Semiconductor quantum-limited amplifier

D. Phan, 1 P. Falthans-Scheinecker, 1 U. Mishra, 1 W.M. Strickland, 2 D. Langone, 2 J. Shabani, 2 and A.P. Higginbotham 1

1 IST Austria, Am Campus 1, 3400 Klosterneuburg, Austria
2 Center for Quantum Information Physics, Department of Physics, New York University, New York, NY, 10003, USA

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We have built a parametric amplifier with a Josephson field effect transistor (JoFET) as the active element. The device’s resonant frequency is field-effect tunable over a range of 2 GHz. The JoFET amplifier has 20 dB of gain, 4 MHz of instantaneous bandwidth, and a 1 dB compression point of several photons when operated at a fixed resonance frequency. The amplifier’s noise performance approaches the limits imposed by quantum mechanics. Magnetic-field compatibility and opportunities for sensing are discussed.

I. INTRODUCTION

Quantum-limited amplifiers are, for many experimental platforms, the first link in the quantum signal processing chain, allowing minute signals to be measured by noisy, classical electronics [1–3]. Whereas later parts of the chain are dominated by semiconductor-based devices, the quantum-limited step can currently only be performed using metallic superconductors [4–12].

Aluminum-oxide based tunnel junctions have proven to be more reliable and stable than any other platform, thanks to the formation of pristine Al-AlO interfaces through the natural oxidization of Al.

Here, we introduce Al-InAs as the basis for quantum-limited signal processing devices. We demonstrate a quantum-limited amplifier using an Al-InAs Josephson field-effect transistor (JoFET) as the active element. Our device has a resonant frequency tunable over 2 GHz via the field effect. In optimal operating ranges, the JoFET amplifier has 20 dB of gain with a 4 MHz instantaneous bandwidth. The gain is sufficient for integration into a measurement chain with conventional semiconductor amplifiers. Accordingly, we demonstrate a total added noise that approaches the fundamental limits placed by quantum mechanics. In contrast to metallic superconducting amplifiers, our approach is compatible with parallel magnetic fields and can be used as a detector for voltages from a high-impedance source. The JoFET amplifier completes the suite of semiconductor-based options for quantum control, information processing, and readout.

II. DEVICE DESIGN

The JoFET amplifier is implemented as a half-wave coplanar waveguide (CPW) resonator with a gated superconductor-semiconductor hybrid Josephson field-effect-transistor positioned at the voltage node [Fig.1(a)].

Ground planes of the device are formed from 50 nm thin film Nb with 1 μm-squared flux-pinning holes near the edges to improve magnetic-field resilience [35, 36]. The entire center pin of the resonator is made of Al-InAs heterostructure, defined by chemical etch. The JoFET is defined by selective removal of the Al, followed by atomic layer deposition of alumina, and finally an electrostatic gate is defined with electron beam lithography and Au evaporation. Eventually, large Au chip-to-chip bond pads are also co-deposited in the final step to improve wire-bonding yield. The characteristic impedance of the CPW is designed to be near 50 Ω with 25 μm wide center conductor width and a 16 μm gap to the ground plane. The device is capacitively coupled to an open 50 Ω transmission line, which forms the only measurement port.

We targeted a geometric resonant frequency of 6 GHz for compatibility with with the standard 4 – 8 GHz band used in circuit quantum electrodynamics experiments. In designing the circuit, it is important to account for the kinetic inductance of the Al-InAs heterostructure [37], which typically causes a 20% reduction in the circuit resonant frequency. The device is therefore designed with a geometric resonant frequency of 7.2 GHz, corresponding to a CPW resonator length of 8 mm.

The external coupling κ ex and the critical current Ic determine the operating bandwidth and dynamic range of the amplifier [38]. While κ ex can be estimated from the circuit geometry, Ic is more subtle because it depends on material details. Based on independent transport tests,
we found that a JoFET with a w ith of 25 \( \mu m \) have an expected critical current of 10 \( \mu A \) at positive gate voltages. Combined with our designed external coupling of 2\( \pi \times 14 \) MHz this is expected to give a Kerr nonlinearity of approximately \( -3 \) kHz and a 1 dB compression point of a few photons with 20 dB of gain, which is suitable for parametric amplification [38].

**III. GATE TUNABILITY**

The microwave reflection coefficient \( R = \Gamma e^{i\phi} \) of a small incident signal is measured from the sample in a dilution refrigerator with a standard measurement chain, including a cryogenic commercial high-electron mobility transistor amplifier (HEMT). The measured reflection coefficient displays a small dip in magnitude and a 360° winding of phase, signaling that the resonator is strongly coupled to the measurement port and that dissipation is weak [Fig. 1(b)]. Indeed, fitting the reflection coefficient to a one-port model gives an external coupling efficiency \( \kappa_{ex} / \kappa_{tot} \sim 0.85 \) at high gate voltages.

Application of negative gate voltage results in dramatic changes in resonator frequency [Fig. 1(b)], indicating that the JoFET contributes a gate-tunable inductance to the resonator. The resonant frequency can be tuned over more than 2 GHz of bandwidth [Fig. 1(c)], exhibiting weak voltage dependence at extremal values, characteristic of a transistor reaching pinch-off at negative voltage and saturation at positive voltage.

The internal dissipation rate \( \kappa_i \) and external coupling rates \( \kappa_{ex} \) also evolve with gate voltage. Near JoFET pinch off \( \kappa_i \) increases sharply, and \( \kappa_{ex} \) decreases over the same range [Fig. 2(a),2(b)]. The net effect is therefore a decrease in coupling efficiency near pinch-off, which limits the usable frequency range of the JoFET amplifier.

A plausible origin or the increase in \( \kappa_i \) is dissipation in the JoFET region. To test this hypothesis, we consider a minimal model of the JoFET as a parallel resistor \( R_J \) and inductor \( L_J \) [Fig. 2(b),inset], introducing an effective resistance \( R_J/(\delta \phi)^2 \) where \( \delta \phi \) is the flux drop across the parallel combination. The flux drop can be inferred based on the measured resonant frequency, and couplings can then be calculated within an equivalent RLC model (see Appendix C).

Fitting \( \kappa_i \) to the circuit model results in satisfactory agreement with the data, with a best-fit shunt resistance \( R_J = 14.0 \pm 0.2 \) k\( \Omega \). Based on this agreement, we conclude that the junction presents dissipation to the circuit, with the practical effect of limiting performance at high inductances. We speculate that the dissipation is due to our use of a normal-conducting electrostatic gate, and is therefore not intrinsic to the Al-InAs system. The simple RLC model captures gate dependence of \( \kappa_{ex} \) at a qualitative level [Fig. 2(b)], although observed gate-dependence exceeds theoretical expectations. This discrepancy may reflect the role of extra capacitance in the JoFET region introduced by the electrostatic gate, which is not accounted for in the model.
IV. NONLINEARITIES AND AMPLIFICATION

In addition to tunable inductance, the presence of the JoFET imparts a power-driven nonlinearity to the resonator [38]. Measuring the reflected phase as a function of signal frequency and power reveals that the resonant frequency smoothly decreases with increasing power [Fig. 3(a)]. The output power from microwave sources is related to the resonator input power by an estimated total attenuation of 110 dB for all measurements. The downward shift in resonant frequency is accompanied by “sharpening” of the phase response [Fig. 3(b)]. Downward shift in resonant frequency and an alteration in lineshape are key qualitative signatures of the required Kerr nonlinearity $K$ for parametric amplification. The Kerr nonlinearity is estimated by measuring the change in resonant frequency per incident power in the low power limit, as shown by the linear fit in Fig. 3(c). Following this procedure at different gate voltages reveal that the nonlinearity is tunable with voltage, increasing in magnitude as gate voltage is decreased Fig. 3(d), mirroring the decrease in circuit resonant frequency observed in Fig. 1c. Calculating the expected Kerr nonlinearity based on the frequency shift in Fig. 1(c) adequately accounts for the observed gate-voltage dependence. The resulting calibration factor of $1.5 \times 10^5$ photons per Watt of vector network analyzer (VNA) power is, however, not consistent with our independently measured attenuation or power calibration. We speculate that this discrepancy is dominated by the propagation of systematic errors in our circuit-parameter estimate, which can cause large effects because the Kerr nonlinearity is a fourth-order effect.

Parametric amplification is generated by applying a strong pump tone red detuned from the bare resonant frequency, in the vicinity of the phase “sharpening” features already identified close to the critical power Fig. 3(b). Measuring scattering parameters with a weak probe signal reveals in excess of 20 dB of gain with a 4 MHz bandwidth, and a sharp phase response typical of a parametric amplifier [Fig. 4(a),(b)]. Modest detunings in the...
pump frequency do not substantially affect the amplifier response, but for large deviations the gain decreases to unity [Fig. 4(b)]. Measuring the amplifier gain \( G \) as a function of pump frequency \( f_{\text{pump}} \) and power \( P_{\text{pump}} \) reveals a continuous region of maximum gain with an easily identifiable optimum operating region, as expected for a parametric amplifier [Fig. 4(c)]. The gate voltages used for the gain optimization and measurement shown in Fig. 4(a),4(b) and in Fig. 4(c),4(d) are different due to gate instability and hysteresis. After setting the gate voltages, it typically takes 15 minutes for the resonance to stabilize. Consequently, the resonant frequency remains stable for circa an hour which is just sufficient for our gain and noise measurements. When the resonant frequency drifts away, retrieval of amplification requires tuning up the pump frequency and amplitude again. We anticipate that optimization of the gate dielectric can greatly improve stability in future devices. The power handling capability of the amplifier is quantified by measuring the gain for different signal powers [Fig. 4(d)]. At very low input signal power where the average intracavity photon number is 0.01 [Fig. 4(d)], the gain saturates at 20.3 dB. Large input powers cause the amplifier gain to decrease, giving a 1 dB compression point of 2.9 intracavity photons [input power of \(-125\) dBm, Fig. 4(d)]. The gain, bandwidth, and compression point are comparable to values obtained with early, practically useful parametric amplifiers based on metallic Al/AlOx Josephson junctions \([7, 8, 39]\).

Quantum signals typically consist of few photons, making it crucial to achieve noise performance near the quantum limit. To assess the noise performance of the JoFET amplifier, a weak pilot signal is measured first with a commercial HEMT amplifier, and then with parametric gain activated [Fig. 5(a)]. The JoFET amplifier dramatically improves the signal-to-noise ratio (SNR) of the pilot-signal measurement. The total input-referred noise of the JoFET amplifier and subsequent measurement chain approaches the limits placed by quantum mechanics of a photon from nondegenerate amplification \([2, 3]\). At small detunings we find a noise temperature of 0.38 ± 0.2 K, which is consistent with the expected total value 0.40 ± 0.1 K from vacuum fluctuations (0.29 K), nonzero cavity loss at this gate voltage (0.09 ± 0.01 K), and input-referred noise from the classical measurement chain (0.02 K).

This noise measurement was carefully calibrated by varying the temperature of the mixing-chamber stage of the dilution refrigerator and measuring noise at various frequencies with the JoFET amplifier off [Fig. 5(b)]. In the high temperature limit, the output noise is linear in temperature, with an offset that reflects the added noise of the chain referred to the mixing chamber plate, giving \( T_{\text{H,MC}} \) = 1.61 K at the JoFET operating frequency. At low temperature input-referred noise saturates. Calibrating over a wide range of frequencies reveals that noise saturation is pronounced only for high frequencies, confirming that the saturation is due to quantum, as opposed to thermal, fluctuations \([3]\), and that these fluctuations are faithfully resolved by our calibration procedure. To find the noise referred to the JoFET input, \( T_{\text{IN}} \), we divide by the independently measured insertion loss of all components in-between the JoFET input and the mixing chamber plate, including resonator loss (see Appendix H). This calibration procedure counts resonator
V. MAGNETIC FIELD OPERATION

A technical advantage conferred by the Al-InAs hybrid platform is compatibility with external magnetic fields. To explore the magnetic-field compatibility of the JoFET amplifier, we have steadily increased the parallel external magnetic field to 15 mT while compensating for small field misalignments with a perpendicular magnetic coil. Even with compensation, the resonator parameters evolve slightly in magnetic field, and a slight increase in loss is observed. Applying a pump tone gives an optimized gain of 10 dB [Fig. 6(a)-6(b)]. This demonstrates parametric amplification at a magnetic field order of magnitudes larger than the typical operating value for tunable metallic superconducting parametric amplifiers. Future experiments need to improve the resonator stability in magnetic field in order to allow a reliable noise measurement; we were unable to demonstrate quantum-limited operation in a magnetic field. The most likely reason for this is the depinning of trapped flux by high pump powers. This effect can be removed by forming the resonator center pin from a field-compatible superconductor that connects to a smaller Al-InAs microstructure, as was done with recent field-compatible superconducting qubits [40].

VI. OUTLOOK

Summarizing, we have demonstrated a quantum-limited and JoFET amplifier, already obtaining performance comparable to traditional Al/AlOx devices in our initial, unoptimized device. A number of techniques are available to further improve the performance of our device. The participation factor of the superconductor-semiconductor heterostructure can be reduced drastically by working with a small mesa in the JoFET region. This would likely decrease microwave losses, and allow the use of field-resilient superconductors like NbTi, which should allow operations in external magnetic fields of order 1 Tesla. The number of JoFETs and the designed critical current can also be adapted to provide the target nonlinearity at milder gate voltages [38]. Device dissipation can also likely be improved by using a superconducting electrostatic gate with on-chip filters. Finally, superconducting-semiconducting hybrid material systems are being actively developed so material-level improvements can be expected.

Our work opens up the new, general direction of quantum-limited signal processing devices based on semiconductors, and can be expanded to devices such as [39, 41] circulators [42, 43] or signal generators [44]. An appealing advantage of a semiconductor platform is the possibility of tight integration with semiconductor-based quantum devices, for instance spin qubits, or CMOS-based quantum control solutions [25–27]. It is also interesting to note that the electrostatic gate can be viewed as a receiver for high-impedance electrical signals, suggesting applications such as electrometry with an integrated, quantum-limited amplifier.

During preparation of this manuscript we became aware of related works demonstrating parametric amplification with graphene weak links [45, 46].

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Data availability Raw data for all plots in the main text and supplement will be included with the manuscript before publication. Further data available upon reasonable request.
Appendix A: Sample preparation

The JoFET amplifier was fabricated from Alproximitized InAs quantum well which was grown on a semi-insulating, Fe counter-doped (100) InP wafer. All the patterns were transferred with Raith EBPG 5100 electron beam lithography system. The Josephson weak link was made by etching a trench in the aluminum layer in the commercial etchant Transene D at 50 degrees C for 5 seconds. The trench was designed to be 20 nm long and 25 μm wide. SEM imaging later showed that the trench width was about 50 nm. The superconductor-semiconductor mesa forming the center pin of the CPW resonator was formed by masking with PMMA followed by a semiconductor wet etch in a mixture of CH₃COOH : H₂O₂ : H₃PO₄ for 150 seconds at room temperature. The ground plane was constructed by evaporating Ti 5 nm, Nb 50 nm following a short 1-minute Ar ion milling at an accelerating voltage of 400 V, with an ion current of 21 mA in a Plassys UHV evaporator. In the next step, the dielectric layer separating the gate and the junction was deposited with an Oxford ALD system running a thermal ALD alumina process at 150 degrees C in 150 cycles, which gave an approximated thickness of 12 nm. Eventually, the gate that covers just the area of the Josephson weak link was created by evaporating Ti 8 nm and Au 80 nm at a tilt angle of 30 degrees with 5 RPM planetary rotation in a Plassys HV evaporator. Due to low adhesion of Al bond wire to Nb ground plane, co-deposited Au bond pads were used to enhance Al alloy formation during wire bonding, hence, chip-to-chip bond yield. All lift-off and cleaning processes were performed in hot acetone at 50 degrees C and isopropanol.

Appendix B: Measurements

The sample was mounted on a copper bracket on a homemade printed circuit board. We used a vector network analyzer (VNA) model Keysight P9372A to measure R. A Rohde & Schwarz signal generator model SGS100A was used to provide the pump tone. When pumping at a fixed frequency, we used the VNA to provide a small probe signal and measure the gain profile from the reflected monotone at probe frequency. A home-made IVVI voltage source provided the gate bias. To find an effective lumped-element representation of the circuit, and then couple it to transmission lines following the procedure in Ref. [47].

Our circuit can be imagined as two resonators of length l and inductance per unit length L, capacitance per unit length C, and attenuation constant α coupled through the parallel RL circuit. For a large shunt resistance the inductance of the parallel combination is given by L. The model therefore introduces dissipation without changing the resonant frequency. The resonant wave vector k satisfies [48]

\[ 2 \cot(kl) = -\frac{L}{L_{eff}} \cdot kl. \]  

(C1)

The effective capacitance is [38]

\[ C_{eff} = C_l(1 + \text{sinc}(2kl)) . \]  

(C2)

We do not know of an explicit treatment of dissipation in our geometry available in the literature, so in analogy with the definition of Ceff we model it with the effective resistance is \( R_{eff} = \frac{1}{\kappa} \), where \( \kappa \) = 2 cos(kl) is the flux drop across the resistor. The internal dissipation and external coupling rates are then

\[ \kappa_i = \frac{\alpha l}{Z_0 C_{eff}} + \frac{(\delta \phi)^2}{R_{eff} C_{eff}} \]  

(C3)

\[ \kappa_{ex} = \frac{1}{R^* C_{eff}} \]  

(C4)

where \( R^* = (1 + \omega^2 C^2 Z^2)/(\omega^2 C^2 Z_0) \) is the effective parallel resistance from the measurement port with impedance \( Z_0 \) coupled with capacitance \( C_k \).

To fit this model we use the fact that \( kl = (\pi/2)f/f_0 \) where \( f_0 \) is the resonant frequency of the CPW resonator when \( L \) is zero. We estimate \( f_0 = 6 \text{ GHz} \) and then extract \( kl \) directly from the measured data. We fix the characteristic impedance of the resonator as \( Z_0 = 60 \Omega \) based on the difference between the designed and observed resonant frequency, which we attribute to the kinetic inductance of the heterostructure [37]. Knowledge of \( kl, f_0, \) and \( Z_0 \) allows he effective capacitance and Josephson inductance to be calculated. Eq. C3 is fit for \( \alpha \) and \( R_{eff} \) in Fig. 2(a) and Eq. C4 is fit for \( C_k \).
Appendix D: Conversion from signal power to average intracavity photon number

The average intra-cavity photon number $n$ is linearly dependent on the incident power $P_{\text{in}}$ at the input port:

$$n = \frac{1}{h f_s} \frac{4 \kappa_{ex} P_{\text{in}}}{(\kappa_{ex} + \kappa_i)^2 + 4 \Delta_i^2}. \quad (D1)$$

The input power is estimated based on the VNA power and the total attenuation $A$ of the setup.

Appendix E: Total attenuation estimate

The total attenuation $A$ between the output of the VNA and the input of the device is calculated from noise spectrum from the measurement port with a signal tone at $-43$ dBm output power. The noise floor is identified with input-referred noise temperature $T_{\text{H,MC}} = 1.61$ K from the HEMT calibration measurement shown in Fig. 5(b), combined with the resolution bandwidth of the signal analyzer $BW = 3.88$ kHz, giving $P_{\text{noise}} = 10 \log (k_B T_N BW/(1 \text{ mW})) = -160.6$ dBm. The signal is 6.3 dB above the noise floor, so its input referred magnitude is $-153$ dBm. Accounting for the small $-0.84$ dB of loss from the (detuned) cavity then gives the total attenuation $A = -110$ dB quoted in the main text.

Appendix F: Kerr nonlinearity extraction from power sweep

The dependence of resonance frequency on input power was used to estimate the Kerr nonlinearity $K$ similar to [7]. Resonant frequency decreased linearly with increasing average intracavity photon number following the Hamiltonian [38]:

$$H_{\text{JPA}} = \hbar \left( \tilde{\omega}_0 + \frac{K}{2} \left\langle A^\dagger A \right\rangle \right) A^\dagger A. \quad (F1)$$

The slope of the line in Fig. 3(c) measures $K/2$.

Appendix G: JoFET amplifier design

CPW resonator was designed to have a characteristic impedance of 50 Ω with the center conductor with and gap of 25 μm and 16 μm. Resonator length was chosen so that the target resonant frequency of 6 GHz would be achieved at base temperature. To account for the kinetic inductance of the 10 nm-thick Al layer, geometric resonant frequency was designed to be 7.2 GHz so that result resonance would drop to 6 GHz based on previous fabrication experience. The kinetic inductance of the Al film fluctuates from device to device, presumably due to uncontrolled oxidation of the Al. The JoFET was designed for a maximal critical current of 10 μA which gives a zeroth order Kerr nonlinearity $K$ of 1.4 kHz at zero bias and in this CPW configuration, such that the device can be gated into a regime appropriate for amplification [38]. Designed dimensions were chosen based on our previous lithography tests and critical current measurements.

Appendix H: JoFET input noise referral

The temperature sweep in Fig. 5(b) is used to refer HEMT noise to the mixing chamber plate [point ⋄ in Fig. 7]. Noise is then referred to device input [point ⋆ in Fig. 7] by measuring the insertion loss $\eta$ of the components in the signal path: the circulators, coaxial cables, sample board, and device. When the JoFET amplifier is off, noise at ⋄ is related to noise at ⋆ according to $S_\diamond = \eta S_\star + (1 - \eta) V + T_{\text{H,MC}}$, where the second term represents the introduction of vacuum noise $V$ by loss. The input-referred noise is then $S_\diamond/\eta$, which in the special case of vacuum-noise input gives $(T_{\text{H,MC}} + V)/\eta$.

We measured the transmission of the cable-circulator combination to be 0.6 dB and the sample holder insertion reflected insertion loss to be 0.2 dB. With the JoFET amplifier off, there is 0.6 dB of insertion loss from the detuned circuit resonance. This results in a total insertion loss of $\eta = 0.72$ between ⋆ and ⋄. Referring the added noise of the HEMT and vacuum fluctuations $V$.
We determine the JoFET amplifier must be divided out, which is measured using the pilot tone in Fig. 5(a). We determine the JoFET amplifier frequency response from a Lorentzian fit to the measured noise spectrum, and determine the overall gain level from the pilot signals, accounting for their nonzero detuning. Noise is input-referred by dividing by the full JoFET gain profile and the cable-circulator insertion loss.

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