The performance of a quantum device relies on the operating environment it is embedded.
An effective quantum environment must provide isolation from the various sources of noise, while allowing coherent transmission of control signals to implement fast quantum operations.
This implies a careful consideration in the design of the cryogenic setup as well as the room-temperature microwave signal-processing chain.
Superconducting circuits are operated at 20 mK, where the quantum system can be initialized in its ground state and avoid spurious thermal excitations. While such temperatures can be reached using commercially available dilution refrigerators, careful shielding and filtering must also put in place to minimize the exposure

to residual thermal noise and stray electromagnetic radiation.

There are mainly 3 aspects that we need to consider:

1) arrangement of the input microwave lines for low noise control
2) configuration of the output lines to allow optimal extraction of the signal
3) shielding of the device itself at 20 mK.

1 We must consider thermal noise that arises from both passive and active heat loads, propagating down the dilution refrigerator (DR). The former is caused by the flow of energy from upper to lower temperature stages in the DR; the latter is due to the dissipation of applied control signals in the route to the device. Residual thermal photons result in an elevated effective temperature of the device. In particular, the readout resonators, which are typically strongly coupled to the

transmission lines and the environment, can be prone to a non-negligible thermal photon population $\bar{n}_{th}$. This can lead to dephasing of the qubit that couples to the resonator, at a rate given by $\Gamma_\phi^{th} = \frac{\bar{n}_{th} \kappa \chi^2}{\kappa^2 + \chi^2}$, where κ is the linewidth of the resonator and χ is its dispersive coupling strength to the qubit.

It is crucial to reduce the amount of residual thermal photons in our quantum system in order to achieve better coherence time.

In order to suppress thermal noise, $\boxed{\text{attenuators}}$ must be introduced along the microwave lines in the DR. An attenuator of "x dB" reduces both the signal and the noise coming from higher temperature stages, in addition introducing an extra thermal noise component from the temperature stage where the attenuator is anchored.

The thermal energy at a given temperature T_i translates into an influx of photons propagating down to the device with an average number $\langle n_i \rangle$, given by $\langle n_i \rangle = \frac{1}{e^{\hbar\omega/k_B T_i} - 1}$. It must be suppressed to or below the temperature of the subsequent cooling stage in order to minimize the propagation of thermal noise. This can be achieved by introducing appropriate cryogenic attenuators at each temperature stage. An attenuator can be modelled as a beam splitter, which transmits a small fraction of the incident power while, at the same time, adds a certain amount of thermal photons due to the black-body radiation at the temperature stage where it is located.

To illustrate this, let us consider a system with an attenuator A_1 placed at a higher temperature stage and A_2 on a lower stage. The resulting averaged photon number on the lower stage is given by $\langle n_f \rangle = \frac{\langle n_i \rangle}{A_1 A_2} + \left(1 - \frac{1}{A_1}\right)\frac{\langle n_1 \rangle}{A_2} + \left(1 - \frac{1}{A_2}\right)\langle n_2 \rangle$, where n_i is the average number of photons going into

the system and $n_{1,2}$ the photons at each of the stage.
Note that usually the attenuation value is expressed in dBs: $A = 10^{A'/10}$, where A' is the level of attenuation in dBs.
Attenuators are more effective in suppressing thermal noise when they are placed on the lower temperature stage. However, most attenuators employ dissipative elements which could cause additional heating if the dissipation exceeds the cooling power of the DR at the specific temperature.
In addition, to reduce the thermal noise due to active heat loads, precautions should also be taken to filter out spurious frequency components that could cause unwanted transitions or couplings. This can be achieved by adding cryogenic filters at the 20 mK plate. Those filters can provide ~40 dB attenuation above 10-12 GHz. Other kind of filters, called Eccosorb filters, are capable to absorbe signals above 20 GHz to protect the device from high frequency radiation.

(2) The output lines are responsible for carrying the resulting signals from the device, at the base temperature stage of the DR, back to room temperature. These are typically very small sygnals, at the single photon level. This small signal can be easily degraded by noise or dissipation. This signal must be handeled carefully to ensure clean, stable, and effective collection of quantum information from the device.

It is necessary to suppress the thermal noise propagating from the higher temperature stages while allowing the signal to propagate without significant attenuation. This can be achieved by making use of "directional" components such as circulators and isolators placed on the output lines (instead of the attenuators). These nonreciprocal elements provide a low-pass path for the quantum signal extracted from the device to travel to room temperature while reducing the thermal noise flowing in the opposite direction.

It is also necessary to enhance the weak outgoing signal through a well-designed chain of amplifiers such that it can be distinguished from noise as it propagate back to room temperature. The role of each amplifiers along the outgoing signal path is to make the signal stand out against the noise added by the subsequent amplifier as well as the thermal photons from the higher-temperature stages. When designing amplification chains, it is important to note that amplification always comes at the cost of elevated noise temperature. This causes a degradation of the SNR despite the overall increase of the signal power. The noise temperature T_N provides a measure of the noise added by the amplifier. For an amplifier with a power gain of G, we can relate the amplified signal P^S_{out} and noise P^N_{out} to the input powers $P^{S,N}_{in}$, by : $P^S_{out} = G P^S_{in}$; $P^N_{out} = G(T_{in} + T_N) k_B B$

SNR = Signal to noise ratio

where B is the bandwidth of the noise going into the amplifier, T_{in} and T_N the effective temperatures associated with the input signal and the noise of the amplifier. We can now write the SNR at the output of the amplifier as $SNR_{out} = P^S_{out}/P^N_{out} = P^S_{in}/(T_{in}+T_N)\,K_B B$. Comparing this with the input SNR, given by $SNR_{in} = P^S_{in}/T_{in} K_B B$, we arrive at the relation $SNR_{in}/SNR_{out} = 1 + \frac{T_N}{T_{in}} > 1$, indicating that the SNR after the amplifier (out) is always going to be lower than that at the input. This means that on the output line it is crucial to minimize the noise associated with the first gain stage. This can be achieved using a quantum-limited amplifier connected to the output of the resonator with minimal attenuation in between. This amplifier only adds the minimum amount of noise allowed by the laws of of quantum mechanics. A quantum-limited amplifier is followed by a commercial wide-band cryogenic HEMT amplifier at the 4K-stage to further boost the signal before it is acquired at room temperature.

(3) Finally the device must also be shielded from stray magnetic fields. It has been shown that when a device is cooled-down to 10 mK in the presence of a magnetic field of more than 0.1 Gauss, vorticies can be trapped in the thin-film superconductor and cause a reduction of the coherence properties. To mitigate this, we house the sample in Cryoperm shields which are made out of high permeability nickel alloys and treated specifically to ensure robust magnetic screening properties. It is also crucial that we minimize the residual magnetic field inside the shield by using only non-magnetic components.

The system noise temperature for the amplifier chain can be expressed in terms of the individual gain figures "G_n" and noise temperatures "$T_{N,n}$" of each constituent amplifier. $T_{syst} = T_{N,1} + \frac{T_{N,2}}{G_1} + \frac{T_{N,3}}{G_1 G_2} + \cdots$ where $n = 1, 2, 3, \ldots$ denotes the order of the amplifiers starting from the qubit chip. From this expression we can see that the noise temperature T_{syst} is dominated by the noise contribution from the first amplifier. The gain of the first amplifier has the effect of suppressing the noise added by the second amplifier, and so on. If the first amplifier is a low-noise high-electron mobility transistor (HEMT) amplifier ($T_N \simeq 2K$) the system noise temperature T_{syst} will be around ~7-10 K corresponding to around 10-20 added photons of noise per signal photon around 5 GHz.

Measuring the resonator amplitude and phase

The readout circuit can be set up in measuring either reflection or transmission. The best state discrimination is obtained by maximizing the separation between the two states in the (I, Q)-plane (the in-phase and quadrature component of the voltage). It can be shown that this separation is maximal when the resonator is probed just in between the two qubit-state dependent resonance frequencies $\left[\omega_{RF} = \left(\omega_r^{|0\rangle} + \omega_r^{|1\rangle}\right)/2\right]$. In this case, the reflected magnitude is identical for $|0\rangle$ and $|1\rangle$, and all information about the qubit state is encoded in the phase θ. In addition, the qubit-resonator detuning should be designed to obey the criterion for maximal state visibility $[\chi = \kappa/2]$.
The quantum dynamics of the qubit can be mapped onto the phase of the classical microwave response.

A readout event start with a short microwave tone directed to the resonator, at the resonator probe frequency ω_{RO}. After interacting with the resonator the reflected / transmitted signal has the form $s(t) = A_{RO}\cos(\omega_{RO}t + \theta_{RO})$, where ω_{RO} is the "carrier" frequency used to probe the resonator.

A_{RO} and θ_{RO} are the qubit-state dependent amplitude and phase that we want to measure.

This signal can be rewritten in a static "phasor" notation that separates out the time dependance $\omega_{RO}t$

$$s(t) = \mathrm{Re}\left\{\underbrace{A_{RO}\,e^{j\theta_{RO}}}_{\text{phasor}}\,e^{j\omega_{RO}t}\right\}$$, where the phasor $A_{RO}\exp(j\theta_{RO})$ fully specifies an harmonic signal $s(t)$ at a known frequency ω_{RO}.

To perform qubit readout, we want to measure the "in-phase" component I and a "quadrature" component Q of the complex number of the phasor $A_{RO}\,e^{j\theta_{RO}} = A_{RO}\cos\theta_{RO} + jA_{RO}\sin\theta_{RO} \equiv I + jQ$

I-Q mixing: a direct means to extract I and Q is to perform a homodyne or a heterodyne measurement using analog I-Q mixers.

The readout signal $s(t)$ and the reference local-oscillator signal $y(t) = A_{LO} \cos \omega_{LO} t$ are fed into the mixer via the RF and LO mixer ports. The mixer equally splits both these signals into two branches and multiplies them:

- in the I-branch, the signal $s_I = s(t)/2$ is multiplied by the LO signal $y_I(t) = (A_{LO}/2) \cos \omega_{LO} t$
- in the Q-branch, the signal $s_Q(t) = s(t)/2$ is multiplied by a $\pi/2$-phase-shifted version of the LO signal $y_Q(t) = -(A_{LO}/2) \sin \omega_{LO} t$

At the mixer I and Q ports, the output signals $I(t)$ and $Q(t)$ contain terms at the sum and difference frequencies, generally referred to as an intermediate frequency $\omega_{IF} = \omega_{RO} \pm \omega_{LO}$. The resulting signals are low-pass filtered passing only the terms at the difference frequency $I_{IF}(t)$, $Q_{IF}(t)$ which are then digitized.

After digital signal processing, one can obtain the static in-phase (I) and quadrature (Q) components, from which one calculates the amplitude A_{RO} and the phase θ_{RO}.

Heterodyne demodulation: In a heterodyne scheme, a local oscillator at frequency ω_{LO} is offset by an intermediate frequency ω_{IF} to target a unique readout frequency ω_{RO}. Here, we want to extract A_{RO} and θ_{RO} from the reflected/transmitted tone using a heterodyne scheme. The first step is to perform analog I-Q mixing, where the LO and RO frequencies are different $\omega_{IF} = |\omega_{RO} - \omega_{LO}| > 0$. Mixing LO and RO signals yields the signals $I(t)$ and $Q(t)$ with terms with both sum and difference frequencies. By using low-pass filters we can filter out the sum frequencies, yielding the IF signals: $I_{IF}(t) = \frac{1}{T}\int_0^T dt\, S_I(t)\, y_I(t) = A_{RO} A_{LO}/8\, \cos(\omega_{IF} t + \theta_{RO})$

$$Q_{IF}(t) = 1/T \int_0^T dt\, S_Q(t)\, y_Q(t) = A_{RO} A_{LO}/8\, \sin(\omega_{IF} t + \theta_{RO})$$

We notice that the signal that we are interested is the change in A_{RO} and θ_{RO} in correspondence of a change in the qubit states. The analog-demodulated $I_{IF}(t)$ and $Q_{IF}(t)$ are now oscillating at a frequency that is generally low enough to be digitized using commonly available analog-to-digital converters (ADCs). The resulting digital signals are now written as $I_{IF}[n]$ and $Q_{IF}[n]$:

$$I_{IF}[n] = \frac{A_{RO} A_{LO}}{8} \cos(\Omega_{IF} n + \theta_{RO})$$

$$Q_{IF}[n] = \frac{A_{RO} A_{LO}}{8} \sin(\Omega_{IF} n + \theta_{RO})$$

where $n = t/\Delta t$ indexes the sample number of the continuous-time signals $I_{IF}(t)$ and $Q_{IF}(t)$; $\Omega_{IF} = \omega_{IF} \Delta t$ is the digital frequency, and Δt is the sampling period (typically around ~1ns).

Digital demodulation comprises the point-by-point multiplication of I_{IF} and Q_{IF} by $\cos \Omega_{IF} n$ and $\sin \Omega_{IF} n$. Averaging the resulting time series eliminates the $2\Omega_{IF}$ component while retaining the DC component.

One finally obtains:

$$I = \frac{1}{M} \sum_{m_1}^{m_2} I_{IF}[m] \cos[\Omega_{IF} m] = \frac{A_{RO} A_{LO}}{16} \cos \theta_{RO}$$

$$Q = \frac{1}{M} \sum_{m_1}^{m_2} Q_{IF}[m] \sin[\Omega_{IF} m] = \frac{A_{RO} A_{LO}}{16} \sin \theta_{RO}$$

where $M = m_2 - m_1 + 1$.

Now I and Q can be used to find A_{RO} and θ_{RO}.