the inability

**Table S4:** Tripole & adjacent stripline dimensions

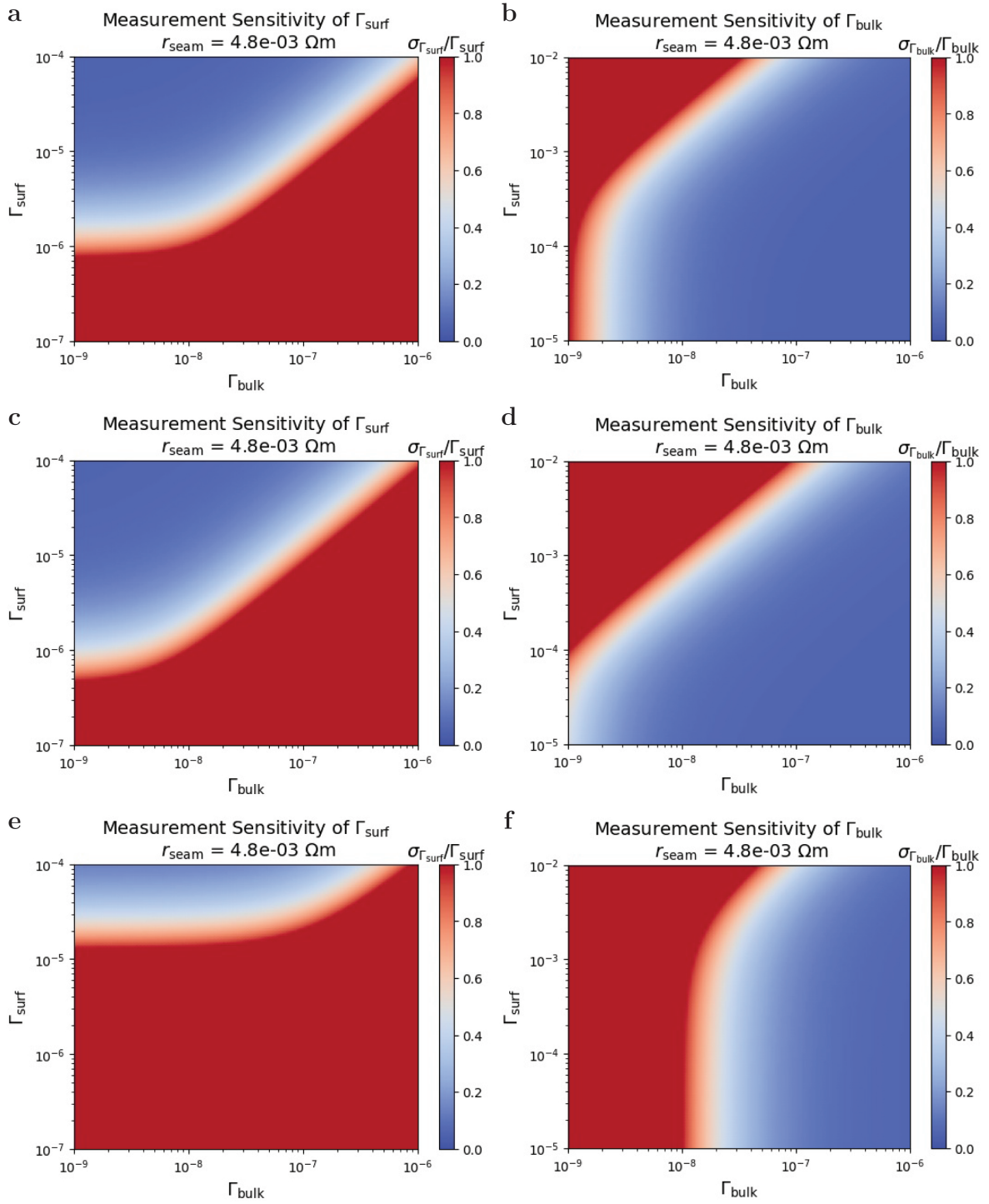
Dimension	Device Type			
	TSLv1	TSLv2	TSLv3	ASLv1
$cpl$	0.5 mm	0.5 mm	0.5 mm	0.3 mm
$L$	14 mm	12 mm	12 mm	14 mm <sup>1</sup>
$w1$	10 $\mu\text{m}$	10 $\mu\text{m}$	10 $\mu\text{m}$	-
$w2$	400 $\mu\text{m}$	400 $\mu\text{m}$	100 $\mu\text{m}$	-
$d1$	10 $\mu\text{m}$	10 $\mu\text{m}$	20 $\mu\text{m}$	-
$d2$	1200 $\mu\text{m}$	1200 $\mu\text{m}$	1200 $\mu\text{m}$	-
$d$	-	-	-	10 $\mu\text{m}$
$w$	-	-	-	150 $\mu\text{m}$

<sup>1</sup>This stripline was meandered.

**Table S5:** TSL and ASL participation matrices

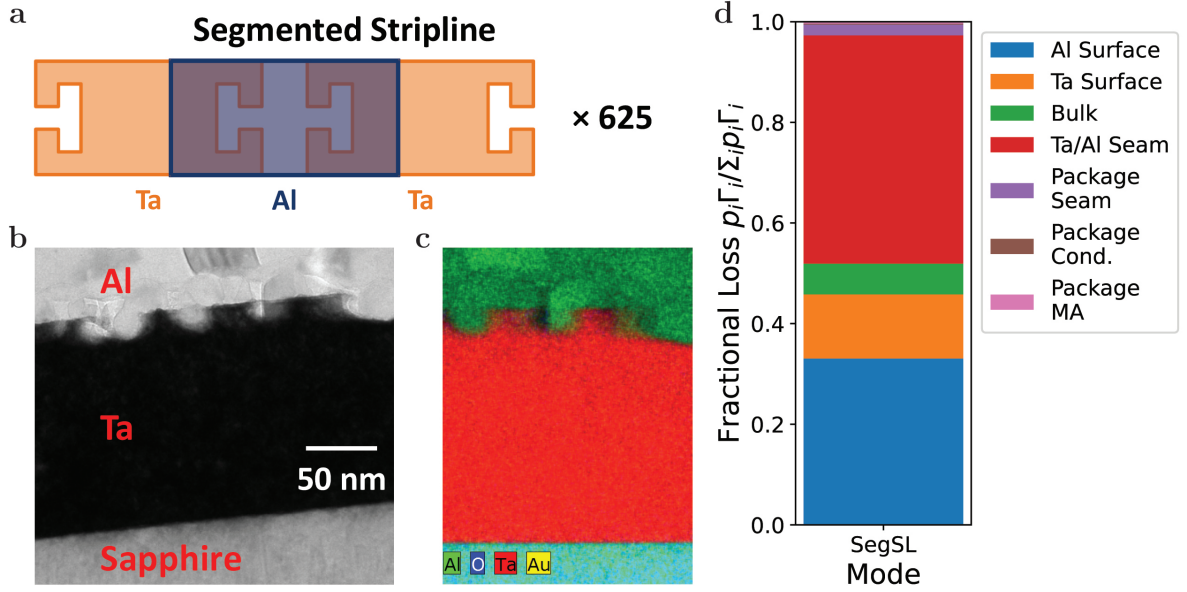
Device	Type	Substrate Thickness ( $\mu\text{m}$ )	Mode	Freq (GHz)	Participation Matrix				
					$p_{\text{surf}}$	$p_{\text{bulk}}$	$p_{\text{pkg}_{\text{cond}}}$	$p_{\text{pkg}_{\text{MA}}}$	$y_{\text{seam}} (\Omega\text{m})^{-1}$
AM22, DZ22	TSLv1	430	D1	4.52	$1.2 \times 10^{-3}$	0.90	$4.3 \times 10^{-8}$	$5.9 \times 10^{-10}$	$7.9 \times 10^{-9}$
			D2	5.34	$3.5 \times 10^{-5}$	0.73	$3.4 \times 10^{-6}$	$2.9 \times 10^{-8}$	$1.3 \times 10^{-8}$
			C	6.64	$2.2 \times 10^{-5}$	0.38	$1.2 \times 10^{-5}$	$1.2 \times 10^{-7}$	$3.1 \times 10^{-6}$
A23A1	TSLv3	650	D1	5.24	$9.0 \times 10^{-4}$	0.90	$1.5 \times 10^{-7}$	$1.6 \times 10^{-9}$	$8.6 \times 10^{-6}$
			D2	5.51	$8.9 \times 10^{-5}$	0.85	$1.7 \times 10^{-6}$	$1.4 \times 10^{-8}$	$3.4 \times 10^{-8}$
			C	6.68	$5.2 \times 10^{-5}$	0.54	$8.7 \times 10^{-6}$	$9.7 \times 10^{-8}$	$9.3 \times 10^{-6}$
EF21	TSLv1	530	D1	4.52	$1.2 \times 10^{-3}$	0.90	$4.9 \times 10^{-8}$	$6.1 \times 10^{-10}$	$1.5 \times 10^{-8}$
			D2	5.14	$3.4 \times 10^{-5}$	0.76	$3.6 \times 10^{-6}$	$2.9 \times 10^{-8}$	$2.0 \times 10^{-8}$
			C	6.45	$2.1 \times 10^{-5}$	0.41	$1.2 \times 10^{-5}$	$1.2 \times 10^{-7}$	$5.8 \times 10^{-6}$
EC21	ASLv1	530	D	3.68	$6.7 \times 10^{-4}$	0.90	$6.5 \times 10^{-8}$	$4.8 \times 10^{-10}$	$3.2 \times 10^{-8}$
			C	6.03	$2.8 \times 10^{-5}$	0.55	$4.4 \times 10^{-6}$	$7.5 \times 10^{-8}$	$1.3 \times 10^{-5}$
R22	TSLv1	650	D1	4.52	$1.2 \times 10^{-3}$	0.90	$6.0 \times 10^{-8}$	$6.5 \times 10^{-10}$	$2.5 \times 10^{-8}$
			D2	4.97	$3.4 \times 10^{-5}$	0.79	$3.9 \times 10^{-6}$	$2.9 \times 10^{-8}$	$2.6 \times 10^{-8}$
			C	6.27	$2.1 \times 10^{-5}$	0.43	$1.8 \times 10^{-5}$	$1.3 \times 10^{-7}$	$9.4 \times 10^{-6}$
BF22	TSLv2	650	D1	5.25	$1.2 \times 10^{-3}$	0.90	$8.1 \times 10^{-8}$	$8.5 \times 10^{-10}$	$4.0 \times 10^{-8}$
			D2	5.74	$3.5 \times 10^{-5}$	0.80	$3.8 \times 10^{-6}$	$2.7 \times 10^{-8}$	$3.6 \times 10^{-8}$
			C	7.13	$2.2 \times 10^{-5}$	0.45	$1.3 \times 10^{-5}$	$1.3 \times 10^{-7}$	$1.4 \times 10^{-5}$

for the ASL to distinguish between bulk and seam loss reduces its sensitivity by an order of magnitude. Regardless, these sensitivities are still below what we actually resolve in measured devices, indicating that these loss characterization devices are well conditioned to probe losses in our regime of interest.



**Fig. S11: Surface and bulk loss measurement sensitivity for TSLv1 (a, b), TSLv3 (c, d), and ASLv1 (e, f). Here,  $r_{\text{seam}} = 1/g_{\text{seam}}$ .**

## 2 Extracting Ta/Al contact loss



**Fig. S12: Segmented stripline to measure Ta/Al contact loss.** **a** Segmented stripline design; alternating segmented of Ta and Al are repeated to make the full length of the stripline. **b** TEM of Ta/Al interface, showing damage to the Ta film caused by the ion beam cleaning prior to Al deposition. **c** Elemental map formed by energy-dispersive X-ray spectroscopy of the TEM sample shows no oxide in the Ta/Al interface, indicating that good metal-to-metal contact is present. **d** Single-photon loss budget for a representative segmented stripline, showing that 50% of the device’s total loss is due to Ta/Al contact loss.

Contact loss due to the interface between the Ta and Al films may contribute significantly to the total loss of a transmon qubit. Tantalum oxide or other contaminants located in the tantalum/aluminum interface may contribute to an effective resistance in series with the transmon’s Josephson junction. To alleviate this loss, we employ an argon ion beam cleaning step prior to aluminum deposition to remove the tantalum oxide and any potential contamination. TEM of the resulting Ta/Al interface reveals no oxide between Al and Ta, showing that ion beam clean effectively removes the tantalum oxide (Fig. S12b, c). However, the ion beam also appears to have damaged and roughened the Ta film, possibly introducing lossy defects. To characterize and quantify this loss channel, we designed the segmented stripline, a thin-film resonator made up of around 625 alternating segments of Ta and Al that contact each other, resulting in a 10  $\mu\text{m}$  wide, 12.5 mm long stripline that is over 100 times more sensitive to contact loss than a regular transmon (Fig. S12a)[50]. We model the participations of this device by simulating the surface, bulk, and package participations using an electromagnetic simulation with the same method as was used for the tripole striplines. We use a seam loss model to describe the Ta/Al contact loss[55] similar to the package seam loss, where the geometric component of the loss is described by a seam admittance per unit length,  $y_{\text{seamTa/Al}}$ , and an intrinsic loss factor described as the inverse of the seam conductance per unit length,  $1/g_{\text{seamTa/Al}}$ . Since the current mostly flows along the propagation axis of the stripline, we can assume a seam length equal to the stripline width  $w = 10 \mu\text{m}$ , and convert the seam conductance into a contact resistance,  $R_{\text{Ta/Al}} = (w \cdot g_{\text{seamTa/Al}})^{-1}$ . To calculate  $y_{\text{seamTa/Al}}$ , we use an analytical model that assumes a sinusoidal current distribution throughout the stripline:



$$y_{\text{seam}_{\text{Ta}/\text{Al}}} = \frac{2 \sum_i \sin^2(\pi z_i/l)}{\pi w Z_0}, \quad (12)$$

where  $l$  is the total length of the stripline,  $z_i$  is the position along the propagation axis of the  $i$ th Ta/Al contact,  $w$  is the width of the stripline, and  $Z_0$  is the characteristic impedance of the stripline mode. The participations of the segmented stripline are given in Table S6. The device was fabricated on annealed HEM sapphire using the same process as the tantalum-based transmon. Since we have already characterized the losses associated with the Al surface, Ta surface, bulk substrate, and package, we only need the quality factor measurement of the segmented stripline to extract the Ta/Al contact loss. Two nominally identical segmented striplines were measured at single photon powers. Their quality factors were very similar, differing by less than 10% (Table S7). Loss extraction for both devices revealed that around 50% of the total loss from the segmented stripline was due to the Ta/Al contact loss, which verifies that our loss characterization device is sensitive to the loss channel of interest (Fig. S12d). We therefore calculated an average seam resistance of  $260 \pm 47$  n $\Omega$ .

**Table S6:** Segmented stripline participations

Freq (GHz)	5.74
$p_{\text{surfTa}}$	$1.6 \times 10^{-4}$
$p_{\text{surfAl}}$	$1.6 \times 10^{-4}$
$p_{\text{bulk}}$	0.72
$y_{\text{seam}_{\text{Ta}/\text{Al}}}$	$9.4 \times 10^4$
$p_{\text{pkgcond}}$	$3.3 \times 10^{-6}$
$p_{\text{pkgMA}}$	$4.8 \times 10^{-8}$
$y_{\text{pkgseam}}$	$2.3 \times 10^{-6}$

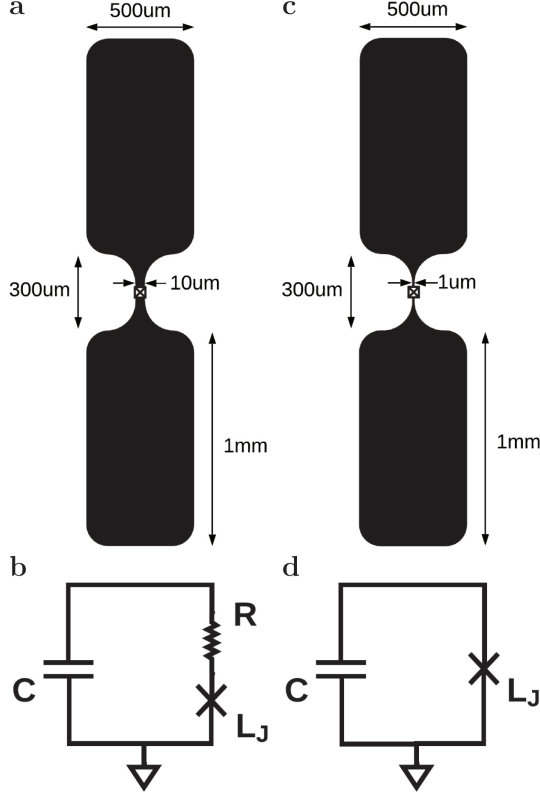
**Table S7:** Segmented stripline loss

Device	$Q_{\text{int}}(\bar{n} = 1)$	$R_{\text{Ta}/\text{Al}}$ (n $\Omega$ )
SegSL1	$1.97 \times 10^6$	$246 \pm 59$
SegSL2	$1.88 \times 10^6$	$272 \pm 73$

As with the segmented stripline, the Ta/Al contact participation of the tantalum-based transmon  $y_{\text{seam}_{\text{Ta}/\text{Al}}}$  was calculated analytically. We assume a lumped-element model, where the contact resistance is modeled as a capacitor shunting a resistor in series with the junction. The quality factor due to the contact loss can then be straightforwardly written down as  $Q_{\text{Ta}/\text{Al}} = Z_0/2R_{\text{Ta}/\text{Al}}$ , where  $Z_0 = \sqrt{L_J/C}$  is the characteristic impedance of the transmon mode and the factor of 2 in the denominator accounts for the presence of two Ta/Al contacts in the tantalum-based transmon. Therefore, the Ta/Al seam admittance becomes  $y_{\text{seam}_{\text{Ta}/\text{Al}}} = 2/Z_0 w$ , where  $w = 10$   $\mu\text{m}$  is the length of the seam. For our transmon design, this gives  $y_{\text{seam}_{\text{Ta}/\text{Al}}} = 737.86$  ( $\Omega\text{m}$ ) $^{-1}$ . Using the extracted value of  $R_{\text{Ta}/\text{Al}} = 260 \pm 47$  n $\Omega$  from the segmented striplines, we estimate the loss due to Ta/Al contact resistance to limit the transmon  $Q_{\text{int}}$  to a maximum of approximately  $5 \times 10^8$ .

### 3 Transmon qubit device design

We utilized a standard 3D transmon design used for cavity-based cQED experiments[39] (Fig. S13). Each device contains a transmon coupled to a stripline readout resonator patterned on a chip that is



**Fig. S13: Transmon qubit device design.** **a, b** Ta-based transmon and circuit model. The resistor is used to represent Ta/Al contact loss. **c, d** Al-based transmon and circuit model. Junction leads are thinner to improve reliability and reproducibility of the electron-beam lithography process.

inserted into a cylindrical tunnel. The chip also contains a bandpass Purcell filter to obtain a large external coupling of the drive line to the readout resonator while preserving a low external coupling of the drive line to the transmon (Table S9). Tantalum- and aluminum-based transmons had very similar designs, with some exceptions. In the tantalum-based transmon, the tantalum leads from the capacitor to the Josephson junction are wider at 10  $\mu\text{m}$  (Fig. S13a). For the aluminum-based transmon, patterning such a large feature so close to the junction using only electron-beam lithography resulted in significant proximity dosing which severely impacted device yield. As a result, Al-based transmons were patterned with 1  $\mu\text{m}$ -wide junction leads (Fig. S13b). This design change resulted in slightly higher overall surface participation for the aluminum-based transmon (Table S8).

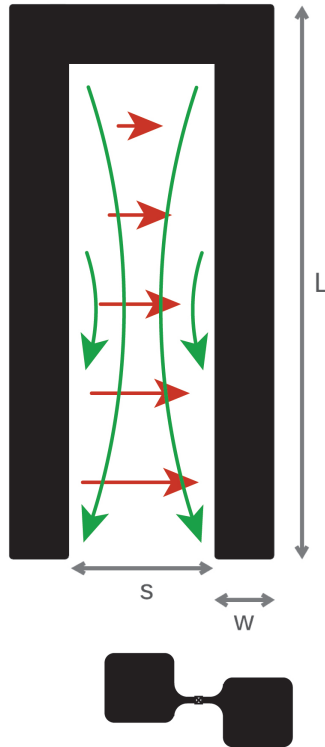
**Table S8: Transmon participations**

	Al transmon	Ta transmon
$p_{\text{surfTa}}$	-	$8.1 \times 10^{-5}$
$p_{\text{surfAl}}$	$1.5 \times 10^{-4}$	$5.5 \times 10^{-5}$
$p_{\text{bulk}}$	0.84	0.84
$y_{\text{seamTa/Al}}$	-	$7.2 \times 10^2$
$p_{\text{pkgcond}}$	$9.3 \times 10^{-8}$	$9.3 \times 10^{-8}$
$p_{\text{pkgMA}}$	$5.1 \times 10^{-9}$	$5.1 \times 10^{-9}$
$y_{\text{pkgseam}}$	$3.0 \times 10^{-9}$	$3.0 \times 10^{-9}$

**Table S9:** Typical transmon & readout parameters

$\omega_t/2\pi$ (GHz)	4.5-5.1
$\omega_r/2\pi$ (GHz)	9.0-9.3
$\chi_{tt}/2\pi$ (MHz)	170-180
$\chi_{tr}/2\pi$ (MHz)	0.5-1.1
$\kappa_r/2\pi$ (MHz)	0.5-1.1

## 4 Hairpin stripline device design



**Fig. S14: Hairpin stripline device design.** Field behaviors of the memory mode (red arrows) and readout mode (green arrows) are shown. The ancilla transmon has a staggered capacitor pad design that allows coupling to both modes.

The hairpin stripline (Fig. S14) is optimized to minimize package loss in order to maximize quantum memory coherence. The insensitivity to package loss is obtained by folding a half-wave (length  $2L$ ) stripline resonator into itself, resulting in a fundamental mode that is in some sense “differential”; electric field lines do not terminate at the walls of the package. Additionally, the current antinode is positioned at a location where the current flow is perpendicular to the axis of the cylindrical tunnel, resulting in minimal induced current along the walls of the package. Maximizing the coherence of the hairpin stripline requires using the materials and processes that yield the lowest intrinsic loss as well as minimizing surface and

bulk dielectric participation by optimizing design. The latter is accomplished by optimizing the width of the stripline  $w$ , the spacing between the arms of the hairpin  $s$ , and the tunnel radius. Increasing  $w$  slightly reduces surface participation, while increasing  $s$  reduces both bulk and surface participation. However, increasing either while keeping the tunnel radius fixed results in increasing the package participation. For the losses extracted in this work and a fixed tunnel radius of 2.5 mm, we found the optimal values for  $w$  and  $s$  to be 800  $\mu\text{m}$  and 1200  $\mu\text{m}$ , respectively (Table S10).

The ancilla transmon couples to both the memory mode and the higher order stripline mode that acts as a readout mode. The readout mode is equivalent to a full-wave resonance mode, where electric field antinodes are located at both ends of the hairpin. The opposite polarities at either end of the hairpin gives rise to an electric field pattern that is orthogonal to that of the memory mode. To couple to both modes simultaneously, we stagger the transmon’s capacitor pads, giving it a net dipole moment that points diagonally with respect to the memory and readout mode’s fields.

**Table S10:** Hairpin stripline participations

$p_{\text{surf}}$	$2.4 \times 10^{-5}$
$p_{\text{bulk}}$	0.72
$p_{\text{pkg}_{\text{cond}}}$	$6.7 \times 10^{-6}$
$p_{\text{pkg}_{\text{MA}}}$	$5.4 \times 10^{-8}$
$y_{\text{seam}}$	$3.9 \times 10^{-8}$

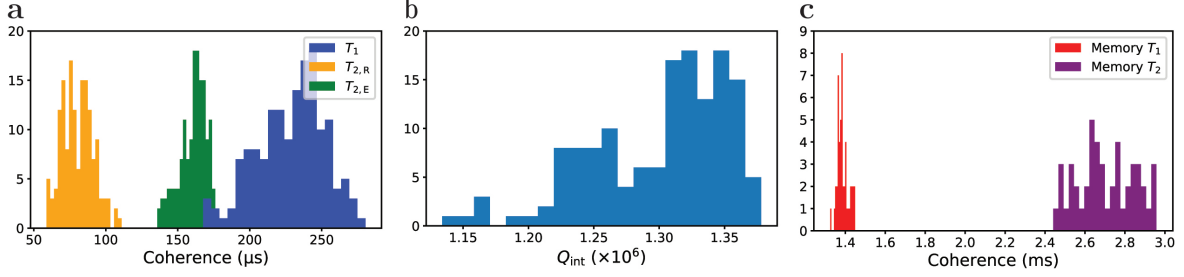
**Table S11:** Typical quantum memory parameters

$\omega_{\text{m}}/2\pi$ (GHz)	3.9-4.0
$\omega_{\text{t}}/2\pi$ (GHz)	5.7-6.8
$\omega_{\text{r}}/2\pi$ (GHz)	9.0-9.3
$\chi_{\text{tt}}/2\pi$ (MHz)	201-217
$\chi_{\text{tm}}/2\pi$ (MHz)	0.1-0.4
$\chi_{\text{tr}}/2\pi$ (MHz)	0.3-0.6
$\kappa_{\text{r}}/2\pi$ (MHz)	0.2-0.5

## 5 Temporal fluctuations of coherence in transmons, quantum memories, and resonators

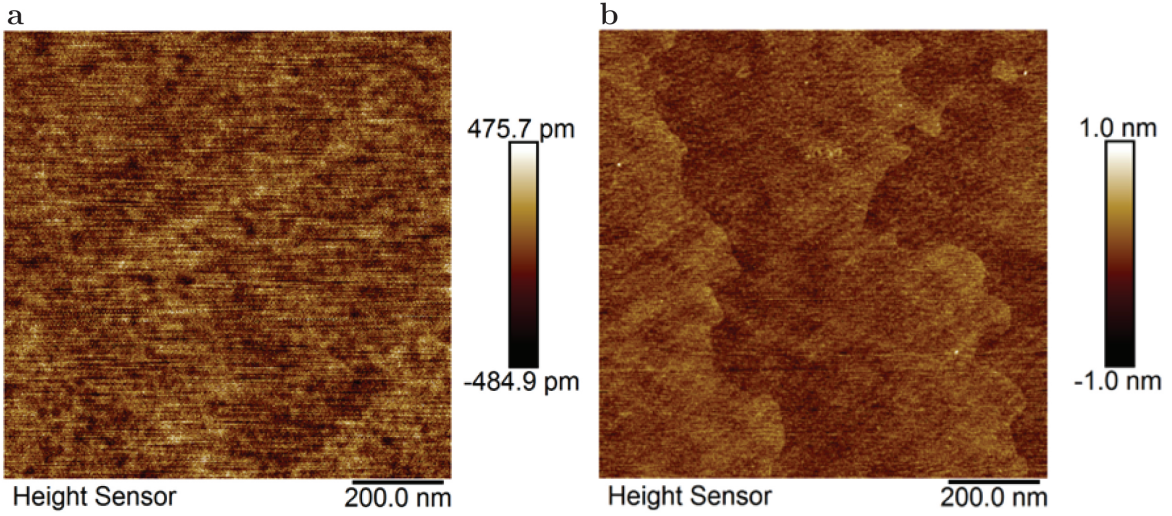
Transmon qubit coherence was measured over a period of at least 10 hours to capture temporal fluctuations (Fig. S15a), with some devices being measured over two days. We observed significant fluctuation in both  $T_1$  and  $T_2$  over long timescales, with  $T_1$  fluctuating by around  $\pm 30\%$  about the mean. In most devices,  $T_{2,\text{E}}$  was almost a factor of 2 higher than  $T_{2,\text{R}}$  but not as high as  $2T_1$ , indicating that low frequency and high frequency noise are present; we attribute the high frequency noise to thermal photon shot noise from the dispersively-coupled readout resonator[74]. We attribute temporal behavior of  $T_1$  to fluctuating TLSs near and inside the Josephson junction, where the electric field densities due to the parasitic capacitance of the junction electrodes can be very high. We compare this behavior to that of the D1 mode Q over time of an aluminum tripole stripline (Fig. S15b). The internal Q of the D1 mode, despite having higher surface participation than the transmon, fluctuates by only  $\pm 10\%$  around its average value,

due to the resonator’s much larger area and more uniformly distributed electric field. Finally, we measure quantum memory coherence over time (Fig. S15c), and see that similar to the resonator and in sharp contrast with the transmon, the quantum memory  $T_1$  and  $T_2$  are quite stable over long timescales, with  $T_1$  and  $T_1$ -limited  $T_2$  fluctuating by only around 10% of their average values. From this, we can estimate that the majority of the transmon’s  $T_1$  fluctuation comes from interactions with TLSs inside or within 100 nm from the junction, while around 10% of the fluctuations can be attributed to TLSs interacting with the rest of the transmon circuit, behavior that is already captured by resonator measurements.



**Fig. S15: Temporal fluctuations in coherence.** **a** Representative histogram of temporal fluctuations for coherence in a Ta-based transmon device measured over 48 hours. **b** Representative histogram of temporal fluctuations for  $Q_{\text{int}}$  in a tripole stripline D1 mode measured over 35 hours. Data was taken at low power in the TLS-dominated regime, with  $\bar{n} \sim 100$ . **c** Representative histogram of temporal fluctuations in coherence in a hairpin stripline quantum memory device measured over 30 hours.

## 6 Sapphire annealing

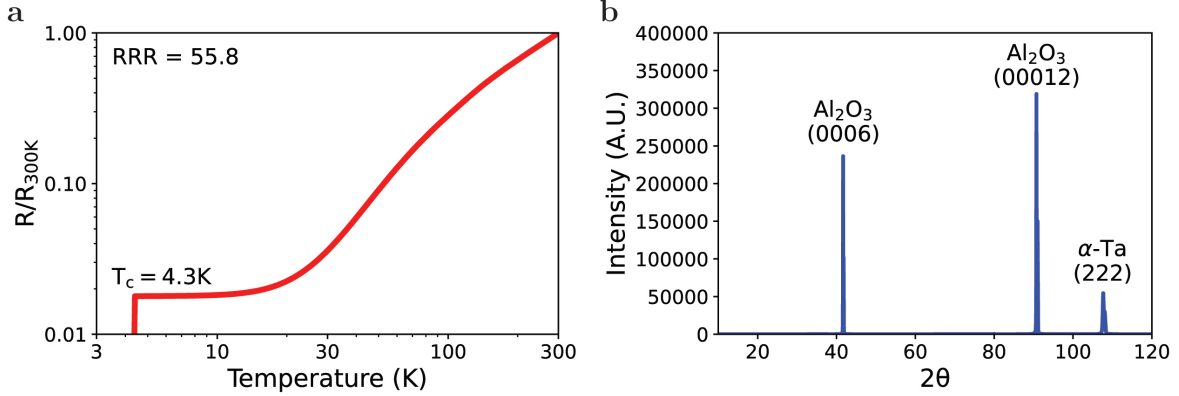


**Fig. S16: Atomic force microscopy of sapphire surface before annealing (a) and after annealing (b).**

Atomic Force Microscopy (AFM) on sapphire substrates was conducted before and after annealing using a Bruker Dimension Fastscan AFM. Surfaces before annealing had sub-nanometer roughness and uniform

surface topology with no distinguishable features (Fig. S16a). After annealing, surfaces were atomically flat and displayed a terraced structure typically seen for annealed c-plane sapphire (Fig. S16b)[38, 52]. The terraces have step height of around 220 pm, approximately equal to the inter-atomic spacing in the c-axis ( $c/6 = 216$  pm), and width 420 nm related to the miscut angle of the wafer, which in this case is calculated to be approximately  $0.03^\circ$ .

## 7 Tantalum film characterization

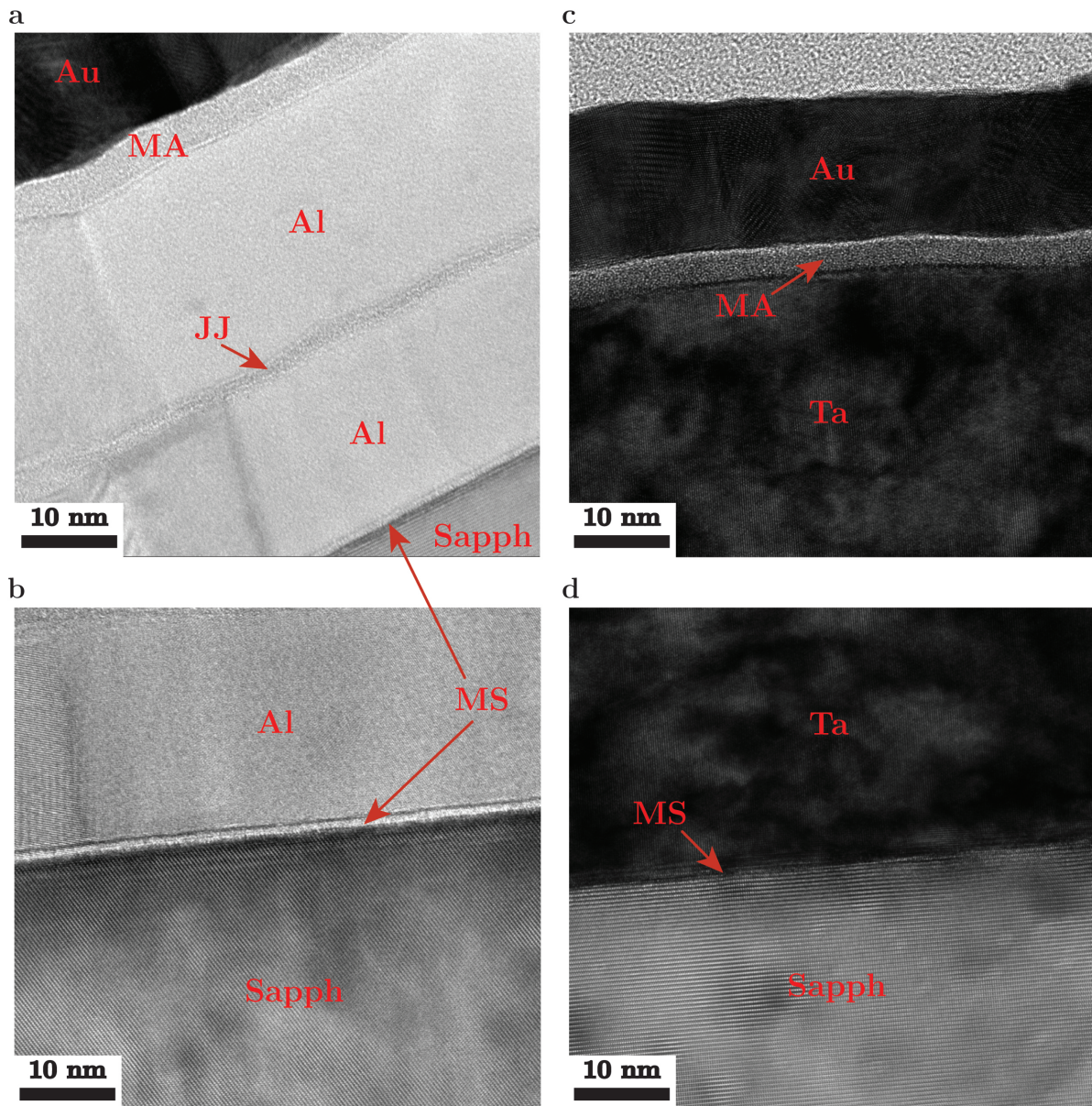


**Fig. S17: Tantalum film characterization** (a) DC resistance as a function of temperature. The sharp drop in resistance at  $T = 4.3$  K indicates the emergence of the superconducting state. (b) X-ray diffractometry (XRD) of a Ta film. This particular film was entirely in the (111) orientation; other films were either entirely (110) or a mixture of (111) and (110) (not shown).

Tantalum films sputtered at  $800^\circ\text{C}$  were consistently in the  $\alpha$  phase. Resistance as a function of temperature was measured using a Quantum Design PPMS DynaCool for multiple samples, all of which had  $T_c > 4.17$  K and  $\text{RRR} > 15$ , with our best sample having  $T_c = 4.3$  K and  $\text{RRR} = 55.8$  (Fig. S17a). All films grown this way whose crystal structure was measured using a Rigaku Miniflex II XRD confirmed the dominant presence of  $\alpha$ -Ta growing in either the (111) or (110) orientation, while the  $\beta$  phase was not observed (Fig. S17b).

## 8 TEM film characterization

TEM of Al/ $\text{AlO}_x$ /Al junctions and Ta thin films was performed using an FEI Talos F200X (Fig. S18). The measured samples had a thin layer of gold sputtered on them as part of the sample preparation process. Significant differences were observed in the metal-substrate (MS) interface of the two films. While the tantalum/sapphire interface was free of amorphous material and displayed nearly epitaxial growth, the aluminum/sapphire interface has a thin ( $\approx 2$  nm) amorphous region. Additionally, the metal-air (MA) interface of tantalum has a thin ( $\approx 3$  nm) oxide layer, while the MA interface of aluminum has an  $\approx 5$  nm oxide layer. Between the two layers of aluminum lies the Josephson junction oxide, which is approximately  $\approx 2$  nm and looks amorphous, similar to the aluminum MA and MS interfaces.



**Fig. S18: TEM of Al and Ta films.** **a, b** TEM of the MS and MA interfaces of a typical Al/ $\text{AlO}_x$ /Al film. The MA interface appears amorphous and has general stoichiometry  $\text{AlO}_x$ . The lower and upper Al layers are the two Josephson junction electrodes, and the  $\text{AlO}_x$  between them act as the tunnel barrier. **c, d** TEM of the MS and MA interfaces of a typical Ta thin film. Like the MA interface of the Al films, the MA interface of the Ta film also appears amorphous, with general stoichiometry  $\text{TaO}_x$ . Gold is sputtered on top of the sample to protect the films from the TEM sample preparation process.

## 9 Efficient frequency-domain sampling for resonator measurements in a vector network analyzer (VNA)

The default sweep type of a standard VNA is linear, i.e. it samples frequency points uniformly across the desired frequency span. In the standard circle fitting algorithm[56], a point in frequency space translates to a point in the complex plane according to the frequency-phase relation:

$$\tan\theta = 2Q_L \left( \frac{f}{f_r} - 1 \right). \quad (13)$$

Due to the nonlinearity of this relation, a set of  $S_{21}$  values uniformly parameterized by frequency is not uniformly parameterized by phase; the density of points is highest at the off-resonant point of the circle and lowest at the on-resonant point. Given a fixed number of points, we can improve fitting reliability by customizing the distribution of points in frequency space such that we realize a uniform distribution of points in phase space[75]. Experimentally, we employ the alternative segment sweep type provided by the standard VNA.

To counteract the bunching of points towards the off-resonant point, we can sample a higher density of frequency points around resonance. A simple resolution, which is adopted for many of our sweeps, is a frequency spacing that increases quadratically with each point away from the center frequency, yielding a more uniform phase distribution.

We also present the framework for a more optimal frequency distribution. Determining the  $N$  optimal frequency points is not as straightforward as taking  $\theta_n = 2\pi n/N$ , however. Firstly, traversing the full circumference of the circle requires sweeping from  $f = -\infty$  to  $f = +\infty$ , which is of course unphysical; secondly, we have little knowledge of  $Q_L$  prior to measurement and circle fitting. To this end, we devise a formula built around one parameter that can be observed qualitatively before fitting, the frequency span of measurement expressed in terms of the number of linewidths.

Suppose that we will eventually sweep a phase span  $\Delta\theta$ , which is related to the frequency span  $\Delta f$  by:

$$\tan\Delta\theta = \frac{2 \frac{Q_L \Delta f}{f_r}}{1 - \left( \frac{Q_L \Delta f}{f_r} \right)^2}. \quad (14)$$

Given that the center frequency is set to be equal to the resonant frequency. By defining the linewidth as  $\kappa = f_r/Q_L$  and introducing a parameter that we call the weight  $W = \Delta f/\kappa$ , we can rewrite Eq. (14) in terms of only the weight:

$$\tan \Delta\theta = \frac{2W}{1 - W^2}. \quad (15)$$

We note that the weight is the aforementioned frequency-span-to-linewidth ratio, and is named so because it ultimately determines the weighting of the point distribution towards the on-resonant point of the circle. We find that aiming for  $W \approx 5$  sweeps enough of the circle for a reliable fitting.

Using the frequency-phase relation to express the  $n$ th frequency point  $f_n$ :

$$f_n = f_r + \frac{f_r}{2Q_L} \tan \left( \frac{n}{N} \Delta\theta \right) = f_r + \frac{\Delta f}{2W} \tan \left( \frac{n}{N-1} \tan^{-1} \frac{2W}{1 - W^2} \right). \quad (16)$$

The points are indexed such that for an odd number of points  $N$ ,  $n$  runs from  $-(N-1)/2$  to  $(N-1)/2$ . Eq. (16) contains three parameters that can be fixed prior to fitting: the center frequency, the frequency span and the weight. Because the weight is likely to be an qualitative estimate, the frequency points  $f_n$  need to be rescaled such that the highest frequency point is  $f_{(N-1)/2} = f_r + \Delta f/2$ , and similarly for the lowest frequency point. We introduce the ratio of the desired frequency span to the erroneous span as a result of  $W$  being an estimate:

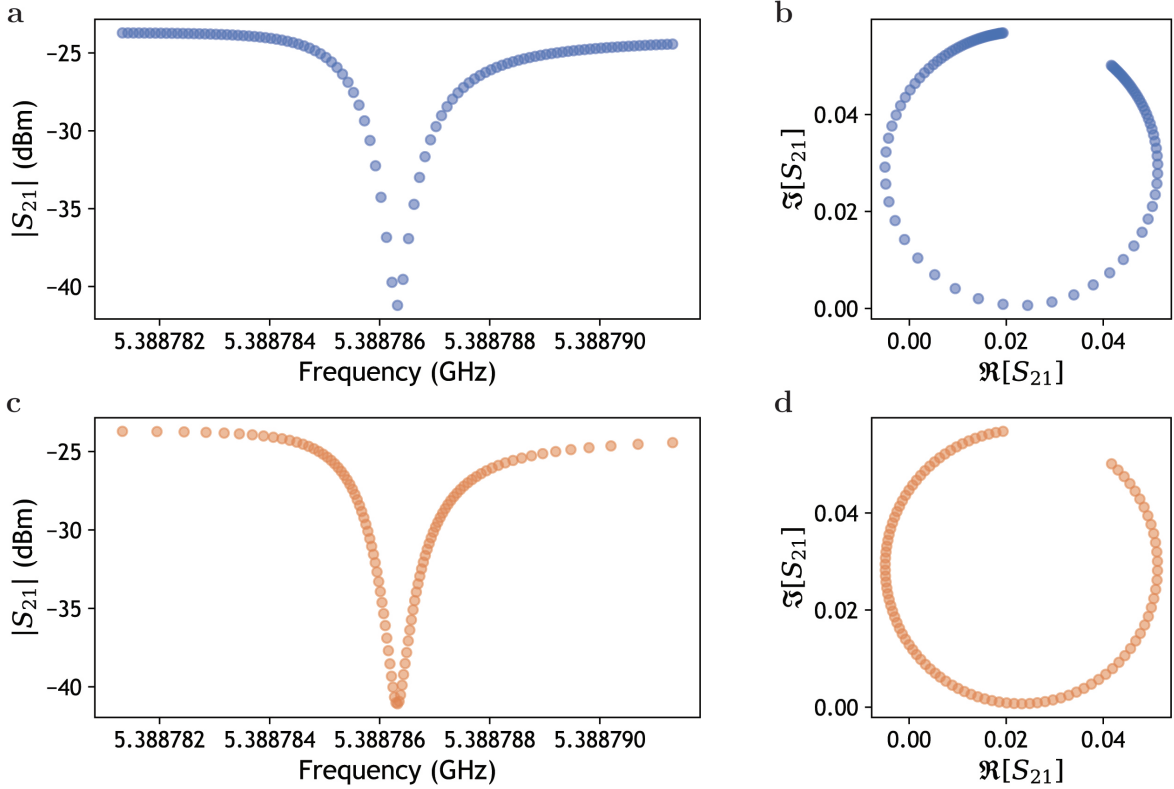


$$R = \frac{\Delta f/2}{\frac{\Delta f}{2W} \tan\left(\frac{1}{2} \tan^{-1} \frac{2W}{1-W^2}\right)} = \frac{W}{\tan\left(\frac{1}{2} \tan^{-1} \frac{2W}{1-W^2}\right)}, \quad (17)$$

which modifies Eq. (17) to become:

$$f_n = f_r + R \frac{\Delta f}{2W} \tan\left(\frac{n}{N-1} \tan^{-1} \frac{2W}{1-W^2}\right). \quad (18)$$

This rescaling offers the flexibility of adjusting the nonlinearity of the frequency distribution through  $W$ . The larger the weight, the higher the density of points at the on-resonance point; the limit of  $W \rightarrow 0$  recovers the linear sweep. We find that the number of points required for a reliable fitting can be reduced by a factor of  $\approx 5$  by using an appropriate segment sweep, thereby shortening measurement duration by the same factor.



**Fig. S19:  $S_{21}(f)$  amplitude and complex response with different sampling.** **a, b** Linear frequency sweep with  $N = 101$  points, where sampled points are equally spaced in frequency. **c, d** A  $W = 5$  segment sweep using  $N = 101$  points. The sampled points are now more densely packed about the resonance point and equally spaced in the complex plane. Data is taken at high power from the D2 mode of an aluminum tripole stripline device fabricated on annealed HEM sapphire.

## 10 Derivation of resonator average photon number

We derive the average photon number in a resonator in hanger configuration using input-output theory. For a system characterized by stationary mode operator  $\hat{a}$  coupled to an environment characterized by

propagating mode operators  $\hat{a}_{\text{in}}$  and  $\hat{a}_{\text{out}}$  with energy loss rate  $\gamma$ , the time evolution of  $\hat{a}$  is determined by the quantum Langevin equation:

$$\frac{d}{dt}\hat{a}(t) = \frac{i}{\hbar}[\hat{H}, \hat{a}(t)] - \frac{\gamma}{2}\hat{a}(t) + \sqrt{\gamma}\hat{a}_{\text{in}}(t), \quad (19)$$

where  $\hat{H}$  is the Hamiltonian of the system. The three terms on the right-hand side denote conservative time evolution, dissipation and drive respectively, and without dissipation and drive we recover the Heisenberg equation of motion. To apply input-output theory to a resonator in the hanger configuration, we introduce an internal loss channel with energy loss rate  $\gamma_{\text{int}}$ , and two coupling channels that describe coupling to right- and left-propagating waves with energy loss rates  $\gamma_+$  and  $\gamma_-$  respectively. The rates are related to their respective Q-factors by  $Q_{\text{int}} = \omega_r/\gamma_{\text{int}}$ ,  $Q_+ = \omega_r/\gamma_+$  and  $Q_- = \omega_r/\gamma_-$ .

The quantum Langevin equation has one dissipation term and one drive term for each of the three channels:

$$\frac{d}{dt}\hat{a}(t) = \frac{i}{\hbar}[\hat{H}, \hat{a}(t)] - \frac{\gamma_{\text{L}}}{2}\hat{a}(t) + \sqrt{\gamma_{\text{int}}}\hat{a}_{\text{in,int}}(t) + \sqrt{\gamma_+}\hat{a}_{\text{in,+}}(t) + \sqrt{\gamma_-}\hat{a}_{\text{in,-}}(t), \quad (20)$$

where we have combined the three dissipation terms into one by defining the total (loaded) energy loss rate  $\gamma_{\text{L}} = \gamma_{\text{int}} + \gamma_+ + \gamma_-$ . Because we have imposed the existence of three channels, there are three boundary conditions, obtained by matching the amplitudes of the mode operators between the system and each of the three channels:

$$\begin{aligned} \hat{a}_{\text{in,int}}(t) - \hat{a}_{\text{out,int}}(t) &= \sqrt{\gamma_{\text{int}}}\hat{a}(t) \\ \hat{a}_{\text{in,+}}(t) - \hat{a}_{\text{out,+}}(t) &= \sqrt{\gamma_+}\hat{a}(t) \\ \hat{a}_{\text{in,-}}(t) - \hat{a}_{\text{out,-}}(t) &= \sqrt{\gamma_-}\hat{a}(t). \end{aligned} \quad (21)$$

Since we drive only via the right-propagating coupling channel, Eq. (20) reduces to:

$$\frac{d}{dt}\hat{a}(t) = \frac{i}{\hbar}[\hat{H}, \hat{a}(t)] - \frac{\gamma_{\text{L}}}{2}\hat{a}(t) + \sqrt{\gamma_+}\hat{a}_{\text{in,+}}(t). \quad (22)$$

For a harmonic oscillator, the frequency domain expression is:

$$-i\omega\hat{a}(\omega) = -i\omega_r\hat{a}(\omega) - \frac{\gamma_{\text{L}}}{2}\hat{a}(\omega) + \sqrt{\gamma_+}\hat{a}_{\text{in,+}}(\omega) \implies \hat{a} = \frac{\sqrt{\gamma_+}}{\frac{\gamma_{\text{L}}}{2} - i(\omega - \omega_r)}\hat{a}_{\text{in,+}}, \quad (23)$$

where the explicit dependence of the operators on  $\omega$  has been omitted for clarity. From Eq. (23), the number operator is:

$$\hat{n} = \hat{a}^\dagger\hat{a} = \frac{\gamma_+}{\frac{\gamma_{\text{L}}^2}{4} + (\omega - \omega_r)^2}\hat{a}_{\text{in,+}}^\dagger\hat{a}_{\text{in,+}}. \quad (24)$$

For energy to be conserved, the input power must be equal to the power dissipated. We assume that we drive at a single frequency  $\omega$ :

$$P_{\text{in}} = \hbar\omega \langle \hat{a}_{\text{in,+}}^\dagger\hat{a}_{\text{in,+}} \rangle. \quad (25)$$

Using Eqs. (24)-(25), the average photon number  $\bar{n} = \langle \hat{n} \rangle$  is:

$$\bar{n} = \frac{\gamma_+}{\frac{\gamma_{\text{L}}^2}{4} + (\omega - \omega_r)^2} \frac{P_{\text{in}}}{\hbar\omega}. \quad (26)$$

Rewriting the rates in terms of their respective quality factors:

$$\bar{n} = \frac{\frac{4Q_{\text{L}}^2}{\omega_r Q_+}}{1 + 4Q_{\text{L}}^2\left(\frac{\omega}{\omega_r} - 1\right)^2} \frac{P_{\text{in}}}{\hbar\omega}. \quad (27)$$

If the drive is on-resonance  $\omega = \omega_r$  and the coupling is symmetric  $Q_+ = 2Q_{\text{c}}$ :

$$\bar{n} = \frac{2}{\hbar\omega_r^2} \frac{Q_L^2}{Q_c} P_{\text{in}}. \quad (28)$$

The input power  $P_{\text{in}}$  is defined as the power applied to the device under test, and not the output power of the VNA. Precise determination of  $P_{\text{in}}$  necessitates proper characterization of the total attenuation of input line that extends from the VNA to the device in the dilution refrigerator. The input line (Fig. S7) consists of a number of discrete attenuators and several feet of SMA cabling whose frequency-dependent line attenuation must also be characterized. We carefully measure the input line attenuation as a function of frequency including the discrete attenuators at room temperature by measuring the attenuation using a VNA. This line attenuation is then applied to determine  $P_{\text{in}}$  at the device. Because the majority of the input line cabling uses stainless steel conductors whose attenuation changes very little with temperature, we estimate that the deviation in line attenuation when cooled to cryogenic temperatures is small; we estimate a  $\pm 1$  dB uncertainty in the determination of  $P_{\text{in}}$ .