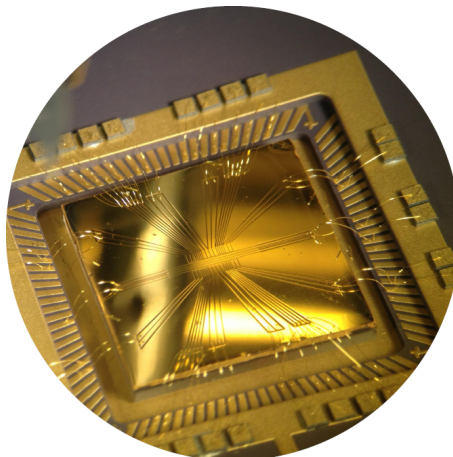


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# Near-field microwave addressing of trapped-ion qubits for scalable quantum computation

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A thesis submitted for the degree of  
Doctor of Philosophy



Diana Prado Lopes Aude Craik

Trinity Term  
2016

Christ Church  
Oxford



# Abstract

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## **Near-field microwave addressing of trapped-ion qubits for scalable quantum computation**

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Christ Church, Oxford

This thesis reports high-fidelity near-field spatial microwave addressing of long-lived  $^{43}\text{Ca}^+$  “atomic clock” qubits performed in a two-zone single-layer surface-electrode ion trap. Addressing is implemented by using two of the trap’s integrated microwave electrodes, one in each zone, to drive single-qubit rotations in the zone we choose to address whilst interferometrically cancelling the microwave field at the neighbour (non-addressed) zone. Using this field-nulling scheme, we measure a Rabi frequency ratio between addressed and non-addressed zones of up to 1400, from which we calculate an addressing error (or a spin-flip probability on the qubit transition) of  $10^{-6}$ . Off-resonant excitation out of the qubit state is a more significant source of error in this experiment, but we also demonstrate polarization control of the microwave field at an error level of  $2 \times 10^{-5}$ , which, if combined with individual-ion addressing, would be sufficient to suppress off-resonant excitation errors to the  $10^{-9}$  level.

Further, this thesis presents preliminary results obtained with a micron-scale coupled-microstrip differential antenna probe that can be scanned over an ion-trap chip to map microwave magnetic near fields. The probe is designed to enable the measurement of fields at tens of microns above electrode surfaces and to act as an effective characterization tool, speeding up design-fabrication-characterization cycles in the production of new prototype microwave ion-trap chips.

Finally, a new multi-layer design for an ion-trap chip which displays, in simulations, a 100-fold improvement in addressing performance, is presented. The chip electrode structure is designed to use the cancelling effect of microwave return currents to produce Rabi frequency ratios of order 1000 between trap zones using a single microwave electrode (i.e. without the need for nulling fields). If realized, this chip could be used to drive individually addressed single-qubit operations on arrays of memory qubits in parallel and with high fidelity.





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In memory of my wonderful parents,  
*Eliana Prado Lopes Aude & Julio Salek Aude*



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# 1

## Introduction

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The work described in this thesis aims to contribute to the current scientific effort dedicated to the experimental realization of universal quantum computation. This chapter will motivate the drive to build a quantum computer by underlining the impressive scale of the computational power such a machine would yield and listing some of the applications which have already been envisioned for it, while noting that many more are likely to emerge. Sections 1.2 and 1.3 outline what must be achieved experimentally in order to realize a quantum computer and highlight the main challenges faced by experimentalists today: namely, the reduction of errors in elementary operations to a level which permits fault tolerant quantum computation to be performed in platforms which are scalable to many-qubit systems. Section 1.4 serves as a brief introduction to the basic principles, phenomena and terminology which are relevant to ion trapping and which are used in the later chapters of the thesis. Finally, the structure of this thesis is outlined in section 1.5.

### 1.1 Why build a quantum computer?

Both classical and quantum computing involve performing operations on elements of a computational basis, usually integers represented as binary numbers, with each digit encoded in a physical bit or qubit. The intrinsic difference be-

tween the type of computation available to a quantum computer (QC) and that which is possible in a classical machine emerges from multipartite quantum entanglement, specifically from the fact that the entangled states manipulated by a QC physically represent correlations between elements of the computational basis. For example, the state  $\frac{|1_A\rangle|0_B\rangle+|0_A\rangle|1_B\rangle}{\sqrt{2}}$  is a physical representation of the fact that when the one-bit integer encoded in qubit A is  $|0\rangle$ , the integer stored in qubit B is  $|1\rangle$  and vice-versa. More generally, a quantum computer is capable of producing entangled states via processes of the following form:

$$\sum_{x=0}^{2^n-1} |x\rangle|0\rangle \rightarrow \sum_{x=0}^{2^n-1} |x\rangle|f(x)\rangle \quad (1.1)$$

where, as is pointed out in [Ste03a], the superposition on the right-hand side fully represents the correlations between the integer  $x$  and the output of the function  $f(x)$  (in the sense that a measurement of this state will always produce a pair of values correlated such that if one is  $|x\rangle$  the other is  $|f(x)\rangle$ , for any  $x$  in the superposition) without fully representing the  $2^N - 1$  values which constitute the complete  $(x, f(x))$  truth table (since there is only a small probability that a measurement will extract any given pair of  $(x, f(x))$  from the state). Efficient quantum algorithms make use of entangled states as a compact, yet complete, representation of correlations to solve certain problems in a targeted way, greatly reducing the number of computational steps required to arrive at a solution (as compared to classical algorithms) by avoiding the calculation of extraneous results. [Ste03a] gives the example of Shor's algorithm [Sho97], which can calculate the period of a function  $f(x)$  without evaluating its value at every  $x$  and compares the exponential speed-up the algorithm achieves through the exploitation of correlations represented in entangled states to the speed-up implemented by the binary search algorithm of classical computing, where logical reasoning is used to avoid searching through every element of an ordered list.

The implications of the algorithmic speed-ups made possible by quantum computers are extraordinary. Applications extend to the fields of medicine



(where classically intractable protein folding simulations are required for drug design [KWPO<sup>+</sup>11]), cryptography (where RSA encryption relies on the eavesdropper's inability to factor large prime numbers, a classically difficult problem which can be efficiently solved in polynomial time by a QC running Shor's algorithm), precision measurement (where a QC could be used to perform entanglement-enhanced spectroscopy [WBB<sup>+</sup>02]), materials science (quantum simulations of interactions in solids would help us understand high temperature superconductivity [Blo08] and model topological insulators [MBG<sup>+</sup>12]), and many others [CZ12, LWG<sup>+</sup>10, BN09]. Indeed, it is expected that the arrival of large scale quantum computation will unlock a vast range of yet unimagined applications, opening the doors to the expansion of research in many areas of scientific investigation.

## 1.2 Requirements for Quantum Computation

The essential components of a quantum computer were described by Deutsch in 1989 [Deu85]. Based on Deutsch's concept, DiVincenzo set out the criteria which must be satisfied by any physical implementation of such a computer [Div00]. They are as follows:

1. *A scalable physical system with well characterized qubits*
2. *The ability to initialize the state of the qubits to a simple fiducial state such as  $|000\dots\rangle$*
3. *Long relevant decoherence times, much longer than the gate operation time*
4. *A "universal" set of quantum gates*
5. *A qubit-specific measurement capability*

Candidate qubit systems being actively researched include superconducting Josephson junctions, nitrogen and silicon-vacancy centres in diamond, silicon

quantum dots, single photons, neutral atoms and trapped ions. Several of the listed implementations have already been experimentally shown to satisfy most of the DiVincenzo criteria in small scale few-qubit realizations. However, one important requirement that is not explicitly mentioned in the above list is the need for fault-tolerance, i.e. the operations performed by our QC must be robust to errors introduced by the environment. Errors must not accumulate during computations and must be correctable with the available resources. In the next section, we will turn our attention to how fault tolerance can be achieved with the use of quantum error correction and discuss how the resources required to perform fault tolerant quantum computing scale with gate error rates.

### 1.3 Quantum error correction and fault tolerance

In classical computing, errors are routinely corrected with the use of error correcting codes. One of the simplest classical error correcting codes is the repetition code whereby information is protected by keeping copies of each bit. Let's imagine, for instance, that our computer is subjected to noise that flips a bit with probability  $p$ . By keeping three copies of each bit, we can reduce the probability of error to the probability that at least two bits have been corrupted,  ${}^3C_2p^2(1 - p) + p^3$  (which is smaller than  $p$  for  $p < 0.5$ ).

The repetition code is not particularly efficient in terms of the trade-off between reduction of the error probability and resources required (i.e. the number of ancilla (copy) bits needed and number of necessary comparisons), but it serves to illustrate the redundancy principle upon which all classical error-correcting codes are based.

For a long time, it was believed that performing error correction in the quantum domain would not be possible, precisely because we cannot make copies of qubits, a fact which is formalized by the 'no-cloning' theorem. To add to this difficulty, errors are continuous in the quantum world: noise can rotate the

qubit around any axis and by any angle in the Bloch sphere, in stark contrast to the discrete space inhabited by classical errors. There was also uncertainty as to whether one could even detect the errors without derailing the computation, since it is not possible to observe a quantum state without destroying its coherence.

Without error correction, it was clear that universal quantum computation could not be performed. There would be successive accumulation of errors and the computer would quickly decohere. But in 1995, Andrew Steane, Peter Shor and Robert Calderbank put forward a revolutionary idea. They suggested that, if we entangle a set of ancilla qubits with the qubit we want to protect, we can make parity measurements on the entangled state which allow us to detect and correct for errors [Ste96, CS96]. These parity measurements give no information about the state of any individual qubit, and hence preserve the coherence of the state. They are, however, projective measurements and will hence project a general error undergone by the qubit onto one of a discrete set of errors, which can then be corrected for. To see this, note that any interaction between a set of qubits and an environment can be written as:

$$|\phi\rangle|\psi_0\rangle_e \rightarrow \sum_i (E_i|\phi\rangle)|\psi_i\rangle_e \quad (1.2)$$

where  $|\phi\rangle$  is the unadulterated state of our qubit,  $|\psi_0\rangle_e$  is the initial state of the environment and  $E_i = e_{i0}I + e_{i1}X + e_{i2}Y + e_{i3}Z$  is a general error operator which represents an arbitrary rotation on the Bloch sphere (we have defined  $X \equiv \sigma_x$ ,  $Y \equiv -i\sigma_y$  and  $Z \equiv \sigma_z$  and  $e_{i[0,1,2,3]}$  are constants).

If one then entangles a set of ancilla qubits with this noisy state, the state of the total system  $\Psi$  can be written as:

$$|\Psi\rangle = \sum_i |s_i\rangle_a (E_i|\phi\rangle)|\psi_i\rangle_e \quad (1.3)$$

where the state of the ancilla qubits has been written as a superposition of elements  $|s_i\rangle_a$  with  $s_i$  denoting one of the possible error syndromes (sets of eigen-

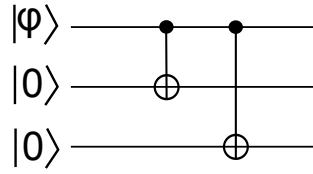


Figure 1.1: Encoding circuit for bit-flip code. Controlled-not gates with the ancilla qubits as targets and  $|\phi\rangle$  as the control produce an entangled state given by  $a|000\rangle + b|111\rangle$  where  $|\phi\rangle = a|0\rangle + b|1\rangle$ .

Table 1.1: Bit flip code. To detect the error which occurred, we extract two error syndromes by making two projective measurements:  $Z_1Z_2$  and  $Z_2Z_3$ .

| Error | Erroneous code word           | $Z_1Z_2$ eigen-value | $Z_2Z_3$ eigen-value | Correction operation |
|-------|-------------------------------|----------------------|----------------------|----------------------|
| $X_1$ | $a 100\rangle + b 011\rangle$ | -1                   | +1                   | $X_1$                |
| $X_2$ | $a 010\rangle + b 101\rangle$ | -1                   | -1                   | $X_2$                |
| $X_3$ | $a 001\rangle + b 110\rangle$ | +1                   | -1                   | $X_3$                |
| $I$   | $a 000\rangle + b 111\rangle$ | +1                   | +1                   | $I$                  |

values which are results of projective parity measurements) which may be extracted from the ancilla. If we do a projective measurement on the ancillas,  $|\Psi\rangle$  will collapse to a single element of the sum  $|s_j\rangle_a (E_j|\phi\rangle)|\psi_j\rangle_e$ , yielding  $s_j$  as the measured syndrome. Assuming that there is only one  $E_j$  for a given  $s_j$  (there is actually a coset of error operators corresponding to  $s_j$  and we choose one from each coset to correct), we can then reverse this rotation and correct the error. The measurement of the ancilla has therefore collapsed a general noisy state into a particular discrete error that we can reverse.

To illustrate the process of syndrome extraction and error correction, we briefly consider the simplest error correcting code: the bit-flip code. Here, a logical 0 is represented as an entangled state of three qubits (our computation qubit and two ancillas),  $|0_L\rangle \equiv |000\rangle$  and, similarly, logical 1 is encoded as  $|1_L\rangle \equiv |111\rangle$ . The encoding circuit is shown in fig. 1.1 and table. 1.1 illustrates how spin-flip ( $X$ ) errors can be corrected for. A very similar code can be used to correct for phase-flip errors ( $Z$ ), and it can be shown that, if one is able to correct for both spin-flip and phase-flip errors, one can correct for any general error [Ste06].

QEC can, in principle, enable fault-tolerant quantum computation. However, for any given error correcting code, the number of resources (i.e. the number of ancilla qubits) required to do fault-tolerant quantum computation quickly becomes impractically large if the error incurred per operation of our QC is above a certain threshold. Fig. 1.2 illustrates, for the class of codes called the CSS codes (named after its inventors, Calderbank, Shor and Steane), how the number of operations which can be performed before the QC decoheres scales with the error per gate (EPG) and with the number of ancilla qubits. We see that there is a clear threshold at an EPG just under  $10^{-3}$ . Above this error level, no matter how many resources we invest (i.e. how many physical qubits we use to encode a logical qubit), we can perform very few operations before the QC decoheres - we are sitting at the base of the 'mount' depicted in the plot. If we can decrease the EPG by a factor of 10 to  $\sim 10^{-4}$ , we are able to do a significant number of operations with a modest resource investment (e.g. at an EPG of  $\sim 10^{-4}$ , we can do about 10 billion operations if using just 10 ancilla qubits to represent one logical qubit). At the threshold EPG (the mountain face), the error is low enough to allow us to keep the coherence alive but not to do much else - all resources are invested on correcting for errors and no useful computations can be performed.

This threshold EPG is termed the 'fault tolerant threshold' and is different for different error correcting codes. More recent codes put forward for QEC, such as the surface code, set the thresholds at an error level of up to  $10^{-2}$  [WFSH09, RH07, BK98]. Regardless of the code being used, reducing the EPG can only increase the number of operations a QC can perform per invested resource. Much of the recent experimental effort in quantum computation has been dedicated to increasing the fidelity of elementary computing operations, since the lower we can push the errors, the more feasible non-trivial computation with a QC becomes.

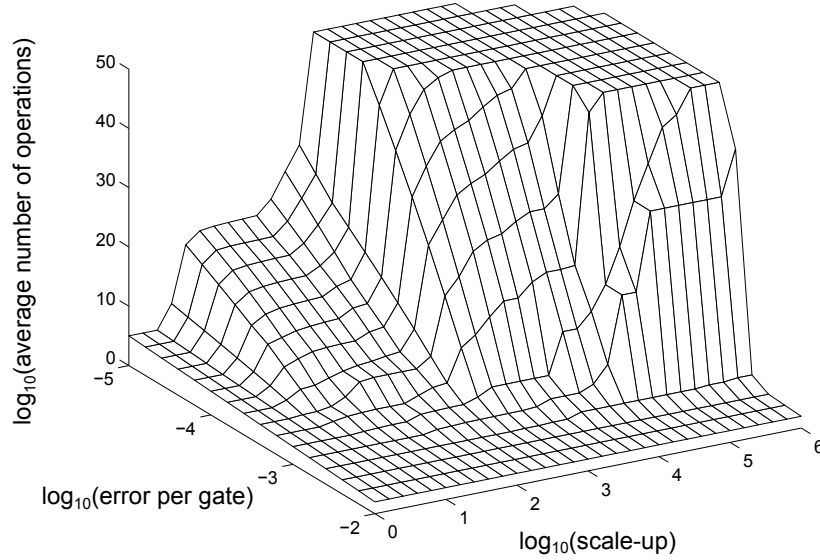


Figure 1.2: Plot from [Ste03b] showing the average number of operations that can be performed with a QC (on which a CSS code is being used to do quantum error correction) as a function of error per gate and number of ancilla qubits required per logical qubit (scale-up). The plot has been truncated at  $10^{50}$  operations.

## 1.4 Ion-trap quantum computers

In this work, we will pursue the trapped ion implementation of quantum computation. Subsection 1.4.1 will motivate this choice by outlining the attractive features of trapped ions as a candidate platform for quantum computation. A short introduction to the experimental methods used in this implementation is given in subsection 1.4.2, as well as a discussion of relevant phenomena observed in trapped ions.

### 1.4.1 Why trapped ions?

It is useful to assess the suitability of trapped ions as a quantum computing platform in the context of the Di Vincenzo criteria. We now re-visit each criterion in turn and describe the degree to which the state-of-the-art trapped-ion systems fulfil it, whilst highlighting advantageous features of the trapped-ion implementation.

1. *A scalable physical system with well characterized qubits:* We encode information in two hyperfine levels of a trapped ion. Coherent manipulation of the population in the qubit states is made possible by experimental techniques which have matured over several decades, having originated from the field of precision spectroscopy. Using RF trapping (see section 1.4.2) and laser cooling techniques (see section 3.1.2), we can trap and cool single ions. This enables us to work with well isolated and relatively simple quantum systems. Furthermore, encoding information into atomic energy levels ensures that, when subjected to the same environment, every qubit behaves identically - this kind of uniformity is inherent in atom and ion-based quantum computation, but must be engineered in other platforms (such as solid-state qubits). Scalable implementations of ion trapping technology have already been developed (in the form of microfabricated surface ion traps, as discussed in section 1.4.3) and significant effort is being invested into performing high-fidelity operations in scalable prototype traps [ACLS<sup>+</sup>17, HAB<sup>+</sup>14, HSA<sup>+</sup>16, GTL<sup>+</sup>16].
  
2. *The ability to initialize the state of the qubits to a simple fiducial state such as  $|000\dots\rangle$  and a qubit-specific measurement capability:* State preparation is performed by optically pumping population to a desired qubit state (see section 3.1.3). Individual-ion readout can be performed by using a laser beam to selectively couple one of the two qubit states to a third energy level via a cycling transition on which many photons can be scattered (see section 3.1.4). When the readout laser is pulsed, an ion will fluoresce if it is in the state which couples to the cycling transition, but will appear dark if it is in the other qubit state [MSW<sup>+</sup>08]. This state-dependent fluorescence can be collected with a camera or PMT. For a string of trapped-ions, the fluorescence can be spatially resolved on a camera so that the state of each ion can be individually read-out. Our group has manipulated  $^{43}\text{Ca}^+$  qubits

with combined state-preparation and readout errors as low as  $7 \times 10^{-4}$  [HAB<sup>+</sup>14].

3. *Long relevant decoherence times, much longer than the gate operation time:* One can choose to work with qubit transitions which are first order insensitive to magnetic field fluctuations (so called “atomic clock” transitions). Such transitions exist between hyperfine states of the ground level of an ion and have resonance frequencies in the microwave regime. They offer essentially infinite  $T_1$  times as well as  $T_2^*$  times on the order of tens of seconds [LOJ<sup>+</sup>05, HAB<sup>+</sup>14]. Typical gate operation times for trapped ions range between  $1\mu\text{s}$  and  $1\text{ms}$ , implying that decoherence time to gate time ratios of up to  $10^7$  are currently achievable.
4. *A “universal” set of quantum gates:* single qubit gates can be performed in ions by driving transitions between the two qubit states using laser or microwave fields. Two-qubit gates are performed by inducing two-ion entanglement via the Coulomb interaction, which couples the motion of the ions in a trap. Both single qubit and two-qubit gates have been implemented with very high fidelities using trapped ions, with our group having performed microwave-driven single qubit gates and two-qubit gates at the  $10^{-6}$  and  $10^{-3}$  error level respectively [HAB<sup>+</sup>14, HSA<sup>+</sup>16].

### 1.4.2 The Paul trap

Fig. 1.3a depicts a traditional linear Paul trap. Named in honour of its inventor, Wolfgang Paul, it is the main instrument now used to trap individual ions for quantum computation. The trap consists of four metal rods: an RF potential is applied to two of them and the other two are segmented DC electrodes, with each segment held at a different static voltage. The confinement along the  $y$  axis - which we refer to as the trap axis - is achieved by setting the static voltages



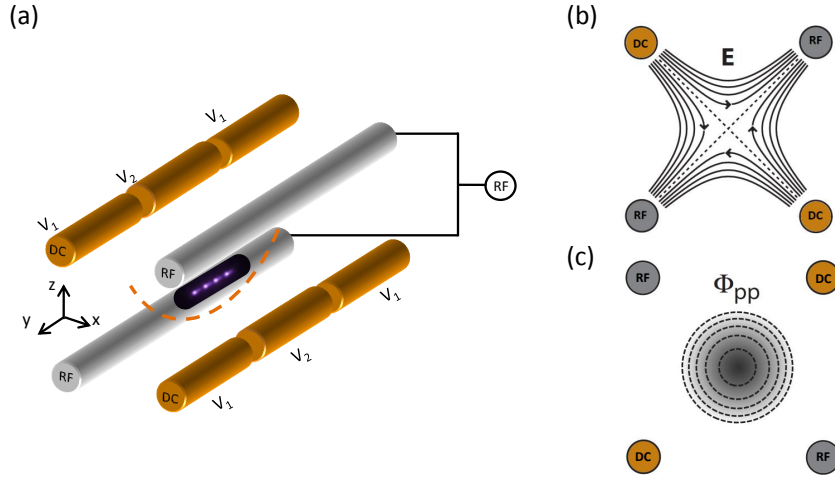


Figure 1.3: (a) Linear Paul trap showing segmented DC electrodes (orange) and RF electrodes (grey). The dashed orange line shows the DC harmonic well potential which provides confinement along the trap axis. In the centre of the well is a string of four ions. Figures (b) and (c) are taken from [ABLW08] and show (b) the instantaneous radial RF potential seen by the ion in the radial ( $xz$ ) plane and (c) the pseudo-potential (the effective time-averaged potential experienced by the ion in the radial plane).

$V_1$  and  $V_2$  so as to generate a DC harmonic well along this axis. The total DC potential is:

$$\phi_{dc}(x, y, z) = \alpha_{dc} V_{dc} \left( y^2 - \frac{x^2 + z^2}{2} \right) \quad (1.4)$$

where  $\alpha_{dc}$  is a geometric factor which has units of  $[1/\text{distance}^2]$  and  $V_{dc} = V_1 - V_2$ . The first term is the trapping harmonic well along the trap axis. The motion along this direction is simply a harmonic oscillation at the frequency  $\omega_{\text{axial}} = \sqrt{\frac{2\alpha_{dc}V_{dc}}{m}}$ . The second term represents the equal and opposite anti-trapping in two orthogonal directions (inverted harmonic wells along  $x$  and  $z$ ) as required by Laplace's equation. To confine in the  $xz$  plane (or the radial plane), we use the two RF electrodes to generate the RF quadrupole potential shown in fig. 1.3b. If we apply a voltage  $V_{\text{rf}}\cos(\Omega_{\text{rf}}t)$  to the RF electrodes, we obtain the following potential:

$$\phi_{\text{rf}}(x, y, z, t) = \alpha_{\text{rf}} V_{\text{rf}}(x^2 - z^2) \cos(\Omega_{\text{rf}}(t)) \quad (1.5)$$

where  $\alpha_{\text{rf}}$  is again a geometric factor which has units of  $[1/\text{distance}^2]$ . This potential can be visualized as a saddle which flips sign at the RF frequency, keeping the ion trapped at its centre turning point. Superposing the RF and DC potentials, we obtain the total potential seen by the ion:

$$\phi(x, y, z, t) = \alpha_{\text{rf}} V_{\text{rf}}(x^2 - z^2) \cos(\Omega_{\text{rf}} t) + \alpha_{\text{dc}} V_{\text{dc}} \left( y^2 - \frac{x^2 + z^2}{2} \right) \quad (1.6)$$

If  $\omega_{\text{axial}} \ll \Omega_{\text{rf}}$  we can, to a good approximation, ignore the effect of the anti-trapping DC potentials and turn our attention to how the RF potential determines the motion of the ion in the radial plane.

#### 1.4.2.1 Ion motion in the Paul trap

To derive the classical trajectory of the ion in a Paul trap, one can solve the relevant differential equations, the Mathieu equations. This treatment is carried out in several textbooks, such as [Gho95]. Here, we will summarize the results of a simplified approach taken in [ABLW08], whereby the motion of the ion is approximately calculated based on the pseudo-potential, a time-averaged version of the radio frequency potential produced by the trap RF electrodes.

#### 1.4.2.2 The pseudo-potential approximation

We approximate motion of the ion in the RF potential - which we will refer to as the radial motion from now on - as a superposition of a fast oscillation at the RF frequency, called micromotion, and a slow orbit, termed secular motion. The micromotion is simply written down as:

$$r(t) = -A_{\mu\text{m}} \cos(\Omega_{\text{rf}} t) \quad (1.7)$$

where  $A_{\mu m} = \frac{eE_0}{m\Omega_{\text{rf}}^2}$  is the amplitude of the motion and  $E_0$  is the amplitude of the electric field oscillation at the ion. The secular motion is calculated by first assuming that the ion motion is purely micromotion and then calculating the net force felt by the ion in the RF potential over one cycle of this micromotion. This net, time-averaged, force can be written as the gradient of static 2D harmonic well potential, the pseudo-potential, which is given by:

$$\phi_{\text{pp}} = \frac{e|E|^2}{4m\Omega_{\text{rf}}^2} = \frac{1}{2e}m\omega_{\text{sec}}^2(x^2 + z^2) \quad (1.8)$$

where  $|E|^2 = |-\nabla\phi_{\text{rf}}|^2$  and  $\omega_{\text{sec}} \approx \sqrt{2\frac{e\alpha_{\text{rf}}V_{\text{rf}}}{m\Omega_{\text{rf}}}}$ . In this approximation, the secular radial motion of the ion is simply a harmonic oscillation, at a frequency  $\omega_{\text{sec}}$ .

### 1.4.2.3 Trap stability

In order for the pseudopotential approximation to hold, we must have that  $\omega_{\text{sec}} \ll \Omega_{\text{rf}}$ . This is easy to see if we imagine the ion as a spherical mass in a finite saddle which flips at the RF frequency: if the saddle flips too slowly compared to the time it takes the mass to explore it (which is given by the frequency of its secular orbit,  $\omega_{\text{sec}}$ ), the mass will roll off the saddle, escaping confinement. When the Mathieu equations are solved to give the exact ion trajectory, the ratios of the secular and axial frequencies to the RF frequency are quantified in two dimensionless stability parameters,  $a$  and  $q$ . These stability parameters can be expressed in terms of the axial and secular frequencies as follows:

$$q = \frac{4e\alpha_{\text{rf}}V_{\text{rf}}}{m\Omega_{\text{rf}}^2} = 2\sqrt{2}\frac{\omega_{\text{sec}}}{\Omega_{\text{rf}}} \quad (1.9)$$

$$a = -\frac{4e\alpha_{\text{dc}}V_{\text{dc}}}{m\Omega_{\text{rf}}^2} = -2\frac{\omega_{\text{axial}}^2}{\Omega_{\text{rf}}^2} \quad (1.10)$$

The trap is only stable when  $a$  and  $q$  are within certain ranges; one may plot stability zones for the trap in the  $a$ - $q$  plane. The ion trap used in this thesis is

operated in what is called the first stability zone, where  $a$  and  $q$  are required to be less than 1.

### 1.4.2.4 Micromotion

Along the trap axis, the ion experiences no micromotion because this axis lies along the null of the RF potential. Once the ion is trapped, we Doppler cool its motion so that it only explores a very small area of the trap around the RF null and experiences no measurable micromotion - i.e. its radial motion is then well described by a harmonic oscillation at the secular frequency. However, stray electric fields can easily push the ion away from the RF null, leading to micromotion. Micromotion is generally considered an undesirable effect (except in cases where one purposefully induces it in order to, for instance, produce Doppler shifts which facilitate individual-ion addressing [Lei99, ANK<sup>+</sup>15]); this is due to the following:

1. Micromotion will introduce sidebands (at  $\pm n\Omega_{\text{rf}}$  from the carrier frequency, where  $n$  is an integer) to the spectra of any radiation impinging upon the ion, reducing the intensity in the carrier drive and, hence, the ion fluorescence on the transition being driven. This could lead to, for instance, slower Doppler cooling rates and single qubit gates, as well as off-resonant excitation if spectator transitions exist close to the sideband frequencies;
2. Micromotion should not heat the motion of a single trapped ion since it drives the ion at the RF frequency, which is much larger than the secular, or motional, frequency. However, parametric heating can occur if multiple ions are stored in the same trap [WMI<sup>+</sup>98].

One method of detecting micromotion utilizes the fact that it will modulate the fluorescence emitted by the ion in synchrony with the oscillation cycle of

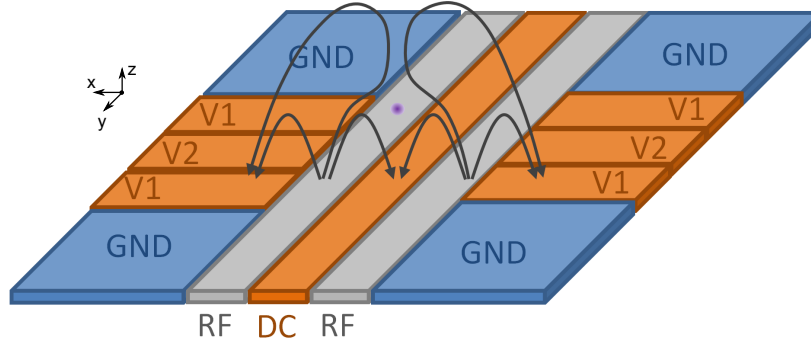


Figure 1.4: Surface trap version of the linear Paul trap depicted in fig. 1.3a in a “five-wire” geometry. DC electrodes are depicted in orange, RF electrodes in grey and ground plane in blue. The DC finger electrodes labelled with voltages  $V_1$  and  $V_2$  replace the two segmented DC rails in the macroscopic linear Paul trap. The RF field produced above the electrodes by the RF rails is outlined with black arrows. Just as in the macroscopic trap, the ion is trapped in the null of the RF quadrupole potential, indicated here by a purple dot.

the trap RF drive. By illuminating the ion with a laser beam along a direction  $\mathbf{k}$  and correlating the arrival time of photons emitted by the ion with the trap RF oscillation cycle, one can obtain a measure of the micromotion amplitude along  $\mathbf{k}$ . By applying shim fields to the trap’s segmented DC electrodes it is then possible to push the ion back towards the trap axis along  $\mathbf{k}$ . To completely eliminate radial micromotion, one must repeat this process for two orthogonal directions.

### 1.4.3 Surface-electrode ion traps

With a view to making ion-trapping technology more scalable, micro-fabrication techniques began to be employed to produce surface-electrode ion traps in the early 2000’s [CBB<sup>+</sup>05]. These traps are planar versions of the traditional Paul trap which can generate the same confinement potentials tens of microns above a set of microfabricated electrodes RF and DC electrodes. One configuration of surface electrodes which generates the required potentials - called the “five-wire geometry” - is depicted in fig. 1.4.

The potentials generated are significantly shallower than those produced by macroscopic traps (surface trap depths are on the order of meV, compared to

eV depths achieved with macroscopic traps), which means that ion lifetimes are typically reduced from weeks to several hours. However, surface trap lifetimes are still orders of magnitude longer than gate times (microseconds) and should not, therefore, significantly limit computational capabilities [BMJS16].

Micro-fabricated trap designs are easily scalable. The photo-lithography techniques employed to produce them have already been used to fabricate trap chips with of order 100 trap zones [AUW<sup>+</sup>10]. The small ion-electrode distances in surface traps also opens the door to the exploration of near-field microwave manipulations of ions: by integrating microwave electrodes into the trap structure, one can directly drive hyperfine transitions. The advantages of near-field microwave-driven ion-trap quantum computation are detailed in chapter 2 and the experiments in this thesis are dedicated towards the advancement of this implementation of quantum computation.

### 1.4.3.1 Heating rate in surface traps

One significant disadvantage of small ion-electrode distances is increased anomalous motional heating. This type of motional heating is postulated to be caused by fluctuating patch potentials on the surfaces of the electrodes. These potentials oscillate at MHz frequencies, producing electric fields which drive the ion's secular motion at a rate which drops off with the fourth power of the ion-electrode distance (assuming the patches are small in diameter compared to the ion-electrode distance) [TKK<sup>+</sup>00].

Motional heating reduces the fidelity of two-ion entangling gates (which rely on the excitation of collective motional modes to generate entanglement) and thus present a significant obstacle in the development of a fault-tolerant QC. Over the past decade, there has been some experimental success on the reduction of heating rates by ablating trap surfaces under vacuum with high-powered lasers [AGH<sup>+</sup>11] or focused-ion beams [HCW<sup>+</sup>12]. Such techniques aim to re-

duce organic surface contamination which, it has been hypothesized, may be source of fluctuating patch potentials. It has also been shown that traps operated at cryogenic temperatures generally have lower heating rates than those operated at room temperature [LGA<sup>+</sup>08].

Section 3.4.3 of this thesis describes a gold oven which can be used to evaporate a clean layer of gold onto the trap electrodes under vacuum, with the aim shielding the ion from any surface organic contamination - it is hoped that this may be a useful method of reducing anomalous heating.

## 1.5 Structure of this thesis

This thesis is structured as follows:

- Chapter 2 motivates the development of a near-field microwave-driven ion trap quantum computer based on the quantum CCD (quantum charge coupled device) architecture for surface-electrode ion trap chips. The chapter highlights single-qubit addressing as one of the main areas which require investigation for the realization of such a QC and outlines the approach this work takes to tackle it;
- Chapter 3 describes all the experimental apparatus used to perform the experiments discussed in this thesis. The chapter begins by introducing the Calcium-43 qubit and the optical and microwave transitions used to manipulate it and then proceeds to describe the apparatus used in this work including the laser setup, the vacuum system, trap chip fabrication methods and the trap DC, RF and microwave drive systems;
- Chapter 4 begins by discussing the design, simulation and packaging of MAT (Microwave Addressing Trap), the single-layer ion trap chip used in this work. The chip is a prototype two-zone surface ion trap which was used to implement high-fidelity single qubit addressing with near-field

microwaves. The chapter then presents experimental results and discusses future developments;

- Chapter 5 presents a new multi-layer design for a chip ion trap which incorporates the potential improvements identified after the experiments described in chapter 4 were performed with MAT. The chapter also presents simulations which indicate a 100-fold improvement in addressing performance is possible with this design, as compared to MAT;
- Chapter 6 introduces a near-field scanning system to be used to map microwave near-fields over ion-trap chips. The system comprises of a home-made microfabricated PCB antenna which is scanned over a device under test using a commercial translation stage. Until now, the microwave performance of ion trap chips has only been assessed by using the ion as a field probe once the chip is under vacuum - it is hoped that this characterization system will provide a simpler and more comprehensive way to test prototype ion trap chips, speeding up design-test-optimization cycles. This chapter describes the design and implementation of this characterization system and reports initial field-mapping results;
- Chapter 7 is the concluding chapter of the thesis and discusses the contribution this work has made to the field and the outlook for near-field microwave quantum computation.



# 2

## The near-field microwave quantum processor

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### 2.1 The Quantum CCD (QCCD) architecture

In 2002 D. Kielpinsky et al [KMW02] proposed the quantum CCD (quantum charge-coupled device) architecture for scaling up ion trap quantum computers to many-qubit processors. The architecture is represented schematically in fig. 2.1 and is based on a surface trap design where ions can be shuttled to different sections of a processor. Shuttling is implemented by applying time varying potentials to DC ‘finger’ electrodes which line the trap axis. Information is stored in short ion chains, which can be split apart for individual-ion addressing and brought back together for entangling operations. Junctions, where the trap axis forks or bends around corners, interconnect the 2D array of trapping zones. The QCCD architecture offers several attractive features:

1. Short ion chains are more robust than longer chains. Long ion chains are stored in traps with lower axial trap frequency for a given radial frequency. These tend to have higher heating rates, since lower energy quanta are required to promote the ions to higher energy motional modes. Further, as the number of ions in a single trap increases, so does the density of mo-

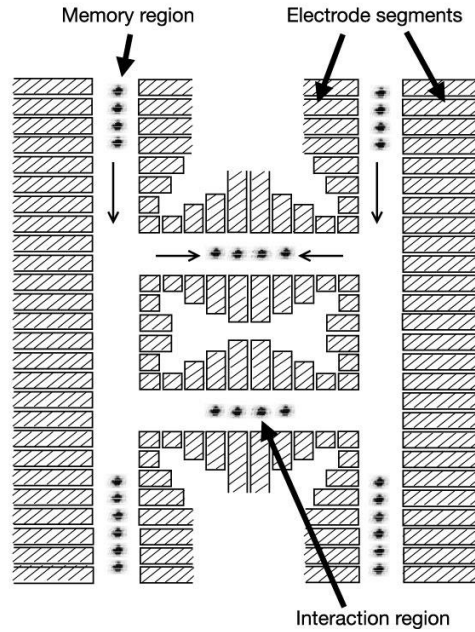


Figure 2.1: The QCCD architecture. Figure from [KMW02] (Reprinted by permission of Macmillan Publishers Ltd: Nature 417, 709–711, copyright 2002).

tional modes of the ion chain, making it much more difficult to address specific motional modes in order to perform entangling gates. This is owing to the fact that, if one wants to avoid zigzag arrangement, the centre-of-mass (COM) axial frequency falls as the number of ions increases;

2. There are no fundamental limitations which prevent scaling this architecture to arbitrary numbers of qubits - the challenges are all technical;
3. Several major aspects of the technical implementation of this architecture have already been demonstrated, i.e. a chip with hundreds of trapping zones can be easily fabricated using standard photolithographic techniques [AUW<sup>+</sup>10], ion shuttling without significant heating has already been shown (and can be done on a similar time scale as a two-qubit gate) [WZR<sup>+</sup>12] and the fundamental single-qubit and two-qubit operations required for quantum computation have been performed with ions in prototype surface traps with one or two trapping zones [BWC<sup>+</sup>11,

OWC<sup>+</sup>11, ACLS<sup>+</sup>17, HAB<sup>+</sup>14, HSA<sup>+</sup>16].

In this thesis, we will adopt the QCCD architecture as a backbone structure to guide the development of scalable prototype ion-trap quantum computing chips. This section has reviewed the architecture's merits; the next section will consider its vulnerabilities and discuss how using near-field microwave radiation to drive qubit operations can produce a more robust QCCD processor.

## 2.2 Microwave QCCD

When the QCCD architecture was first proposed, quantum information operations on trapped ions were mostly done using laser light. Thus, it is unsurprising that the paper postulates driving qubit operations using lasers. In particular, the authors envision the use of an optical-frequency qubit transition (giving, as an example choice, the  $S_{1/2} \leftrightarrow D_{5/2}$  quadrupole transition in  $^{40}\text{Ca}^+$ ) and discuss the technical challenges associated with implementing the QCCD architecture with optical qubits. In this section, we review these challenges and extend the discussion to consider laser-driven Raman transitions between ground state hyperfine levels. We will see that many of the technical challenges associated with laser-driven implementations can be entirely avoided if we instead use near-field microwave radiation to directly drive hyperfine qubits.

### 2.2.1 Integrating the driving field source into the ion trap chip

Driving transitions with lasers would require integrating optical elements into the trap structure. Progress has been made towards developing this technology [MBM<sup>+</sup>16], but significant challenges still remain, particularly for blue and UV wavelengths, which are required to drive Raman transitions in most alkaline earth metal ions [BHL<sup>+</sup>16]. In contrast, microwave electrodes can be straightforwardly incorporated into the trap structure [OWC<sup>+</sup>11, ACLH<sup>+</sup>14, ACLS<sup>+</sup>17,

AHB<sup>+</sup>13]. Transitioning from lasers to microwaves would replace the technically daunting problem of bringing hundreds of highly stabilised laser beams onto a surface trap by the more tractable issue of multiplying up the number of microwave lines feeding the chip.

### 2.2.2 De-phasing during shuttling

One of the main error sources discussed in the QCCD proposal paper is the de-phasing of the qubit during shuttling. Small spatial variations of the static magnetic field will change the energy splitting between qubit states via the Zeeman effect (whether at first or higher order) and cause the qubit to pick up a phase. The authors concede that calibrating this phase over paths across the entire chip would be a very demanding task and propose instead to encode information into a decoherence-free subspace of two ions. Dephasing would then be heavily suppressed (as it would now only occur as a result of small differences in the B-field experienced by each ion as they are both shuttled through the chip), but at the cost of an added layer of complexity.

The qubits we will consider for use in the near-field microwave QCCD are immune to this kind of dephasing at first order. They are so-called ‘clock’ qubits: at a certain value of the static magnetic field,  $B_0$ , the Zeeman shifts of both upper and lower qubit states are the same for small excursions away from  $B_0$ . At  $B_0$ , the qubit transition frequency is hence first-order insensitive to small changes in the magnetic field. These qubits have  $T_2^*$  times on the order of tens of seconds [HAB<sup>+</sup>14] and essentially infinite  $T_1$  times.

It is possible to drive such qubits using pairs of Raman laser beams, but this technique will suffer from photon scattering errors, which present a fundamental limit to gate fidelity [OIB<sup>+</sup>07] and from other technical problems discussed in this chapter.

### 2.2.3 Phase stability in logic gate sequences

Another source of decoherence in the optical-qubit QCCD arises from the fact that, in order to retain phase coherence during logic operations, we must position the ions in the laser beam with sub-wavelength spatial accuracy (i.e. to an accuracy of tens to hundreds of nm). In [KMW02], the authors note that while it is possible to stabilize the voltage sources driving the DC electrodes so that this is achievable, mechanical vibrations of the chip and stray electric fields can easily push the ion off-target by a fraction of an optical wavelength. The optical path length leading up to the ion must also be stabilized to the same degree, placing tough restrictions on the mechanical stability of all optical components from the laser source to the ion.

We can avoid this problem in a hyperfine-qubit-based architecture by either using co-propagating Raman laser beams to drive the qubit (as phase fluctuations which are common-mode to both beams do not affect the qubit drive) or by using microwave driving fields, which have wavelengths in the of tens of cm. For a given level of path-length oscillation (i.e. for a given level of thermal, mechanical and electrical noise), the phase stability achievable with microwaves is a factor  $10^5$  better than that which is attainable using lasers. That is, the microwave near-field generated by a given electrode at two points within any area of the chip one could reasonably hope to explore with a trapped ion are in phase.

### 2.2.4 Amplitude and polarisation stability of the driving field

Mechanical and thermal fluctuations on the optical beam path will lead to less light being coupled into the chip's integrated fibers, causing amplitude fluctuations at the ion (often termed 'beam pointing' error). Changes in the dielectric constant of fiber cladding due to thermal fluctuations will also lead to polarisation noise.

Coupling to microwave waveguides is not affected by these fluctuations

since the wave travels through a hardwired path all the way up to the ion: there is little opportunity for thermal or mechanical fluctuations to change the angle of the  $k$  vector of any given wavefront. The polarisation of the microwave field produced by a given trap electrode will depend only on the geometry of the path the microwave current takes along that electrode, which is registered to the trap surface. This is again largely insensitive to thermal or mechanical noise.

Transient thermal effects on active microwave devices, such as amplifiers and switches, will, however produce some amplitude noise.

### 2.3 Individual-ion addressing in the QCCD architecture

In the laser-driven QCCD, high-fidelity addressing will require separating the ions into small chains of order two ions long. The diffraction limit prevents a laser beam from being focused to a sub-wavelength waist and crosstalk will therefore be large if many ions are in the same trap. Even for a two-ion chain, with ion-ion separation set at  $7.4\mu\text{m}$ , crosstalk experimentally achieved with laser addressing is at the  $10^{-4}$  level [CMBK14]. Separating long ion chains into chains of one or two ions will mean pushing the individual trap minima apart by a distance of at least the same order of magnitude as the distance between the ion and the trap surface ( $\sim 100\mu\text{m}$ ). Although holding the ions closer together than an ion-surface distance is fundamentally possible, generating the DC potentials required to maintain separate deep traps whilst holding the ions this close together would be very technically challenging (it would require reducing the width of the DC electrodes to around  $10\mu\text{m}$  and applying large, stable, voltage differentials to neighbouring electrodes). It is therefore reasonable to design prototype QCCD systems where ions have to be separated by of order a few hundred microns in order to be individually addressed with high fidelity.

The question which arises now is whether it is possible to perform high fidelity individual-ion addressing in the microwave-driven QCCD. With far-field

microwaves, spatial addressing on the hundred micron length scale is not possible since the radiation wavelength is on the order of 10cm (addressing in frequency space can still be done with high fidelity [PSVW14], but typically requires the application of a static magnetic field gradient and the use of field-sensitive qubit transitions which have naturally shorter coherence times than clock transitions).

Fortunately, our ions are in the near field of the microwave electrodes, at a distance of one thousandth of a wavelength from the microwave source. Hence, over the entire area of the chip that is reasonably explorable with one ion, the microwave field oscillates in phase and the shape of the field amplitude envelopes depend on the geometry of the microwave source (the integrated microwave electrodes) and the surrounding ground plane and trap electrodes, not on the microwave wavelength. In fact, if we were to freeze time at the point where the amplitude of the microwave current being fed to an isolated waveguide were at a maximum, the microwave near fields generated by the centre conductor would look the same as the field which would be generated by DC current of the same amplitude. However, there is one crucial difference between the near-field landscape for static and microwave currents: a microwave current flowing through a metal wire will generate a time varying B field which will induce return currents on any surrounding metal which it permeates, unlike a static DC current.

This effect is both a blessing and a curse in terms of crosstalk. The blessing ensues from the fact that the field produced by a microwave current-carrying electrode will generally decay in amplitude faster than that produced by an electrode carrying DC current if appropriate sections of ground plane are placed immediately around it. This is because opposing return currents will be induced in the surrounding ground plane and these currents will partially cancel the field as we move away from the microwave electrode.

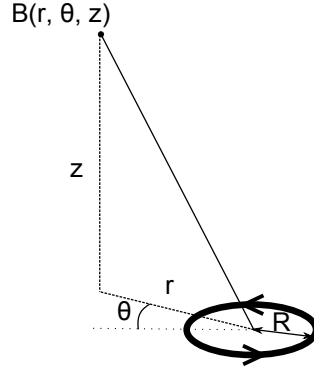


Figure 2.2: DC current loop in cylindrical coordinate system.

However, any induced microwave return currents make microwave near-fields generated by anything but the simplest waveguides highly non-trivial to calculate: the fields must be numerically computed using software which solves Maxwell's equations over a mesh of points over the microwave device. The computation is intensive and field solutions can be disconcertingly sensitive to small changes in electrode geometry. Hence, great care must be taken to make robust designs with well known impedances, or stray return currents coupling between zones of the chip could work to increase crosstalk, rather than attenuate it.

If we imagine designing the microwave electrodes as small current loops, we can obtain a rough idea of how quickly the field will decay between trap zones by looking at how the field from a DC current loop decays. Considering the cylindrical coordinate system depicted in fig. 2.2, the field components at a point  $(r, \theta, z)$  are given by:

$$B_r(r, \theta, z) = \frac{\mu_0 I_z}{2\pi r} \frac{1}{\sqrt{(R+r)^2 + z^2}} \left\{ -K(k) + \frac{R^2 + r^2 + z^2}{(R-r)^2 + z^2} E(k) \right\} \quad (2.1)$$

$$B_z(r, \theta, z) = \frac{\mu_0 I_z}{2\pi} \frac{1}{\sqrt{(R+r)^2 + z^2}} \left\{ -K(k) + \frac{R^2 - r^2 - z^2}{(R-r)^2 + z^2} E(k) \right\} \quad (2.2)$$

where  $k = \sqrt{4rR(z^2 + (r+R)^2)^{-1}}$ ,  $R$  is the radius of the current loop,  $K(k)$  is the complete elliptic integral of the first kind and  $E(k)$  is the complete elliptic



integral of the second kind. Supposing the static B-field is aligned parallel to the trap surface and we adopt a  $\pi$ -polarised qubit transition, we are most interested in the component of the field in the plane parallel to the trap surface,  $B_r$ . Let us set the radius,  $R$  of our microwave loop electrode to be approximately one ion-surface distance, approximately  $100\mu\text{m}$ . Then the ion height above the trap is  $R$  and we are therefore most interested in the field at  $z = R$ . Let's say we will space our trap zones out by a distance of five loop radii,  $5R$ . Let us compare the field at a distance of one radius  $R$  away from the loop electrode along the trap axis,  $B_r(R, \theta, R)$  to that in the neighbouring trap zone, five radii away  $B_r(5R, \theta, R)$ . For simplicity, let's define  $B_0 = \frac{\mu_0 I_z}{2\pi}$  and drop the  $\theta$  argument since it does not enter in any of our expressions, as the field is cylindrically symmetrical:

$$B_r(R, R) = \frac{B_0}{\sqrt{5}R^2} \left\{ -K(\sqrt{0.8}) + 3E(\sqrt{0.8}) \right\} \quad (2.3)$$

$$B_r(5R, R) = \frac{B_0}{\sqrt{37}R^2} \left\{ -K(\sqrt{20/37}) + \frac{7}{17}E(\sqrt{20/37}) \right\} \quad (2.4)$$

$$\therefore \left| \frac{B_r(R, R)}{B_r(5R, R)} \right| \approx 46 \quad (2.5)$$

Based on this rough calculation, it is therefore reasonable to expect that the field for a microwave loop electrode will decay by a factor of  $\sim 10$  to  $50$  between trap zones. This will not be enough to perform high-fidelity addressing. The ratio could be made better by designing return current paths which aid in field cancellation, but it could also be made worse if these stray return currents couple to neighbour zones.

One way to address this problem is to very carefully construct the geometry of the trap electrodes with an aim to guide the return currents generated by the microwave electrodes so as to make full use of their canceling effect. Unfortunately, the electrode geometry is heavily constrained by trapping parameters and by fabrication capabilities. For instance, the trapping RF electrodes must run uninterrupted along the trap axis so, in a single-layer fabrication process, it

is only feasible to place microwave electrodes outside the RF rail (as in fig. 2.3). In such a configuration, the microwave source is not directly below the ion and will hence produce a smaller field differential between zones. To maintain shuttling capabilities along the entire trap axis, we must place DC electrodes along the entire length of the trap axis, a requirement which prevents the insertion of low inductance ground plane segments which could help to isolate trap zones and better guide return currents. We must also keep the amount of exposed dielectric on the trap surface to a minimum in order to avoid charging effects which could heat the ions - this is yet another restriction which hampers our ability to design return current paths.

Multi-layer techniques will relax some of these constraints - and chapter 5 of this thesis presents a low-crosstalk trap design which could be fabricated using two such techniques - but we would like to adopt a robust scheme which allows us to do high-fidelity addressing even when optimum return current routing is not possible. One such scheme would involve actively canceling out residual crosstalk in each trap zone by generating fields in that zone which are equal and opposite to the crosstalk fields. The next section discusses this scheme in detail.

### **2.4 The nulling scheme for near-field microwave addressing**

There will be four integrated microwave electrodes per trap zone. This gives us enough degrees of freedom to completely control the field amplitude and polarisation in three dimensions in each zone (in fact, only three electrodes would be required for this level of control - one for each spatial dimension - but we will choose to use four in order to keep the trap design symmetric). Each electrode will produce a field in its resident zone and a crosstalk field in neighbour zones. By choosing the appropriate drive currents to be fed to each electrode, it should

be possible to cancel out unwanted crosstalk fields and apply to each zone, independently and in parallel, fields of any desired amplitude, polarisation and phase.

Consider an array of  $N$  zones, each with four integrated microwave electrodes. We can relate the  $x$ -component of the B-field produced in each zone to the currents applied to the  $4N$  electrodes on the array by the following equation:

$$\begin{pmatrix} \mathbf{B}_1^x \\ \mathbf{B}_2^x \\ \vdots \\ \mathbf{B}_N^x \end{pmatrix} = \mathbf{X} \times \begin{pmatrix} \mathbf{I}_1 \\ \mathbf{I}_2 \\ \vdots \\ \mathbf{I}_{4N} \end{pmatrix} \quad (2.6)$$

and similarly for the  $y$  and  $z$  field components. Here,  $\mathbf{I}_j = I_j e^{i\phi_j}$  denotes the current applied to electrode  $j$  and  $\mathbf{B}_n^x = B_n^x e^{i\phi_n^x}$  denotes the  $x$ -component of the B field generated in zone  $n$  when a set of currents  $\{\mathbf{I}_1, \dots, \mathbf{I}_{4N}\}$  is applied to the microwave electrodes.  $\mathbf{X}$  is the  $N \times 4N$  coupling matrix whose elements describe the  $x$ -components of the field generated by each microwave electrode in each zone. Specifically,  $\mathbf{X}_{i,j} = X_{i,j} e^{i\theta_{i,j}}$  gives the  $x$ -component of the B-field generated in zone  $i$  when electrode  $j$  is powered with a unit microwave current. Similarly, we denote the coupling matrices describing the  $y$  and  $z$ -components of the field as  $\mathbf{Y}$  and  $\mathbf{Z}$ .

The complex elements of  $\mathbf{X}$  can be determined experimentally, as follows:

- The magnitude  $X_{i,j}$  of each element can be determined by feeding a unit current to electrode  $j$  and, whilst keeping all other electrodes grounded, measuring the Rabi frequency for an ion trapped at zone  $i$  on a transition that couples to the  $x$ -component of the microwave magnetic field. For example, if one aligns the static B-field - and hence the quantisation axis - along  $x$ , one would measure the Rabi frequency on a  $\pi$ -polarised transition of an ion in zone  $i$  driven by electrode  $j$  in order to obtain a measure of  $X_{i,j}$ .

- The phase  $\theta_{i,j}$  of each element can be determined by driving electrodes pairwise on a transition that couples to the  $x$ -component of the microwave field so as to establish the dependence of the transition Rabi frequency on the relative phase between the electrodes (i.e. a maximum Rabi frequency will be achieved when the  $x$ -components of the fields generated by the two electrodes are in phase and the minimum Rabi frequency will occur when the two electrodes' driving fields are out of phase). More generally,  $X$  can be determined by any sufficiently large and linearly independent set of measurements.

Once we have measured all elements of the coupling matrices  $\mathbf{X}$ ,  $\mathbf{Y}$  and  $\mathbf{Z}$ , we can collate them into a single  $3N \times 4N$  matrix  $\mathbf{M}$ , which describes the coupling between all electrodes and all field components, as follows:

$$\mathbf{M} = \begin{pmatrix} \mathbf{X}_{1,1} & \mathbf{X}_{1,2} & \dots & \mathbf{X}_{1,N} \\ \mathbf{Y}_{1,1} & \mathbf{Y}_{1,2} & \dots & \mathbf{Y}_{1,N} \\ \mathbf{Z}_{1,1} & \mathbf{Z}_{1,2} & \dots & \mathbf{Z}_{1,N} \\ \vdots & \vdots & \vdots & \vdots \\ \mathbf{X}_{N,4N} & \mathbf{X}_{N,4N} & \dots & \mathbf{X}_{N,4N} \\ \mathbf{Y}_{N,4N} & \mathbf{Y}_{N,4N} & \dots & \mathbf{Y}_{N,4N} \\ \mathbf{Z}_{N,4N} & \mathbf{Z}_{N,4N} & \dots & \mathbf{Z}_{N,4N} \end{pmatrix} \quad (2.7)$$

so that

$$\mathbf{B} = \begin{pmatrix} \mathbf{B}_1^x \\ \mathbf{B}_1^y \\ \mathbf{B}_1^z \\ \vdots \\ \mathbf{B}_N^x \\ \mathbf{B}_N^y \\ \mathbf{B}_N^z \end{pmatrix} = \mathbf{M} \times \begin{pmatrix} \mathbf{I}_1 \\ \mathbf{I}_2 \\ \vdots \\ \mathbf{I}_{4N} \end{pmatrix} = \mathbf{M} \times \mathbf{I} \quad (2.8)$$

We can now calculate the currents which should be applied to each electrode in order to generate the desired fields at each zone by inverting this matrix equation. Recall that we are using  $4N$  currents to define  $3N$  field components, so the problem is underconstrained, i.e. we can find infinitely many current vectors  $\mathbf{I}$  which produce the desired field vector  $\mathbf{B}$ . We can, for instance, choose the particular solution given by:

$$\mathbf{I} = \text{pinv}(\mathbf{M}) \cdot \mathbf{B} \quad (2.9)$$

where  $\text{pinv}(\mathbf{M})$  is the Moore-Penrose pseudo inverse of  $\mathbf{M}$ : this solution minimises  $\|\mathbf{I}\|$  and hence the power consumption of the chip for a given set of desired fields.

Fig. 2.3 schematically depicts the scheme under operation for two ions: one ion undergoes a single-qubit rotation driven with a  $\pi$ -polarised microwave field, whilst the identity operator is applied to its neighbor ion (i.e. the microwave field is nulled at the neighbour ion). This addressing scheme will, in principle, give us the ability to perform arbitrary single-qubit operations independently and in parallel on all ions on the chip.

## 2.5 Scaling up the nulling scheme

We now consider both how phase and amplitude errors in the currents being fed to the electrodes propagate into errors in the field seen by the ions, and how these errors scale as we increase the number of trap zones.

If the fluctuations in current fed to each electrode are uncorrelated, one may estimate that the total error in the microwave B-field produced in a given zone will be the quadrature sum of a constant B-field error,  $\epsilon_0$ , generated by each electrode. If we have  $N$  zones in our processor, the total error will then scale as  $\sqrt{N}$  and the more zones we add, the larger the error we will incur.

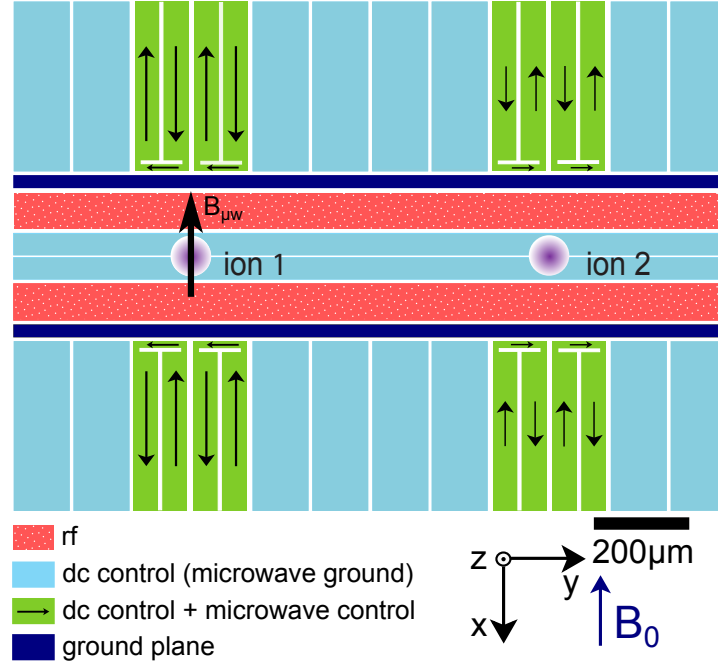


Figure 2.3: Individual-ion addressing using the nulling scheme in a two-zone processor. A set of currents are applied to the eight microwave electrodes on the processor so that the field is nulled at ion 2 and a  $\pi$ -polarised microwave field is generated at ion 1. Hence, we drive a single qubit rotation on ion 1 whilst applying the identity operator to ion 2.

This analysis is slightly naive, however, since it does not take into account that the effect of any microwave electrode on a given trap zone decays with distance. Suppose that a microwave electrode produces a field of amplitude  $\epsilon_0$  in its resident zone. Let us assume that the field obeys a power-law decay of the form  $\frac{1}{r^k}$ , where  $r$  is the distance from the source microwave electrode. Then, the field error produced by a given microwave electrode at a zone a distance  $r$  away is  $\frac{\epsilon_0}{r^k}$ . Consider a zone  $i$  in a large 2D array of trap zones spaced by a unit distance. The number of zones forming a perimeter at a distance  $R$  from zone  $i$  will be proportional to  $R$  and will contribute a maximum crosstalk field of  $\chi_R = R \times \frac{\epsilon_0}{R^k} = \frac{\epsilon_0}{R^{k-1}}$  at zone  $i$ . By summing over all distances  $R$ , we obtain an upper limit on the crosstalk at zone  $i$  of  $\chi_{\text{tot}} = \sum_{R=1}^{R=\infty} \chi_R = \sum_{R=0}^{R=\infty} \frac{\epsilon_0}{R^{k-1}}$ , which converges for  $k > 2$ .

As discussed in section 2.3, the fields produced by a DC current element

would decay as  $\frac{1}{r^2}$  and we can expect a faster decay from microwave fields if we make use of the canceling effect of return currents. Hence, with appropriate trap electrode design, it should be possible to obtain  $k > 2$ . This means that the fields in any given zone would only be affected by electrodes within a finite perimeter around that zone: beyond that perimeter, more zones can be added without an increase in field error.

Finally, we turn our attention to the feasibility of performing the calibration measurements required to determine the matrix  $\mathbf{M}$  in eq. 2.7 as we scale up the number of zones in our array. If  $m$  calibration measurements are required to calibrate the effect of the electrodes in a given zone on an ion trapped in zone  $z$ , then we will need to perform  $m \cdot N$  calibration measurements for that zone, where  $N$  is, as usual, the number of zones in the processor. Since we have to perform this calibration routine for all zones, the total number of calibration measurements required will be given by  $m \cdot N^2$  - an unfavourable quadratic scaling.

However, if we are able to read out all  $N$  zones of our array simultaneously, the measurements can be performed in parallel, which means that the number of independent calibration measurements required is reduced to  $m \cdot N$  and now scales linearly with the number of trap zones. Our group has previously performed simultaneous readout of many ions by imaging ion fluorescence with a camera [MSW<sup>+</sup>08, BSWL10]; other parallel readout schemes involving integrated detectors can also be implemented, as in [DGK<sup>+</sup>14, VCA<sup>+</sup>10, EWA<sup>+</sup>13].

It seems, therefore, that the nulling scheme for near-field microwave addressing is scalable to many-ion processors. In the next chapters, we pursue a prototype implementation of this scheme with the aim to develop designs amenable to be scaled up into many-qubit microwave-driven QCCD processors.





# 3

## Experimental apparatus

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This chapter describes the experimental apparatus used to perform the experiments reported in this thesis. Section 3.1 describes our chosen qubit transition in the  $^{43}\text{Ca}^+$  ion and the techniques used to photoionise and Doppler cool the ion and to perform qubit state preparation and readout. Laser beam paths and other aspects of the optical setup are outlined in section 3.2, the procedure used to fabricate the chip ion-trap used in the experiments is detailed in section 3.3 and the vacuum system that houses it is described in section 3.4. Finally, sections 3.5, 3.6 and 3.7 discuss, respectively, the systems used to drive the DC, RF and microwave supplies to the trap.

### 3.1 A hyperfine qubit in $^{43}\text{Ca}^+$

Our ion of choice is  $^{43}\text{Ca}^+$ . In this work, we use a qubit formed by the two  $M_f = 0$  Zeeman sub-levels of the  $F = 3$  and  $F = 4$  hyperfine manifolds in the  $S_{1/2}$  ground state (the qubit transition is indicated by the green line in fig. 3.1). This qubit is first-order insensitive to magnetic field fluctuations at low static fields (here, our static magnetic field is set to  $B = 2.8\text{G}$ ). Such qubits are commonly referred to as ‘clock’ qubits and offer long coherence times: the coherence time for this qubit at a static field of  $B = 2.0\text{G}$  was measured elsewhere to be  $T_2^* = 6(1)\text{s}$  [BHL<sup>+</sup>16]. Fig. 3.1 shows the transitions we use to manipulate the qubit.

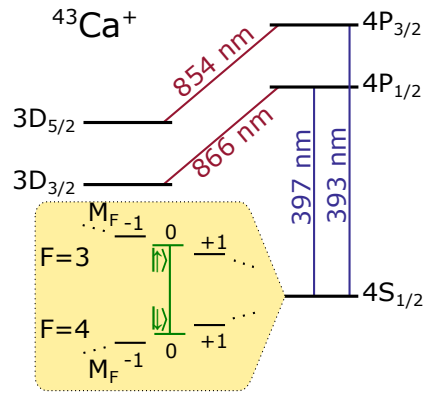


Figure 3.1: Level structure of the  $^{43}\text{Ca}^+$  ion indicating the laser frequencies used for qubit manipulation and Doppler cooling. Inset: a detailed view of the  $F = 3$  and  $F = 4$  hyperfine levels of  $^2S_{1/2}$ , showing the Zeeman states around  $|\uparrow\rangle$  and  $|\downarrow\rangle$ , with the qubit transition highlighted in green.

The following subsections describe how the Calcium atom is ionised and the schemes used for qubit initialisation and read-out.

### 3.1.1 Photoionisation

Singly-ionised Calcium is obtained in this experiment by photoionisation. A beam of neutral Calcium atoms is produced by an oven (see section 3.4.2) placed inside the vacuum system which houses the ion trap. Fig. 3.2 shows the two transitions we use to photoionise Calcium atoms. The atom is first excited from the  $4^1S_0$  ground state to  $4^1P_1$  by a 423nm laser. To ionise the atom, one electron is then promoted from this state to the continuum by a 378nm laser. The two photoionisation lasers beams are aligned at  $90^\circ$  to the atomic beam, in order to avoid Doppler shifting the transition frequencies.

### 3.1.2 Doppler Cooling

Once an ion is captured in the trap, its motion must be slowed in order to reduce the probability that it will gain enough kinetic energy to leave the trap via a collision with background gas or due to anomalous heating caused by proximity to the surface of the trap electrodes. We Doppler cool the ion's motion by using

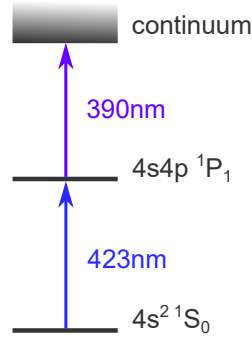


Figure 3.2: Level diagram showing the two transitions we use to photoionise neutral Calcium. The excitation to the continuum requires light of wavelength shorter than 389.81nm; we currently use a 378nm diode laser for this step.

laser light at 397nm to scatter photons on the  $S_{1/2} \leftrightarrow P_{1/2}$  transition, as illustrated in fig. 3.1. An Electro Optical Modulator (EOM) is used to put 3.2GHz sidebands on the 397 light so that population in both the  $F=3$  and  $F=4$  manifolds can be excited to the  $F=3$  manifold in  $P_{1/2}$ . An infrared repumper laser at 866nm is also required to close the Doppler cooling cycle since there is a 5% chance that population in  $P_{1/2}$  will decay to the long-lived  $D_{5/2}$  state.

### 3.1.3 State preparation

We prepare the  $|\downarrow\rangle$  qubit state by optically pumping on the Doppler cooling transition using a  $\pi$ -polarised 397nm laser beam and the 866nm repumper beam. The selection rule that forbids  $M_F = 0 \leftrightarrow M_{F'} = 0$  when  $\Delta F = 0$  causes population to build up in  $F = 4, M_F = 0$ , the lower qubit state, as indicated in fig. 3.3. State preparation fidelities using this scheme are limited to  $\leq 99\%$  due to off-resonant excitation of population in the  $S_{1/2}$  hyperfine states to the  $F = 3$  manifold of  $P_{1/2}$ : the  $S_{1/2}^{4,0} \leftrightarrow P_{1/2}^{3,0}$  transition is not forbidden by the aforementioned selection rule and will hence off-resonantly depump population from  $|\downarrow\rangle$ . Our group has shown that preparation of the clock qubit can be performed with fidelities of  $\gtrsim 99.8\%$  if population is instead pumped to the  $|F = 4, M_F = +4\rangle$  stretch state and transferred to  $|\downarrow\rangle = |F = 4, M_F = 0\rangle$  using

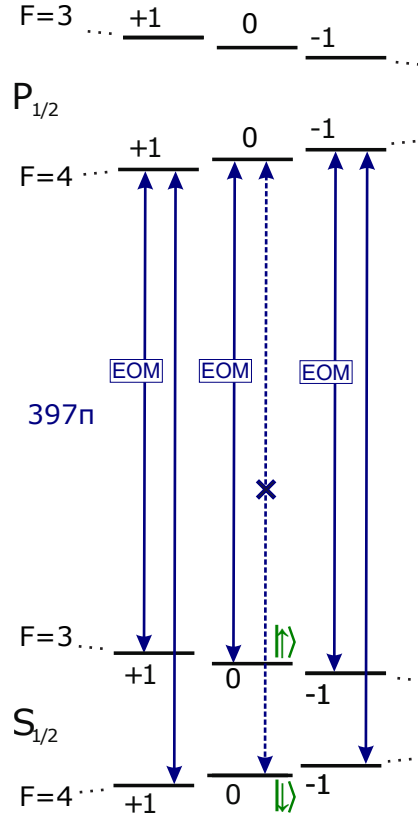


Figure 3.3: Preparation of the  $|\downarrow\rangle$  qubit state using a  $\pi$ -polarised 397nm beam. An EOM ensures that the beam addresses both  $F=3$  and  $F=4$  hyperfine levels of the ground state and repumps population from all of the  $M_F$  levels in  $S_{1/2}$  to their counterparts in  $P_{3/2}$ , except for population in  $|\downarrow\rangle = |F=4, M_F=0\rangle$ , which is not repumped due to the selection rule which forbids  $\Delta M_F = 0$  when  $\Delta F = 0$ . This configuration hence pumps population to  $|\downarrow\rangle$ . Figure modified from [Lin12].

microwave pulses [HAB<sup>+</sup>14].

### 3.1.4 Readout

Readout is performed by first using a  $\sigma^+$ -polarised 393nm pulse to selectively transfer any population in  $|\downarrow\rangle$  to the long lived  $D_{5/2}$  “shelf” via  $P_{3/2}$  (see fig. 3.4). The qubit state is then determined by using the Doppler cooling lasers to look for fluorescence on the  $S_{1/2}^{4,0} \leftrightarrow P_{1/2}^{3,0}$  transition: the population transferred from  $|\downarrow\rangle$  to the shelf will not fluoresce, whilst the population left behind in  $|\uparrow\rangle$  will.

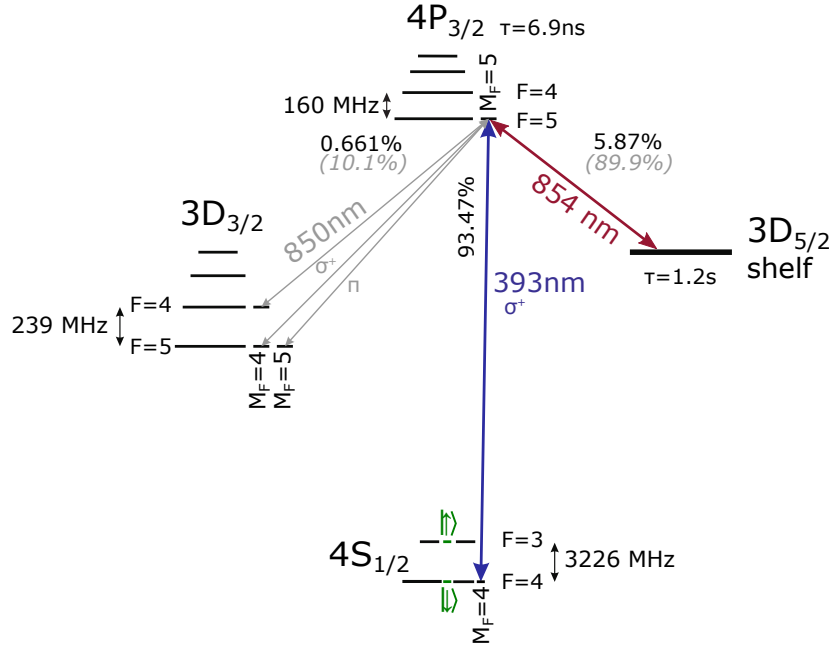


Figure 3.4: Shelving scheme for readout. A  $\sigma^+$ -polarised 393nm beam first pumps any population in  $|\downarrow\rangle$  to the  $|F = 4, M_F = +4\rangle$  state in the  $4S_{1/2}$  ground state, henceforth referred to as the ‘stretch+’ state. The 393 beam selectively excites population to the  $F=5$  level in  $4P_{3/2}$ ; off-resonant excitation to  $F=4$  is strongly suppressed due to the 160MHz frequency splitting between  $F=4$  and  $F=5$ . The population decays predominantly back to  $F=4$  in the ground state (with 93.47% probability). From the stretch+ state, the 393 $\sigma^+$  beam can only excite population to  $|F = 5, M_F = +5\rangle$  in  $4P_{3/2}$  (assuming no polarisation impurity). From there, it will decay to the long-lived  $3D_{5/2}$  ‘shelf’ with 5.87% probability, to  $3D_{3/2}$  with 0.661% probability, or back to the stretch+ state with 93.47% probability. However, the 393 $\sigma^+$  constantly repumps any population in the stretch+ back to  $|F = 5, M_F = +5\rangle$  in  $4P_{3/2}$ , so the relevant branching ratios from this state are the relative probabilities of decay to the  $3D_{5/2}$  shelf and to  $3D_{3/2}$ : 89.9% and 10.1% respectively, as indicated in italicised parenthesis in the figure. The latter unwanted decay can be addressed by using  $\pi$  and  $\sigma^+$ -polarised 850 beams to repump population from  $3D_{3/2}$  back to  $|F = 5, M_F = +5\rangle$ . These beams were not used in the experiments performed here since high-fidelity readout was not required. Readout fidelity was hence limited to  $\sim 90\%$ . Fidelity can be further improved by using microwave pulses to transfer population from  $|\downarrow\rangle$  to the stretch+ state, rather than relying on optical pumping with the 393, which is subject to errors due to unwanted decay of  $|\downarrow\rangle$  population to  $|\uparrow\rangle$  via the D-states [HAB<sup>+</sup>14, MSW<sup>+</sup>08]. Figure modified from [Szw09].

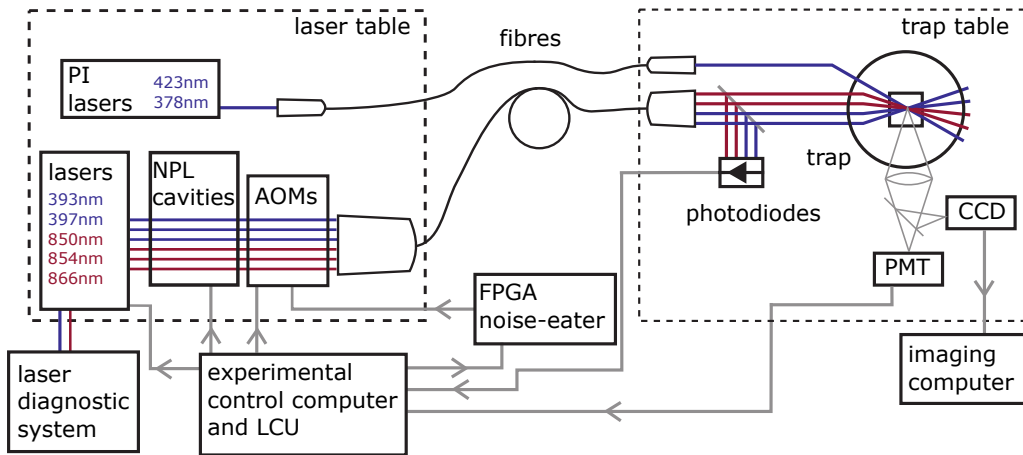


Figure 3.5: Schematic showing optical setup. Figure modified from [All11].

## 3.2 Optical setup

Fig. 3.5 summarises the optical setup for this experiment. There are three optical tables in the lab which we refer to as the two ‘laser’ tables and the ‘experiment’ table. For simplicity, we will refer to the two laser tables collectively as the ‘laser table’. The laser table carries all of the diode lasers and any optics required for laser frequency stabilisation and diagnostics. Optical fibres take the laser light from the laser table to the experiment table, where the beam paths required for two experiments (the experiment described in this thesis and one other) are set up around the respective vacuum systems. The sections which follow describe the optics and electronics on each table in more detail.

### 3.2.1 Laser table

We use a total of seven diode lasers to manipulate the Calcium-43 ion: four at ultraviolet (UV) wavelengths (378nm, 393nm, 397nm and 423nm) and three at infra-red wavelengths (850nm, 854nm and 866nm) - though the 850nm laser is not used in the experiments described in this thesis. The following subsections discuss the elements of the beam paths, the diode lasers used in this experiment, laser frequency stabilisation methods and our laser diagnostic system. These

laser systems were set up over several years by previous and current students in the group (see, for example, [Szw09, All11, Sep16]) and are described here for reference.

### 3.2.1.1 Beam path elements

Fig. 3.6 depicts typical beam paths for the UV and infrared lasers. An optical isolator is placed directly after the laser to reduce optical feedback. To prevent amplified spontaneous emission (ASE) at wavelengths close to the laser wavelength from reaching the experiment, a blazed diffraction grating is placed in each beam path: 393nm ASE from the 397nm laser could, for instance, shelve the ion if the grating were not present and 854nm ASE from the 850nm laser could de-shelve it.

To switch the beams on and off, acousto optic modulators (AOMs) are used. An AOM crystal diffracts a beam off sound waves produced by applying a radio frequency (RF) drive to the piezo-electric transducer on which it sits. The AOM can be switched on and off by TTL signals issued by the experimental control computer. When the AOM is on, the first order diffracted beam reaches the experiment. When it is off, RF leakage can allow a small amount of light to still be diffracted. Extinction is improved in two ways: firstly, a DC signal is applied to the frequency modulation (FM) input of the AOM when it is off [Szw09]. This detunes any leaked RF by a few MHz, which in turn causes any light diffracted in the first order beam to be detuned by the same amount. Secondly, the AOMs are used in a double-pass configuration, which means that the beam must be diffracted by an AOM twice before it can reach the experiment. The double-pass configuration also prevents the final output beam direction from changing when the AOM frequency is changed.

### 3. EXPERIMENTAL APPARATUS

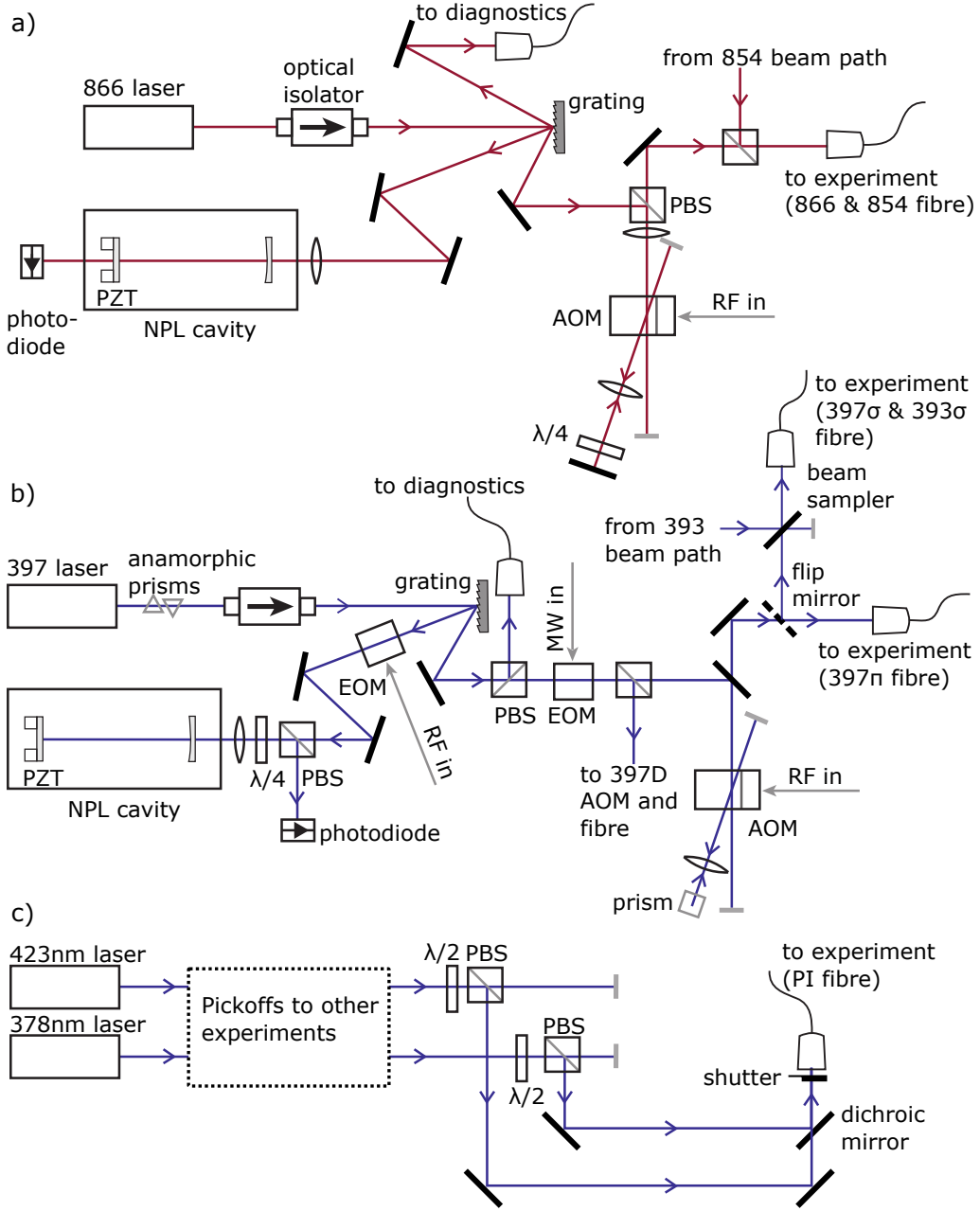


Figure 3.6: Beam paths on the laser table. a) IR laser beam paths: the 866 beam path is shown schematically and the PBS which is used to superimpose the 854 and 866 is depicted on the right of the figure. As indicated, a single IR fibre carries the light from these two lasers to the experiment table. b) UV beam paths (except for PI lasers): there are three fibres carrying 397 light between the laser and experiment table, one for each of the 397 beam paths on the experiment (397 $\pi$ , 397 $\sigma$  and 397 Doppler - see fig. 3.9). Since 397 $\sigma$  and 397 $\pi$  light are never required concurrently in the experiments performed here (397 $\sigma$  is only required to prepare the ‘stretch’ states  $|F = 4, M_F = \pm 4\rangle$  and  $|F = 3, M_F = \pm 3\rangle$  for the experiments described in section 4.2.2, whilst 397 $\pi$  is needed to prepare the  $|\downarrow\rangle$  clock qubit state  $|F = 4, M_F = 0\rangle$ ), we select between them using a flip mirror. When in use, the 397 $\sigma$  is superimposed with the 393 $\sigma$  beam path on a blue beam sampler, indicated on the top right-hand corner of the figure. Also indicated is a PBS used to pickoff a separate beam path for the 397 Doppler cooling beam (denoted ‘397D’). c) The PI beam paths. Figures (a) and (b) were modified from [All11].



### 3.2.1.2 The diode lasers

All seven lasers used here are commercial external cavity diode lasers, produced by Toptica Photonics AG. The photoionisation lasers are DL Pro models and all others are DL 100 models. The lasers' Littrow-type external cavity is formed by a grating, which reflects the first-order beam back into the laser diode. The grating angle and position can be coarsely adjusted with micrometer screws. Fine tuning of the laser frequency is achieved by making small changes to the cavity length. This can be done either by moving the grating along the cavity axis with a scannable piezo-electric actuator or by adjusting the diode's supply current, which in turn changes the index of refraction of the diode. Our laser lock systems control the laser frequencies by feeding back both on the piezo and on the diode current, with piezo control being suitable to correct for slow drifts and current control counteracting fast frequency noise. The laser resonator's temperature is also tunable and is stabilised by a Peltier thermoelectric cooler.

### 3.2.1.3 Laser frequency stabilisation

To stabilise the laser frequencies, some of the light on each beam path is coupled into a low-drift ( $\leq 500\text{kHz}$  per hour) NPL Fabry-Perot reference cavity. If the laser is tuned to the side of a cavity fringe, any frequency fluctuations are translated into intensity fluctuations in the cavity's transmission signal. This transmission signal is recorded by a photodiode and fed into a PID loop which controls the laser frequency by feeding back on both the laser's internal piezo and the laser diode's current. This method of frequency stabilisation is often referred to as a 'side-of-fringe' lock and is used on the 850nm and 854nm lasers.

A more intricate frequency locking method, called the Pound-Drever-Hall (PDH) lock [DHK<sup>+</sup>83, Pou46], is used to frequency-stabilise the 397nm, 393nm and 866nm laser light. This method is insensitive to intensity fluctuations on the laser and offers a more robust lock. It involves modulating the light coupled

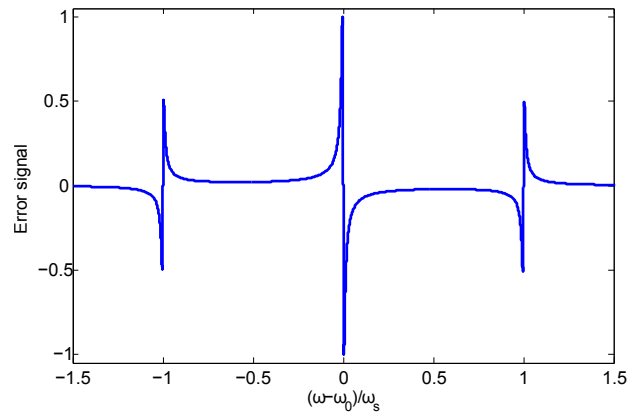


Figure 3.7: PDH error signal plotted against detuning from the cavity resonance ( $\omega_0$ ). The  $x$ -axis is plotted in units of the sideband frequency  $\omega_s$

into the cavity using an Electro Optical Modulator (EOM) at 81MHz, which introduces sidebands to the cavity transfer function. The photodiode signal is then mixed with the modulation signal to give an error signal that has several desirable properties (refer to fig. 3.7):

- i Close to resonance, it changes linearly with laser frequency detuning and has a large slope (inversely proportional to the cavity linewidth);
- ii Regardless of laser intensity, it is zero on resonance;
- iii While the side-of-fringe lock can only correct for small frequency excursions away from the side of the cavity fringe, the PDH lock can correct fluctuations in frequency of up to  $\pm\omega_s$  (where  $\omega_s$  is the sideband frequency) away from resonance. This is because the PDH error signal changes sign across the resonance and retains its sign up to one sideband frequency ( $\omega_s$ ) away from resonance. Over a large frequency range of  $\pm\omega_s$ , the error signal's sign hence functions as an indicator to the feedback electronics as to whether the frequency needs to be increased or decreased.

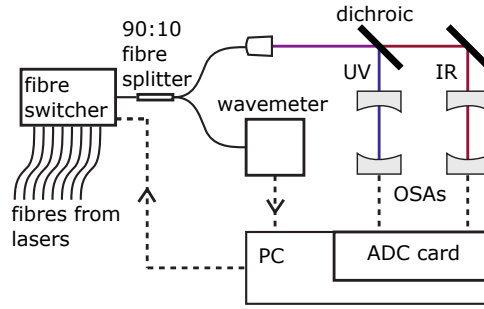


Figure 3.8: Schematic of our group’s laser diagnostic system. A beam of  $200\mu\text{W}$  of light is picked off from each laser and directed to our diagnostics system. The light is split by a 90:10 fibre splitter between an optical spectrum analyser (OSA) and a wavemeter, the readings from which are displayed in a Labview GUI on an auxiliary PC. Figure modified from [All11].

### 3.2.1.4 Laser diagnostics

Fig. 3.8 illustrates our laser diagnostic system. A beam of  $200\mu\text{W}$  of light is picked off from each laser, directed to a fibre switcher and split by a 90:10 fibre splitter between a scanning confocal cavity optical spectrum analyser (OSA) and a Fizeau interferometer-type wavemeter. The outputs of both these instruments are recorded by the diagnostics computer on an analogue to digital card and displayed on a Labview graphical user interface (GUI). The Labview GUI allows the user to select which channels the fibre switcher switches between and displays the wavelength readings, detunings from atomic transitions and frequency spectra for those lasers. This allows us both to tune the lasers to the correct frequency and to examine their mode structure, adjusting the current of any lasers which are lasing in multiple modes.

### 3.2.2 Experiment table

The beam paths on the experiment table are illustrated in fig. 3.9. Section 3.2.2.1 describes the purpose of each beam path, section 3.2.2.2 outlines our laser intensity stabilisation system and section 3.2.2.3 discusses the choice of beam spot size at the ion.

#### 3.2.2.1 Beam paths

We shall refer to the Doppler cooling 397nm and 866nm beams as 397D and 866D. These two beams are superimposed on a blue mirror that is transparent to IR light and are focused onto the trap using a achromatic lens. They travel parallel to the optical table and are in a superposition of  $\sigma$  and  $\pi$  polarisation and enter through the left window of the vacuum system.

Before the 866D beam is superimposed with the 397D beam, a portion of it is picked-off using a polarising beam splitter (PBS) cube preceded by a half wave plate. This portion, which we will refer to as the 866vert beam, is used to compensate for micromotion out of the plane of the trap (using the method outlined in [ASS<sup>+</sup>10]) and enters the vacuum system through the front viewport, crashing into the trap at a  $45^\circ$  angle.

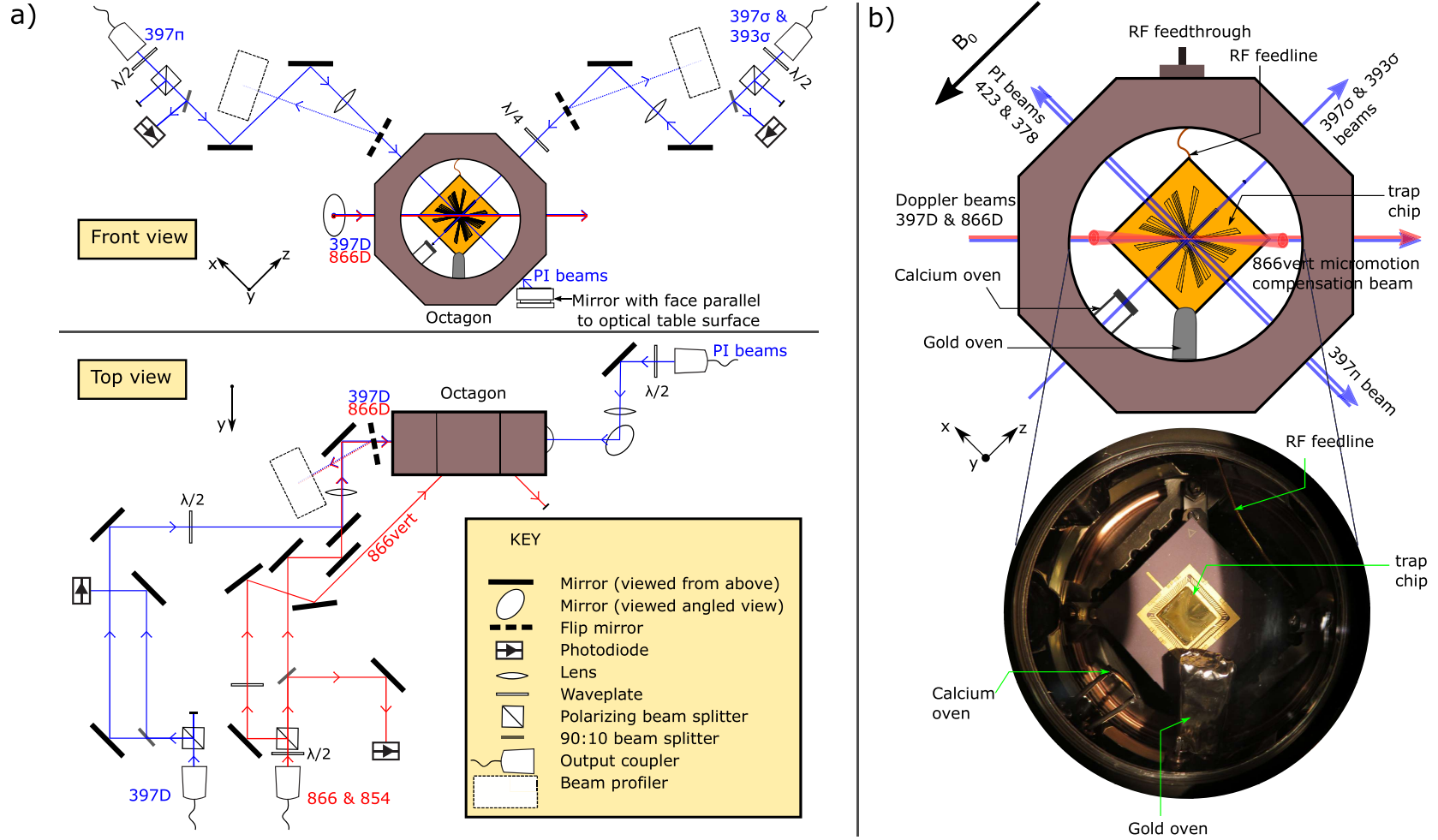


Figure 3.9: a) Schematic of beam paths around the experiment. The ‘front view’ (top) shows the 397 $\pi$  and 397 $\sigma$ &393 $\sigma$  beam paths, which are set up on breadboards 90° to the plane of the optics table. The ‘top view’ shows the PI, 397D and 866 & 854 beam paths, which are set up on the optics table, as viewed by an observer looking down on the optics table. Flip mirrors placed just before each beam path enters the octagon are used to redirect the beams towards a beam profiler during alignment (see section 3.2.4). Note that the last mirror on the PI beam path is a fixed mirror whose face is parallel to the optics table. It reflects the PI beams upward through the vacuum system at 45° to the plane of the optics table. b) Magnified diagram (top) and photograph (bottom) of the octagon housing the trap chip. The direction of the static B-field when it is aligned parallel to the trap axis is indicated by the arrow labelled  $B_0$ .

The  $\sigma$ -polarised 393nm and 397nm, henceforth referred to as  $397\sigma$  and  $393\sigma$ , are required for preparation of the stretch  $|F = 4, M_F = \pm 4\rangle$  states and for read-out, respectively. They are superimposed on the trap table and co-propagate along a breadboard at  $90^\circ$  to the trap table (see fig. 3.15, entering the vacuum system through the top right window. A quarter wave-plate is used to set the  $397\sigma$  polarisation to  $\sigma^+$  or  $\sigma^-$  depending on whether the  $|F = 4, M_F = +4\rangle$  and the  $|F = 4, M_F = -4\rangle$  state is being prepared (if the clock state is being prepared, the  $397\sigma$  beam is not used for preparation and the waveplate can be set to either  $\sigma^+$  or  $\sigma^-$  polarisation).

The  $\pi$ -polarised 397nm beam, henceforth referred to as  $397\pi$ , which is required for preparation of the clock qubit, enters the vacuum system from the top left window. This beam path is therefore also set up on a breadboard at  $90^\circ$  to the optics table.

The 378nm and 423nm photoionisation beams, henceforth referred to as the PI beams, are superimposed on the trap table and enter the vacuum system through the bottom right window. The PI beams traverse the vacuum system at right angles to the atomic beam emitted by the calcium oven, in order to minimise the Doppler shift seen by the atoms on the photoionisation transitions.

#### 3.2.2.2 The noise-eater: Laser intensity stabilisation

To counter intensity noise introduced by the optical fibres that carry light between the laser and experiment tables, a ‘fibre-noise-eating’ system is in place in our laboratory. This works by sampling the power of a beam after the fibre (on the trap table) and feeding back on the RF power driving the AOM before the fibre. The RF drive amplitude controls the amplitude of the sound waves produced in the AOM crystal, which sets the intensity of the diffracted beam (recall that we use the first-order diffracted beams for the experiment). Therefore, feeding back on the AOM RF drive power allows the system to correct for

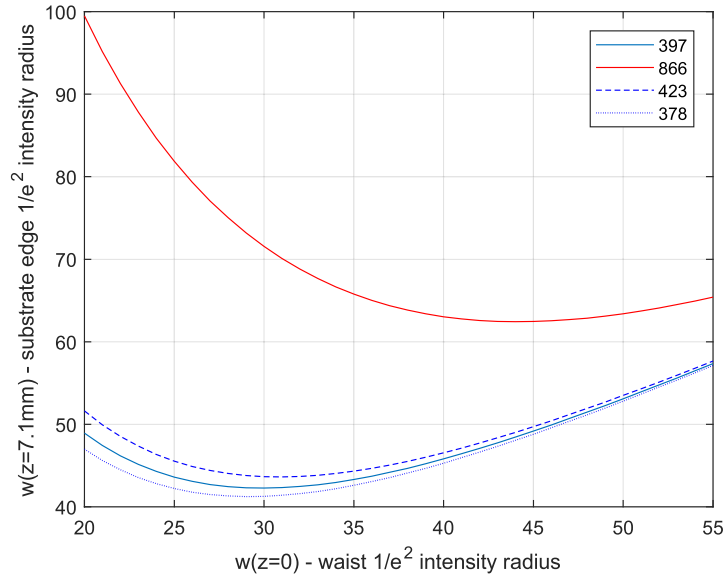


Figure 3.10: Plot of the  $1/e^2$  beam radius at the edge of the trap chip (i.e.  $2\sqrt{5} \approx 7.1\text{mm}$  away from the centre of the trap) versus the waist at the beam focus.

intensity fluctuations after the fibre. Ten percent of the power of each beam on the trap table (except for the PI beams, since photoionisation does not require very stable intensities) is picked off with a 90:10 beam splitter and measured with a photodiode, whose voltage is fed into the noise-eater system.

### 3.2.2.3 Choosing the spot size at the ion

The beam spot size at the ion was chosen so as to reduce the light intensity clipping the edge of the trap surface, since UV light can charge the trap electrodes via the photoelectric effect. Equation 3.1 gives the  $1/e^2$  beam radius at a distance  $z$  from the focus as a function of the beam radius at the focus,  $\omega_0$  and of the Rayleigh distance  $z_R = \frac{\pi\omega_0^2}{\lambda}$ , where  $\lambda$  is the wavelength of the light. Fig. 3.10 plots the waist at the edge of the trap chip versus the waist at the focus and indicates an optimum waist radius of  $30\mu\text{m}$ , i.e. a spot size, or diameter, of  $60\mu\text{m}$ .

$$\omega(z) = \omega_0 \sqrt{1 + (z/z_R)^2} \quad (3.1)$$

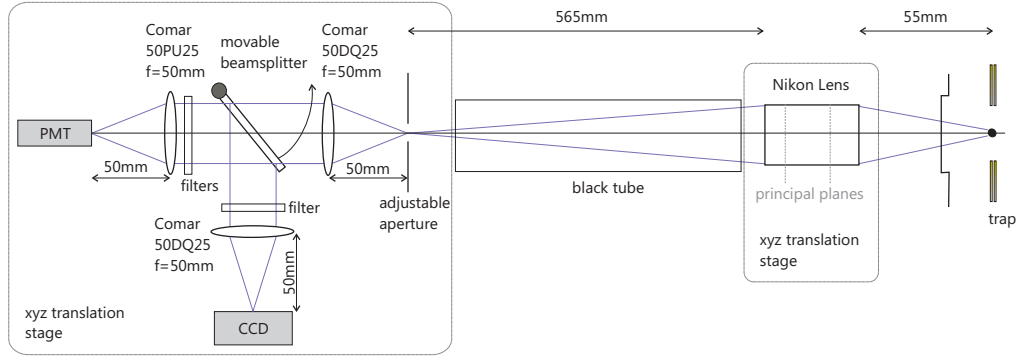


Figure 3.11: Schematic depiction of the imaging system used in these experiments. Figure modified from [Lin12].

#### 3.2.3 Imaging system

The imaging system used in this experiment is schematically depicted in fig. 3.11, and was constructed by Alice Burrell [Bur10] for use in a previous experiment. It features a Nikon lens of large numerical aperture ( $\sin\theta = 0.29$ ), which images the ion fluorescence with  $7.7\times$  magnification. The image is then projected either onto a photomultiplier tube (PMT) or onto a CCD camera (light can be redirected from one to the other by a 90% reflection beam splitter on a flippable stage). The camera is used to align the imaging system with respect to the trap electrodes (see fig. 3.12). When preparing to run the experiment, we use the set of movable shutters on the focal plane of the Nikon lens to trim the imaged area so as to cut down the amount of scattered light (scattered from the surface of the trap) which reaches the PMT.

#### 3.2.4 Beam alignment

To align the beams onto surface traps, we use both the camera in the imaging system and a beam profiler. If a beam is grazed across the trap surface, it is easily visible on the camera (due to scatter from the trap's gold electrodes) and can be steered relative to the trap electrodes to traverse the relevant trapping zone (see fig. 3.12c). On any given beam path, the final mirror before the octagon controls



the angle of the beam with respect to the trap plane and the second-to-last mirror controls the position of the beam's centre on the last mirror. We would like to align all of the beams (with the exception of the 866vert) so that they traverse the octagon travelling parallel to the trap plane and at a distance of one ion height away from the trap surface. This can be accomplished by using the process described in fig. 3.12 to align the beam so that it is parallel to the trap and then using the beam profiler to position it at one ion height above the trap surface, as follows: a flip mirror placed just before the beam enters the octagon can be used to redirect the light to the beam profiler, which is positioned so that it intersects the focus of the beam (see fig. 3.9), i.e. it is placed at the same distance away from the flip mirror's surface as is the centre of the trap. The second-to-last mirror on the beam path (not counting the flip mirror) is then used to crash the beam onto the trap, glazing it across the surface (we can judge when the centre of the beam intersects the surface of the trap by looking for maximum scatter from the beam on the camera). Flipping in the flip mirror allows us to record the coordinates of this beam position on the beam profiler. We can then use the second-to-last mirror to translate the spot on the beam profiler by a distance of one ion height along the direction which, when the flip mirror is taken out, corresponds to the normal to the trap plane. In this way, we can translate beam away from the trap surface, positioning it at the ion height with micrometer accuracy.

The 397D is the first beam to be aligned in this experiment, using the procedure we have just outlined. The 866D can then be straightforwardly superimposed with it at two positions along the beam path (e.g. on a card before the achromatic lens and on the beam profiler). The  $397\sigma$  and  $397\pi$  beams are the next to be aligned using the procedure described. This latter is counter-propagating with the PI beam path and hence serves as a useful guide for the alignment of the PI beams, which would otherwise be slightly complicated by the fact that the last steerable mirror face on the PI path is angled in such a way that its steer-

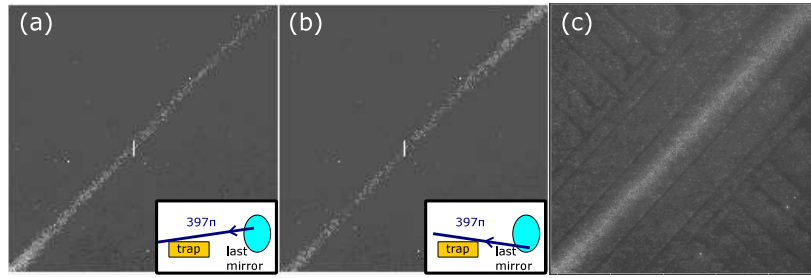


Figure 3.12: Camera images of the  $397\pi$  beam glazing the trap surface during beam alignment. To align the beam parallel to the trap surface, successive small adjustments to the position of the beam on the last mirror in the beam path were made using the second-to-last mirror. After each adjustment, the last mirror on the beam path was used to tilt the beam away from the trap surface until it could no longer be seen on the camera and then to tilt it back towards the trap surface until the first glimpse of scatter could be seen on the camera. An asymmetric beam image (i.e. an image where the beam is clearly brighter on one of the edges of the chip) indicates that it is not parallel to the trap, so the iterative process is continued until the beam image is seen to be as symmetric as possible. a) Here, the  $397\pi$  beam is too high on the last mirror, since the corner of the chip which is furthest from where the beam enters the octagon appears brighter than the one that is closest to its entry point (recall that the  $397\pi$  beam enters the octagon from the top right). Inset: schematic cross-sectional view of the trap showing this alignment scenario. b) The  $397\pi$  beam is now too low on the last mirror, since the trap corner that is closest to where it enters the octagon is appears brighter. c) The beam is aligned parallel to the trap structure and appears symmetric in the camera image. The trap structure is also made visible in this image (it is illuminated by a torch shone through the front viewport of the octagon).

ing axes are not aligned with the in and out-of-trap-plane directions (refer to fig. 3.9). Coarse alignment of the PI beams is done by superimposing them with an already-aligned  $397\pi$  on a card in two places along the beam path. Fine alignment is then done by steering the PI beams so as to maximise their coupling into  $397\pi$  fibre output coupler. Finally, the 866vert beam is simply aligned by using the camera to view the spot it creates on the trap surface (where it crashes onto the trap, at  $45^\circ$  to the surface) and steering this spot onto the relevant trapping zone.

### 3.3 Chip trap fabrication

I fabricated the chip ion-trap used in this work in Oxford University cleanrooms using a slight variation of the process developed by our group in [ASS<sup>+</sup>10] (based on methods used at MIT [Lab08] and electroplating techniques devel-

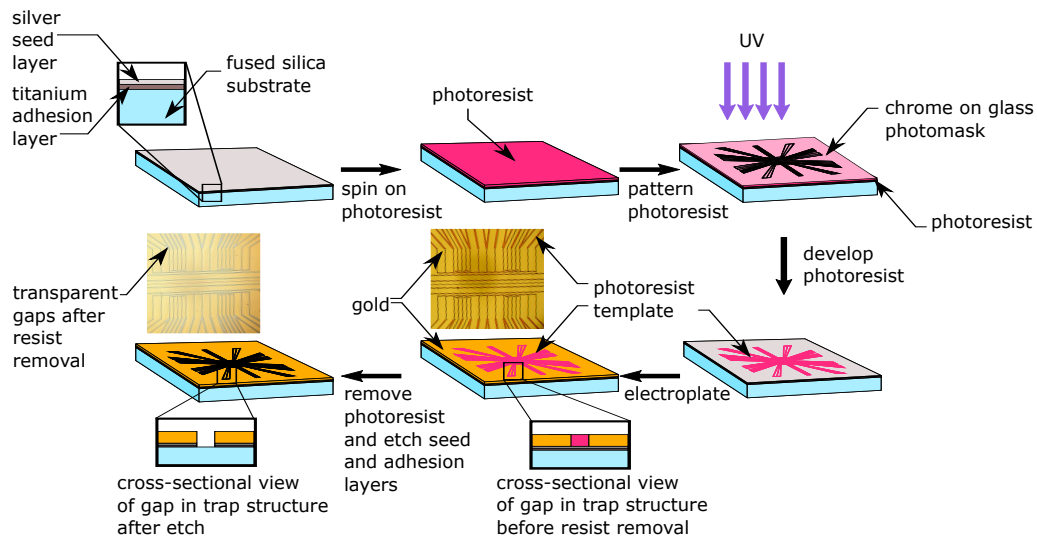


Figure 3.13: Flow chart depicting process used to fabricated the ion trap chip used in the experiments described in this thesis. The images in the bottom left-hand corner were taken during one of the fabrication runs using a light microscope.

oped in [KMK<sup>+</sup>04]). Fig. 3.13 illustrates the process used, which is summarised below:

1. Twenty 5mm × 5mm × 0.5mm fused silica trap substrates were cleaned by soaking them in ‘Pirana etch’ (a 10:1 mixture of sulphuric acid and 30% concentration hydrogen peroxide) at 100°C for three minutes. The substrates are then rinsed in de-ionised water.
2. In order to deposit inert metals such as gold and silver on a silica substrate, an adhesion layer is required. A clean, thin layer of a reactive metal such as titanium or chromium, adheres both to silica substrate and the inert metal layer that follows it. In a high-vacuum environment, a titanium adhesion layer (a few monolayers thick) is evaporated onto the substrates, followed by a 200nm silver seed layer. Both layers are deposited without breaking vacuum, to prevent the titanium from oxidising and losing its adhesive properties.

3. The substrates are evaporatively primed with hexamethyldisilazane (HMDS) primer. Each substrate is then, in turn, mounted on a calibrated spinner and the photoresist is puddled onto it. The photoresist is spun onto the substrate by running the spinner at a speed of 4000rpm for 45 s with a ramp time of 20 s to give a photoresist thickness of  $5\mu\text{m}$ .
4. Each substrate is placed on a mask aligner under a chrome-on-glass photomask on which the pattern of the trap electrodes has been imprinted. The photomask is transparent in the areas of the design to be metallised in the final chip and opaque along the areas which will be gaps. When placed in the mask aligner, the photomask allows the UV light to degrade the photoresist in the areas of the substrate which are to become metallised electrodes. After exposure to the UV light, each substrate is baked at  $115^{\circ}\text{C}$  for 2 minutes on a hotplate and then left to rest for 30min so that the photoresist can re-hydrate.
5. The substrates are then rinsed in a developer which washes away the degraded areas of the photoresist. This leaves behind a photoresist template, to be filled with electroplated gold to form the trap electrodes.
6. Each substrate is electroplated using the setup depicted in fig. 3.14 for 31.5 minutes with 0.8V applied across the cathode and anode, giving a current of 2mA. The electroplating solution<sup>1</sup> is kept at a temperature of  $50^{\circ}\text{C}$  throughout the plating process. These settings were chosen to give a final gold thickness of  $5\mu\text{m}$ , according to the plating solution's data sheet. The resist is then washed off by rinsing the substrates off in acetone, isopropanol and deionised water.
7. The silver seed layer is etched by submerging each substrate in a solution composed of 1 part ammonium to 1 part 30% concentration hydrogen per-

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<sup>1</sup>Metalor ECF 63

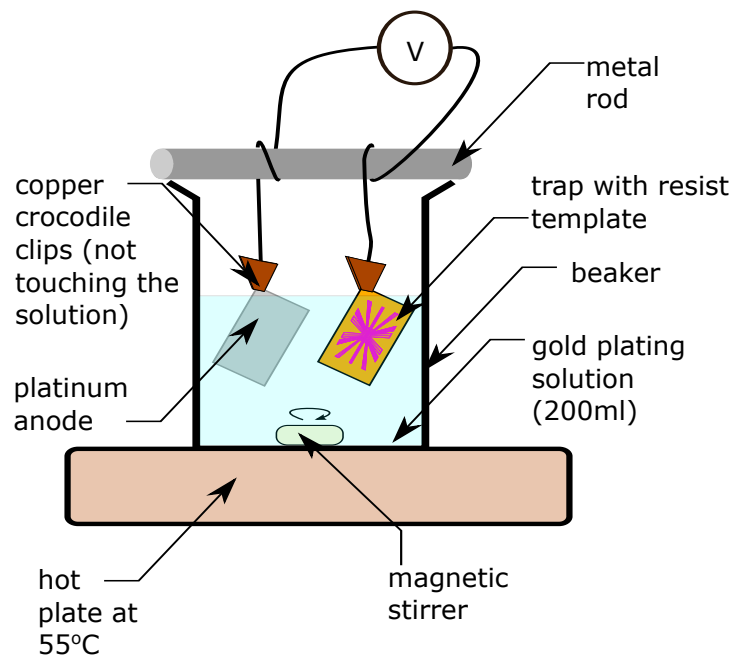


Figure 3.14: Schematic of electroplating setup used to plate the trap's gold electrodes.

oxide to 4 parts water for 15 s.

8. The titanium adhesion layer is etched by placing the substrate in a 10% concentration solution of hydrofluoric acid (HF) for 15 s.
9. A final step had to be added to this process during the fabrication of the ion trap used in this work. After the HF etch, the gaps between the electrodes were still not completely transparent when viewed through a back-lit light microscope: a fine, grey layer could be seen along most of the gaps. This layer is believed to have been titanium oxide. It was removed by first rinsing each substrate in acetone, isopropanol and de-ionised water (to remove any organic deposits which might protect the grey layer from etchants) and then submerging it in 30% concentration hydrogen peroxide for 20 s.

## 3.4 Vacuum system

Fig. 3.15 shows the vacuum system and surrounding experimental fixtures. The trap is housed in a stainless steel spherical octagon<sup>2</sup> which is depicted in fig. 3.9b: it has eight 1.33" CF flanges and two 4.5" CF flanges. Fused silica viewports are bolted to three of the 1.33" flanges and to the front 4.5" CF flange, with the remaining 1.33" flanges being used for a single-pin RF feedthrough and two four-pin feedthroughs for the Calcium oven and the gold oven. The octagon is attached to one of the 4.5" flanges of a 5-way cross. The cross' two side flanges attach to microwave feedthroughs, its top flange attaches to a 4-way cross which supports an ion gauge<sup>3</sup>, a 25-way D-sub DC feedthrough and a getter pump<sup>4</sup> and its back flange connects to a 3-way cross which supports a 27l/s ion pump<sup>5</sup> and two serially-connected all-metal valves<sup>6</sup> that are used to close the system off from the surroundings (after it is baked out and pumped down using external pumps). The subsequent sections describe the cleaning procedure used on the ultra high vacuum (UHV) parts of the vacuum system, the assembly of the calcium and gold ovens and the bake-out and pump-down procedures.

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<sup>2</sup>Kimball Physics MCF450-SphOct-Ec2A8, 316L stainless steel

<sup>3</sup>Varian UHV-24P with x-ray limit of  $5 \times 10^{-12}$  Torr (1 Torr = 1.33 mbar)

<sup>4</sup>SAES Getter GP50 Mk2

<sup>5</sup>Varian VacIon Plus 20 Diode

<sup>6</sup>Kurt J Lesker part VZCR40R



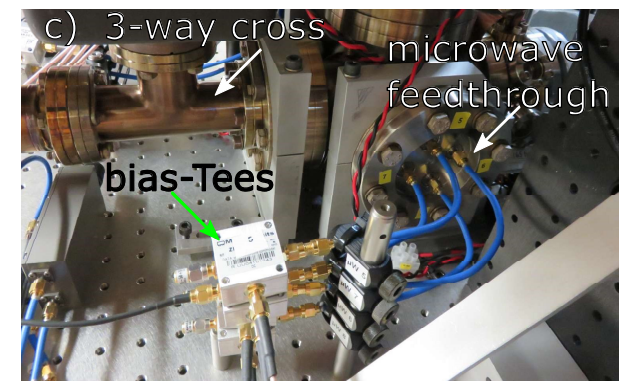
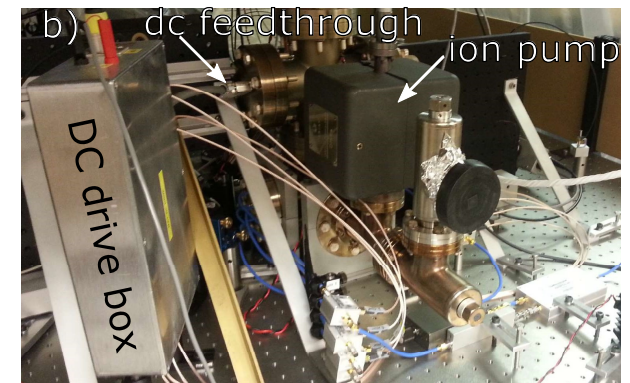
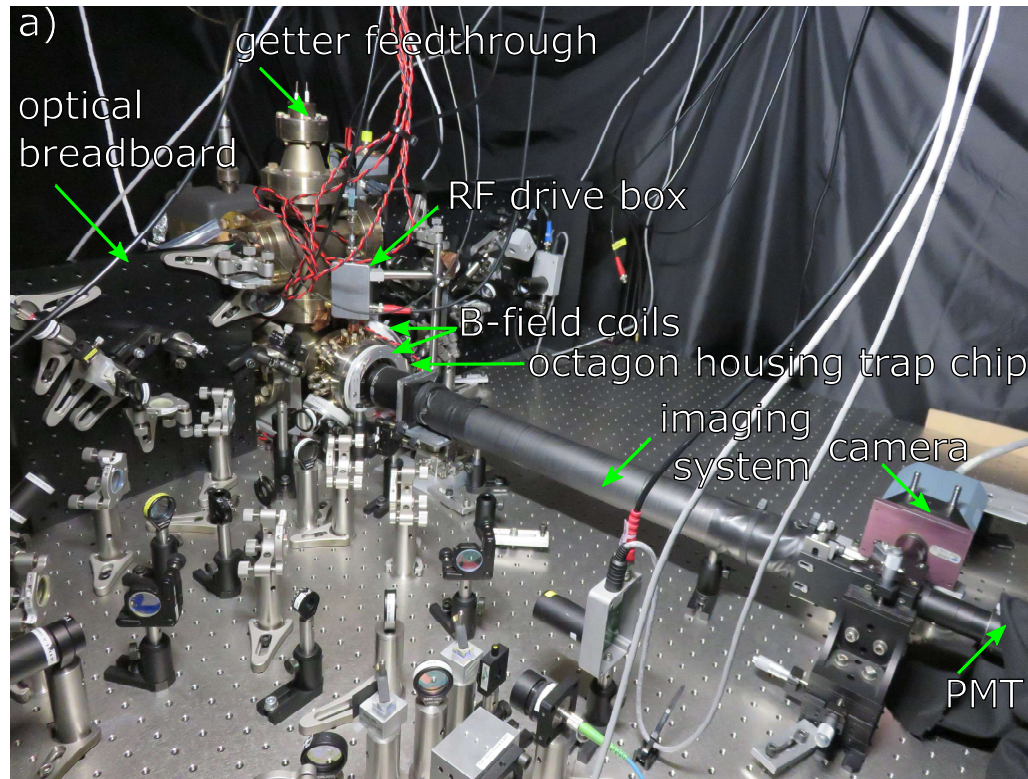


Figure 3.15: Overview of experimental setup. a) Front view of the vacuum system, with the octagon housing the trap chip at its base. Two optical breadboards either side of the octagon support the  $397\sigma$  and  $397\pi$  beam paths. The imaging system and the Doppler beam paths are visible on the optics table. b) Back view of vacuum system showing the ion pump, DC feedthrough, all metal valves and DC drive system (described in section 3.5). c) Side-view of vacuum system, showing one of the microwave feedthroughs.

#### 3.4.1 Vacuum cleaning procedure

To minimise the organic residue on the inside of the vacuum system, each metal vacuum part was submitted to the following cleaning procedure:

1. Any visible stains or grease marks are removed using a sponge soaked in Neutracon, an industrial grade detergent.
2. The part is placed in a clean container (for example, a large beaker or metal container which has been thoroughly cleaned with neutracon, rinsed in de-ionised water and rinsed in acetone) filled with warm deionised water and Neutracon detergent. The container is placed in an ultrasonic bath for 5 minutes. A foil lid is put on the container to prevent contamination during ultrasonic cleaning. After ultrasonic cleaning, the part is then rinsed in de-ionised water.
3. The part is submerged in a container of deoxidine solution, which is placed in the ultrasonic bath for a further 5 minutes. The part is again rinsed in de-ionised water.
4. Finally, the part is submerged in a container of acetone, which is in turn placed in an ultrasonic bath for a further 5 minutes. The part undergoes a final rinse in de-ionised water and is then blow-dried with argon. Blow drying is preferable to allowing the water film to evaporate since, when we blow-dry the part, organic contaminants suspended in the water film are removed with the water. If the water is allowed to evaporate however, these contaminants will be left behind on the surface of the part since they have much higher boiling points than water.
5. If the part is not going to be used immediately, it is wrapped in aluminium foil. Any foil used to wrap vacuum parts is previously cleaned by wiping it with a cleanroom wipe soaked in acetone.



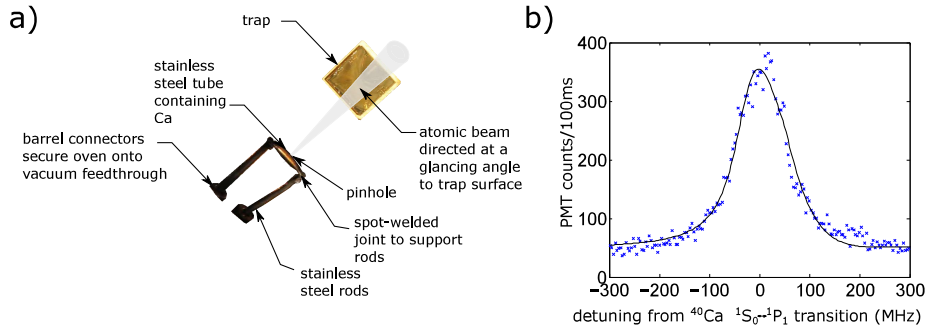


Figure 3.16: a) Depiction of calcium oven under operation. b) Neutral fluorescence signal on the  $^1S_0$  to  $^1P_1$  transition, observed with the oven running at 5.8A (data is fitted with a Voigt profile).

To minimise organic contamination, we wore disposable masks, hair nets and gloves to manipulate all vacuum parts.

### 3.4.2 Calcium oven

Fig. 3.16a shows the calcium oven installed in the vacuum system. It consists of a hollow stainless steel tube filled with pellets of isotopically enriched calcium (12%  $^{43}\text{Ca}^+$ , 88%  $^{40}\text{Ca}^+$ ). The tube was then spot-welded to two stainless steel rods, which are connected via barrel connectors to a two-pin vacuum feedthrough. When a voltage is applied across the feedthrough pins, current runs through the oven, heating it up and causing an atomic beam to be ejected through a pin-hole in the stainless steel tube (see fig. 3.16). The procedure followed to construct the oven was as follows:

1.  $\sim 1\text{cm}$  of hollow 316L stainless steel tube, and two  $\sim 3\text{cm}$  lengths of 316L stainless steel rod were cleaned (as described in section 3.4.1) and air-baked at  $400^\circ$  on a hot plate for 1 hour;
2. One end of the hollow stainless steel tube was crimped with a clean tool and a hypodermic needle was used to pierce a hole half-way across its length

3. The crimped end of the tube was spot-welded to one of the stainless steel rods;
4. The tube was filled with six to seven  $1 \times 1 \times 0.65\text{mm}$  calcium pellets (from a freshly opened vial which had been kept under the manufacturer's vacuum seal until this point to prevent oxidation), each pellet having been cut into four pieces with a clean blade. While this was done and for the rest of this procedure, a constant stream of argon was blown over the oven in order to prevent the calcium from oxidising in air (this procedure can also be performed in a glove box back-filled with argon if one is available);
5. The open end of the hollow tube was crimped and spot-welded to the other stainless steel rod;
6. The oven was installed in the octagon by fastening it onto a two-pin feed-through using barrel connectors.

When firing the oven for the first time, we gradually turn up the supply current whilst illuminating the trap with a 423nm laser until we see fluorescence from the atomic beam on the  $^1S_0$  to  $^1P_1$  transition in neutral calcium. Fig. 3.16 shows the neutral fluorescence signal observed when a current of 5.8A was fed to the calcium oven used in these experiments. Due to oxidation of the outer layer of the calcium pellets, a higher current is often needed to observe fluorescence when the oven is first fired under vacuum, since the pellets must be heated enough for a crack to develop in the oxidised layer, through which a neutral calcium beam can then be ejected. We observed neutral fluorescence at 6A when first firing the oven, but a much lower supply current of 5.2-5.3A is sufficient for day-to-day loading of ions into the trap.

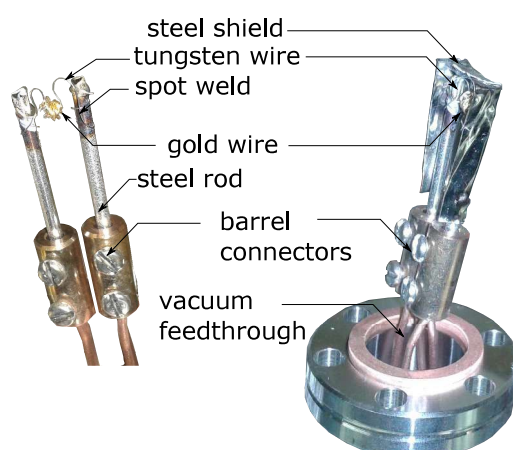


Figure 3.17: Right: front view of the gold oven before the shield was spot-welded onto one of the stainless steel rods. Left: side view of gold oven with shield.

### 3.4.3 Gold oven

With the aim of investigating the effect on the trap heating rate of depositing a new layer of gold over trap electrodes under vacuum, we installed a gold oven in this vacuum system. The oven is depicted in fig. 3.17 and consists of a length of gold wire tightly coiled around a tungsten filament which has been spot-welded to two supporting stainless steel rods. The oven was constructed as follows:

1. A 5cm<sup>2</sup> piece of 316L stainless steel sheet, 5cm of tungsten wire (diameter 0.15mm), 4.5cm of platinum wire (diameter 0.05mm, 99.99% purity), 8cm of gold wire (diameter 0.2mm, 99.99% purity), and two 3cm lengths of 316L stainless steel rod were cleaned (as described in section 3.4.1;
2. The tungsten wire was bent into a spring-like shape (see fig. 3.17). This shape gave it room to expand, preventing it from snapping when heated;
3. The tungsten wire was spot-welded onto two stainless steel rods;
4. The thin platinum wire was wound onto the tungsten: this would provide a surface onto which the gold could wet when melted;

5. The gold wire was wound in a tight coil over the platinum wire;
6. The stainless steel sheet was bent into the shape shown in fig. 3.17 and one corner of it was spot welded to one of the stainless steel rods. This sheet acts as a shield to protect the view-ports from being sprayed with gold when the oven is fired;
7. The oven was installed in a test vacuum chamber (pumped down to  $10^{-5}$  mbar) and heated until the gold melted into a sphere. The voltage applied to the oven feed-through and the current through it were monitored as it was heated, and the power required to melt the gold noted. The gold was melted under vacuum in order to minimise the amount of organics absorbed into the molten sphere. Once cool, the oven and feed-through were removed from the test chamber and installed in the octagon.

To calibrate the amount of gold we expect will be deposited onto the trap surface when we heat the gold oven for a given length of time, a twin gold oven was built and fired in the test chamber shown in fig. 3.18. The following experiment was performed:

1. A glass slide was positioned at the same distance away from the oven as the trap was to be placed in the octagon. An 'X' was drawn using a permanent marker on the backside of the glass slide to mark the position of the centre of the trap;
2. The test chamber was pumped down and the oven was heated until the gold melted and left on for a time  $t$ , after which the glass slide was replaced and the experiment repeated with the oven left on for varying amounts of time;
3. A clean cotton swab was swiped through the middle of each glass slide, over the 'X' position. This produced a step in the thickness of gold on

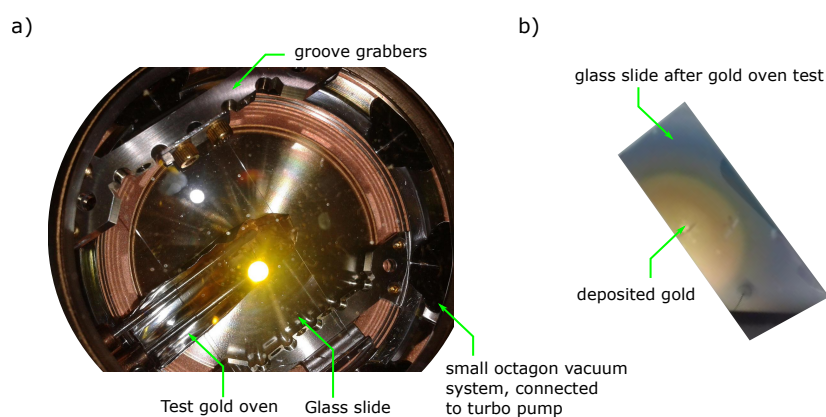


Figure 3.18: a) Setup for testing gold ovens. A small octagon vacuum system pumped down with a turbo pump houses the test gold oven and a glass slide onto which gold is deposited when the oven is fired. b) A layer of gold deposited on a glass slide after a test gold oven is fired using the setup shown in (a).

the glass slide, allowing thickness to be measured using a Dektak surface profiler (this machine measured the change in vertical position of a needle scanned across a step on a surface). By controlling the time for which the oven was on and the supply current, I found that the test gold oven could be used to deposit between 20 and 400nm of gold onto a glass slide.

### 3.4.4 Bake-out procedure

As it was assembled, our ion-trap vacuum system underwent three bakes, which we now describe.

#### 3.4.4.1 Air bake

All stainless steel parts of the vacuum system were put through an air bake. This involved separately wrapping the vacuum system parts in foil after cleaning and baking them at 400° for 8 hours. The bake promotes oxidation of the surface of the steel, forming an oxide barrier which contains hydrogen outgassing from the bulk.

#### 3.4.4.2 Hard bake

This is a high-temperature bake, at 390°C; it is done on all parts of the vacuum system which can withstand such temperatures. To perform it, we assembled the vacuum system as it would be used in the experiment, but replacing all feedthroughs and viewports by blank flanges, removing the ion-pump magnets and keeping the octagon empty since the trap it was to house should not be subjected to such high temperatures. The system was then placed in the oven and pumped down with a turbo pump to  $7 \times 10^{-7}$  mbar. With the turbo pump still running, the oven temperature was ramped up to 390°C over a few hours. After a day baking into the turbo pump, a valve to an external 60l/s ion pump was opened and the turbo pump was valved off. The system was baked into the ion pump for a few days before the temperature was ramped down.

#### 3.4.4.3 Soft bake

The soft bake is a 2-3 week long bake performed on the fully assembled vacuum system. Fig. 3.19 show the vacuum system in the oven, encased in many layers of foil in preparation for the soft bake. Here, the vacuum system was soft baked for a total of 20 days at a temperature of 200°C, first into a turbo pump (for five days, until the pressure dropped to  $\sim 10^{-5}$ ) and then into the external 60l/s ion pump. Before the bake the calcium and gold ovens were degassed into the turbo pump at 0.2A and 3A supply current respectively, so that any surface organics would not contaminate the ion pumps. Similarly, the vacuum system's getter pump was activated with a supply current of 2.5A which heated it to 450°C before the bake so that it might release any adsorbed organics into the turbo pump. Towards the end of the bake, just before the cool down process began, the getter was once again activated, now into the external ion pump. The system's 27l/s ion pump was also briefly turned on (for 1-2mins) at this stage to degas into the more powerful external pump. Once the system was cooled to around 140°C,



Figure 3.19: vacuum system in the oven, ready for the soft bake. Many loose layers of aluminium foil are used ensure that the heat permeates the system slowly, without creating large temperature gradients which could damage viewports or crack delicate ceramic feedthroughs.

the vacuum system's 27l/s ion pump was turned on permanently (this pump remains on for the entire lifetime of the vacuum system). At 110°, the outer metal valve on the vacuum system was closed to prevent re-entry of the water vapour which was pumped out during the bake. Finally, once the system was cool, the getter pump received a final activation (the system ion pump was briefly turned off during this procedure, to prevent it from absorbing any substances outgassed during the getter's high-temperature activation).

### 3.5 DC control system

The system used to deliver the required DC voltages to trapping electrodes is shown schematically in fig. 3.20a and photographically in fig. 3.20b. The voltages are produced by an Analogue Devices 32-channel Digital to Analogue Card (DAC), the AD5372. Each DAC channel is controlled via a set of programmable registers. An ATmega28 microcontroller (on a ChipMax 32 evaluation board) is used to program these registers via a Serial Peripheral Interface (SPI) link. I wrote a Python graphical user interface (GUI) which allows a user to set the desired trap voltages and converts them into text commands sent by serial link

to the microcontroller) (see fig. 3.20). The microcontroller is loaded with a C program (which I wrote based on a previous version by Thomas Harty) that interprets the text commands and converts them into 8 to 16 bit words to be sent to the DAC. Finally, voltages from the DAC are passed through a low-pass filter, which I designed to minimise noise on the trap electrodes (see fig. 3.20a).



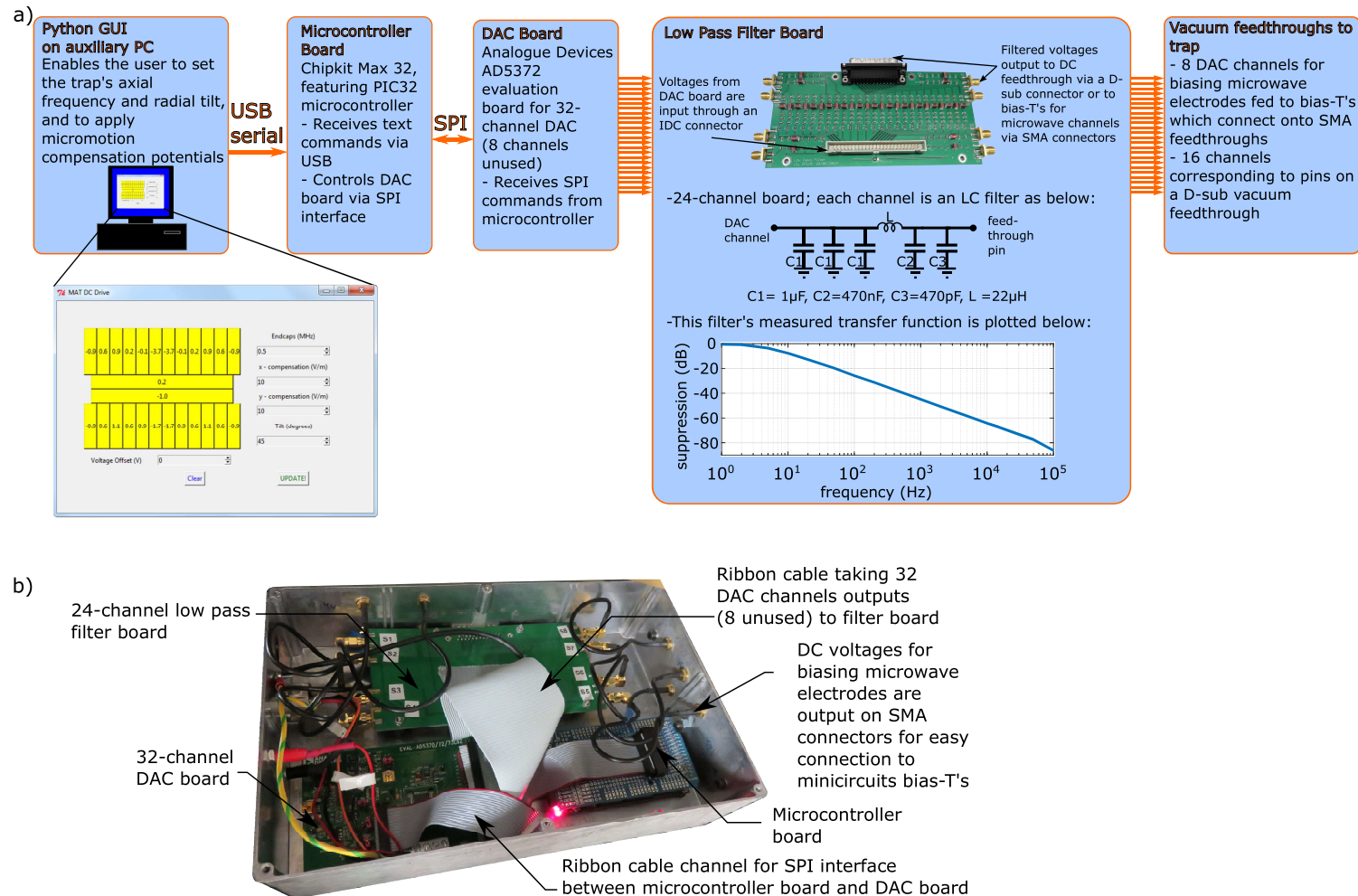


Figure 3.20: The DC drive system. a) Schematic representation of the DC drive chain featuring a screenshot of the Python GUI and a detailed description of the low pass filter board (including a plot of its measured transfer function). b) Photograph of the assembled DC drive box.

### 3.6 Trap RF drive

The 176V of radio frequency required to trap ions are delivered to the trap via a single-pin feedthrough and are produced using the drive chain outlined in fig. 3.20. A toroidal resonator<sup>7</sup> is used to provide a 24-fold step-up in voltage and also acts as a low-Q band-pass filter.

The resonator circuit is depicted in fig. 3.21. The number of turns in the secondary coil ( $N_{\text{secondary}} = 9$ ) was chosen so that the  $LC$  circuit formed by the capacitance across the RF feedthrough,  $C_{\text{trap}}$ , and inductance of the coil,  $L_{\text{coil}}$ , would resonate at the desired RF trapping frequency of 40MHz. For the measured  $C_{\text{trap}}$  of  $\sim 30\text{pF}$ , the inductance required is  $0.58\mu\text{H}$  and, since the datasheet of the toroid used here specifies inductances of  $7\text{nH/turn}$ , the appropriate number of turns in the secondary coil is approximately 8.

A capacitive tap was included at the output of the resonator so that we can monitor the voltage being delivered to the trap without loading the resonator with an oscilloscope probe. The tap consists of a capacitive divider built from a  $1\text{pF}$  capacitor in series with a  $1\text{nF}$  capacitor: if a voltage is applied across the chain, it will be divided in a ratio of 1:1000 across the  $1\text{pF}$  capacitor and the  $1\text{nF}$  capacitor respectively (since the impedance of the former is  $1000\times$  larger than that of the latter), and will hence appear mostly across the  $1\text{nF}$  capacitor. The combined capacitance of the chain is however, approximately  $1\text{pF}$  - small enough not to significantly change the capacitance across the trap to which the resonator is tuned. A permanent BNC connector soldered across the  $1\text{nF}$  capacitor (see fig. 3.21) hence allows us to probe the output voltage of resonator without shifting its resonance frequency. To establish the relationship between the voltage measured at the capacitive tap and the output voltage of the resonator, we perform the following calibration measurement: with the resonator connected to the trap, a  $10\times$  oscilloscope probe is connected across the

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<sup>7</sup>Micrometals T94-6 iron powder toroid

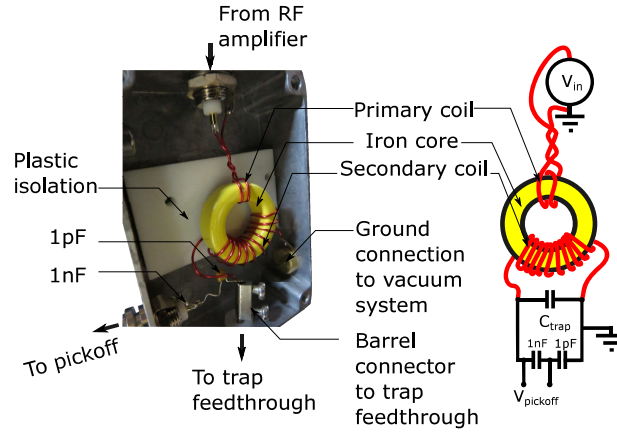


Figure 3.21: Photograph (left) and schematic representation (right) of the trap's RF-drive system.

RF feedthrough to measure the output voltage. The oscilloscope probe loads the resonator, but, from this measurement we could establish that the ratio between the voltage at the capacitive tap and the output voltage was 761. Hence, we can monitor the resonator's output voltage,  $V_{out}$ , by simply monitoring the pickoff voltage,  $V_{pickoff}$ , and using the relationship  $V_{out} = 761 \times V_{pickoff}$ .

### 3.7 Microwave DDS drive

With the aim of developing a microwave drive system which is scalable to many digitally controllable channels, we use a direct digital synthesiser (DDS) chip produced by Analogue Devices, the AD9910. The chip can be controlled via SPI and can produce output waveforms with digitally controllable frequency, phase and amplitude at a frequency of up to 400MHz. This output can then be mixed with a local oscillator at 2.8-2.9GHz to produce a microwaves at around 3.2GHz.

To control the chip, I wrote C code which was loaded into an ATmega28 microcontroller. The program allows the microcontroller to receive user-issued text commands, interpret them and write to or read from the DDS's control and data registers.

The first prototype of the drive system is depicted in fig. 3.22. It consisted

of an evaluation board for the AD9910 connected via a ribbon cable to an evaluation board for the ATmega28 microcontroller (the ChipKitMax32 board by Digilent). The output of the DDS evaluation board was mixed with an external 2.9GHz local oscillator in a single sideband (SSB) mixing configuration.

Our group later began using a custom-made board by Enterpoint as an integrated solution to produce many channels of controllable microwave waveforms. The board features four AD9910 DDS chips and a Spartan 6 FPGA, which can be programmed to control them. To drive a single DDS channel, the on-board FPGA is configured to act like a microcontroller (using, for instance, the Verilog based microBlaze architecture) and loaded with a slightly modified version of the code I developed. In [Bal14], Chris Ballance implements a 4-channel system using one of the Enterpoint boards.

The prototype I developed has been in use as a single-qubit microwave gate drive by another experiment in our group, which uses a macroscopic blade-type trap to perform high-fidelity mixed-species entangling gates [BSH<sup>+</sup>15]. In the experiments performed here, the (single-channel) microwave drive was provided either by a modified version of this prototype, where the DDS output frequency is octupled instead of being SSB-mixed with a local oscillator (modifications by D.T.C. Allcock and T.P. Harty), or simply by a GPIB-controlled synthesiser.

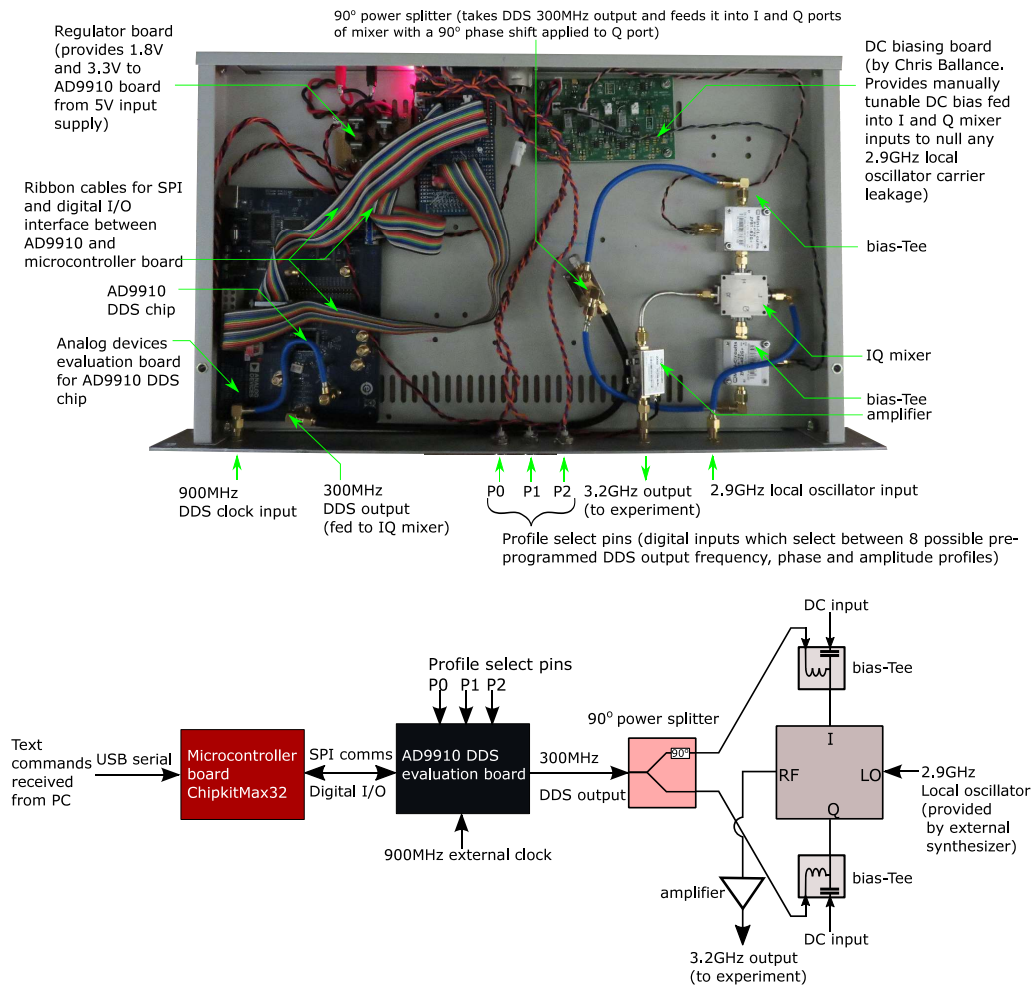


Figure 3.22: Photograph (top) and schematic representation (bottom) of the prototype DDS microwave-drive system.



# 4

## MAT: Microwave Addressing Trap

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### 4.1 Trap design

MAT (Microwave Addressing Trap) was designed to be the simplest implementation of the addressing scheme described in chapter 2 which could be easily, quickly and inexpensively built in-house. The prototype is a planar chip ion trap and consists of gold electrodes patterned onto a fused silica substrate using standard photolithographic techniques. It features two addressing zones,  $960\mu\text{m}$  apart, each with four integrated microwave electrodes. The electrode layout is described in subsection 4.1.1, the design of the radio-frequency and static Paul-trap confinement potentials is detailed in subsection 4.1.2 and 4.1.3 respectively. The socket used to feed microwaves into the trap is described in subsection 4.1.4. Finally, microwave design and simulations are discussed in subsection 4.1.5.

#### 4.1.1 Electrode layout

MAT is a linear surface-electrode ion-trap with electrodes laid out as in fig. 4.1. The RF rails provide confinement in the radial ( $xy$ )–plane. In order to prevent the introduction of unwanted phase-shifts between these rails, they are joined

at one end of the chip so that they can both be driven by a single RF feed. DC voltages applied to the segmented DC electrodes on either side of the RF rails provide axial confinement and can also be used to shuttle the ion along the trap axis. The two unsegmented DC electrodes that lie between the RF rails are used to tilt the radial potential to allow for efficient Doppler cooling, as explained in section 4.1.3.

MAT's two microwave-addressable trapping zones, which we label zone A and zone B, are each fitted with two pairs of integrated microwave electrodes. These are simply DC electrodes that have been patterned with a T-shaped slot around which a microwave current can flow: one side of the slotted electrode is driven with a microwave source and the other side is grounded via a capacitor, creating a microwave current loop which in turn produces a microwave magnetic field at the ion. Note that such electrodes must also continue to function as DC control electrodes (in order to keep the ion axially confined), so we must be able to bias the electrode with a DC voltage, as well as driving it at microwave frequencies. Grounding the electrode's current loop via a capacitor permits the required DC voltages to be sustained and bias-T's at each electrode's feedthroughs are used to apply both DC and microwave drives.

In order to reduce crosstalk between zones, the microwave electrodes are separated by widening sections of ground plane as we move away from the centre of the trap. The gaps between the electrodes in the chip's centre are kept small (5 and 10  $\mu\text{m}$ ) in order to reduce charging effects, but are increased to 25  $\mu\text{m}$  as we move away from the trapping zones in order to improve fabrication yield (dust can bridge small gaps and cause shorts in the electroplating stage of fabrication).



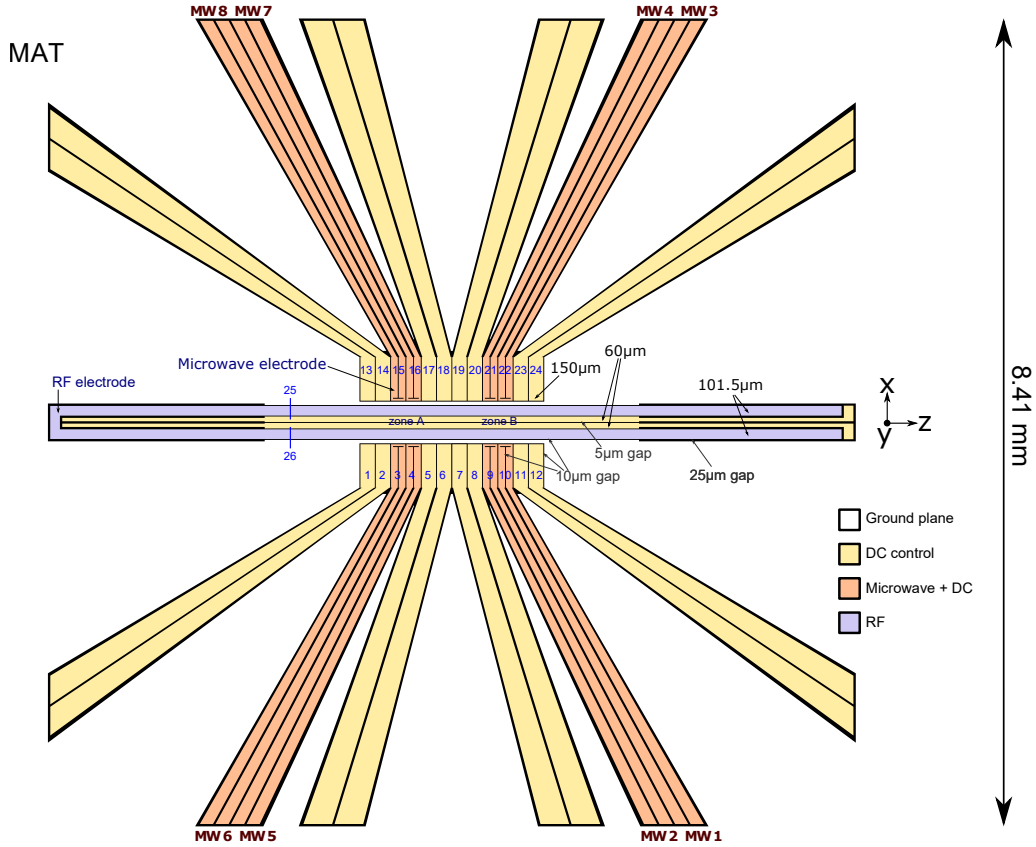


Figure 4.1: To scale schematic of MAT's electrode layout showing DC control electrodes (yellow, numbered from 1–26), microwave control electrodes (orange, labelled MW1–MW8) and the RF rail (purple).

#### 4.1.2 RF trapping potentials

In [Hou08], House derives analytic expressions for the electrostatic potential field above surface electrodes arranged in a 5-wire geometry (fig. 4.2). We use these expressions to choose the dimensions of the trap's electrodes. With electrode dimensions defined as in fig. 4.2 and denoting the angular frequency of the trapping RF as  $\Omega_{\text{RF}}$ , the trap's stability parameters as  $a_s$  and  $q_s$ , the trap depth in eV as  $T_d$ , the ion height above the surface of the trap as  $h$  and the voltage amplitude applied to the trap's RF electrodes as  $V_{\text{RF}}$ , the relevant equations are:

$$b = \frac{4h^2 - a^2}{2a} \quad (4.1)$$

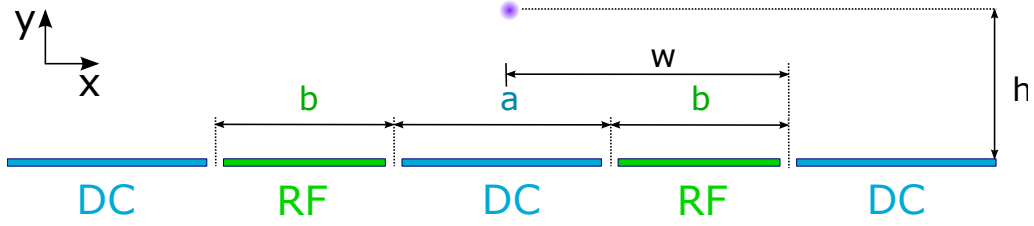


Figure 4.2: Schematic of the symmetric version of the five-wire surface-electrode geometry (cross section perpendicular to the trap axis) labelled with the dimensions relevant to House's analytic equations describing the electrostatic potential field above the trap. DC electrodes are depicted in blue and RF electrodes in green.

$$V_{\text{RF}} = \frac{-q_s \pi m \Omega_{\text{RF}}^2}{16e} \cdot \frac{(a+b)^2 \sqrt{a(a+2b)}}{b} \quad (4.2)$$

$$T_d = \left( \frac{e V_{\text{RF}}}{\pi \Omega_{\text{RF}} \sqrt{m}} \cdot \frac{b}{(a+b)^2 + (a+b) \sqrt{2ab+a}} \right)^2 \quad (4.3)$$

We choose  $a$  with three criteria in mind:

1. To reduce crosstalk, we would like the microwave electrodes in a given zone to be much closer to the ion trapped in that zone than to the neighbour ions. Hence, we want  $w$  to be small compared to the inter-zone distance;
2. The trap must be deep enough to prevent rapid collision-induced ion loss; we would like  $T_d \gtrsim 100 \text{ meV}$ ;
3. The RF voltage applied across the  $10 \mu\text{m}$  gaps between the RF and DC electrodes must be comfortably below the breakdown voltage of the fused silica substrate, which is  $250 \text{ V}/10 \mu\text{m}$ . Hence, we need  $V_{\text{RF}} < 250 \text{ V}$ .

To plot the dependence of  $w$ ,  $T_d$  and  $V_{\text{RF}}$  on  $a$ , we must first set  $\Omega_{\text{RF}}$  and the ion height,  $h$ . We choose an ion height of  $h \sim 110 \mu\text{m}$ , axial frequency  $\omega_{\text{axial}} = 1 \text{ MHz}$ , radial secular frequency  $\omega_{\text{radial}} = 4 \text{ MHz}$  and stability parameter  $q_s = 0.3$  and  $a_s \approx 0$ . Using the equations derived in chapter 1, this gives  $\Omega_{\text{RF}} = 38 \text{ MHz}$ . With these values, we plot the dependence of  $b, w$ ,  $T_d$  and  $V_{\text{RF}}$  on  $a$  in fig 4.3.

Choosing  $a = 135\mu\text{m}$  gives  $w = 190.75\mu\text{m}$ ,  $T_d = 119\text{meV}$  and  $V_{\text{RF}} = 177$ ; a good compromise between minimising  $w$  and maximising  $T_d$ .

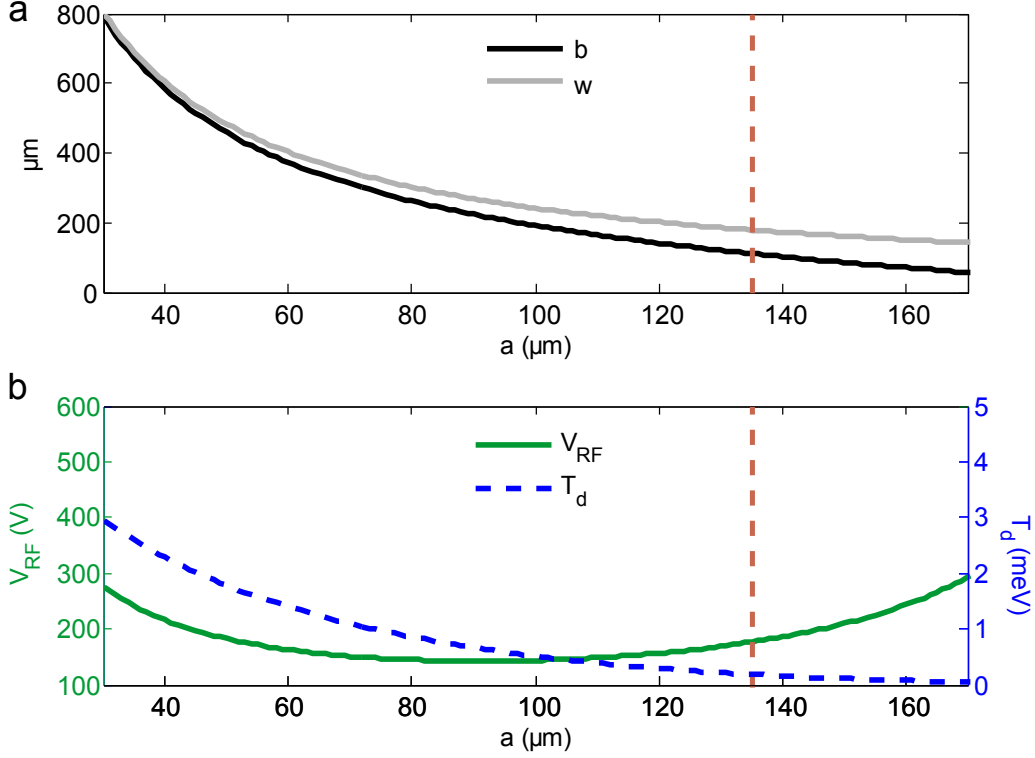


Figure 4.3: Plot of dependence of trap parameters on the centre electrode width,  $a$ , as given by eqns. 4.1, 4.2, 4.3: a) plots the RF electrode width,  $b$ , and the distance between the outer edge of the RF electrodes and the trap axis,  $w$  (see fig. 4.3). b) plots the required RF voltage amplitude,  $V_{\text{RF}}$ , and the trap depth,  $T_d$ , as a function of  $a$ . The vertical red dashed line indicates the chosen value of  $a = 135\mu\text{m}$ .

### 4.1.3 DC trapping potentials

In [Hou08], House also provides a formula for the electrostatic potential above any rectangular surface-electrode, which we can use to calculate which DC voltages to apply to our DC control electrodes. If we take the trap electrodes to lie on the  $y = 0$  plane and take  $(x_{i,1}, z_{i,1})$ ,  $(x_{i,2}, z_{i,2})$  to be the coordinates for electrode  $i$ 's diagonally opposing corners, the potential  $\phi_i$  generated at point  $(x, y, z)$  above the trap when voltage  $V$  is applied to DC electrode  $i$  is given by:

$$\begin{aligned}
 \phi_i(x, y, z) = \frac{V}{2\pi} \{ & \arctan \left[ \frac{(x_{i,2} - x)(z_{i,2} - z)}{y\sqrt{y^2 + (x_{i,2} - x)^2 + (z_{i,2} - z)^2}} \right] \\
 & - \arctan \left[ \frac{(x_{i,1} - x)(z_{i,2} - z)}{y\sqrt{y^2 + (x_{i,1} - x)^2 + (z_{i,2} - z)^2}} \right] \\
 & - \arctan \left[ \frac{(x_{i,2} - x)(z_{i,1} - z)}{y\sqrt{y^2 + (x_{i,2} - x)^2 + (z_{i,1} - z)^2}} \right] \\
 & + \arctan \left[ \frac{(x_{i,1} - x)(z_{i,1} - z)}{y\sqrt{y^2 + (x_{i,1} - x)^2 + (z_{i,1} - z)^2}} \right] \} \quad (4.4)
 \end{aligned}$$

Following [All11], we can take the potential to be approximately quadratic around the ion and fit it with a curve of the form:

$$\phi_i(x, y, z) = \alpha_{x,i}x^2 + \alpha_{y,i}y^2 + \alpha_{z,i}z^2 + \beta_{x,i}x + \beta_{y,i}y + \beta_{z,i}z + \text{const} \quad (4.5)$$

We can hence write a vector of quadratic coefficients  $\mathbf{c}_i = (\alpha_{x,i}, \alpha_{y,i}, \alpha_{z,i}, \beta_{x,i}, \beta_{y,i}, \beta_{z,i})^\top$  that defines the magnitude of the potential produced at point  $(x, y, z)$  when electrode  $i$  is held at 1V (we ignore the constant term, as it has no physical importance). The voltages that need to be applied to each of MAT's DC control electrodes in order to generate a desired DC potential around an ion trapped at  $(x, y, z)$  are then determined by solving the following matrix equation:

$$\boldsymbol{\phi}_{\text{coeffs}} = \mathbf{M}_{\text{dc}} \mathbf{v} \quad (4.6)$$

where  $\mathbf{M}_{\text{dc}} = [\mathbf{c}_1 \dots \mathbf{c}_n]$  is the  $6 \times n$  matrix whose columns are the coefficient vectors for of the  $n$  DC control electrodes we are using to apply the potential,  $\boldsymbol{\phi}_{\text{coeffs}}$  is the coefficient vector for the desired total DC potential around  $(x, y, z)$  and  $\mathbf{v} = (v_1, v_2, \dots, v_n)^\top$  is the vector of voltages that must be applied to the  $n$  DC electrodes in order to generate the desired potential. Since most of the time we will use more than six DC electrodes to apply the potential (i.e.  $n > 6$ ), solving eq 4.6 for  $\mathbf{v}$  is an under-constrained inverse problem. Following [SPM<sup>+</sup>10],

we solve it by employing the Tikhonov method, i.e. finding the  $\mathbf{v}$  which minimises  $(\|\mathbf{M}_{\text{dc}}\mathbf{v} - \boldsymbol{\phi}_{\text{coeffs}}\|^2 + \epsilon\|\mathbf{v}\|^2)$ , where  $\epsilon$  is a tolerance parameter. This method is straightforwardly implemented using MATLAB's *fmincon* function, which we use with the tolerance parameter set to  $\epsilon = 10^{-3}$  and the elements of  $\mathbf{v}$  constrained to lie between -10V and 10V, to conform to the range of voltages our DAC can output.

To operate the trap, we must determine the electrode voltage configurations required to apply four forms of DC potentials to the ion:

- **End-cap potential:** the axial confinement potential. This will harmonically confine the ion along the  $z$ -direction, while providing the equal and opposite anti-trapping in the radial plane required by Laplace's equation. It will have the form  $\phi_{\text{endcap}}(x, y, z) = \alpha_{\text{endcap}}z^2 - \frac{\alpha_{\text{endcap}}}{2}(x^2 + y^2)$  and a coefficient vector  $\boldsymbol{\phi}_{\text{endcap}} = (\frac{\alpha_{\text{endcap}}}{2}, \frac{\alpha_{\text{endcap}}}{2}, \alpha_{\text{endcap}}, 0, 0, 0)^T$ ;
- **Micromotion compensation potentials:** linear potentials in  $x$  and  $y$  that compensate for any stray electric fields that might shift the ion off the RF null. We label these potentials 'x-comp' and 'y-comp', and they have coefficient vectors of the form  $\boldsymbol{\phi}_{\text{x-comp}} = (0, 0, 0, \beta_{\text{x-comp}}, 0, 0)^T$  and  $\boldsymbol{\phi}_{\text{y-comp}} = (0, 0, 0, 0, \beta_{\text{y-comp}}, 0)^T$ , respectively;
- **Tilt potential:** a quadrupole potential in the  $xy$ -plane used to tilt the normal mode axes of the ion's motion in the RF trap. The need for this arises due to the following: in order to simplify nulling of the microwave field and the choice of DC control voltages, all of MAT's electrodes were made to be symmetric about the trap axis. However, symmetric RF surface electrodes produce an almost entirely cylindrically symmetric RF potential: the degeneracy is only lifted by the fact that the RF trap generated by surface electrodes is not as deep in the direction perpendicular to plane of the electrodes, and hence the potential is very slightly elliptic, with one nor-

mal mode axis perpendicular to the trap surface, and the other parallel to it. To Doppler cool the ion, we must ensure that its normal modes of motion both have components that lie along the cooling laser's  $k$  vector. Since we cannot crash our UV cooling laser onto the surface of the trap without both charging the surface -via the photoelectric effect- and scattering photons that would disrupt readout, we must align our cooling laser's  $k$ -vector parallel to the plane of the trap's electrodes, precluding the cooling of any motion normal to the surface of the trap. Following [ASS<sup>+</sup>10], we resolve this by splitting the centre DC electrode into two electrodes and applying to them a DC voltage differential: this produces a static, tilted quadrupole potential which, when superimposed with the time-averaged RF potential (the 'pseudo-potential') seen by the ion, results in an elliptical pseudopotential with tilted normal mode axes. We choose to tilt the potential by  $45^\circ$  and, to avoid quadratic cross terms in  $xy$  (which would make solving for the electrode voltages needed to generate this tilt potential a non-linear problem), we write the tilt potential's coefficient vector in a basis that has been rotated by the same angle:  $(x' = \frac{(x+y)}{2}, y' = \frac{(x-y)}{2}, z' = z)$ :  $\phi_{\text{tilt}} = (\alpha_{\text{tilt}}, -\alpha_{\text{tilt}}, 0, 0, 0, 0)^T$ . To solve equation 4.6 for the electrode voltages needed to generate the tilt potential, we must therefore use a version of the matrix  $M_{\text{dc}}$  that has been calculated in the rotated basis.

Let us label the voltage vectors that generate the above potentials for a given choice of  $\alpha_{\text{endcap}}$ ,  $\alpha_{\text{xcomp}}$ ,  $\alpha_{\text{ycomp}}$  and  $\alpha_{\text{tilt}}$  as  $\mathbf{v}_{\text{endcaps}}$ ,  $\mathbf{v}_{\text{xcomp}}$ ,  $\mathbf{v}_{\text{ycomp}}$ ,  $\mathbf{v}_{\text{tilt}}$ , respectively. When operating the trap, we generally need to apply and scale all four of these potentials simultaneously. The total voltage  $v_i$  applied to DC electrode  $i$  will be equal to the  $i$ 'th element of the vector  $\mathbf{v}_{\text{total}} = k_1 \mathbf{v}_{\text{endcaps}} + k_2 \mathbf{v}_{\text{xcomp}} + k_3 \mathbf{v}_{\text{ycomp}} + k_4 \mathbf{v}_{\text{tilt}}$ , where  $k_1, \dots, k_4$  are scale factors adjusted during the experiment (for example,  $k_2$  and  $k_3$  may need to be adjusted as the stray electric fields that are pushing the ion off the RF null drift). We can think of  $\mathbf{v}_{\text{endcaps}}$ ,  $\mathbf{v}_{\text{xcomp}}$ ,

Table 4.1: A possible set of DC basis voltages, calculated for  $\alpha_{\text{endcap}} = 1.02 \times 10^6 \text{V/m}^2$  (giving a DC axial confinement frequency of 500kHz),  $\alpha_{\text{xcomp}} = \alpha_{\text{ycomp}} = 1 \text{V/m}$  and  $\alpha_{\text{tilt}} = 1.00 \times 10^7 \text{V/m}^2$ . This basis set is for an ion trapped close to the center of the trap, with only the surrounding three pairs of DC electrodes (6-8 and 18-20) and the split centre DC electrodes (25,26) in use (all other electrodes are held at 0V).

| DC electrode number | $v_{\text{endcaps}}$ (mV) | $v_{\text{xcomp}}$ (mV) | $v_{\text{ycomp}}$ (mV) | $v_{\text{tilt}}$ (mV) |
|---------------------|---------------------------|-------------------------|-------------------------|------------------------|
| 6                   | 818                       | 0.097                   | 5.47                    | 790                    |
| 7                   | -4571                     | 0.16                    | -0.66                   | 1210                   |
| 8                   | 814                       | 0.097                   | 5.48                    | 790                    |
| 18                  | 818                       | -0.097                  | 5.47                    | -790                   |
| 19                  | -4571                     | -0.16                   | -0.66                   | -1210                  |
| 20                  | 814                       | -0.097                  | 5.48                    | -790                   |
| 25                  | -325                      | 0                       | 0.46                    | -523                   |
| 26                  | -325                      | 0                       | 0.46                    | 523                    |

$v_{\text{ycomp}}$ ,  $v_{\text{tilt}}$  as a voltage basis set for trap operation. Table 4.1 shows an example voltage basis set for an ion trapped in zone B, with only the surrounding four pairs of DC electrodes and the split centre DC electrodes in use. Figure 4.4 shows plots of the potentials generated at the ion when each of the basis voltage vectors given in the table are applied.

#### 4.1.4 Trap packaging

In this initial experiment, the trap was packaged using a readily available Ceramic Pin Grid Array (CPGA)<sup>1</sup>. This is a chip holder with 100 gold-plated pads on the top side that are wired to an array of gold pins on the bottom side (see fig. 4.7a). This package had been previously used by several ion-trapping groups [Sti07], who developed UHV-compatible Vespel sockets to connect the CPGA pins to in-vacuum DC cables. These consist of two Vespel frames which sandwich together gold pin receptacles that have been crimped onto in-vacuum cables (see figs. 4.7c and 4.9). One such Vespel socket frame had been previously used in our group and was available for this experiment, providing a quick, albeit non-ideal, solution for a first proof-of-principle prototype. In sub-

<sup>1</sup>Kyocera CPG10018

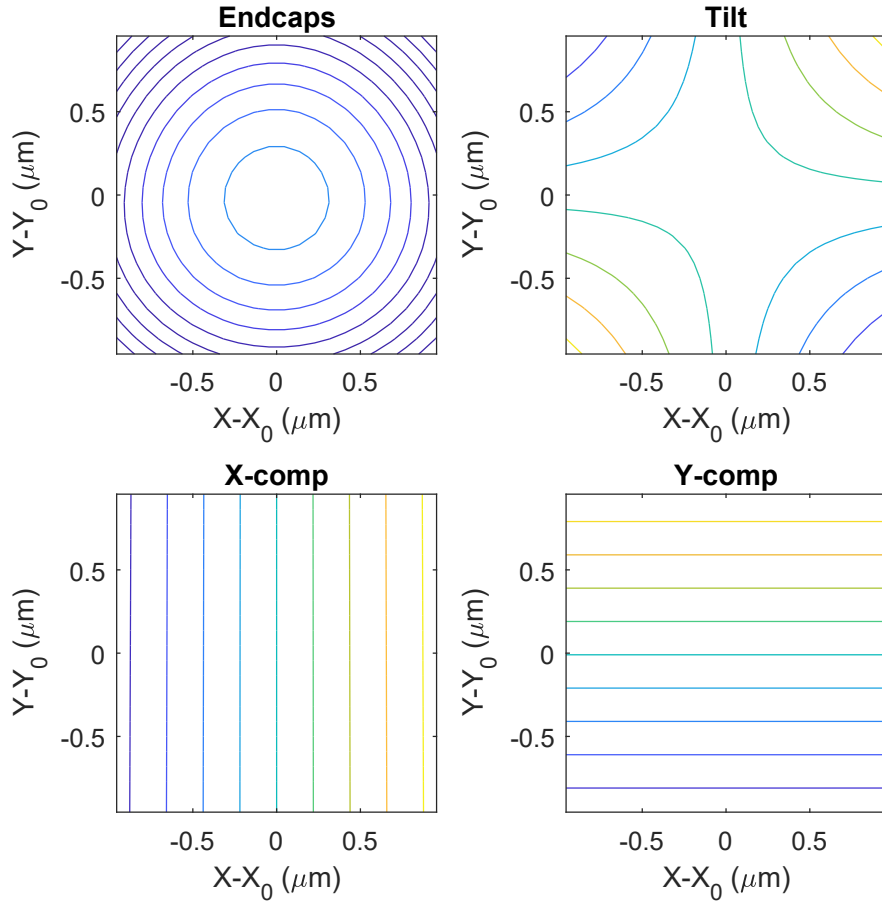


Figure 4.4: DC potentials produced in the  $xy$ -plane (perpendicular to the trap axis) when the basis set voltages given in table 4.1 are applied to MAT's electrodes. The plots are centered around the position where the ion is trapped,  $(x = x_0, y = y_0)$ . Note that the contour scales are not the same in each plot - these merely give an idea of the shape of the potentials generated by each voltage basis set.

sequent work, we invested time in designing a fully impedance-matched, well-characterised printed-circuit-board socket which could be fully modelled in our finite element simulations (see chapter 5).

To obtain a rough measure of the expected power loss through the CPGA pins and inner wiring, the test rig depicted in fig. 4.5 was built. A short section of  $50\Omega$  transmission line was glued to the metallised top of the CPGA and wire-bonded onto two pads, one connected to port A and the other connected to port B. To measure the loss through the package, microwaves were fed into port A



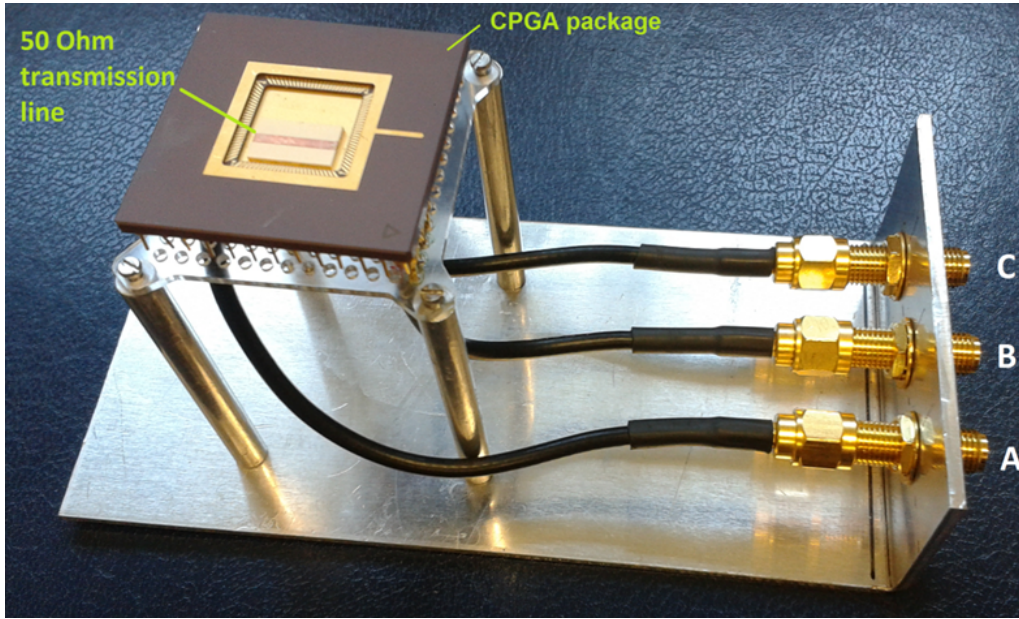


Figure 4.5: Test rig for obtaining rough measures of the loss through the CPGA package and the crosstalk between CPGA pins. In this test rig, a small section of  $50\Omega$  transmission line is placed where the trap would sit. The coaxial cables powering the rig are crimp connected to the CPGA pins in the same way as was done in the socket used for MAT.

and the power at port B was recorded. Microwave power flowed from port A through the package, onto the transmission line, through the package again and out through port B, which was connected to a power meter. The loss through the package is therefore approximately half the loss recorded at the power meter. The recorded powers across the relevant frequency range are plotted in fig. 4.6. The measurement indicates that the package has an insertion loss of between 4dB and 7dB. Since we planned to perform only single qubit operations with MAT, we required only a few milliwatts of power to reach the ion and could therefore tolerate this insertion loss.

In an attempt to gauge the levels of crosstalk that could be expected between nearby the CPGA pins, Port C was connected to a pin located three pins away from the pin connected to port B. Power was recorded at port C when microwaves were fed into port A with port B  $50\Omega$  terminated. Accounting for the insertion loss of the coaxial cables on the test-rig, the crosstalk power at the pin

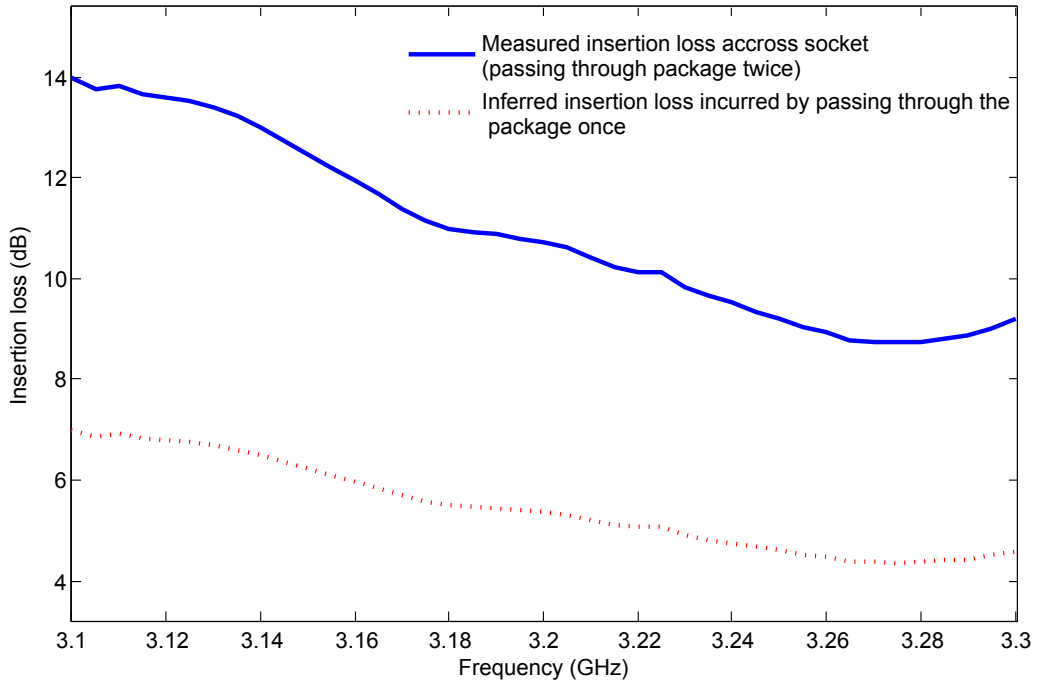


Figure 4.6: Plot of measured insertion loss of the test trap socket (blue solid line), excluding the coaxial cables' insertion loss (the cables' insertion loss was determined by multiplying the length of coaxial cable used by the dB/m insertion loss measured on an intact length of this type of coaxial cable). The red dashed line is the inferred insertion loss of the CPGA package, i.e. half the insertion loss of the socket. This measure of insertion loss should include the loss through the crimp connections of the coaxial cable to the CPGA pins, the loss through the CPGA pins and inner wiring and the loss through a wire-bond connecting a CPGA pad to the  $50\Omega$  transmission line.

connected to port C across 3.1 to 3.3 GHz was recorded to be  $\lesssim 2\%$  of the power at the pin connected to port B. This is, however, only a rough estimate since the pickup at each pin will also depend the load impedance connected to it.

It was decided that this initial experiment could be performed using the CPGA as a trap package since enough power would reach the ion and the crosstalk between CPGA pins was at few % level.

The trap chip was glued to the metallised top surface of the CPGA with a small amount of UHV-compatible epoxy<sup>2</sup> and the trap electrodes wire-bonded to the CPGA pads with  $25\mu\text{m}$  diameter gold wire (see fig. 4.8). Microwave signals are brought to the CPGA by in-vacuum coaxial SMA cables: the connection

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<sup>2</sup>Epotek 353ND

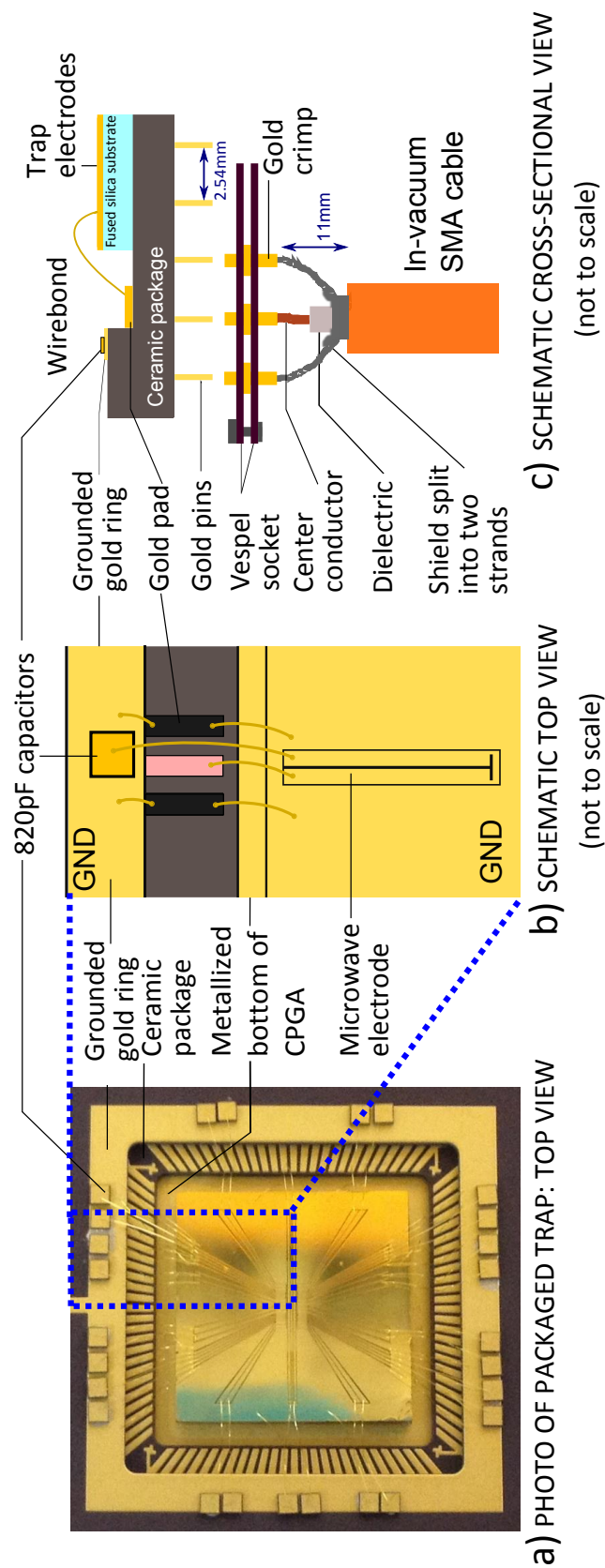


Figure 4.7: Trap packaging. a) Top-view photograph of trap chip mounted on the CPGA. Chip capacitors can be seen glued onto the metallised outer ring of the CPGA. b) Schematic representation of the wire-bond connections between a microwave electrode on the trap chip, the CPGA pads and the capacitors on the outer ring. The pads shaded black are connected to the grounded shield of the coaxial cable on the back side of the CPGA and the pad shaded pink is connected to the cable's centre conductor. c) Schematic diagram of the connection between the in-vacuum coaxial microwave cable and the CPGA socket.

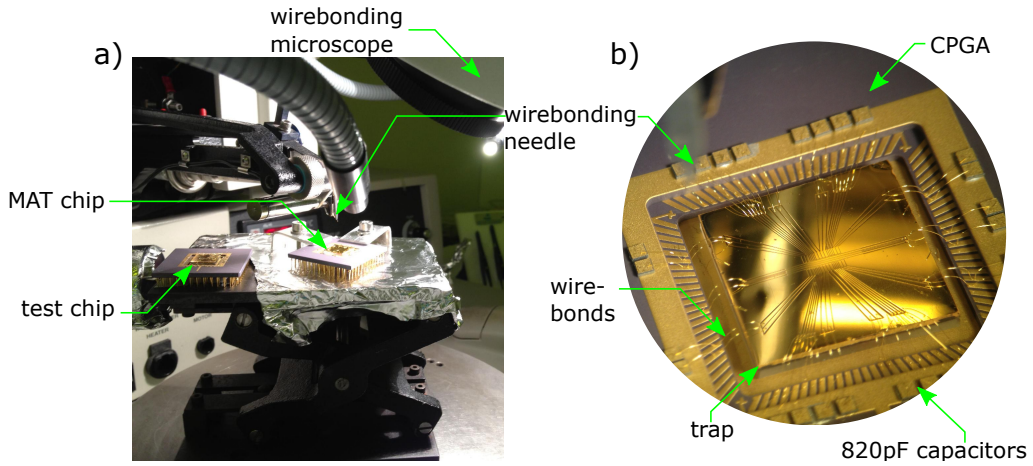


Figure 4.8: a) Wire-bonding setup: the trap chip is mounted on clean foil on a wire-bonding machine (a test chip, used for wire-bonding practice and to calibrate the machine parameters, is seen on the left). b) View of the trap through the wire-bonding machine's microscope.

to the pins is made by splitting the SMA shield and connecting the shield halves and the centre conductor to three neighbouring CPGA pins, which in turn connect onto adjacent pads on the top surface. This ensures that the microwave centre conductor is always surrounded by two ground wires in a make-shift coplanar waveguide (CPW) configuration. This is an attempt to minimise ground discontinuities and reduce reflections in the coaxial cable-to-CPGA transition (see fig. 4.7c). The pads that are connected to microwave ground are wire-bonded both to the ground plane of the trap and to the metallised gold ring which surrounds the pads on the top surface of the CPGA (see fig. 4.7b). The ring is also used to ground one side of the 820pF chip capacitors<sup>3</sup> that are used to filter the DC electrodes and to allow DC biasing of the microwave electrodes - we glue the capacitors onto the ring using conductive silver epoxy<sup>4</sup>, taking care not to short the top and bottom surfaces. The metallised CPGA surface underneath the trap is also grounded via eight wire-bonds that connect it to the trap ground plane and are uniformly distributed around the chip.

To prevent the very fragile crimp connections between the coaxial cables and

<sup>3</sup>Complex Corp CSM - 200 - 35 x 35 x 8 - G - 821 - M

<sup>4</sup>Epotek H20E silver epoxy

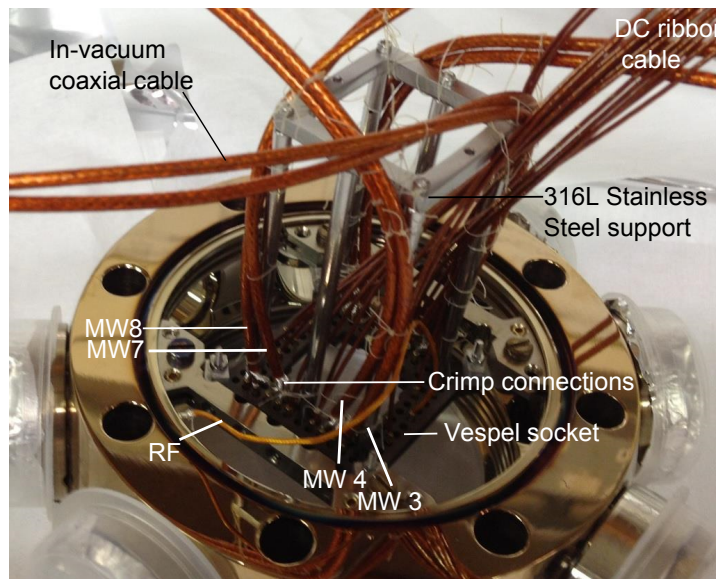


Figure 4.9: Vespel socket assembly installed on the vacuum system's CF-flanged octagon. The in-vacuum SMA cables are tied with Kapton string to a 316L stainless steel support which provides strain relief, preventing the crimp connections to the socket from breaking when the cables are bent into the vacuum system.

the socket from rupturing under the small amount of strain they would experience during the installation of the socket assembly in the vacuum system, a 316L stainless steel support rig was built and fastened to the socket (see fig. 4.9). The rig provides strain relief to the cables, which are fastened to it using Kapton string.

#### 4.1.5 Microwave simulation

The trap was simulated using Ansys HFSS 3D full wave electromagnetic finite element simulation software. The tool works by allowing the user to draw or import a geometry which is then meshed by the software. Maxwell's equations are solved on the mesh points and the mesh is subsequently refined between points where the solution appears to be changing quickly. Such mesh refinements are termed "adaptive passes": HFSS repeats the process of adaptive mesh refinement and solving until the convergence criteria are met or the computing resources are exhausted.



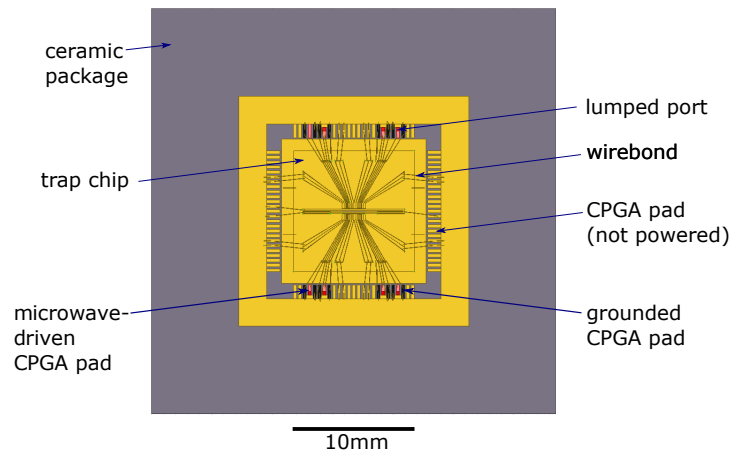


Figure 4.10: Trap model in HFSS. The model includes the top surface of the CPGA and the wire-bonds, but does not include the wiring between the CPGA pads and pins. Power is instead delivered to the CPGA pads via lumped ports (see fig. 4.11), which are the red squares in this model. CPGA pads carrying microwave signal are coloured pink and the grounded pads are coloured black.

The HFSS model includes the entire trap electrode geometry, the top side of the CPGA and all wire-bonds (see fig. 4.10). However, lack of information about the inner wiring of the CPGA package between the pads on the top side and the pins on the bottom meant that these connections could not be modelled. Similarly, the hand-made crimp connections between the coaxial cable and the CPGA could not be faithfully reproduced in the model: instead, power is delivered directly to the CPGA pads using “lumped ports”, which act as voltage sources of  $50\Omega$  characteristic impedance (see fig. 4.11). The chip capacitors are modelled as shorts since they have very low impedance at microwave frequencies. The RF rail was connected by a wire-bond to a pad that was left floating in the model, because the impedance seen by microwaves picked up by the RF rail was not designed for and was hence difficult to establish before the system was assembled. As explained in chapter 3, the RF rail is connected to the resonator by a single-pin feedthrough into the vacuum system and a copper wire crimp-connected to one of the CPGA pins. Therefore, the microwaves will see a combination of the inductance of the CPGA package wiring, the inductance of the copper wire to the feedthrough, the capacitance between this wire and the

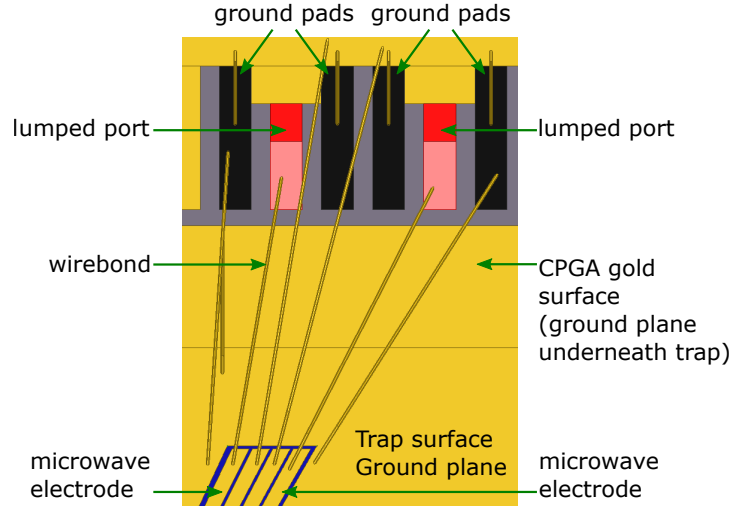


Figure 4.11: Lumped ports in HFSS. These ports apply a microwave voltage to the CPGA pads (pink) with respect to the grounded pads (black). Recall that we place one grounded pad either side of the live microwave pad in order to maintain a coplanar waveguide configuration and reduce reflections. In the model, the two grounded pads are shorted by rectangular gold section (seen above the red lumped ports in this figure) to ensure that they are at the same potential: this allows the lumped port to be drawn in such a way that it can apply a microwave voltage to the pink pad with respect to the common ground between the black pads.

vacuum can, the feedthrough capacitance and the RF resonator load, none of which could be faithfully modelled.

From simulation results, we can plot the predicted magnitude of the qubit-driving  $\pi$ -polarised component of microwave magnetic field generated along the trap axis when each microwave control electrode is powered in turn and all others are  $50\Omega$  terminated. Figures. 4.12 and 4.13 show plots of the field component that would couple to the qubit transition when the static  $B$  field (and hence the quantisation axis) is aligned perpendicular to and along the trap axis respectively.

As described in section 2, to predict the currents  $I_1 \dots I_8$  that must be fed to the eight microwave electrodes in order produce a desired field at the ions in two trap zones, we must solve equation 2.9 using the  $6 \times 8$  matrix  $\mathbf{M}$  - extracted from the HFSS model - that describes the fields generated at each ion by each control electrode. For this trap, the problem is under-constrained because we use four microwave electrodes per zone when only three would be required

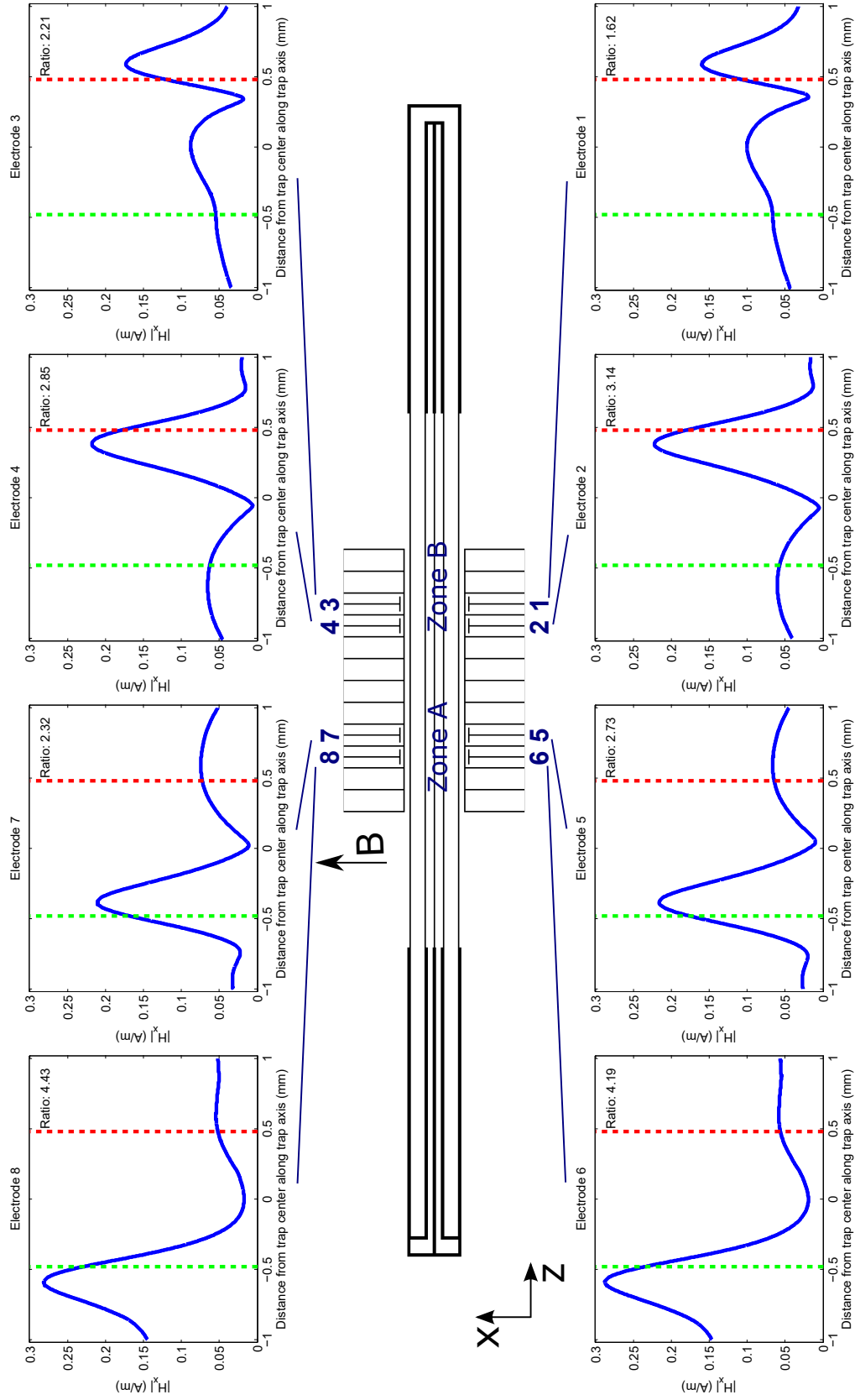


Figure 4.12: Here, the static  $B$  field is aligned along the  $x$ -direction, perpendicular to the trap axis. The component of the microwave field that couples to the qubit is therefore the  $x$ -component, since the qubit transition is  $\pi$  polarised. This figure shows plots of the modelled magnitudes of the  $x$ -component of the  $H$ -field versus distance along the trap axis ( $z$ -axis) when each electrode is, in turn, powered with 1mW of microwaves at 3.2GHz. The dashed lines in each plot indicate the  $z$ -coordinates where the ion would be trapped in addressing zones A (green) and B (red).



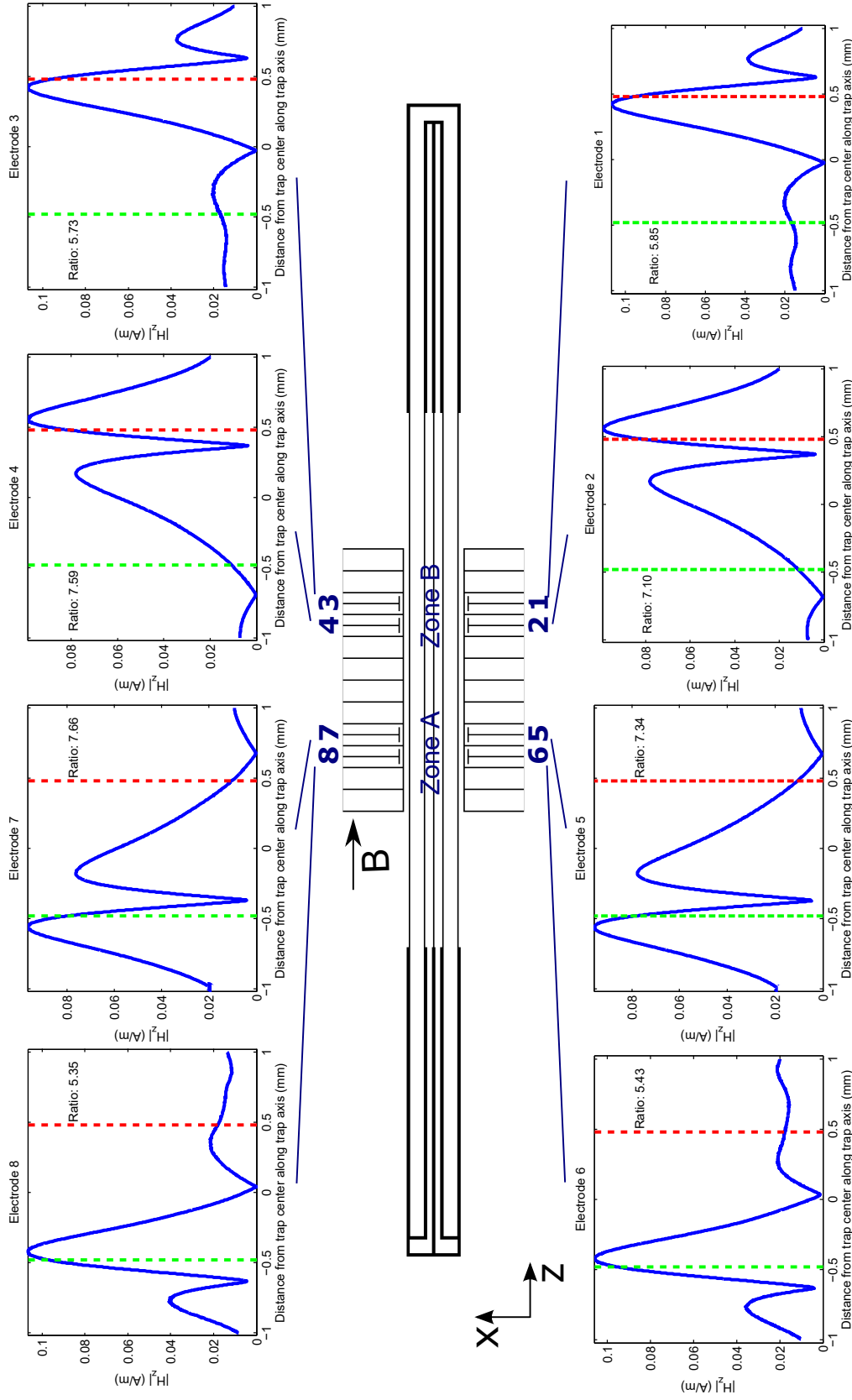


Figure 4.13: With the static  $B$  field aligned along the  $z$ -direction (parallel to the trap axis), the component of the microwave field that couples to the qubit is the  $z$ -component. This figure shows plots of the modelled magnitudes of the  $H$ -field versus distance along the trap axis ( $z$ -axis) when each electrode is, in turn, powered with 1mW of microwaves at 3.2GHz. The dashed lines in each plot indicate the  $z$ -coordinates where the ion would be trapped in addressing zones A (green) and B (red).

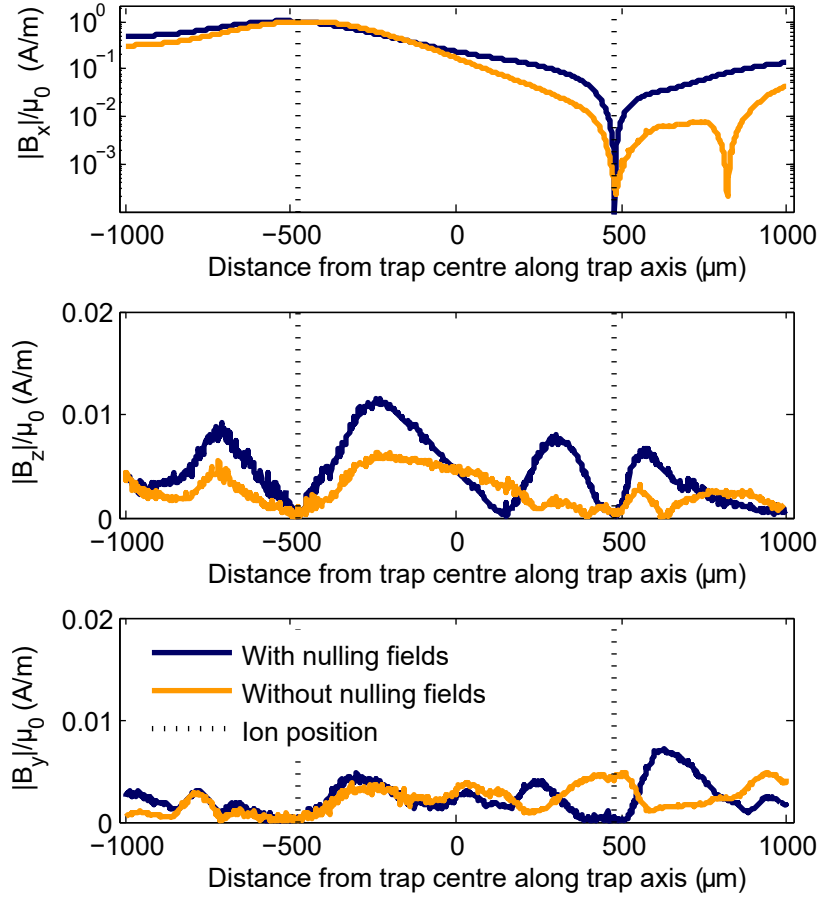


Figure 4.14: Simulated  $x$  (top),  $z$  (middle) and  $y$  (bottom) components of the microwave  $B$ -field amplitude plotted along the trap axis ( $110 \mu\text{m}$  above the electrode surface) when ion 1 is addressed with a field amplitude  $B/\mu_0 = 1 \text{ A/m}$  ( $B \approx 13 \text{ mG}$ ) in the  $x$  direction with (dark blue) and without (orange) the use of nulling fields. Note that the plot for the  $x$  component is on a log scale.

to fully control the polarisation of the microwave field applied to each ion, i.e. the matrix  $\mathbf{M}$  has a two-dimensional null-space and there are infinitely many choices of current vectors  $\mathbf{I} = (I_1 \dots I_8)^\top$  that would produce a desired field at the ions. We choose the particular solution which minimises the magnitude of the currents needed (i.e. the solution that minimises  $\|\mathbf{I}\| = \sqrt{|I_1|^2 + \dots + |I_8|^2}$ ), given by:

$$\mathbf{I} = \text{pinv}(\mathbf{M}) \cdot \mathbf{B} \quad (4.7)$$

where  $\text{pinv}(\mathbf{M})$  is the Moore-Penrose pseudo-inverse of  $\mathbf{M}$ .

To establish a measure of the intrinsic crosstalk between trap zones, without the use of nulling fields, we can calculate the minimum field generated in the neighbour zone when we use only the electrodes in the addressed zone to apply a field to the addressed ion. We now solve equation 2.9 for a reduced  $3 \times 4$  matrix,  $\mathbf{M}_{3 \times 4}$ , which describes the three field components generated at the addressed ion by the 4 electrodes in the addressed zone. The problem is again under-constrained (4 electrodes define 3 field components) and will have solutions of the form:

$$\mathbf{I} = \mathbf{I}_p + \tilde{c}_n \mathbf{I}_n \quad (4.8)$$

where  $\mathbf{I}_p = \text{pinv}(\mathbf{M}_{3 \times 4}) \cdot \mathbf{B}$  is a particular solution,  $\tilde{c}_n$  is a complex number and  $\mathbf{I}_n$  is a vector in the 1 dimensional null space of  $\mathbf{M}_{3 \times 4}$ . We want the solution which produces minimum crosstalk, i.e. the solution which minimises the magnitude of the total field vector ( $\|\mathbf{B}\| = \sqrt{|B_x|^2 + |B_y|^2 + |B_z|^2}$ ) at the neighbour zone. We find such a solution by doing a constrained search through the nullspace of  $\mathbf{M}_{3 \times 4}$ , which involves stepping through values of the magnitude and phase of  $\tilde{c}_n$  between 0 and 25 and 0 and  $2\pi$  respectively. Fig. 4.14 shows the simulated fields along the trap axis generated when a field  $B/\mu_0 = 1 \text{ A/m}$  is applied to the ion trapped in zone A along the  $x$  direction, with and without the use of nulling fields. From the plot, the simulations predicted a crosstalk of  $\approx 0.5\%$ , even without the use of nulling fields.

## 4.2 Trap characterisation

### 4.2.1 Ion lifetime and heating rate

Fig. 4.15 shows a camera photograph of the fluorescence from one of the first ions trapped in MAT. After the soft bake, the vacuum system reached a pressure of  $3 \times 10^{-11}$  mbar and, even before fully optimising trapping parameters, we observed ion lifetimes of order 1.5 hours. During a period of 1 month between

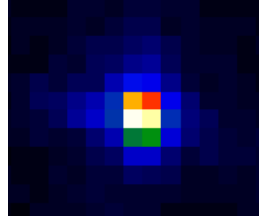


Figure 4.15: Camera photograph of the fluorescence emitted by one of the first ions trapped in MAT.

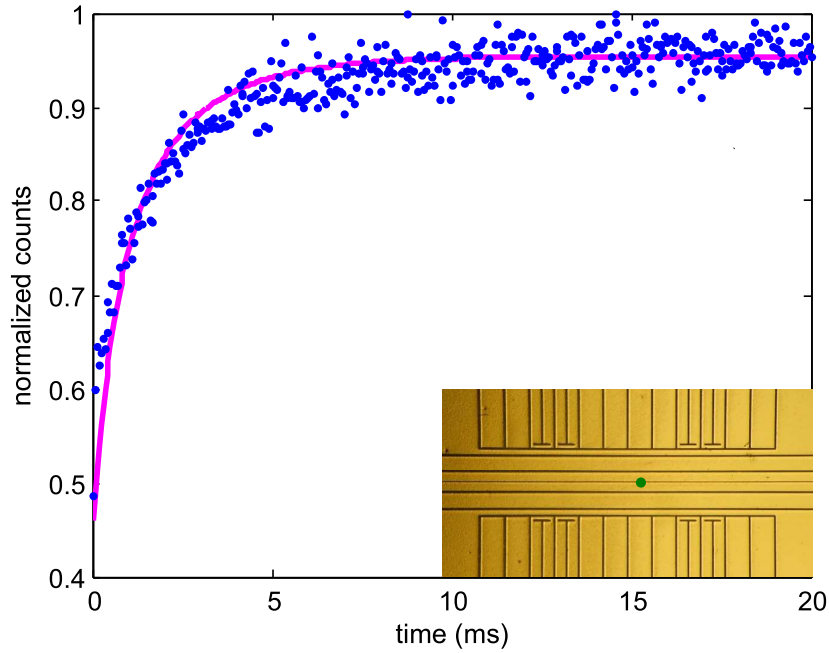


Figure 4.16: Doppler re-cool fluorescence profile measured with a heating delay of 400ms and an axial trap frequency of 0.5MHz. The fit to the data indicates a heating rate of 0.4 quanta/ $\mu$ s. Inset: Top view of trap electrodes with a green dot indicating the position where the ion was trapped when the heating rate data in this plot was taken.

trapping the first ion in MAT and Christmas break, the pressure stayed in the low  $10^{-11}$  mbar and, as expected, was slowly decreasing. Unfortunately, after returning from Christmas break in January, we found the pressure had risen by an order of magnitude to  $4 \times 10^{-10}$  mbar and ion lifetimes were reduced to between 5 and 15 minutes. The reason for this increase was not apparent, but we believe a very slow leak must have developed due to a stress fracture in one of the vacuum feedthroughs or viewports. For the next year, the pressure remained constant.

We recently made preliminary measurements of the axial-mode heating rate of an ion trapped in the centre zone of the trap. We measure the heating rate by observing the fluorescence profile of the ion as it is Doppler cooled after an initial period of heating (with the Doppler cooling lasers turned off), a method proposed and described in [WEL<sup>+</sup>07]. Fig. 4.16 depicts the fitted fluorescence profile measured for an axial trap of frequency of 0.5 MHz and a heating period of 400 ms. The fit indicates a heating rate of 0.4 quanta/ $\mu$ s. This heating rate may not be limited by the surface quality of the trap, but rather by other sources of technical noise, since we did not perform an exhaustive search for such sources.

#### 4.2.2 Microwave characterisation

Using the microwave drive chain shown in fig. 4.18, we powered each microwave electrode in turn ( $50\ \Omega$  terminating all other electrodes) and drove Rabi flops in the two trap zones. The Rabi frequencies recorded on the qubit transition and on the ‘stretch’ transitions (indicated on fig. 4.22) gave a direct measurement of the amplitude of the  $\pi$  and  $\sigma_{\pm}$  microwave field components generated by each electrode. We refer to the  $(F = 4, m_F = -4) \leftrightarrow (F = 3, m_F = -3)$  transition the  $\sigma^+$ -stretch transition and the  $(F = 4, m_F = -4) \leftrightarrow (F = 3, m_F = +3)$  transition the  $\sigma^-$ -stretch transition because they are driven by  $\sigma^+$  and  $\sigma^-$ -polarised light, respectively (the name ‘stretch’ deriving from the fact that the transition frequency increases linearly - or ‘stretches’- with increasing static B-field).

Recall that, as explained in section 3.1.3, the preparation pulses used when we want to drive the qubit transition leave the ion in state  $|\downarrow\rangle = (F = 4, m_F = 0)$ . To drive the  $\sigma^+$  and  $\sigma^-$  stretch transitions, we prepare the  $(F = 4, m_F = +4)$  or the  $(F = 4, m_F = -4)$  stretch states by optically pumping on the  $4S_{1/2} \leftrightarrow 4P_{1/2}$  transition using  $\sigma^+$  or  $\sigma^-$ -polarised 397 light respectively. Recall also from section 3.1.4 that readout is done by selectively transferring population from  $|\downarrow\rangle$

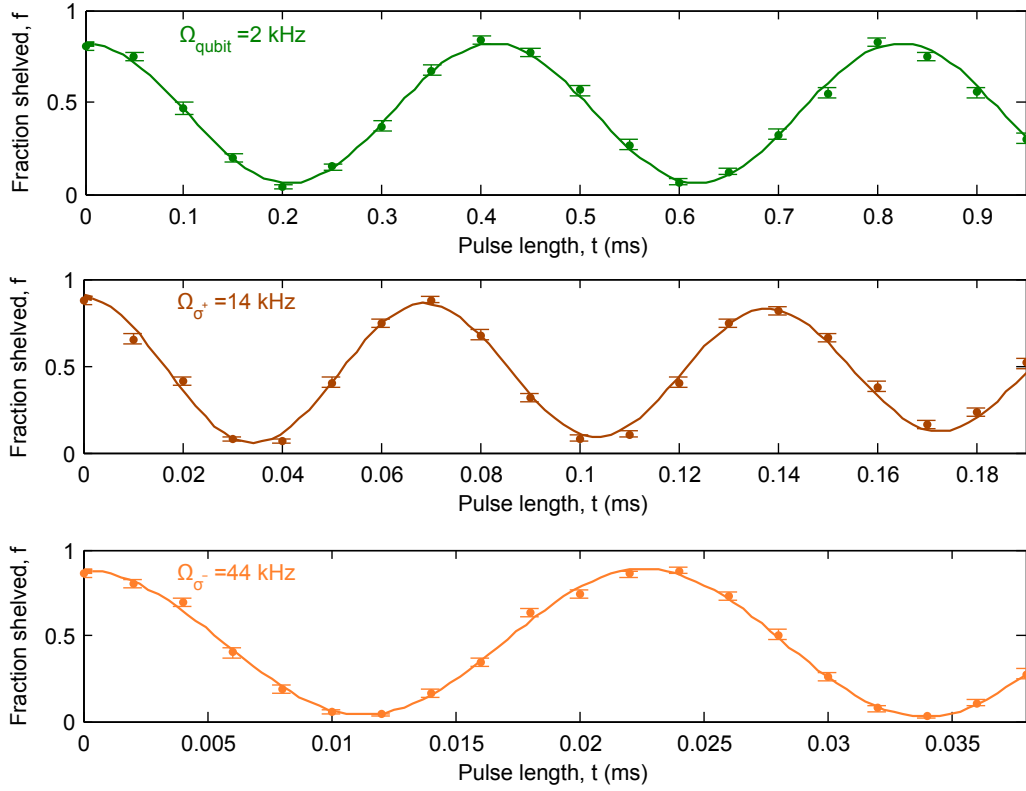


Figure 4.17: Rabi flops driven by powering microwave electrode 2 (see fig. 4.1) with microwaves tuned to resonance with the  $\pi$ -polarised qubit transition (top, green), the  $\sigma^+$  stretch transition (middle, brown) and the  $\sigma^-$  stretch transition (bottom, orange). These transitions are illustrated in the same colour code in fig. 4.22. Note that the three datasets were taken at different microwave power levels:  $\approx 50\text{mW}$  (top), and  $70\text{mW}$  (middle and bottom).

to the metastable  $3D_{5/2}$  ‘shelf’ state and looking for fluorescence on the  $397\text{nm}$   $4S_{1/2} \leftrightarrow 4P_{1/2}$  transition.

To measure Rabi frequencies in a given zone we trap an ion in that zone and apply to it a pulse sequence which consists of a microwave pulse of length  $t$  preceded by preparation laser pulses and followed by readout laser pulses. For each microwave pulse length  $t$ , we determine the final state of the ion by performing between 200 and 500 shots of the experiment and recording the fraction of the shots where the ion was shelved. We apply this experimental pulse sequence with values of  $t$  stepped between 0 and a few hundred microseconds, in what we term a ‘pulse length scan’. To be sure that none of the data taken is aliased, we follow each pulse length scan with an ‘anti-aliasing scan’ where we

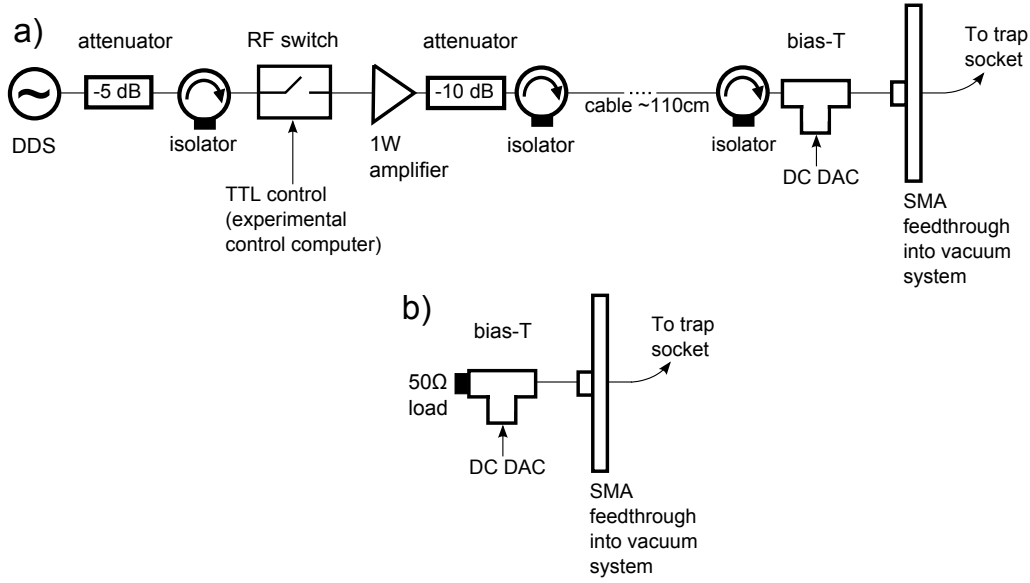


Figure 4.18: Drive chain used to characterise MAT's microwave electrode. To power a given electrode and measure Rabi frequencies, chain (a) was connected to that electrode's bias-T while all other electrodes' bias-Ts were 50Ω terminated as shown in (b).

use the same number of steps in  $t$  but divide the range of  $t$  values by an irrational number, such as  $\pi$  (e.g. having done a pulse length scan where  $t$  varied between 0 and  $200\mu\text{s}$  in 30 steps, we would follow it with an anti-aliasing scan where  $t$  varied between 0 and  $\frac{200\mu\text{s}}{\pi} \approx 64\mu\text{s}$  in 30 steps). Fig. 4.17 shows examples of Rabi flopping driven by pulse length scans with microwaves tuned to the qubit transition and the stretch transitions.

Table 4.2: Part numbers of microwave components used in the microwave drive chain shown schematically in fig. 4.18a.

| Component name | Part number                                              |
|----------------|----------------------------------------------------------|
| DDS            | AD9910 evaluation board driven by Digilent Chip-kit32Max |
| Attenuators    | Minicircuits K2-BW3+                                     |
| Isolators      | MICA Microwave T - 602S01, 20dB, 2-4GHz                  |
| Bias-T         | ZFBT-4R2G+                                               |
| Amplifier      | ZVE-8G+                                                  |
| RF switch      | ZASWA-2-50DR                                             |

We first measured Rabi frequencies on the qubit transition with the static

$B$  field aligned perpendicular to the trap axis: this gave an indication of how microwave current coupled onto the RF electrode. Currents flowing along the RF electrode would produce microwave fields perpendicular to the trap axis and would hence couple to the qubit transition if the static  $B$  field is also aligned in that direction. We see from the results in table 4.3 that, in this configuration, the ratio of the Rabi frequencies on the qubit transition measured in the zone where the powered electrode resides to those measured in the neighbour zone is  $\approx 1$  for all electrodes. We suspect this may be because the fields produced by currents coupled to the RF electrode are of approximately the same magnitude in both zones and overwhelm those produced by currents on the microwave electrodes because the RF electrode is much closer to the ion. The simulation results do not accurately match the measured ratios. This may be because it was not possible to faithfully model the RF electrode's load impedance.

Table 4.3: Rabi frequencies measured on the qubit transition with the static B-field oriented perpendicular to the trap axis, along the  $x$ -direction (see fig.4.12). Columns 2 and 3 give, for each electrode, the Rabi frequencies measured in zones A and B when each electrode is powered in turn, with all others  $50\Omega$  terminated. All Rabi frequencies have been normalised to 1mW input power. The fourth column gives, for each electrode, the ratio of the Rabi frequency measured when the ion was in the zone closest to that electrode, to the Rabi frequency measured when the ion was in the neighbour zone.

| Electrode | Rabi Frequency<br>zone A (kHz) | Rabi Frequency<br>zone B (kHz) | Rabi Frequency Ratio<br>(Near zone/ Far zone) |
|-----------|--------------------------------|--------------------------------|-----------------------------------------------|
| 1         | 6.19(3)                        | 6.17(6)                        | 1.00(1)                                       |
| 2         | 2.19(1)                        | 3.04(2)                        | 1.39(1)                                       |
| 3         | 14.5(1)                        | 10.34(1)                       | 0.713(5)                                      |
| 4         | 4.54(2)                        | 4.50(6)                        | 0.98(1)                                       |
| 5         | 13.6(1)                        | 11.39(4)                       | 1.19(1)                                       |
| 6         | 15.22(6)                       | 13.51(4)                       | 1.126(6)                                      |
| 7         | 6.33(2)                        | 4.49(1)                        | 1.410(5)                                      |
| 8         | 2.13(1)                        | 2.30(2)                        | 0.93(1)                                       |

Rotating the static magnetic field by  $90^\circ$  (so that it was aligned with the trap axis) produced the Rabi frequencies listed in tables 4.4, 4.5 and 4.6 on the qubit transition and on the  $\sigma^+$  and  $\sigma^-$  driven stretch transitions, respectively. The Rabi frequency ratios on the qubit transition were now significantly different from 1



for most electrodes, suggesting that stray currents flowing down the RF electrode were no longer coupling to the qubit. The ratios did not, however, match the simulated values listed in fig. 4.13. This is again very likely because of the limitations as to how faithfully and completely the physical apparatus could be included in the simulations: e.g. crosstalk picked up in the wiring inside CPGA package was not accounted for in the simulation and the fact that the RF electrode impedance was poorly defined will have affected the current distribution over the entire trap structure. Furthermore, because the microwave electrodes in this trap design were not directly underneath the ion, the field amplitudes produced by these electrodes at the ion have decayed to about one-tenth of their maximum value (attained directly above the electrode): at the ion, these fields are of the same order of magnitude as the fields produced by crosstalk currents induced across the trap and are hence very sensitive to the precise distribution of induced crosstalk currents, which are difficult to accurately simulate and are very sensitive to fabrication inaccuracies and any other discrepancies between model and experiment.

The fact that there were significant Rabi frequency differentials between zones when the static B-field was aligned along the trap axis allowed us to perform a demonstration of near-field addressing with nulling fields, as described in the next section.

## **4.3 Demonstration of high-fidelity addressing with nulling fields**

We performed a proof-of-principle addressing experiment using two electrodes, one in each zone, to simultaneously drive Rabi flops in the zone we chose to address and null the qubit-driving microwave field polarisation component (the  $\pi$ -polarised component) in the neighbour zone. This was done using the mi-

#### 4. MAT: MICROWAVE ADDRESSING TRAP

Table 4.4: Rabi frequencies measured on the qubit transition with the static B-field oriented parallel to the trap axis, along the  $z$ -direction (see fig.4.13). Columns 2 and 3 give, for each electrode, the Rabi frequencies measured in zones A and B when each electrode is powered in turn, with all others  $50\Omega$  terminated. All Rabi frequencies have been normalised to 1mW input power. The fourth column gives, for each electrode, the ratio of the Rabi frequency measured when the ion was in the zone closest to that electrode, to the Rabi frequency measured when the ion was in the neighbour zone.

| Electrode | Rabi Frequency<br>zone A (Hz per<br>mW) | Rabi Frequency<br>zone B (Hz per<br>mW) | Rabi Frequency Ratio<br>(Near zone/ Far zone) |
|-----------|-----------------------------------------|-----------------------------------------|-----------------------------------------------|
| 1         | 337(1)                                  | 1201(8)                                 | 3.56(3)                                       |
| 2         | 337(1)                                  | 313(5)                                  | 0.93(2)                                       |
| 3         | 311(2)                                  | 818(5)                                  | 2.63(3)                                       |
| 4         | 253.3(8)                                | 437(6)                                  | 1.72(3)                                       |
| 5         | 159(1)                                  | 371(3)                                  | 0.430(7)                                      |
| 6         | 814(4)                                  | 553(6)                                  | 1.47(2)                                       |
| 7         | 304.4(6)                                | 391(5)                                  | 0.78(1)                                       |
| 8         | 734(3)                                  | 253(3)                                  | 2.91(5)                                       |

Table 4.5: Rabi frequencies measured on the  $\sigma^+$  stretch transition ( $(F = 4, m_F = -4) \leftrightarrow (F = 3, m_F = -3)$ ) with the static B-field oriented parallel to the trap axis, along the  $z$ -direction. All Rabi frequencies have been normalised to 1mW input power.

| Electrode | Rabi Frequency<br>zone A (kHz per<br>mW) | Rabi Frequency<br>zone B (kHz per<br>mW) | Rabi Frequency Ratio<br>(Near zone/ Far zone) |
|-----------|------------------------------------------|------------------------------------------|-----------------------------------------------|
| 1         | 6.58(2)                                  | 6.77(2)                                  | 1.028(6)                                      |
| 2         | 5.29(2)                                  | 6.442(8)                                 | 1.220(5)                                      |
| 3         | 19.18(5)                                 | 15.25(5)                                 | 0.795(5)                                      |
| 4         | 1.772(8)                                 | 0.973(3)                                 | 0.549(4)                                      |
| 5         | 8.11(2)                                  | 8.19(2)                                  | 0.990(5)                                      |
| 6         | 15.11(4)                                 | 13.79(2)                                 | 1.095(5)                                      |
| 7         | 5.27(2)                                  | 3.337(9)                                 | 1.578(9)                                      |
| 8         | 3.023(4)                                 | 2.92(1)                                  | 1.035(6)                                      |

crowave interferometer shown in fig. 4.19c to adjust the relative phase and amplitude of the microwaves fed to the two electrodes. Electrodes 1 (in zone B) and 8 (in zone A) were used for this experiment, since they generated the largest single-electrode Rabi frequency differentials between zones on the qubit transition (see table 4.4). We shall refer to the electrode in the addressed zone as the ‘driving electrode’ and to the electrode in the neighbour zone as the ‘nulling electrode’.

### 4.3. Demonstration of high-fidelity addressing with nulling fields

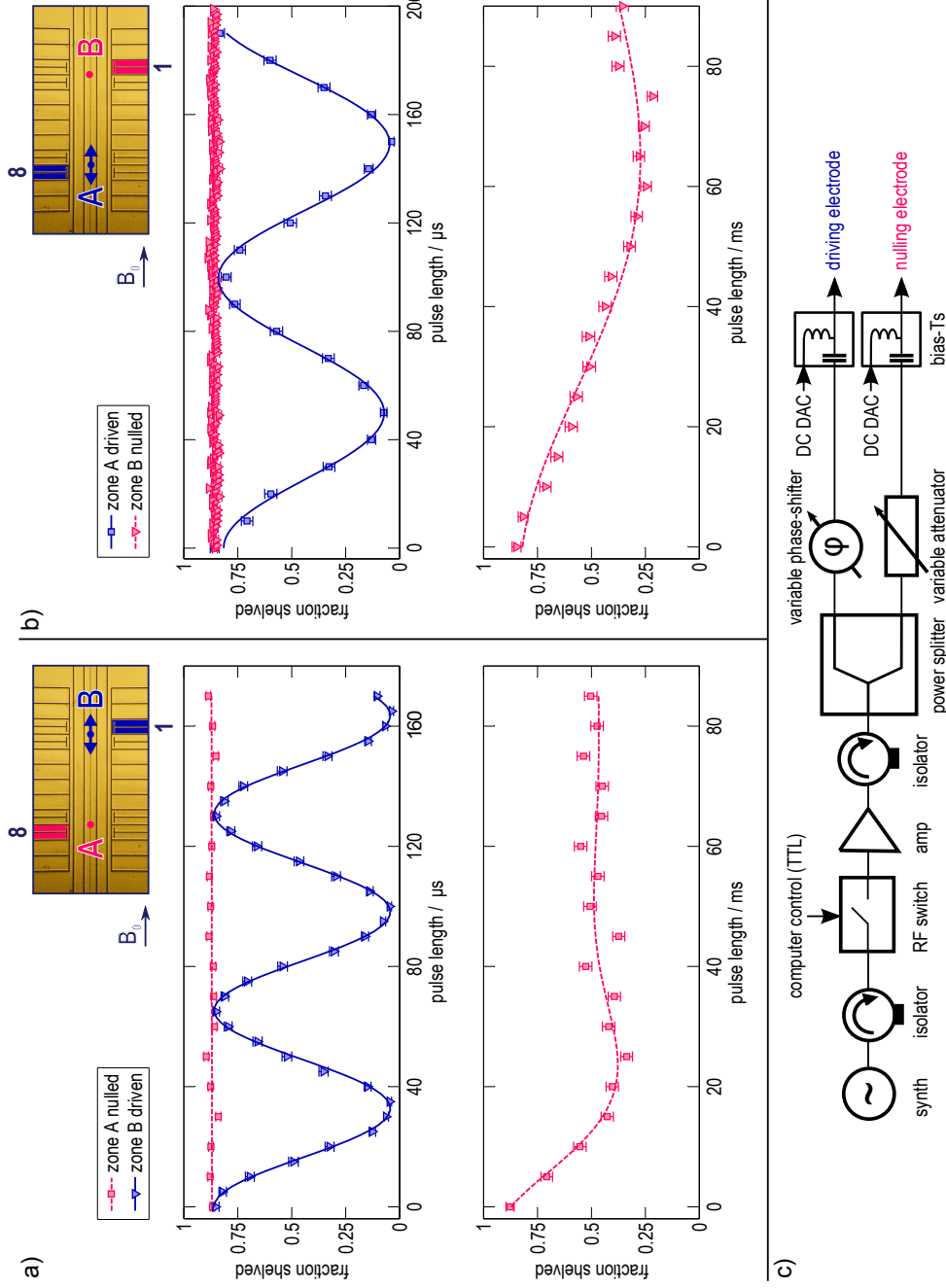


Figure 4.19: a) Zone B is the 'driven' zone, zone A is the 'nulled' zone. Top: Rabi flops in the 'driven' zone (solid, blue curve) with  $\Omega_{\text{driven}} = 15.29(2)\text{kHz}$  are seen when we scan the length of microwave pulses fed into electrodes 1 and 8 from 0 to  $\sim 200\mu\text{s}$ . On this time-scale, the ion in the 'nulled' zone is unaffected (dashed, red line). Bottom: when we scan the microwave pulse length to  $\sim 100\text{ms}$ , we are able to measure a small Rabi frequency in the 'nulled zone' of  $\Omega_{\text{nulled}} = 18(2)\text{Hz}$  before decoherence reduces the contrast of the Rabi flops. b) Similarly, with Zone A as the 'driven' zone and zone B as the 'nulled' zone, we measure  $\Omega_{\text{driven}} = 10.07(3)\text{kHz}$  and  $\Omega_{\text{nulled}} = 7.2(2)\text{Hz}$ . c) The microwave drive chain used to deliver the microwave pulses to electrodes 1 and 8. For reference, the power levels at the vacuum feed-through for the data in plot (b) are  $\approx 170\text{mW}$  at the driving electrode (electrode 8) and  $\approx 9\text{mW}$  at the nulling electrode (electrode 1).

#### 4. MAT: MICROWAVE ADDRESSING TRAP

Table 4.6: Rabi frequencies measured on the  $\sigma^-$  stretch transition ( $(F = 4, m_F = +4) \leftrightarrow (F = 3, m_F = +3)$ ) with the static B-field oriented parallel to the trap axis, along the  $z$ -direction. All Rabi frequencies have been normalised to 1mW input power.

| Electrode | Rabi Frequency<br>zone A (kHz per<br>mW) | Rabi Frequency<br>zone B (kHz per<br>mW) | Rabi Frequency Ratio<br>(Near zone/ Far zone) |
|-----------|------------------------------------------|------------------------------------------|-----------------------------------------------|
| 1         | 4.74(1)                                  | 3.04(1)                                  | 0.640(4)                                      |
| 2         | 1.723(7)                                 | 2.973(5)                                 | 1.73(1)                                       |
| 3         | 16.18(4)                                 | 12.11(2)                                 | 0.7493(3)                                     |
| 4         | 2.419(3)                                 | 3.35(2)                                  | 1.384(8)                                      |
| 5         | 12.54(3)                                 | 1.116(2)                                 | 1.123(4)                                      |
| 6         | 10.22(2)                                 | 7.10(1)                                  | 1.440(6)                                      |
| 7         | 1.927(7)                                 | 0.738(5)                                 | 2.61(3)                                       |
| 8         | 5.23(2)                                  | 3.89(1)                                  | 1.34(1)                                       |

To null the field in the neighbour zone, we first set the attenuation on the arm of the microwave interferometer that feeds the nulling electrode so that the measured individual-electrode Rabi frequencies of the ‘driving’ and ‘nulling’ electrodes were roughly the same in that zone. Then, driving both electrodes simultaneously, we adjusted the variable attenuator and variable phase shifter while monitoring the Rabi frequency in the nulled zone. Once the optimum relative phase and amplitude shifts had been set and a minimised Rabi frequency had been achieved in the nulled zone, we monitored the stability of the Rabi frequency in this zone for  $\approx 1$  hour without making any further adjustments. An ion was then trapped in the addressed zone and the Rabi frequency there was measured. The experiment was then repeated, now addressing the previously nulled zone and nulling the fields in the zone that was previously being addressed.

Figs. 4.19a and b show the measured Rabi flops in both the ‘driven’ and ‘nulled’ zones when nulling was done in zone A and B respectively. The apparent decoherence seen in the slow Rabi flops in the nulled zones is not an intrinsic property of the qubit. It is caused by phase and amplitude noise in the arms of the microwave interferometer which lead to Hz level changes in the Rabi fre-

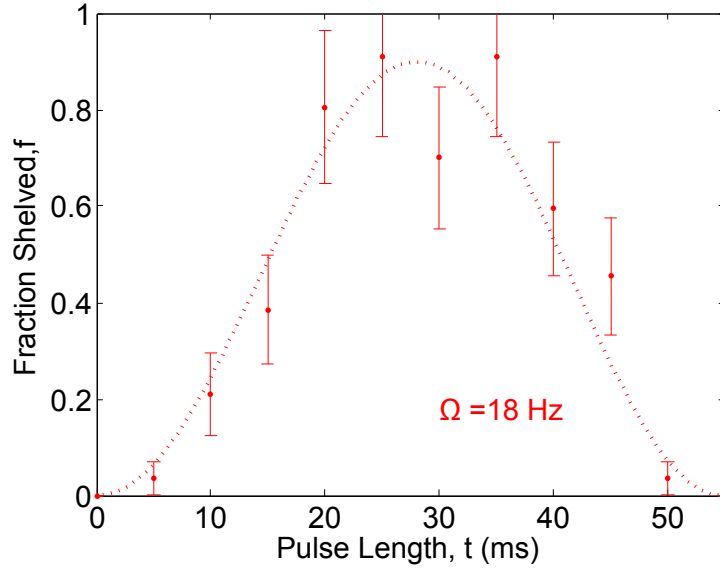


Figure 4.20: Rabi flop on qubit transition driven in Zone B using microwave electrode 1. The microwave drive power was attenuated to produce a  $\pi$  time of 28ms. No loss of contrast is discernible, indicating that any decoherence observed in this timescale in the nulling experiment is not a property of the qubit, but instead a result of phase and amplitude noise in the microwave interferometer. The ‘fraction shelved’ axis has been rescaled (multiplied by 7) here because this data was taken before the installation of a  $397\pi$  beam, when we were preparing only  $1/7^{\text{th}}$  of the population in the  $F = 3; M_F = 0$  qubit state.

quency from shot to shot of the experiment. To verify that this decoherence is not caused by noise in the lab’s static magnetic field, we used a single electrode to drive Rabi flops in the nulled zone and attenuated its input power until the Rabi flops were slowed to having a  $\pi$ -time of 28ms: after a full flop we observed no discernible decay in contrast (fig. 4.20).

#### 4.3.1 Addressing error on the qubit transition

A Rabi frequency ratio between driven and nulled zones  $R_A = \frac{\Omega_{\text{nulled}}}{\Omega_{\text{driven}}} = 1.2(1) \times 10^{-3}$  was achieved when zone A was the nulled zone (fig. 4.19a), and a ratio  $R_B = 7.2(2) \times 10^{-4}$  was obtained when nulling was done in zone B (fig. 4.19b). This implies that the spin-flip probability on the qubit transition of the neighbour ion when we drive a  $\pi$ -pulse (spin-flip) on the addressed ion is  $\epsilon_B = \frac{\pi^2}{4} R_B^2 = 1.27(7) \times 10^{-6}$  and  $\epsilon_A = 3.4(7) \times 10^{-6}$  for zone B nulled and

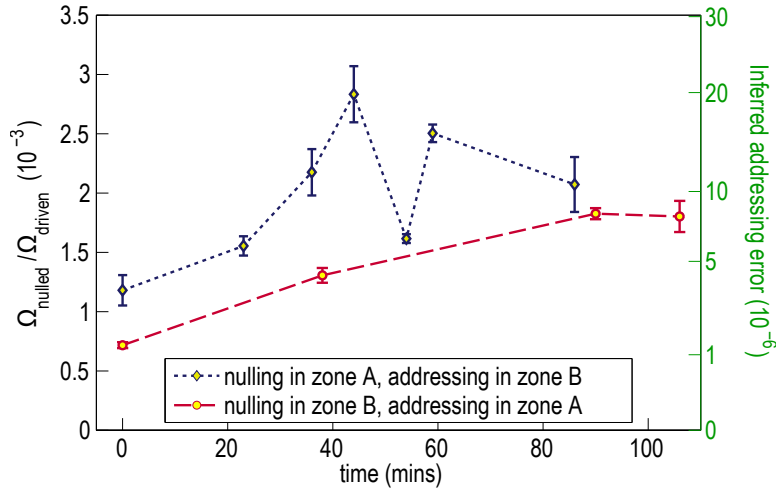


Figure 4.21: The Rabi frequency in the ‘nulled’ zone,  $\Omega_{\text{nulled}}$ , was monitored for more than an hour. For each zone, the nulling parameters were optimised before  $t = 0$ ; thereafter no further adjustments to the nulling parameters were made. The ratio ( $\Omega_{\text{nulled}}/\Omega_{\text{driven}}$ ) is plotted, assuming constant  $\Omega_{\text{driven}}$ , for nulling in zone A (blue, dotted line) and in zone B (red, dashed line). Both ratios remain below  $3 \times 10^{-3}$ , which implies the spin-flip errors on the qubit transition can be kept below  $2 \times 10^{-5}$  without the need to recalibrate nulling over these timescales.

zone A nulled respectively. This is the addressing error introduced by cross-talk between the two zones and it is, encouragingly, significantly below frequently quoted fault-tolerant thresholds of  $10^{-3}$  to  $10^{-4}$ .

From monitoring the nulling stability (fig. 4.21), we see that the both  $R_A$  and  $R_B$  remain below  $3 \times 10^{-3}$  over periods of  $> 1$  hour: the addressing error on the qubit transition should hence remain below  $10^{-5}$  for long periods, without the need for recalibration of nulling parameters.

### 4.3.2 Off-resonant excitation and AC Zeeman Shift Errors

We only used two electrodes in this experiment, so there were only enough degrees of freedom available to null one polarisation of the microwave field, namely, the  $\pi$ -polarised component which drove our qubit transition. Because we did not null other polarisation components, both the neighbour and the addressed qubits could undergo off-resonant excitation on the  $\sigma$  transitions that couple to the qubit states (transitions  $b, c, e$  and  $f$  in fig. 4.22). These  $\sigma$  transi-

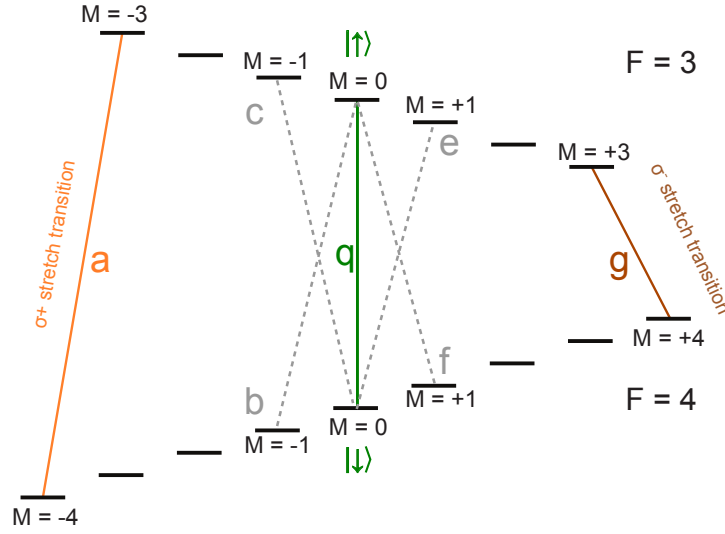


Figure 4.22: Transitions in the  $F = 3$  and  $F = 4$  hyperfine levels of the  $S_{1/2}$  ground state of  $^{43}\text{Ca}^+$ . The qubit transition (green) is labelled  $q$ . Transitions  $b, c, e$  and  $f$  are the transitions which link to the qubit state. They light-shift the qubit transition and are, to a degree, off-resonantly excited during the nulling experiment. The  $\sigma^+$  and  $\sigma^-$  stretch transitions are labelled  $a$  and  $g$  respectively.

tions will also light-shift the qubit transition. In the following subsections we estimate the magnitude of these effects and discuss how to mitigate or eliminate them.

#### 4.3.2.1 Light shift estimate

Referring to fig. 4.22, there are four  $\sigma$ -polarised transitions (labelled  $b, c, e$  and  $f$ ) connecting to the qubit states. They were, to some extent, off-resonantly driven in this experiment. Both the addressed and neighbour qubits will be light-shifted: the derivation that follows applies to either qubit and we state results for both.

At low static magnetic fields, the respective detunings of the relevant transitions from the qubit transition are given by the Zeeman shifts of the  $M_F = \pm 1$  sublevels in the  $F = 3$  and  $F = 4$  manifolds. Conveniently, in  $^{43}\text{Ca}^+$ , the Landé  $g$ -factor has the same magnitude for both  $F = 3$  and  $F = 4$  hyperfine levels of the ground state, i.e.  $|g_{F=3}| = |g_{F=4}| = \frac{1}{4}$ . The detuning is hence the same

for all transitions considered here (up to weak perturbations that give rise to differences on the order of kHz) and is given by:

$$\Delta = \frac{\mu_B B}{4h} \quad (4.9)$$

where  $\mu_B$  is the Bohr magneton,  $h$  is Planck's constant and  $B$  is the static magnetic field. At  $B = 2.8\text{G}$ ,  $\Delta \approx 0.98\text{MHz}$ .

Transitions  $b$  and  $c$  are of higher frequency than the qubit transition and will hence perturb the qubit states in such a way as to push them closer together, light-shifting the qubit transition to the red. Similarly, the lower frequency transitions  $e$  and  $f$  will light-shift the qubit to the blue. Writing the Rabi frequency on transition  $x$  in a given trap zone as  $\Omega_x$ , the total light shift experienced by the ion in that zone will be given by:

$$\delta = \frac{\Omega_e^2 + \Omega_f^2 - \Omega_b^2 - \Omega_c^2}{4\Delta} \quad (4.10)$$

The Rabi frequency on the transition  $x$  between states  $(F = 4, M)$  and  $(F = 3, M')$  in the presence of a microwave magnetic field  $\mathbf{B}_{\mu w}$  is given by:

$$\Omega_x = \langle 4, M | \boldsymbol{\mu} \cdot \mathbf{B}_{\mu w} | 3, M' \rangle = (-1)^{q-M} \frac{3\sqrt{7}}{2} \mu_B B_{\mu w}^q W_x \quad (4.11)$$

$$W_x = \begin{pmatrix} 4 & 1 & 3 \\ -M & -q & M' \end{pmatrix} \quad (4.12)$$

where  $W_x$  is the Wigner  $3j$  symbol for transition  $x$ , and  $q = \Delta M = M' - M$  quantifies the ion's change in angular momentum along the quantisation axis (i.e.  $q = +1$  for a  $\sigma^+$  transition,  $q = -1$  for a  $\sigma^-$  transition and  $q = 0$  for a  $\pi$  transition).  $B_{\mu w}^q$  is the component of the microwave magnetic field polarised along the direction which drives the transition. The  $3j$  symbols for the relevant transitions are given in table 4.7.

Since we did not directly measure the Rabi frequencies on transitions  $b, c, e$  and  $f$ , we must infer them from the Rabi frequency data taken on the stretch



Table 4.7: Evaluated Wigner  $3j$  symbols for transitions in the  $S_{1/2}$  manifold of  $^{43}\text{Ca}$ ,  $43^+$  as labelled in fig. 4.22.

| Transition | Evaluated Wigner $3j$ symbol |
|------------|------------------------------|
| a          | $1/3$                        |
| b          | $\sqrt{1/42}$                |
| c          | $-(1/3)\sqrt{5/14}$          |
| d          | $2/(3\sqrt{7})$              |
| e          | $-(1/3)\sqrt{5/14}$          |
| f          | $\sqrt{1/42}$                |
| g          | $1/3$                        |

transitions of matching polarisation (tables 4.5 and 4.6). To make our notation more clear, we introduce the change in angular momentum  $q$  as a label in our transition names, to indicate their polarisations: e.g. transition b is  $\sigma^+$  polarised, so we can label it  $b_{+1}$ . We can now relate the Rabi frequency produced by electrode  $i$  on transition  $x_q$ ,  $\Omega_{x_q}^i$ , to that produced on the stretch transition of the same polarisation  $\Omega_{\text{stretch}_q}^i$ :

$$\Omega_{x_q}^i = \frac{W_{\text{stretch}_q}}{W_{x_q}} \left( \frac{P_{x_q}^i}{P_{\text{stretch}_q}^i} \right)^{1/2} \Omega_{\text{stretch}_q}^n \quad (4.13)$$

where  $P_{x_q}^i$  is the power in mW fed to electrode  $i$  during the nulling experiment and  $P_{\text{stretch}_q}^i$  is the power fed to electrode  $i$  when we measured the Rabi frequency on the  $\text{stretch}_q$  transition. The stretch transition Rabi frequencies listed in tables 4.5 and 4.6 have already been normalised to 1mW input power, (i.e.  $P_{\text{stretch}_q}^i = 1\text{mW} \forall i, q$ ) so we need only plug into equation 4.13 the powers fed into electrodes 1 and 8 in the nulling experiment to calculate Rabi frequencies generated in the nulled zone on all relevant off-resonantly driven transitions. The calculated Rabi frequencies are given in tables 4.8 and 4.9.

Since we do not know the phase relationship between the microwave electrodes, we cannot predict the Rabi frequencies on the off-resonant transitions when both electrodes are powered. We can, however examine the following scenarios:

#### 4. MAT: MICROWAVE ADDRESSING TRAP

Table 4.8: This table lists the Rabi frequencies produced on the off-resonantly excited  $\sigma$  transitions during the nulling experiment when zone A was the ‘driven’ zone and zone B was the ‘nulled zone’. These Rabi frequencies were deduced from Rabi frequencies measured on stretch transitions and the powers fed to electrodes 1 and 8 in the nulling experiment. The notation used in the column headers of this table should be read as follows:  $\Omega_{\text{driven/nulled}}^i$  is a Rabi frequency measured in the driven or nulled zone when electrode  $i$  is powered. The transition names in column 1 are as labelled in fig. 4.22.

| Transition | $\Omega_{\text{nulled}}^1$<br>(kHz) | $\Omega_{\text{nulled}}^8$<br>(kHz) | $\Omega_{\text{driven}}^1$<br>(kHz) | $\Omega_{\text{driven}}^8$<br>(kHz) |
|------------|-------------------------------------|-------------------------------------|-------------------------------------|-------------------------------------|
| b          | 4                                   | 23                                  | 9                                   | 18                                  |
| c          | 12                                  | 23                                  | 9                                   | 40                                  |
| e          | 6                                   | 30                                  | 12                                  | 23                                  |
| f          | 10                                  | 17                                  | 7                                   | 31                                  |

Table 4.9: This table lists the Rabi frequencies produced on the off-resonantly excited  $\sigma$  transitions during the nulling experiment when zone B was the ‘driven’ zone and zone A was the ‘nulled’ zone. Notation as in table. 4.8.

| Transition | $\Omega_{\text{nulled}}^1$<br>(kHz) | $\Omega_{\text{nulled}}^8$<br>(kHz) | $\Omega_{\text{driven}}^1$<br>(kHz) | $\Omega_{\text{driven}}^8$<br>(kHz) |
|------------|-------------------------------------|-------------------------------------|-------------------------------------|-------------------------------------|
| b          | 37                                  | 9                                   | 38                                  | 9                                   |
| c          | 34                                  | 20                                  | 22                                  | 15                                  |
| e          | 48                                  | 12                                  | 49                                  | 11                                  |
| f          | 27                                  | 16                                  | 17                                  | 12                                  |

1. The  $\sigma^+$  fields from electrodes 1 and 8 are perfectly in phase and interfere constructively, whilst the  $\sigma^-$  fields from electrodes 1 and 8 are perfectly out of phase and interfere destructively. The Rabi frequency on the  $\sigma^+$  transitions will then simply be the sum of the individual electrode Rabi frequencies:  $\Omega_{x+1} = \Omega_{x+1}^1 + \Omega_{x+1}^8$  where  $x \in (b, f)$ . The Rabi frequency on the  $\sigma^-$  transitions will be the difference of the individual electrode Rabi frequencies:  $\Omega_{x-1} = \Omega_{x-1}^1 - \Omega_{x-1}^8$ , where  $x \in (c, e)$ ;
2. The  $\sigma^+$  fields interfere destructively and the  $\sigma^-$  fields interfere constructively:  $\Omega_{x+1} = \Omega_{x+1}^1 - \Omega_{x+1}^8$  where  $x \in (b, f)$  and  $\Omega_{x-1} = \Omega_{x-1}^1 + \Omega_{x-1}^8$ , where  $x \in (c, e)$ ;
3. Both polarisations interfere constructively:  $\Omega_{x\pm 1} = \Omega_{x\pm 1}^1 + \Omega_{x\pm 1}^8$  where  $x \in (b, c, e, f)$ ;

4. Both polarisations interfere destructively:  $\Omega_{x\pm 1} = \Omega_{x\pm 1}^1 - \Omega_{x\pm 1}^8$  where  $x \in (b, c, e, f)$ .

Plugging into eq. 4.10 the Rabi frequencies calculated based on each of the above assumptions produces four values for the light shift, one of which is the maximum light shift that can be felt by the qubit. Table 4.10 lists the maximum light shifts on both the addressed and nulled qubits in both nulling configurations (with zone A nulled and with zone B nulled). As a measure of the phase error caused by these light shifts, table 4.10 also gives the phase error accrued by each qubit when a  $\pi$ -pulse is performed in the addressed zone.

Table 4.10: Estimates of the maximum magnitude light shifts,  $|\delta|$ , on the qubit transition and the corresponding phase errors,  $\epsilon$  accrued by the qubit when a  $\pi$  pulse is applied in the addressed zone. The light shifts and phase errors are given for both the nulled ( $|\delta_{\text{nulled}}|$ ,  $\epsilon_{\text{nulled}}$ ) and driven ( $|\delta_{\text{driven}}|$ ,  $\epsilon_{\text{driven}}$ ) qubits in each of the two nulling configurations.

| Nulling Configuration           | $ \delta_{\text{nulled}} $ , $\epsilon_{\text{nulled}}$ | $ \delta_{\text{driven}} $ , $\epsilon_{\text{driven}}$ |
|---------------------------------|---------------------------------------------------------|---------------------------------------------------------|
| Zone A addressed, zone B nulled | 120Hz, 37mrads                                          | 230Hz, 72mrads                                          |
| Zone B addressed, zone A nulled | 340Hz, 70mrads                                          | 370Hz, 76mrads                                          |

#### 4.3.2.2 Light shift stability

Provided the light-shifts are stable, the accumulated phase on the qubits can be tracked and corrected for. We estimate this stability as follows: we assume that the Rabi frequencies on  $\sigma$  and  $\pi$  transitions fluctuate by similar proportions in the nulled zone and, for simplicity, model this fluctuation as arising from amplitude fluctuations in the Rabi frequency produced by one of the two electrodes in the nulled zone,  $\Omega_{\text{noisy}}$ . From the data in fig. 4.21, we see that fluctuation of the Rabi frequency in the nulled zone over the course of an hour is  $\delta\Omega_{\text{nulled},\pi} \approx 10^{-3} \times \Omega_{\text{driven},\pi}$ . The two electrodes we used in this experiment (electrodes 1 and 8) produce Rabi

frequencies in their neighbour zones that are approximately  $3\times$  smaller than in their resident zones (table 4.4), so that the Rabi frequency of our noisy electrode in the nulled zone is  $\Omega_{\text{noisy},\pi} \approx 1/3 \times \Omega_{\text{driven},\pi}$  (where the last equality follows from the fact that, in the driven zone, most of the contribution to the Rabi frequency comes from the driving electrode). Hence,  $\frac{\delta\Omega_{\text{nulled},\pi}}{\Omega_{\text{noisy},\pi}} \approx 3 \times 10^{-3} = 0.3\% = \frac{\delta\Omega_{\text{nulled},\sigma}}{\Omega_{\text{noisy},\sigma}}$ . The detuning of the off-resonant  $\sigma$  transitions from the qubit transitions also fluctuate by  $\sim 0.1\%$  due to variations of  $\sim 1\text{mG}$  in our static lab B-field. Adding the errors in quadrature, the phase error on the nulled qubit per  $\pi$  pulse on the addressed qubit should fluctuate by:

$$\delta\phi = \sqrt{\left(2 \frac{\delta\Omega_{\text{noisy},\sigma}}{\Omega_{\text{noisy},\sigma}}\right)^2 + \frac{\delta\Delta^2}{\Delta^2}} \approx 0.6\% \quad (4.14)$$

#### 4.3.2.3 Off-resonant excitation error estimate

The probability  $p$  of off-resonant excitation on any one of the  $\sigma$  transitions that couple to the qubit is given by:

$$p = \frac{\Omega^2}{\Omega^2 + \Delta^2} \quad (4.15)$$

where  $\Omega$  is the Rabi frequency on the  $\sigma$  transition. For an order-of-magnitude estimate of the off-resonant excitation errors in the nulling experiment, note that the largest Rabi frequencies on the  $\sigma$  transitions in the nulling experiment (listed in tables 4.8 and 4.9) are of order  $50\text{kHz}$ . Therefore, since  $\Delta \approx 1\text{MHz}$ ,  $p \approx 3 \times 10^{-3}$ .

#### 4.3.2.4 Elimination of light shift and off-resonant excitation errors

Both the off-resonant excitation error and the light shift can be made negligible if we simply increase the static magnetic field and use an intermediate-field clock qubit. In practice, we would wish to do so even

in the absence of these errors, in order to benefit from very long coherence times offered by these qubits (of order 50 s [HAB<sup>+</sup>14]). If the clock qubit at  $B_0 = 146\text{G}$  were used for example, the light shift would be reduced to  $\lesssim 7\text{Hz}$  and the off-resonant excitation error to  $\sim 10^{-6}$ .

Pulse shaping techniques can also be used to suppress off-resonant excitation: one can produce pulses lacking the Fourier components required to excite the off-resonant transitions, significantly reducing the transition probabilities.

Even though these solutions can be easily implemented, the best way to avoid the gate-speed limits set by the presence of off-resonant excitation errors and the qubit-phase-tracking overheads associated with the presence of light shifts is to control the polarisation of the microwave field. MAT's four electrodes per trap zone give us enough degrees of freedom to do this. The next section outlines a first demonstration of such polarisation control in MAT.

## 4.4 Polarisation control

To demonstrate control of the polarisation of the microwave field in each zone, two electrodes in zone A were used to selectively drive one of two nearly degenerate transitions out of the qubit states. The chosen transitions, indicated in fig. 4.23c, are separated by only 1.6kHz in frequency (at a static field of  $B_0 = 2.8\text{G}$ ), but one is driven by  $\sigma^+$  polarisation and the other by  $\sigma^-$  polarisation. Using the microwave interferometer in fig. 4.19c, the relative phase and amplitude of the microwaves fed to electrodes 7 and 8 were adjusted so as to null the  $\sigma^+$  component of the microwave field in Zone A. This was again done by adjusting the variable attenuator and variable phase shifter on the interferometer while monitoring the Rabi fre-

quency on the  $\sigma^+$  transition. With the nulling optimised, we measured the Rabi frequency on the  $\sigma^-$  transition, which required preparing the qubit in the  $|\uparrow\rangle$  state: this was done by using a third microwave trap electrode to apply a  $\pi$ -pulse to the ion after the standard laser preparation pulse sequence.

A Rabi frequency ratio of  $\frac{\Omega_{\sigma^+}}{\Omega_{\sigma^-}} = 2.77(8) \times 10^{-3}$  (fig. 4.23) was achieved, implying an addressing error of  $\epsilon_{pol} = 1.9(1) \times 10^{-5}$ . The polarisation control demonstrated here can be used to suppress off-resonant excitation, allowing for fast operations to be performed on qubits without loss of fidelity. An example of an operation where polarisation control would be useful is high-fidelity read-out in  $^{43}\text{Ca}^+$ , which involves using microwave pulses to selectively transfer population in one of the qubit states to the stretch state ( $F = 4, m_F = +4$ ) [HAB<sup>+</sup>14]: the faster this operation is performed, the larger the Rabi frequencies on the transitions used to transfer the qubit population need to be. Larger Rabi frequencies imply the higher probability of off-resonant excitation and hence larger error per operation if unwanted-polarisations are not nulled; polarisation control removes this speed limit.

## 4.5 Conclusions

Using MAT, we demonstrated independent control of the microwave near-fields at two distinct trap zones. Qubit rotations were driven in a chosen zone, whilst residual crosstalk fields were nulled at the neighbour zone to a level consistent with an addressing error on the qubit transition of  $2(1) \times 10^{-6}$ , significantly below the fault-tolerant threshold. We found the nulling to be robust against phase and amplitude fluctuations when monitored for over an hour, indicating that, after an initial calibration, it is feasible to continuously perform single-qubit addressing operations with

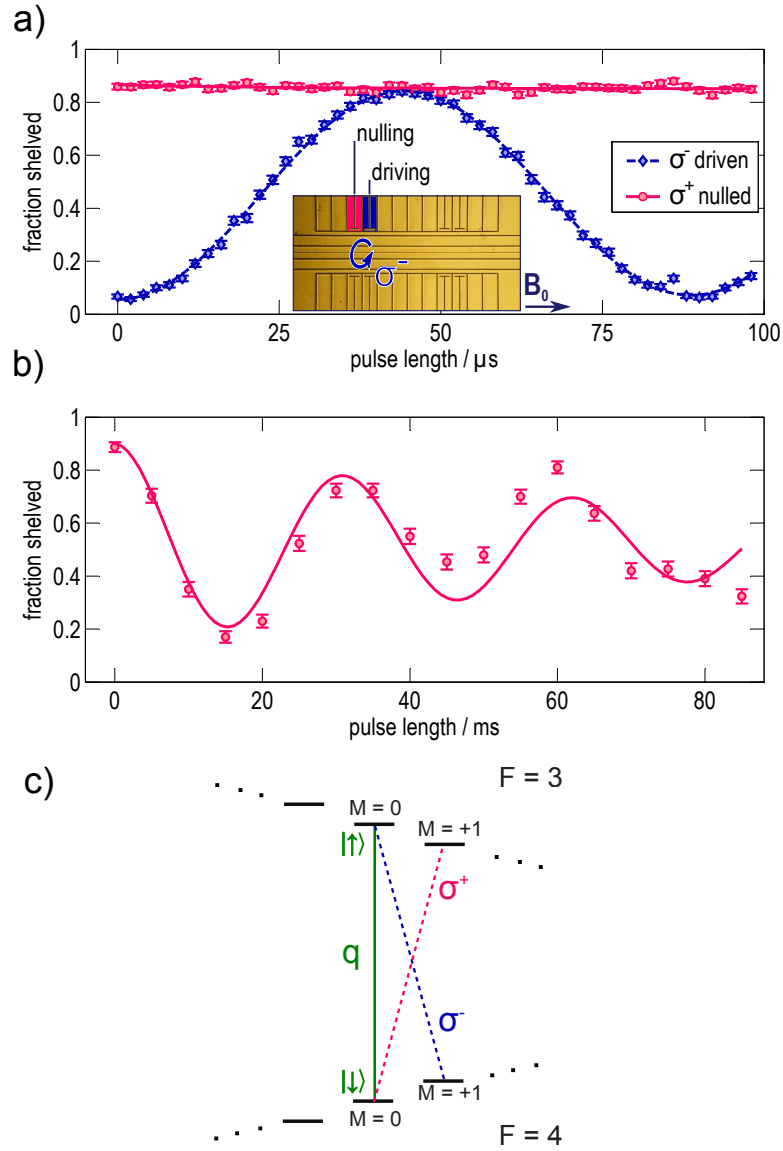


Figure 4.23: To demonstrate polarisation control, we null the  $\sigma^+$  component of the microwave polarisation. We are then able to drive the  $\sigma^-$ -polarised transition that is indicated in blue in (c), without exciting the nearly degenerate  $\sigma^+$ -polarised transition indicated in (c) in red. a) We drive Rabi flops on the 'driven'  $\sigma^-$ -polarised transition (blue, solid curve) with Rabi frequency  $\Omega_{\sigma^-} = 11\text{kHz}$  by scanning the length of the microwave pulse applied to electrodes 7 and 8 from 0 to  $100\mu\text{s}$ . On this time-scale, the 'nulled'  $\sigma^+$ -polarised transition is not visibly excited (solid, red curve). Both datasets on this plot were taken with microwaves resonant with the  $\sigma^+$  transition (which is  $1.6\text{kHz}$  detuned from the  $\sigma^-$  transition). b) On the millisecond timescale, we observe slow Rabi flops on the 'nulled'  $\sigma^+$ -polarised transition with Rabi frequency  $\Omega_{\sigma^+} = 32\text{Hz}$ , driven by residual  $\sigma^+$ -polarisation.

high fidelity for long periods of time. Off-resonant excitation was a more significant source of error in this experiment. We demonstrated how to implement polarisation control of the microwave field at a level sufficient to reduce the off-resonant transition probability to  $1.9(1) \times 10^{-5}$ .

Several improvements can be made to the next iteration trap design.

1 The trap should be designed and powered in such a way as to allow all impedances seen by the microwaves to be faithfully simulated. This requires the following changes to MAT's design:

- i. The CPGA should be replaced with a purpose-built impedance-controlled Printed Circuit Board (PCB). The PCB will be fed by coaxial connectors and carry microwaves to the chip trap via planar transmission lines, designed to be impedance matched to  $50\Omega$  for better power transfer. The board and connectors would be included in the HFSS model so that any crosstalk between feedlines is identified in simulation;
- ii. An on-chip diplexer should be used to terminate the RF: this consists of a chip capacitor and inductor connected in parallel. The component values are chosen so that the 40MHz RF travels mostly down the inductive path and the 3.2GHz microwaves travel down the capacitive path. The latter can be terminated in a known impedance and included in the HFSS model. In this way, the load impedance of the RF electrode as seen by the microwaves will be well defined and any current picked up by it will be more faithfully simulated;

2 Even with the above changes, the microwave fields produced at the ion by a trap with MAT's electrode geometry would be difficult to faithfully simulate. This is because the microwave electrodes are not directly underneath the ion: at the point where the ion is trapped, the



microwave field generated by an electrode in that zone has already decayed to  $\sim 1/10^{\text{th}}$  of its maximum value (which occurs directly above the electrode) and is hence of the same order of magnitude as induced currents coupled across the chip. These induced currents are difficult to model precisely because their presence induces other currents, which in turn induce other currents and so on ad infinitum. The predicted current distribution becomes very sensitive to small changes in trap geometry (which will inevitably occur in fabrication) and to the meshing of the model. These difficulties can be avoided by ensuring that ion is driven directly by the largest fields on the chip.

These improvements were implemented in MiLoT (Microwave Loop Trap), the second iteration design, as described in the next chapter.



# 5

## MiLoT: a new trap design

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The previous chapter described experiments performed with MAT, a single-layer trap with integrated microwave electrodes. The chapter concludes that, whilst Rabi frequency differentials of up to 1300 were achievable when residual fields were nulled using two electrodes, improvements to the electrode layout are necessary to ensure that the Rabi frequency differentials produced by single electrodes (when powered individually) are large and match simulation. This chapter describes the development of a new multi-layer trap design based on routing microwave return currents to produce large single-electrode Rabi frequency differentials between trap zones. Section 5.1 describes the trap's multi-layer architecture proposes two candidate microwave electrode geometries. Section 5.2 presents simulation results obtained for the two geometries in a two-zone trap architecture and proceeds to compare their simulated performance in a four-zone architecture. Finally, section 5.3 outlines future work to be done on MiLoT and comments on the outlook for the design.

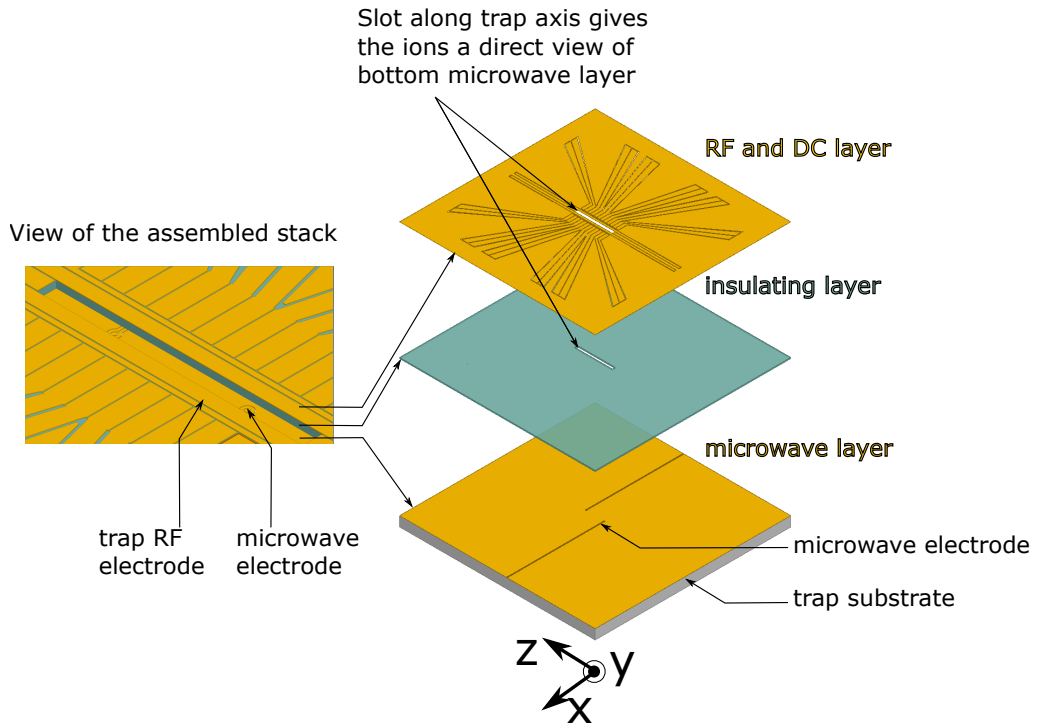


Figure 5.1: Multi-layered stack architecture for new trap design. Left: assembled stack. Right: exploded view of stack layers.

## 5.1 Trap design

### 5.1.1 Three-layer stack architecture

Beginning with the layer furthest away from the ions, the trap's stack layers are the following:

- **Microwave layer:** a  $5\mu\text{m}$  layer of gold patterned on a fused silica substrate which carries all microwave waveguides in the design. It can be fabricated using the same photolithography and electroplating techniques which were used to fabricate MAT (see section 3.3).
- **Insulating layer:** a  $30\mu\text{m}$  layer of polyimide material, a spin-on plastic which has been previously used in atom-chip fabrication to insulate layered electrodes [FTS11, TGH<sup>+</sup>08]. A rectangular slot is patterned into this and the top layer, exposing the microwave electrodes directly

below the ions. These slots ensure that the ions have a direct line of sight to the microwave electrodes: the slot through the top layer prevents its gold electrodes from shielding the microwave fields from the ion and the slot through the insulating polyimide layer serves to minimize the amount of dielectric exposed to the ions.

- RF and DC layer: The top layer of the stack carries all electrodes required to produce the RF and DC trapping potentials. It is a  $5\mu\text{m}$  layer of gold, which, like the microwave layer, can be patterned over the polyimide layer using photolithography and electroplating techniques.

### 5.1.2 Choice of layer thicknesses

According to [FTS11], three layers of polyimide totalling  $7\mu\text{m}$  in thickness are enough to planarize the surface of the plastic to a tolerance of  $300\text{nm}$  when it is spun over a  $5\mu\text{m}$  thick patterned layer. Here, I chose a thickness of  $30\mu\text{m}$  in order to minimize currents induced in the top DC/RF layer: if the gap between the microwave layer and the top layer is made to be much larger than that between the microwave electrodes and their surrounding ground plane (which is  $5\mu\text{m}$ , for the designs that will be presented here), the microwave fields will be encouraged to propagate in coplanar modes, with return currents flowing predominantly along the adjacent microwave-layer ground plane. If, however, the gaps are comparable in size, microwave return currents will travel across the top layer in much the same way as they would across the backing ground plane in a grounded coplanar waveguide. Simulations indicate that a  $30\mu\text{m}$  gap is sufficient to suppress this effect without overly attenuating the microwave field seen by the ions (i.e. when the polyimide layer thickness is increased, ion height above microwave electrodes also increases, so the ions see a

weaker field).

The thickness of the gold microwave layer was chosen to be  $5\mu\text{m}$  because making the electrodes thicker would complicate the fabrication process (since it would require increasing the height:gap aspect ratio of the trap features above 1:1), whilst providing little advantage in terms of mitigating loss in the conductor (since most of the microwave current flows within a layer at the surface of the conductor, with the layer depth given by the skin depth of gold, which is  $1.4\mu\text{m}$  at 3.2GHz).

### 5.1.3 Microwave electrode geometries

Fig. 5.2 depicts two electrode geometries which this section puts forward as suitable candidate structures for producing microwave near-fields which can be confined to individual trap zones: a coplanar-waveguide design (henceforth referred to as the ‘CPW design’) and a design featuring electrodes which resemble loop antennas (henceforth referred to as MiLoT, an acronym for Microwave Loop Trap).

We will first consider two-zone chip designs for both electrode geometries and later scale them up (in simulation) to four-zone architectures. Each trap zone features a single microwave electrode placed directly underneath the ion, which in turn sees the maximum field amplitude delivered by the electrode in its zone. With both electrode geometries, our aim is to exploit the cancelling effect of microwave return currents to passively attenuate inter-zone cross-talk.

#### 5.1.3.1 CPW design: Coplanar waveguide trap

A CPW-based electrode design is a good starting point for this study since its simple electrode structure will allow us to more easily interpret simulation results and develop an intuitive understanding of how to engineer

return-current flow. Current flow in an isolated coplanar waveguide with semi-infinite ground planes is very well understood and can be analytically modelled [GRS93]. Current driven down a centre trace generates an oscillating magnetic field which induces the flow of opposing return currents along the edges of the ground planes on either side of the trace. As discussed in section 2.3, this opposing current flow, if properly routed, should contribute to a faster decay rate of the microwave magnetic field as we move radially away from the centre conductor. In particular, if we orient the CPW trace perpendicular to the trap axis (as shown in fig. 5.2b), an ion trapped over the centre of the CPW trace will see only a field component  $H_z$  along the trap axis ( $H_x$  and  $H_y$ , the field components along the CPW plane and out of the plane of the CPW are zero over the centre of the CPW) and we are interested in using the opposing field generated by return currents along the CPW edge to ensure that  $|H_z|$  decays quickly as possible as we move along the axis to neighbour trap zones.

In the CPW trap design considered here, the coplanar waveguides are surrounded by the trap's complex DC and RF electrode structure which, despite being placed as far as practically possible from the microwave layer to minimize inter-layer coupling, complicates the analysis enough to prevent us from studying the field pattern analytically. A further level of complexity arises from the fact that the finite width of the ground planes surrounding each CPW trace affects the magnitude of eddy the currents induced around the driven zone, and hence influences the degree of nulling these currents provide. We will hence study both the CPW trap design and MiLoT using a finite-element solver (HFSS) to calculate the microwave magnetic field distribution generated when each microwave electrode is driven, as we did with MAT in chapter 4.

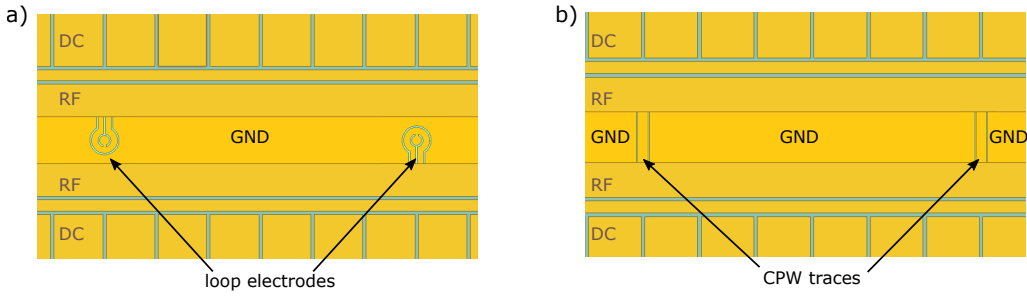


Figure 5.2: Top view of central trapping region of two candidate trap designs: a) MiLoT, a design featuring microwave electrodes resembling planar loop antennas and b) CPW design, featuring coplanar waveguides.

### 5.1.3.2 MiLoT: Microwave loop trap

Instead of CPWs, MiLoT features loop electrodes in every zone. Neighbour loops approach the trap axis from opposite edges of the chip (see fig. 5.6), allowing for wide segments of uninterrupted, centrally connected ground plane between trap zones.

The two feed lines to each loop are driven  $180^\circ$  out of phase so that the differential mode of propagation is excited. This mode produces a large current at the loop tip (underneath the ion), which generates a field which, unlike the CPW field, has similarly-sized components along the  $x$ ,  $y$  and  $z$  directions. We will choose to drive the ion with the  $H_z$  field component (the component aligned with the trap axis), since we will find that this component offers the largest differential in magnitude between trap zones, and will show that the other components will not cause significant off-resonant excitation errors if an intermediate-field qubit is used.

## 5.2 Simulated microwave near-field patterns

### 5.2.1 Two-zone CPW design

Figs. 5.4 and 5.5 plot, for the two-zone versions of the CPW trap and MiLoT respectively, the simulated  $H_z$  field component along the trap axis



produced when we drive each zone in turn. We see that the CPW trap produced  $|H_z|$  differentials of order 100 between trap zones and MiLoT produced differentials of order 1000. These field ratios imply that individual-ion addressing errors at the  $10^{-4}$  and  $10^{-6}$  levels respectively should be achievable with these trap designs (recall that, as in chapter 4, we define the addressing error  $\epsilon$  as the spin-flip probability on the qubit transition in a neighbour zone when we apply a  $\pi$  pulse to a qubit in the addressed zone, where  $\epsilon = \frac{\pi^2}{4} \left( \frac{1}{R_z} \right)^2$  and  $R_z$  is the ratio of the field magnitude produced in the addressed zone to that in the neighbour zone).

To complete the analysis of the fields generated by MiLoT, we plot the non-negligible magnitudes of the  $x$  and  $y$  field components (normalized to  $|H_z|$ ) in fig. 5.3 and list their values at the trap zones in table 5.1. We see the magnitudes of these field components are at most  $1.5 \times |H_z|$ , which implies that the off-resonant excitation errors they will produce are on the order of  $10^{-6}$  for single qubit gates performed at speeds of 100kHz on the 146G clock qubit in  $^{43}\text{Ca}^+$  (spectator transitions in this qubit are at least 50MHz detuned from the qubit transition).

We can see from the field and current plots in figs. 5.4 and 5.5 that the large field differentials produced by MiLoT and the CPW design result not only from the cancelling effect of the return current flowing in the ground plane immediately adjacent to the driven electrode (in the opposite direction to the current in the microwave electrode's centre trace) but also from the presence of field nulls created at the neighbour zones by induced eddy currents. The latter are circulating currents generated by Faraday induction when conductors are permeated by an alternating magnetic flux (here provided by the microwave magnetic field generated by the driven electrode). The magnitude of these currents increases with the area of the conductor they circulate around. Iron cores used for transformers are of-

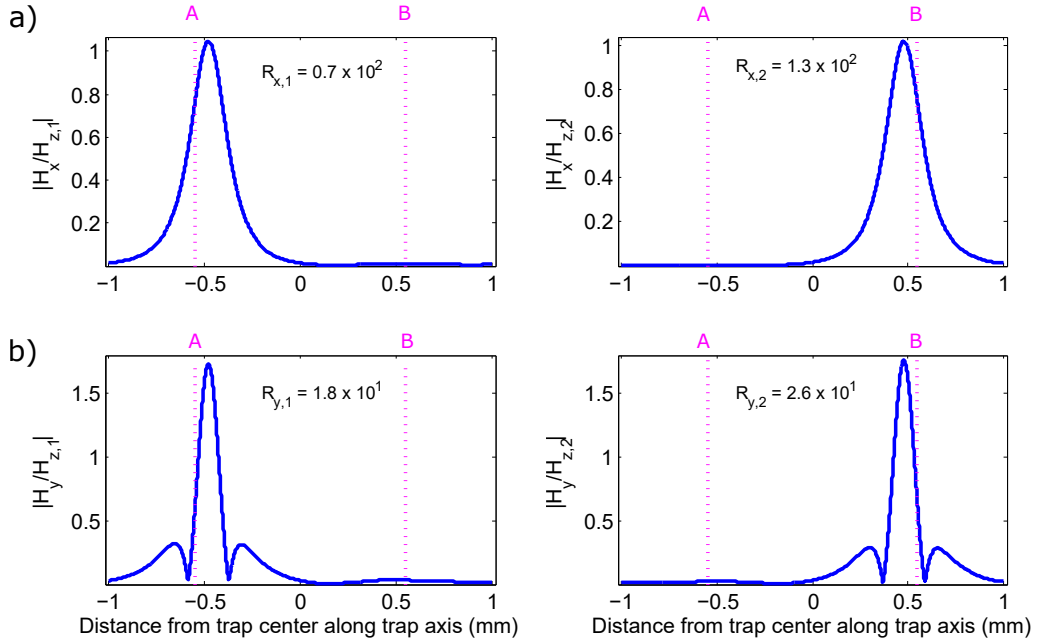


Figure 5.3: Simulated  $|H_x|$  and  $|H_y|$  along the trap axis of two-zone MiLoT. Left:  $|H_x|$  (top) and  $|H_y|$  (bottom) produced when electrode 1 (in zone A) is powered. Right:  $|H_x|$  (top) and  $|H_y|$  (bottom) produced when electrode 2 (in zone B) is powered. The plotted fields magnitudes are normalized to  $|H_z|$  value in the relevant zone. The magenta dotted lines in each plot indicate the positions of the trap zones along the trap axis. The ratios  $R_{x,i}$  and  $R_{y,i}$  of the field magnitude produced by electrode  $i$  in its resident zone to that produced in the neighbour zone are displayed in the plots.

ten laminated to suppress eddy currents since, because they flow in such a way as to oppose magnetic flux, they tend to cause unwanted power dissipation. It is precisely this attenuating quality which we are making use of here and we will see in the next section how careful design of the ground plane geometry is necessary to prevent it from being suppressed.

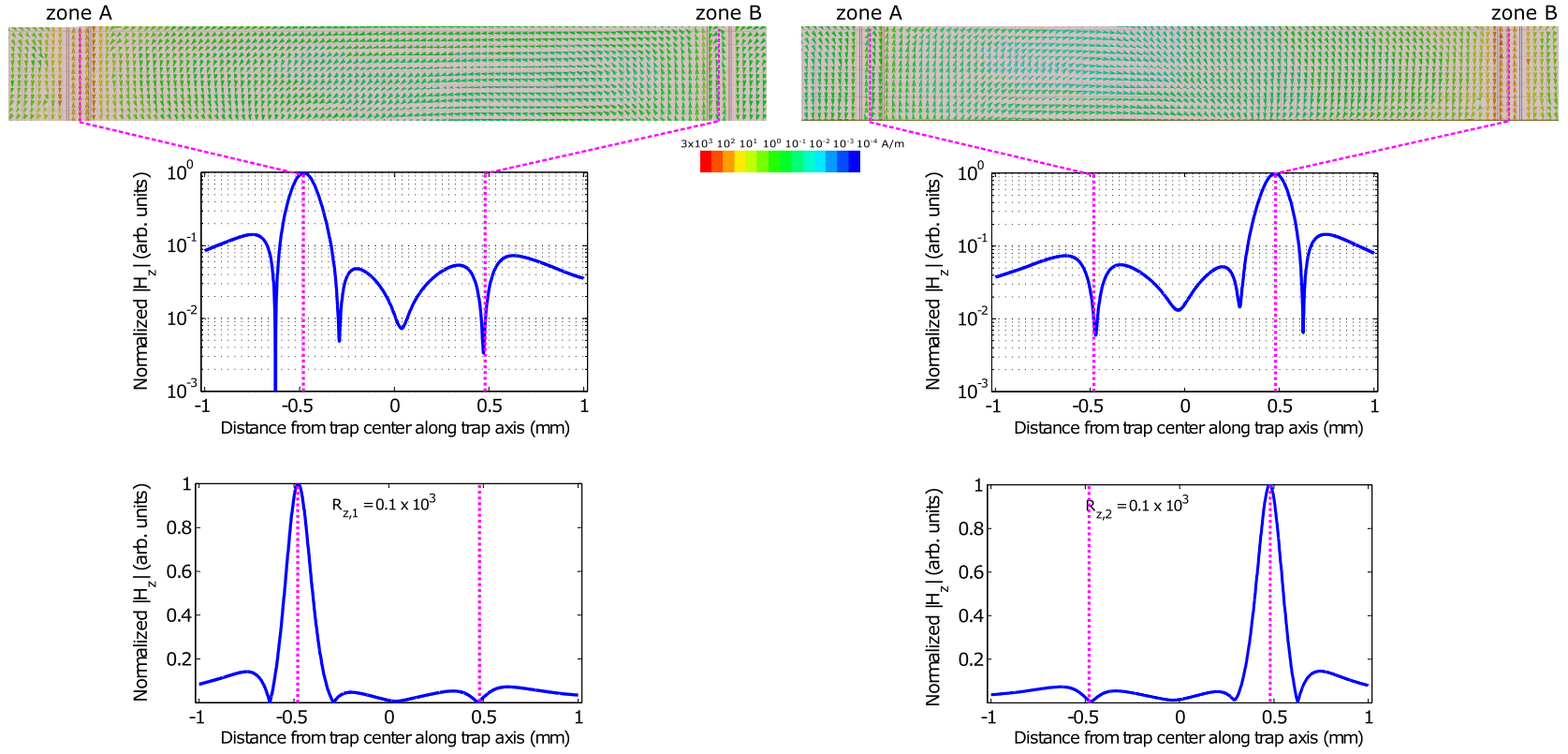


Figure 5.4: Simulated  $|H_z|$  along the trap axis of two-zone CPW trap. Left:  $|H_z|$  produced when electrode 1 (in zone A) is powered. The field magnitude is plotted against distance along trap axis on a logarithmic (top) and a linear (bottom) scale. Right:  $|H_z|$  produced when electrode 2 (in zone B) is powered. The magenta dotted lines in each plot indicate the positions of the trap zones along the trap axis. The ratios  $R_{z,1}$  and  $R_{z,2}$  of the field magnitude produced by electrodes 1 and 2 in their resident zone to that produced in their neighbour zone are displayed in the plots. Surface current maps above each set of plots display the relative magnitude and direction of the surface currents in the central trapping region when the relevant electrode is powered.

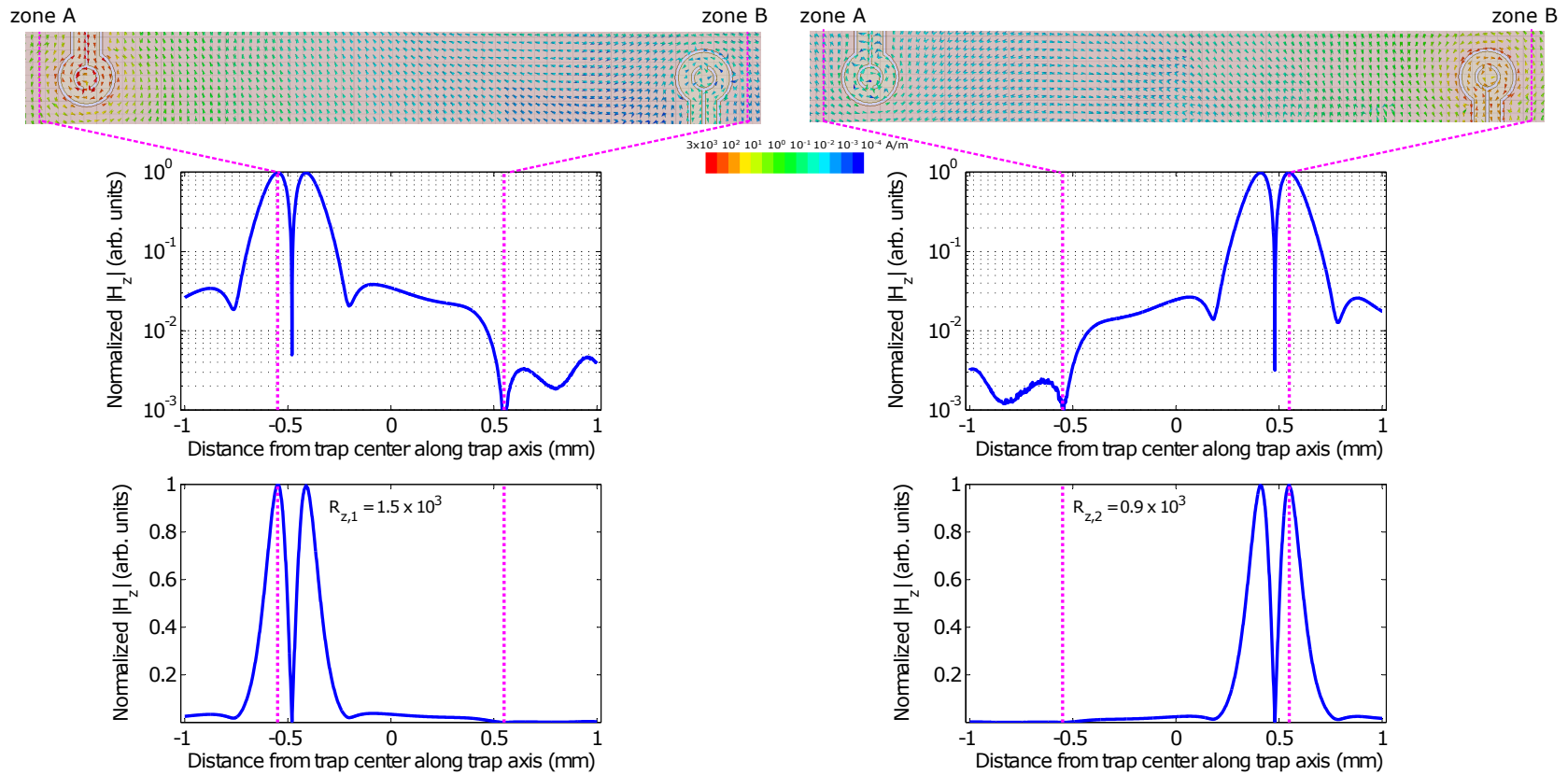


Figure 5.5: Simulated  $|H_z|$  along the trap axis of two-zone MiLoT trap. Left:  $|H_z|$  produced when electrode 1 (in zone A) is powered. The field magnitude is plotted against distance along trap axis on a logarithmic (top) and a linear (bottom) scale. Right:  $|H_z|$  produced when electrode 2 (in zone B) is powered. The magenta dotted lines in each plot indicate the positions of the trap zones along the trap axis. The ratios  $R_{z,1}$  and  $R_{z,2}$  of the field magnitude produced by electrodes 1 and 2 in their resident zone to that produced in their neighbour zone are displayed in the plots. Surface current maps above each set of plots display the relative magnitude and direction of the surface currents in the central trapping region when the relevant electrode is powered.

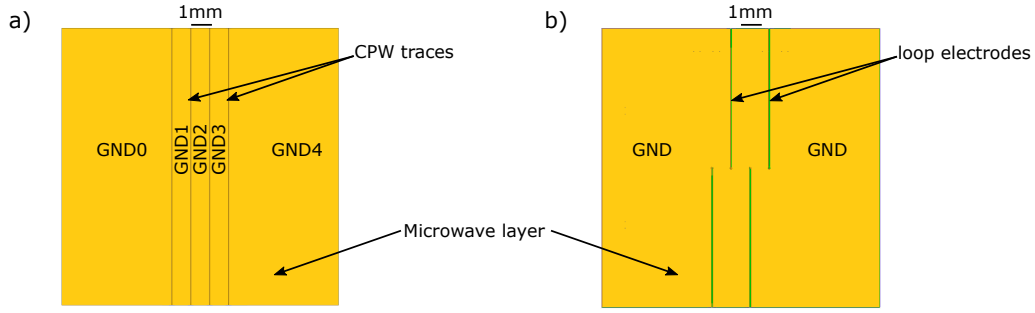


Figure 5.6: Top view of microwave ground planes in four-zone CPW design (a) and in four-zone MiLoT (b). In the CPW design, ground planes GND1 to GND4 are only connected around the edge of the chip, whilst MiLoT's ground plane is also centrally connected.

Table 5.1: Simulated magnitudes of the  $x$ ,  $y$  and  $z$  components of  $H$ -field along the trap axis of two-zone MiLoT, normalized to the  $|H_z|$  value in the relevant zone. Column  $X_i$  refers to the field magnitudes obtained in zone  $X$  when electrode  $i$  is powered (note that electrode 1 is in zone A and electrode 2 is in zone B).

|       | $A_1$ | $B_2$ | $B_1$              | $A_2$                |
|-------|-------|-------|--------------------|----------------------|
| $H_x$ | 0.8   | 0.7   | 0.011              | 0.006                |
| $H_y$ | 0.6   | 0.6   | 0.035              | 0.025                |
| $H_z$ | 1.0   | 1.0   | $7 \times 10^{-4}$ | $1.1 \times 10^{-3}$ |

### 5.2.2 Four-zone designs: MiLoT vs CPW

Fig. 5.7 shows the simulated  $H_z$  field magnitudes produced when we power each microwave electrode in turn on four-zone versions of MiLoT and of the CPW trap design. Let us first turn our attention to the fields produced in the CPW trap. We see that field nulls clearly appear in zones A and D when they are not being addressed (e.g. there is a field null in zone A when electrodes 2, 3 or 4 are powered), but fail to occur in zones B and C, the zones at the centre of the chip. We postulate here that this occurs because the ground planes surrounding the centre zones are much smaller in area than those at the edge of the chip (see fig. 5.6) and hence do not support eddy currents which are large enough to cancel the field in those zones.

The situation is alleviated in MiLoT because the ground plane is centrally

interconnected across trap zones and hence has a larger uninterrupted area. Field nulls still fail to appear in zones B and C, but the design offers cancellation which is better by an order of magnitude than that produced with the CPW trap.

Fig. 5.8 gives a clearer picture of the field patterns in each design, by focusing only on the scenario where electrode 3 in zone C is powered. The figure includes plots of the phase of  $H_z$  along the trap axis for both designs: in these plots, field nulls are clearly demarcated by steps of  $\pi$  in the phase.

Tables 5.2, 5.3, 5.4, 5.5 give the simulated intra-zone field differentials (or ratios) and spin-flip addressing errors calculated from them. Whilst the calculated errors for the CPW design reach  $10^{-2}$ , MiLoT produces field ratios which keep all addressing errors at or the  $10^{-4}$  level for this four-zone prototype architecture.

### 5.3 Future work

Based on the simulation results we have seen in this chapter, MiLoT is a promising architecture for performing high-fidelity near-field microwave addressing without the need for nulling fields. If successfully realized, it would offer a simple platform for implementing parallel single-qubit operations over many trap zones, a required ingredient for the construction of any scalable quantum computer.

To ascertain the robustness of the design and to optimise its performance, it would be prudent both to perform further simulations of architectures scaled up to eight or sixteen zones. If the addressing error is seen to converge at the  $10^{-4}$  level as we scale the design, the next step would be to fabricate a two-zone prototype chip. Using the antenna techniques described

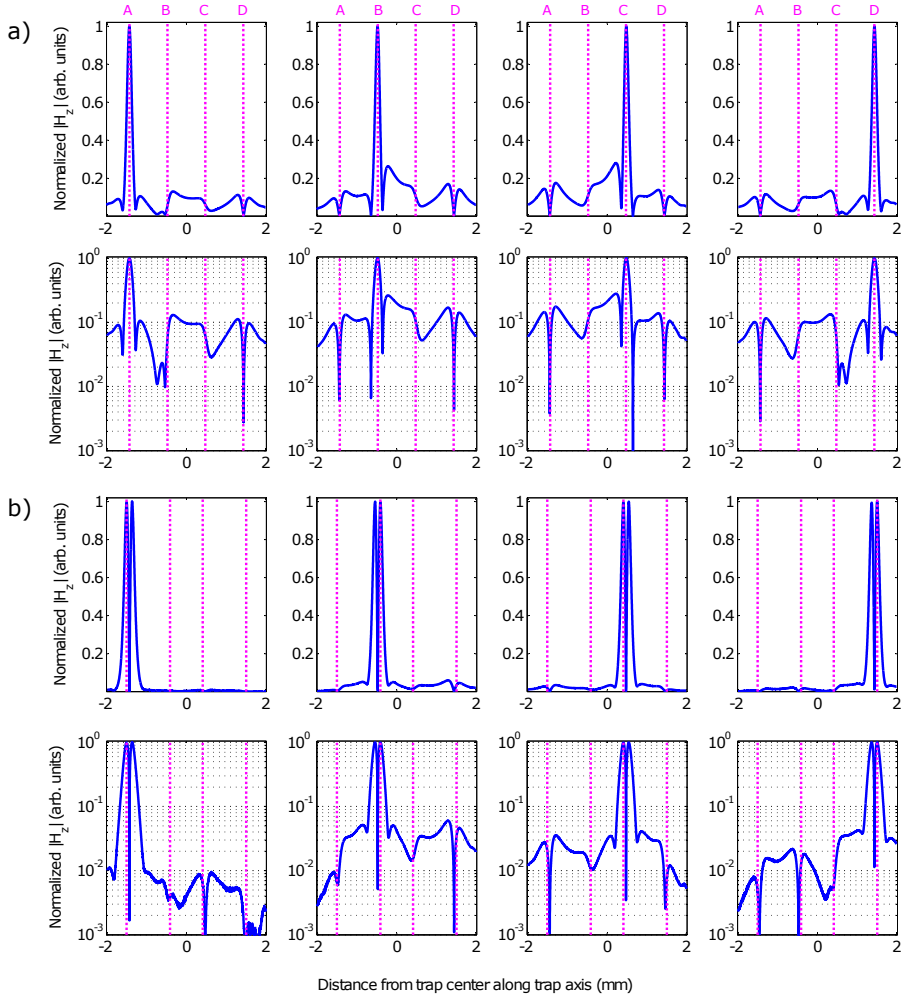


Figure 5.7: Simulated  $|H_z|$  along the trap axis for four-zone CPW (a) and MiLoT (b) trap designs. Left to right: Plots showing  $|H_z|$  produced when electrode 1 (in zone A), 2 (in zone B), 3 (in zone C) and 4 (in zone D) are powered, respectively. The field magnitude is plotted against distance along the trap axis on a logarithmic (top) and a linear (bottom) scale. The magenta dotted lines in each plot indicate the positions of the trap zones along the trap axis.

in the next chapter, the chip's microwave field can be mapped without the need for trapping ions, allowing for a quick comparison between simulation and measurement (and, if required, another design optimization cycle) before the fabrication of scaled-up prototypes.

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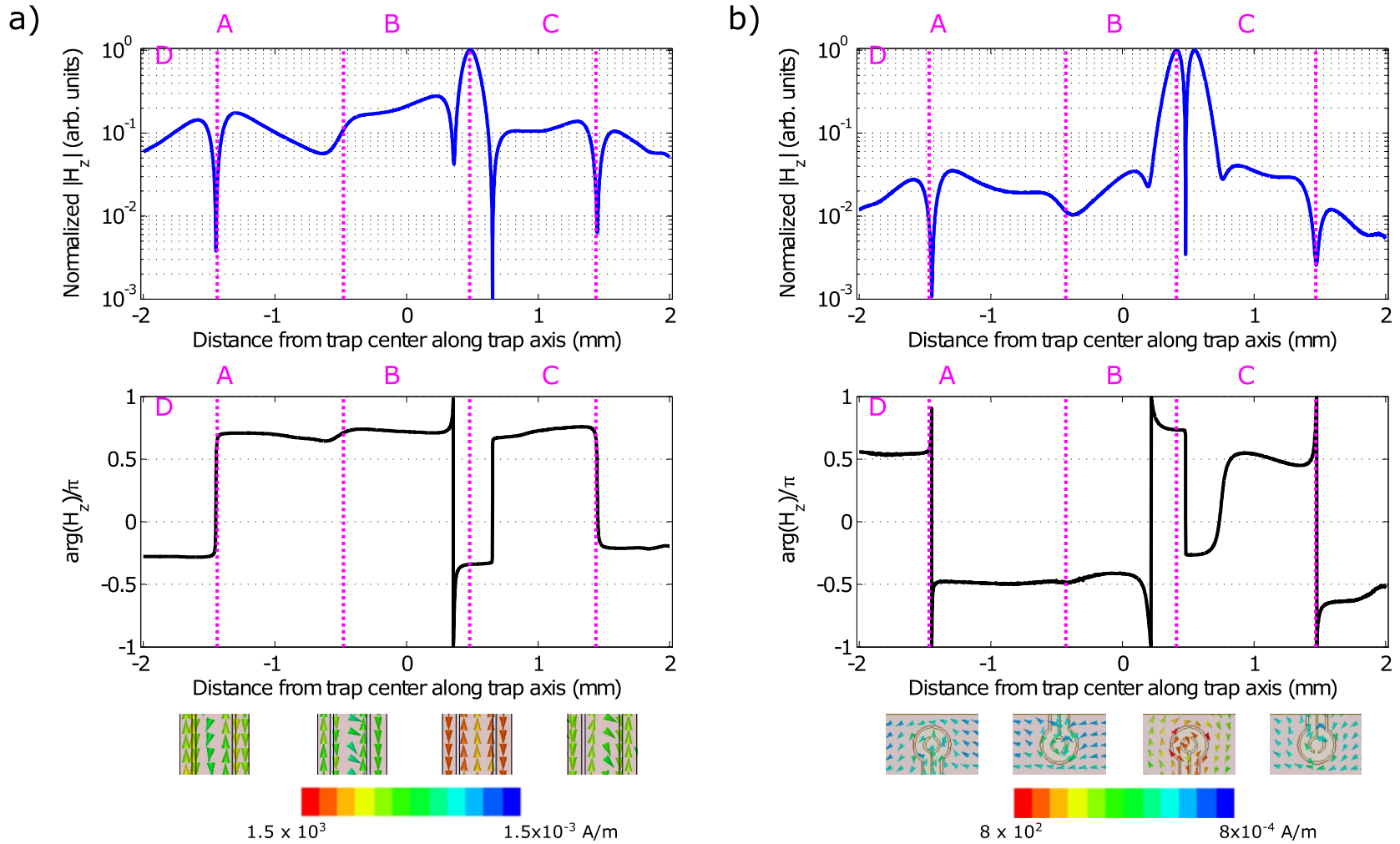


Figure 5.8: Top: Simulated  $|H_z|$  (top) and  $\arg(H_z)$  (bottom) along the trap axis for four-zone CPW (a) and MiLoT (b) trap designs produced when electrode 3 is powered. The magenta dotted lines in each plot indicate the positions of the trap zones along the trap axis. Surface current maps below each set of plots display the relative magnitude and direction of the surface currents around each trap electrode when electrode 3 is powered.



Table 5.2: Simulated  $|H_z|$  produced at each zone of the scaled-up CPW design when electrodes 1 to 4 are powered in turn (normalized  $|H_z|$  in the electrode's zone).

| Electrode | Normalised $H_z$ in zone |                      |                      |                    |
|-----------|--------------------------|----------------------|----------------------|--------------------|
|           | A                        | B                    | C                    | D                  |
| 1         | 1                        | $5 \times 10^{-2}$   | $6 \times 10^{-2}$   | $9 \times 10^{-3}$ |
| 2         | $2 \times 10^{-2}$       | 1                    | $1.0 \times 10^{-1}$ | $2 \times 10^{-2}$ |
| 3         | $2 \times 10^{-2}$       | $1.1 \times 10^{-1}$ | 1                    | $2 \times 10^{-2}$ |
| 4         | $9 \times 10^{-3}$       | $6 \times 10^{-2}$   | $6 \times 10^{-2}$   | 1                  |

Table 5.3: CPW trap addressing errors calculated from field ratios in table 5.2.

| Electrode | Spin flip error in zone ( $10^{-3}$ ) |    |    |     |
|-----------|---------------------------------------|----|----|-----|
|           | A                                     | B  | C  | D   |
| 1         | —                                     | 7  | 9  | 0.2 |
| 2         | 1.0                                   | —  | 30 | 0.7 |
| 3         | 0.8                                   | 30 | —  | 0.9 |
| 4         | 0.2                                   | 10 | 8  | —   |

Table 5.4: Simulated  $|H_z|$  produced at each zone of the scaled-up MiLoT design when electrodes 1 to 4 are powered in turn (normalized  $|H_z|$  in the electrode's zone).

| Electrode | Normalised $H_z$ in zone |                      |                    |                      |
|-----------|--------------------------|----------------------|--------------------|----------------------|
|           | A                        | B                    | C                  | D                    |
| 1         | 1                        | $6 \times 10^{-3}$   | $9 \times 10^{-3}$ | $1.0 \times 10^{-4}$ |
| 2         | $8 \times 10^{-3}$       | 1                    | $2 \times 10^{-2}$ | $1.5 \times 10^{-2}$ |
| 3         | $8 \times 10^{-3}$       | $1.2 \times 10^{-2}$ | 1                  | $3 \times 10^{-3}$   |
| 4         | $8 \times 10^{-4}$       | $1.4 \times 10^{-2}$ | $8 \times 10^{-3}$ | 1                    |

Table 5.5: MiLoT addressing errors calculated from field ratios in table 5.4.

| Electrode | Spin flip error in zone ( $10^{-4}$ ) |     |   |      |
|-----------|---------------------------------------|-----|---|------|
|           | A                                     | B   | C | D    |
| 1         | —                                     | 0.9 | 2 | 0.02 |
| 2         | 2                                     | —   | 7 | 5    |
| 3         | 2                                     | 3   | — | 0.2  |
| 4         | 0.02                                  | 5   | 2 | —    |



# 6

## Near-field mapping for microwave surface ion-traps

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### 6.1 Mapping near-fields over microwave surface ion-traps

As we develop the technology to design and produce chip traps with integrated microwave electrodes that are capable of driving qubits with low cross-talk and efficient power transfer, we require a fast and comprehensive method for characterizing microwave fields over candidate devices. This is crucial if we are to move forwards with efficient design-characterization-optimization research cycles. To date, only trapped ions and neutral atoms have been used as probes to map the microwave near-field over surface ion and atom-traps [HDT15], respectively. Both ions and atoms are accurate and non-invasive probes of magnetic field but, to date, these methods of field mapping remain experimentally and technically challenging.

Using ions to map fields requires significant experimental effort and produces only a (nearly) 1D map of the field (along the trap axis and at points close to the axis). Assuming this type of mapping is to be done in a labora-

tory where the control systems and lasers required to trap ions are already available, it still involves, at the very least, placing the chip under vacuum (along with an oven which is to be the source of the ions), and building up apparatus which is tailored to trapping an ion with the chip currently being tested; i.e. providing the chip with the RF and DC voltages required to trap the ion, setting up the laser beam paths and constructing an imaging system with the appropriate focal length and numerical aperture.

A vapour cell of optically-active atoms in a buffer gas is an interesting alternative way of mapping fields [HDT15], as it can be done without trapping the atoms and when the chip is not under vacuum. This method offers a very promising avenue to the development of robust field mapping techniques but, it has its own technical challenges. For example, the wall-thickness of the thinnest vapour cell developed for this purpose to date prevents the atoms from being brought closer than  $150\mu\text{m}$  from the chip's surface. Most ion-trap chips used for microwave entangling gates trap ions at less than  $100\mu\text{m}$  from the trap surface. The lateral extent of the vapour cells currently being used is in the tens of mm, which is too large for our purpose. A typical ion-trap chip is smaller than this and has wire-bonded pads at the edges. As ultra-thin vapour cell technology develops and the method matures into an off-the-shelf system, these difficulties will likely be alleviated but, at the moment, this method is not suitable for characterizing most microwave ion-trap chips.

This chapter will describe efforts to develop a method of field characterization which can be implemented by ion-trap groups quickly, cheaply and with little experimental complexity. The system involves using a micro-fabricated PCB loop antenna to measure the microwave magnetic near-field pattern over the chip's surface. This type of near-field scanning has been used in the electronics industry for several decades, most notably

to characterize sources of electromagnetic interference in integrated circuits [KWS<sup>+</sup>12, Yag86]. The aim of the work described in this chapter is to introduce this characterization technology to the field of ion (and atom) trapping and to develop prototype antennas which are tailored to the requirements of our area and which research groups can easily fabricate using printed circuit board (PCB) techniques. Section 6.2 gives an overview of the near-field mapping system. Section 6.3 outlines design requirements for a near field antenna tailored to our application and presents a first prototype design. Finally, section 6.4 describes initial field mapping tests and presents data taken in a pilot experiment.

## 6.2 The scanner system

Fig. 6.1 shows the system I use to map the microwave magnetic fields over a device under test (DUT). A micro-fabricated antenna is scanned over the DUT using an automated stepper while a vector network analyser (VNA) is used to record the oscillating B-field flux picked up by the antenna at every position in the scan. The stepper consists of an aluminium arm which can move with sub-micron precision along three axes. Motion along each axis is propelled by a digital motor controlled by a LabVIEW interface written by Prof. Chris Stevens of Oxford's engineering department. Prof. Stevens kindly allowed me to use his purpose-built stepper to test prototype antennas. Research groups who do not have access to such custom-built equipment can reproduce the results using a commercially available motorized translation stage.

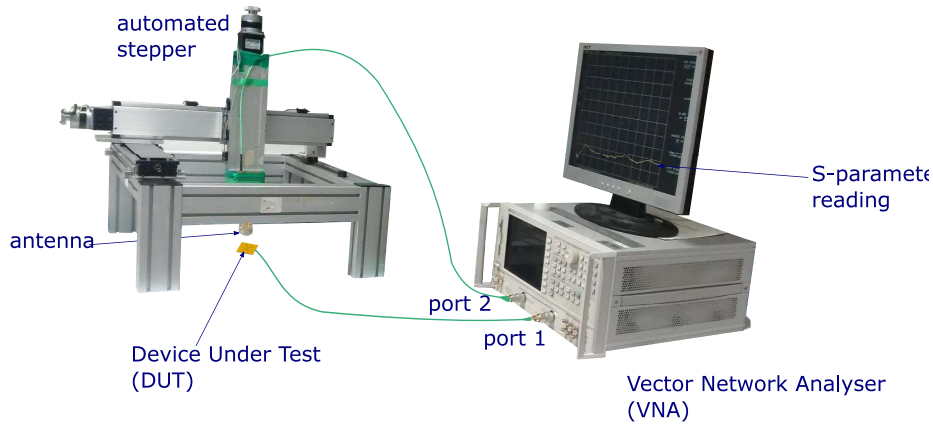


Figure 6.1: A vector network analyser (VNA) applies a constant amplitude oscillating voltage across a port on the DUT and records the voltage measured by the antenna at every position of the scan: the ratio of the measured voltages is the  $S_{21}$  parameter (port 1 of the VNA is connected to the DUT and port 2 to the antenna). At each point in the scan, the VNA sweeps the frequency of the applied voltage over a desired range and displays  $S_{21}$  on the screen as a function of frequency.

### 6.3 The near-field antenna

We would like an antenna that translates microwave B-field flux into a potential difference, or electromotive force (EMF), which appears across the antenna terminals. A simple loop antenna fits this description. In this section, we will introduce the concept of the sensitivity pattern of an antenna and discuss how our application requires us to tailor the properties of this pattern. This discussion will lead us to set out a list of criteria we would like our antenna to fulfil in subsection 6.3.1. A first prototype antenna design will then be introduced in subsection 6.3.2 and important aspects of its design will be examined.

#### 6.3.1 Antenna sensitivity pattern

If we were to use our loop antenna as a transmitter instead of a receiver, and we drove its terminals with a given input oscillating voltage, it would generate a spatially dependent oscillating  $H$ -field pattern. By the reciprocity theorem of electromagnetism, this  $H$ -field pattern is the same as the pattern describing the sensitivity of the antenna to  $H$ -field flux when it

is used as a receiver: i.e. the  $H$ -field sensitivity pattern maps the extent to which a unit oscillating  $H$ -field at a given point in the volume surrounding the antenna contributes to the EMF generated across the antenna.

When we scan the antenna over a device, the EMF we record is the result of the convolution of the antenna's sensitivity pattern with the microwave  $H$ -fields produced by the device around that point. Put more simply, we are, at each point of the scan, averaging the field over an effective volume around the antenna and the extent and shape of this volume is defined by the antenna's sensitivity pattern.

We can hence identify the following desirable characteristics of the sensitivity pattern of the antenna we will use to map microwave near-fields above ion trap chips.

- a) A small effective volume: The sensitivity function must decay quickly with distance from the antenna loop and must only be significantly non-zero in a small volume around the antenna tip, on the order of  $(10\mu\text{m})^3$  to  $(100\mu\text{m})^3$ . If the antenna is sensitive to flux in a large volume around its tip, it will detect stray fields from background sources, leading to uncontrolled contributions to the measured EMF;
- b) Simplicity and balance: The precision with which we can make field measurements is directly related to the confidence with which we know the sensitivity pattern. It is therefore sensible to try to keep the pattern, and hence antenna design, as simple and easy-to-simulate as possible. It would, for instance, be best to maintain a degree of symmetry in the pattern, by terminating both ends of the loop with the same impedance. This would reduce the amount by which the antenna's sensitivity function distorts the measured field, making it easier to deconvolve the measurement from the effect of the antenna

(for example, if the pattern is asymmetric, a field maximum over the DUT will be spatially shifted once it is convolved with the sensitivity pattern). A symmetric design will also ensure that our antenna's feed lines are *balanced*: currents along the feed lines will be equal in amplitude and opposite in phase. This ensures that the feed lines produce no far-field radiation and are hence, by reciprocity, insensitive to EM-sources which are outside the antenna's near-field;

- c) Insensitivity to common-mode noise: Balanced feed lines form a differential pair: the signal we want to extract from the antenna, namely, the induced EMF across the loop, is encoded in the voltage *difference* across the two lines. As long as we only measure signals carried by the differential mode of the feed lines, we are granted insensitivity to common-mode noise, a significant source of background interference (For example, the entire antenna probe could acquire an electric charge with respect to ground, which would appear as an EMF with respect to ground across each of the antenna's two output ports. However, the EMFs appearing at each port would be the same and, therefore do not contribute to the differential signal we measure). To make use of this insensitivity we must convert the differential signal into a single-ended signal that can be fed into the VNA using a *balun*, as explained in section 6.3.2.6.

### 6.3.2 A first prototype antenna

Having discussed desirable features of a sensitivity pattern, we now wish to design a prototype antenna. The following sections describe the important aspects of the design of this first prototype, which is depicted diagrammatically fig. 6.2 and photographically in fig. 6.4c.



### 6.3.2.1 Antenna fabrication

To produce a sensitivity pattern with an effective volume of order  $(100\mu\text{m})^3$ , we will require a micro-fabricated loop, with diameter of order  $100\mu\text{m}$ . For this first prototype, I preferred to keep fabrication as simple as possible, and chose to use only single-layer PCB fabrication techniques. This avoids potential inter-board alignment errors which occur in multi-layer fabrication processes where alignment tolerances are often in the tens of microns. The antenna was fabricated on Rogers 3003 material, a PTFE composite developed for use in high-frequency circuits which offers low losses at microwave frequencies. The dissipation factor, which is defined as the reciprocal of the quality factor, for Rogers 3003 is 0.0013 at 10GHz, a factor of 10 lower than that of FR-4 grade epoxy laminates commonly used for PCB fabrication.

To minimize distortion of the fields due to the antenna, the thinnest commercially available Rogers3003 laminate board was used as our substrate: the board is  $130\mu\text{m}$  thick and is laminated with  $9\mu\text{m}$  of copper on both sides. The fabrication process was done in-house by the Oxford physics photofabrication team, using film masks printed by an external company, JD Phototools.

### 6.3.2.2 Loop dimensions

The dimensions of our loop were chosen to be  $100\mu\text{m}$  by  $50\mu\text{m}$  - small enough to produce a compact sensitivity pattern, but not so small as to sacrifice fabrication yield. For future prototypes, it should be possible to reduce the loop area to  $25\mu\text{m}$  by  $25\mu\text{m}$  whilst still using standard PCB techniques; laser machining or additive fabrication techniques would allow us to reduce the loop area even further. I decided not to invest resources into minimizing the loop dimensions in the fabrication of this first prototype,

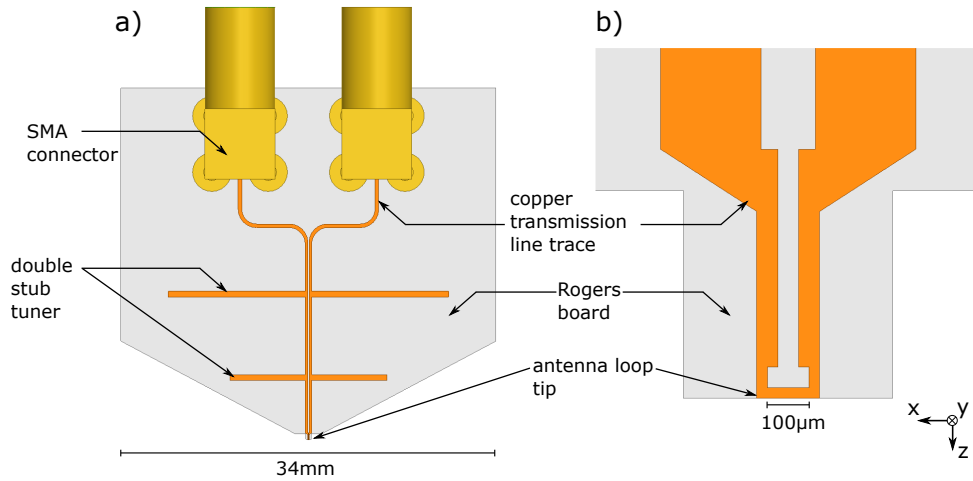


Figure 6.2: Top view of the antenna prototype, as modelled in HFSS. Figure (a) gives an overall view of the antenna PCB and figure (b) is a magnified view of the antenna tip.

since the main aim of this pilot experiment is to check the feasibility of this field mapping method in the context of our application and it seemed to us that any efforts to optimise the choice of loop dimensions would benefit from a better knowledge of the signal-to-noise ratio attainable in with our setup.

### 6.3.2.3 Laser-trimmed antenna tip

We would like to be able to bring the antenna tip as close to the devices we measure as possible, at least one ion height (which, in our case, is  $110\mu\text{m}$ ) away from the surface of the chip. To this end, the substrate was laser-trimmed as close as possible to the loop's top copper trace, as shown in fig. 6.4c. This was done, relatively inexpensively (laser-trimming each board cost on the order of £30) by an Oxford-based laser-machining company, SciTech Precision.

### 6.3.2.4 Feed line design

As discussed in the previous section, it is preferable to keep the feed lines symmetric so that the antenna will be balanced. The simplest way to feed

the antenna in a single-layer board is via a pair of microstrips leading up to two SMA connectors. On the back side of the board, ground plane shields the lines from EM fields, preventing the sensitivity pattern from extending along the length of the feed lines. The ground plane stops short of shielding the underside of the loop, however, so that the sensitivity pattern around the antenna tip can be kept as symmetric as possible. Ideally, ground plane would be placed on top of the feed lines as well, so as to eliminate sensitivity to EM fields along the feed lines on both sides of the board. Future prototypes will incorporate a top ground plane layer, which can be implemented by placing a cover board over the antenna. This cover board's antenna-facing side is bare insulator (having had its copper cladding etched away), whilst its top side is copper clad with  $9\mu\text{m}$  copper. This configuration mimics a buried stripline design whilst preserving fabrication simplicity. Fig. 6.3 shows the simulated sensitivity pattern for the antenna with and without the cover board; the sensitivity pattern for the latter case is confined to the antenna tip.

#### **6.3.2.5 Double stub matching**

In order to pre-empt the possibility that mismatches between the antenna and the VNA might lead to a poor signal-to-noise ratio, a matching network was designed as part of the antenna feed line circuitry. A pair of open-circuited stubs were placed along each of the antenna feed lines, to match the antenna to the  $50\Omega$  SMA inputs, maximising power transfer between the antenna loop and the VNA. Stubs act as impedance transformers: by placing them along certain points of the transmission line, we can cancel out the reactive part of the loop load and match the resistive part to  $50\Omega$ . This technique is routinely used to impedance match microwave devices and is described in detail in many standard texts on microwave

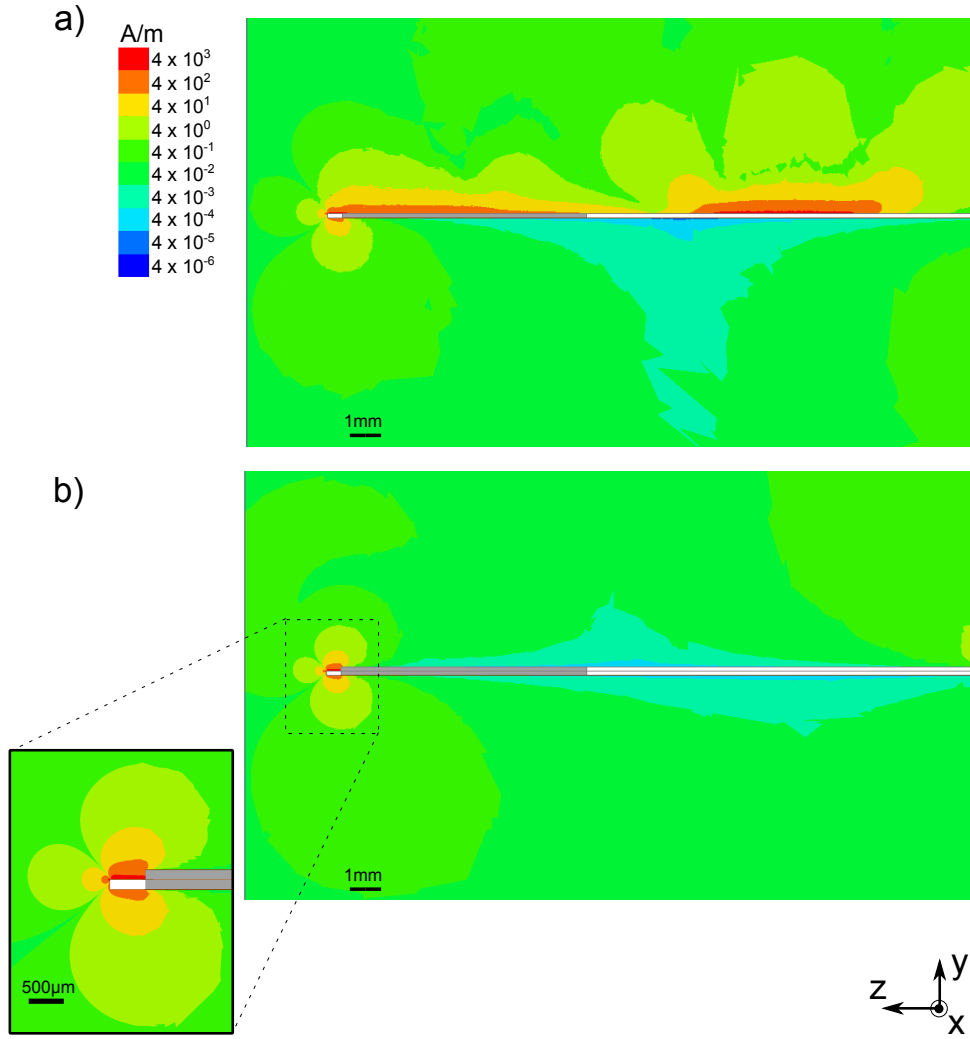


Figure 6.3: Simulated antenna sensitivity patterns for the  $y$  component of the  $H$  field. a) Cross-sectional view of the sensitivity pattern for the antenna design depicted in fig. 6.2. b) Sensitivity pattern for the same antenna design as in (a), but with a cover board added to shield the feed lines between the SMA connectors and the antenna's loop tip (inset is a magnified view of the sensitivity pattern around the antenna's loop tip, which is not shielded by the cover board).

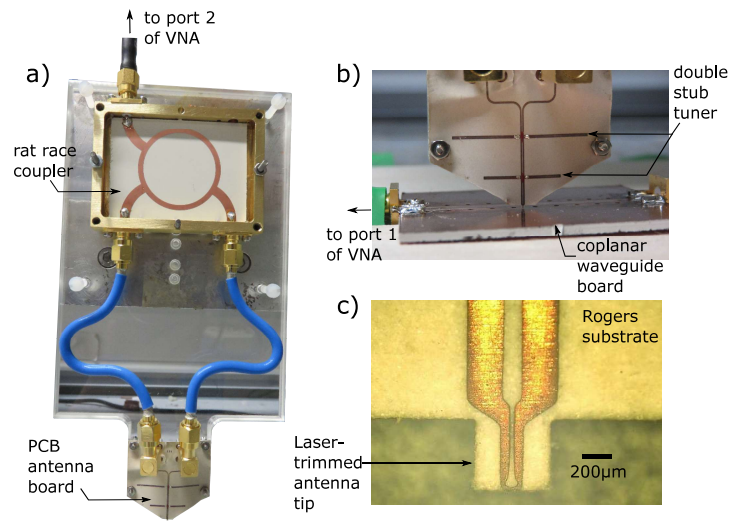


Figure 6.4: a) Assembled antenna probe with rat race coupler. b) Antenna being scanned over a coplanar waveguide. c) Magnified view of laser-trimmed antenna tip.

engineering such as, for instance, [Poz05]. By soldering 1pF tunable capacitors across each pair stubs, we can also tune their impedance to optimize matching and counteract fabrication inaccuracies (PCB fabrication is only accurate to  $\pm 5$  to  $10\mu\text{m}$  and the performance of the matching circuit is sensitive to changes in stub length and width as well as to the spacing of the stubs along the feed lines).

When the first field measurements were taken with this prototype antenna, it became clear that a large signal to noise ratio was attainable without soldering on capacitors to optimize the impedance matching: a signal at 20 to 30dB above the noise floor of the VNA was recorded when I measured fields above a coplanar waveguide over the frequency range of 3 to 4GHz, without the need for an external amplifier.

### 6.3.2.6 Balun design

As discussed in section 6.3.1, microwave flux through the loop excites a differential signal across the two microstrip feed lines and we require a balun-type network to convert this differential signal to a single-ended

signal which can be fed to the VNA. Our balun consists of two stages. The first is the final section of the PCB where the coupled microstrip traces are split apart and each fed to a separate SMA connector. This stage splits the power carried in the coupled microstrip mode into two spatially separate and out-of-phase waveguide modes, each travelling between the ground plane and a single micro-strip line. After this first balun stage, we have converted a differential signal into two single-ended signals which are  $180^\circ$  out-of-phase, each carrying half the power of the differential signal.

In the second stage of the balun, we must combine the two signals into a single waveguide mode which can be coupled to the VNA's SMA connector. Since the signals are out of phase, the largest combined signal will be obtained if we take their difference. We do this by combining the two signals using a rat-race-coupler, a power combiner which phase-shifts one of its inputs by  $180^\circ$ .

There are many ways to implement phase-shifting power combiners. Commercially available models which operate in the 3-4GHz range and provide a  $180^\circ$  phase shift typically cost on the order of a few thousand dollars because they tend to offer large bandwidths, an unnecessary attribute in our application (we are only interested in measuring fields at the frequency of our chosen qubit transition). A rat-race coupler can be easily implemented as a microstrip design and fabricated in-house using PCB techniques. The coupler was designed by tailoring the analytic model MRRCOUP2 in AWR's Microwave Office. The coupler was fabricated on 1.5mm thick Rogers 3003 board with  $17\mu\text{m}$  copper cladding and is shown in fig. 6.5. It is housed in a brass box and contact between the bottom of the box and the bottom copper ground plane of the PCB is made via a thin layer of silver epoxy.

To evaluate the performance of the coupler, the phase imbalance,  $\Delta\phi$ ,

and amplitude imbalance,  $\Delta\alpha$ , across its two input ports were measured. These quantities are defined as follows:

$$\Delta\phi = \text{mod}_{360}(\text{ang}(S_{31}) - \text{ang}(S_{21})) \quad (6.1)$$

$$\Delta\alpha = |S_{31}| - |S_{21}| \quad (6.2)$$

where  $\text{ang}(S_{31})$  is the phase of the complex S-parameter measured between ports 1 and 3 (as numbered in fig. 6.5) of the rat-race coupler.

An ideal rat-race coupler would provide  $180^\circ$  phase imbalance and zero amplitude imbalance between the ports 2 and 3. Fig. 6.6 plots the coupler's measured amplitude and phase imbalance as a function of frequency for frequencies ranging from 3 to 4GHz. We see that measurement and simulation agree to better than 2%: we measure a resonance in the amplitude imbalance at 3.815GHz which was predicted by the HFSS simulation to occur at 3.890GHz (i.e. an error of 75MHz on 3.815GHz). The coupler's performance is closest to ideal around 3GHz, with the best measured performance occurring at 3.01GHz, where the phase imbalance is  $180.11^\circ$  and the amplitude imbalance is 0dB.

Fig. 6.6 shows the measured and simulated amplitude and phase imbalance between the two ports of the coupler.

## 6.4 Pilot experiments with the near-field mapping system

As a first test of the near-field scanning system, the microwave magnetic fields at 1mm above a grounded CoPlanar waveguide (CPW) board and a directional-coupler board were mapped. Fig. 6.7 shows the raw  $S_{21}$  data

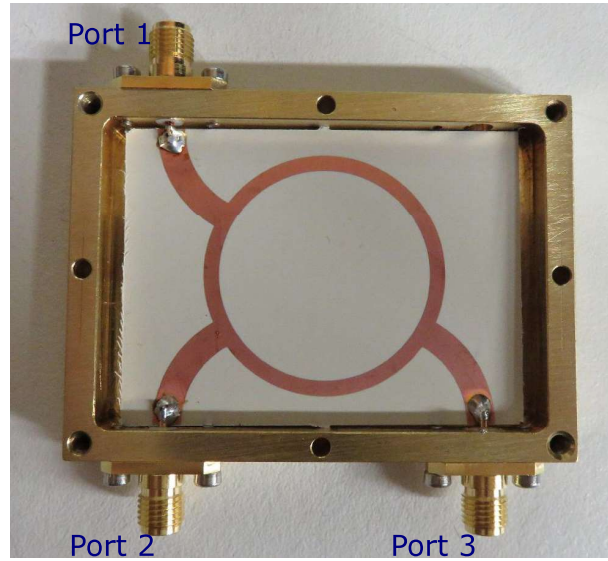


Figure 6.5: Home-made rat race coupler housed in a brass box with lid removed.

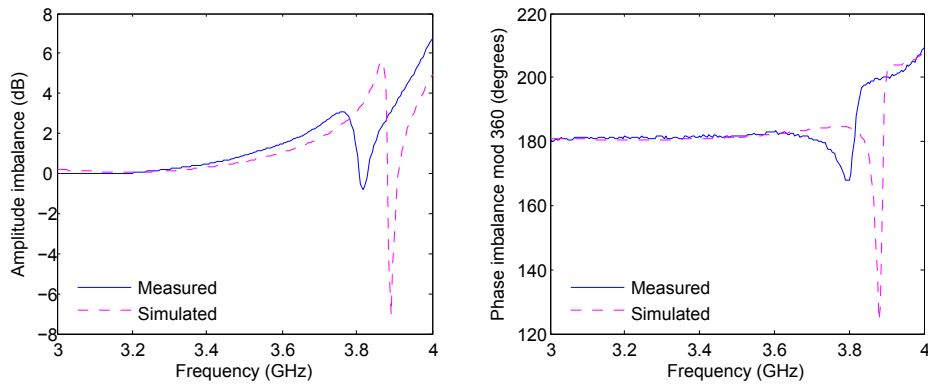


Figure 6.6: Measured and simulated amplitude and phase imbalance for the rat-race coupler.

extracted from the VNA when the antenna was scanned over these test boards.

To compare data to simulation, we must take into account the effect of all components of the EM field on the antenna. To do so, I first extract, from an HFSS model, the antenna's sensitivity to each component of the EM field ( $H_x$ ,  $H_y$ ,  $H_z$ ,  $E_x$ ,  $E_y$ ,  $E_z$ ) in a small volume around its tip. From a model of the test board, I then extract the EM fields generated in a volume of interest around the board. The pattern of the antenna's sensitivity to



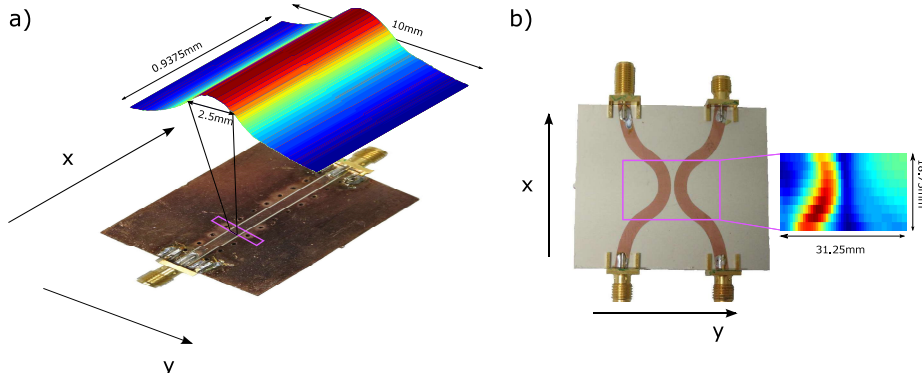


Figure 6.7: Preliminary mapping data taken at 3GHz at  $\approx 1\text{mm}$  above a CPW (a) and a directional-coupler board (b). The pink rectangular outline indicates the area over which the antenna was scanned. This data has not been de-convolved with the antenna sensitivity pattern and is plotted as measured with the VNA.

each EM component is then convolved with the test board's field pattern for that component. The contributions of each EM component to the total field seen by the antenna are then summed (E-field contributions in  $\text{V/m}$  are divided by the impedance of free space,  $377\Omega$ , to give contributions  $\text{A/m}$ , the same units as those of  $H$ -field contributions). Fig. 6.8 shows the results of this analysis for the coplanar waveguide test board. We see from plot b) that the measured field structure is not noticeably wider than the calculated one, which implies the spacial resolution achieved in this test is below  $1\text{mm}$ .

## 6.5 Conclusions and future work

After performing the pilot experiments described here, we conclude that it is feasible to use simple PCB techniques to fabricate loop antenna probes that can map microwave near fields above ion-trap chips with resolutions of a few hundred microns. This resolution is high enough to allow us to obtain a measure of microwave  $H$ -field differentials produced between microwave electrodes in zones which are  $1\text{mm}$  apart. This simple antenna system can hence significantly speed-up the testing of trap chips designed

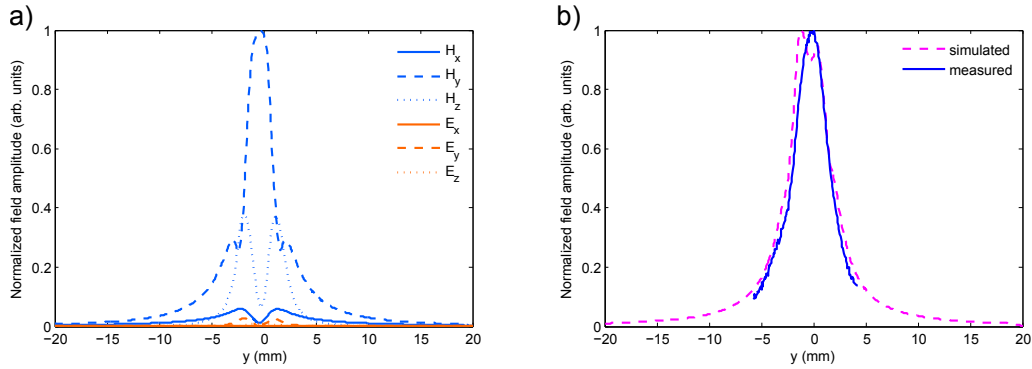


Figure 6.8: a) Calculated contributions from each EM-field component to the signal produced by the antenna when it is scanned over the CPW depicted on the left hand side of fig. 6.7. The strongest contribution is from the  $H$ -field component which is normal to the plane of the loop, whilst E-fields give negligible contributions. b) Measured and simulated fields as we scan the antenna across the  $y$  axis of the CPW (the scan is done with the plane of the antenna oriented normal to the  $y$ -axis) at 1mm above the board's surface. Note: the data has been manually centred in  $y$  so that the maxima of the two curves overlap because a microscope system to record the relative position of the antenna tip and the board during the scan has not yet been set up.

to perform near-field microwave addressing, allowing research groups to efficiently and inexpensively characterize crosstalk, check simulation results and optimize trap designs before placing trap chips under vacuum.

The probe described here was a first prototype which allowed us to ascertain that this method of field mapping is not so technically challenging as to be impractical. It was in fact fairly straightforward to measure small signals with the probe, which was found to have a 30dB dynamic range when used to map fields at 1mm above coplanar waveguide boards. The antenna-scanner system's performance can, however, benefit greatly from the following improvements, which will be implemented to test a second-generation antenna probe:

- A light microscope will be mounted onto the field scanner so that prototype probes can be brought closer to the DUT surface (i.e. 10 to  $100\mu\text{m}$  away from the device). The microscope will also allow us to precisely measure the distance between the antenna tip and the DUT and to accurately position the tip relative to the DUT's electrode

structure;

- The antenna's measurement resolution can be further reduced by decreasing the antenna loop area. Loops with diameters as small as  $50\mu\text{m}$  can be fabricated using commercial PCB techniques. Clean-room-based photolithography techniques can be used to reduce the diameter to under  $10\mu\text{m}$  if even higher resolutions are required. The next prototype antenna design will feature a  $50 \times 50\mu\text{m}$  square loop tip and will be fabricated by a commercial PCB company;
- The new probe will be used to map fields over micro-fabricated ion-trap chips that feature microwave waveguides which are only tens of microns wide. Mapping fields over micron-scale electrode patterns is an important test of the antenna's measurement resolution. Such maps will also provide a detailed view of the near-field close to trap surfaces, which can be used to identify and diagnose discrepancies between measured and simulated field amplitudes.

Once operational, a reliable antenna-based characterisation system for microwave ion-trap chips will significantly speed up and simplify the process of designing new microwave ion and atom chip traps. Progress in the field is currently hampered by our inability to quickly verify simulations, and this simple antenna system will likely prove to be a very useful tool to most ion-trap groups pursuing microwave-driven QIP architectures.



# 7

## Conclusion

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The work presented in this thesis was motivated by the fact that the remarkable potential of microwave-driven clock qubits as platforms for fault-tolerant, scalable quantum computing can only be realized if microwave crosstalk between qubits is efficiently suppressed. The need for a reliable method of crosstalk suppression is in fact common to almost all quantum computing platforms based on qubits which operate in the microwave domain, including superconducting circuits, nitrogen and silicon vacancy centres in diamond and semiconductor quantum dots. Even in implementations where qubits operate at different frequencies, spectral crowding due to a limited bandwidth will eventually mean that spatially structured microwave fields will be required to address individual qubits.

The realization of a microwave-driven QCCD processor architecture with high fidelity hinges on the achievement of large field differentials between processor zones (which are spaced by  $\sim 1\text{mm}$ , a fraction (1%) of the wavelength of the qubit-driving microwaves). These differentials must be created using relatively low microwave powers and an electrode geometry which offers independent field control at each trap zone. Two approaches to achieving this are explored in this thesis. Both are based on driving qubits with oscillating fields produced by microwave electrodes integrated

into each trap zone of a QCCD-type processor. The first approach employs interferometric cancellation of crosstalk fields to produce large single-qubit Rabi frequency differentials between addressed and non-addressed zones. This was implemented using a micro-fabricated two-zone single-layer prototype ion trap (named Microwave Addressing Trap, or MAT) featuring four integrated microwave electrodes per trapping zone.

In a pilot experiment where only a single electrode in each zone was powered, this field-nulling scheme was used to achieve Rabi frequency differentials of up to 1400, from which we calculate a spin-flip probability on the qubit transition of the non-addressed ion of  $1.3 \times 10^{-6}$ . This level of field nulling was found to be stable to environmental fluctuations for over one hour, indicating that, after an initial calibration, it is feasible to perform individual-ion addressing operations with errors  $\lesssim 10^{-5}$  for long periods using this scheme.

Because this pilot experiment was performed using a low-field  $^{43}\text{Ca}^+$  qubit at 2.8G, off-resonant excitation of spectator transitions was a more significant source of error, limiting the total addressing error to the  $10^{-3}$  level. The low-field qubit was used for technical simplicity (to avoid the need for water-cooled magnetic field coils around the vacuum chamber). If the experiment were repeated at a higher static field on the 288G clock qubit in  $^{43}\text{Ca}^+$ , off-resonant excitation would be suppressed to negligible levels and the field ratios of up to 1400 demonstrated here would produce a total addressing error of  $< 4 \times 10^{-6}$ . This error rate is three orders of magnitude below the currently quoted threshold rates necessary for fault-tolerant quantum computation.

MAT was also used here to demonstrate control of the polarisation of the microwave field to a precision which permits off-resonant excitation to be suppressed to the  $10^{-5}$  level. If combined with cross-talk nulling, this po-

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larisation control would enable high-fidelity microwave addressing to be performed even in low-field qubits. The combination was not performed in this work, but can be done by simultaneously driving all four of the microwave electrodes in each of MAT's zones with controlled phase and amplitude relationships.

The work described here was the first demonstration of the use of near-field microwaves to address ions in separate trap zones. Microwave addressing of ions in the same zone had been previously performed at NIST with a spin-flip crosstalk error of  $10^{-3}$  in [WOC<sup>+</sup>13] by selectively shifting one of two ions off the null of a microwave quadrupole field (produced with high-power microwaves) superimposed with the RF trapping potential. This work offers a complementary approach to the NIST scheme by realizing a pilot implementation of individual-ion addressing which can be used to perform arbitrary single-qubit operations in parallel at every zone of a QCCD processor.

This thesis also contributes to the development of microwave-characterization technology by developing an antenna which can be used to map microwave near fields produced by ion-trap chips. The antenna was fabricated in-house using inexpensive PCB fabrication techniques, making this characterization method accessible to any research group hoping to speed up design-fabrication-testing cycles in the production of prototype ion-trap chips. This system can be used to check the accuracy and reliability of EM field simulations, which often give unreliable results due to finite fabrication tolerances and other differences between the modelled and fabricated device.

Finally, a second addressing approach is explored in this thesis. This approach is centred around a multi-layer trap chip architecture (MiLoT) with an electrode structure designed so as to make use of the cancelling effect

of return currents to confine microwave fields. Simulations suggest a single trap electrode in this design can produce intra-zone field differentials of up to 1000, eliminating the need for field nulling (as compared with MAT, which produced single-electrode Rabi frequency differentials of at most 5). If realized, this trap design would enable the implementation of a significantly simpler near-field addressing scheme for performing parallel single-qubit operations in large-scale multi-qubit architectures. This capability to perform parallel operations is something that every quantum computer will need (for example, all the estimates of quantum error correction thresholds assume this capability). Further, the achievement of a better understanding of the microwave methods explored here may lead to insights towards tackling other pertinent problems in quantum computation, such as fast transport of ions around a multizone trap.



# Bibliography

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- [ABLW08] J. M. Amini, J. Britton, D. Leibfried, and D. J. Wineland. Micro-fabricated Chip Traps for Ions. *ArXiv e-prints*, December 2008. [cited at p. 11, 12]
- [ACLH<sup>+</sup>14] D. P. L. Aude Craik, N. M. Linke, T. P. Harty, C. J. Ballance, D. M. Lucas, A. M. Steane, and D. T. C. Allcock. Microwave control electrodes for scalable, parallel, single-qubit operations in a surface-electrode ion trap. *Applied Physics B*, 114(1):3–10, 2014. [cited at p. 21]
- [ACLS<sup>+</sup>17] D. P. L. Aude Craik, N. M. Linke, M. A. Sepiol, T. P. Harty, J. F. Goodwin, C. J. Ballance, D. N. Stacey, A. M. Steane, D. M. Lucas, and D. T. C. Allcock. High-fidelity spatial and polarization addressing of  $^{43}\text{Ca}^+$  qubits using near-field microwave control. *Phys. Rev. A*, 95:022337, Feb 2017. [cited at p. 9, 21]
- [AGH<sup>+</sup>11] D. T. C. Allcock, L. Guidoni, T. P. Harty, C. J. Ballance, M. G. Blain, A. M. Steane, and D. M. Lucas. Reduction of heating rate in a micro-fabricated ion trap by pulsed-laser cleaning. *New Journal of Physics*, 13(12):123023, 2011. [cited at p. 16]
- [AHB<sup>+</sup>13] D. T. C. Allcock, T. P. Harty, C. J. Ballance, B. C. Keitch, N. M. Linke, D. N. Stacey, and D. M. Lucas. A microfabricated ion trap with integrated microwave circuitry. *Applied Physics Letters*, 102(4):044103, 2013. [cited at p. 22]
- [All11] D. T. C. Allcock. *Surface-electrode ion traps for scalable quantum computing*. PhD thesis, University of Oxford, 2011. <http://www.physics.ox.ac.uk/research/ion-trap-quantum-computing-group/publications>. [cited at p. 40, 41, 42, 45, 78]
- [ANK<sup>+</sup>15] N. Akerman, N. Navon, S. Kotler, Y. Glickman, and R. Ozeri. Universal gate-set for trapped-ion qubits using a narrow linewidth diode laser. *New Journal of Physics*, 17(11):113060, 2015. [cited at p. 14]
- [ASS<sup>+</sup>10] D. T. C. Allcock, J. A. Sherman, D. N. Stacey, A. H. Burrell, M. J. Curtis, G. Imreh, N. M. Linke, D. J. Szwer, S. C. Webster, A. M. Steane, and D. M. Lucas. Implementation of a symmetric surface-electrode ion trap with field compensation using a modulated raman effect. *New Journal of Physics*, 12(5):053026, 2010. [cited at p. 46, 52, 80]
- [AUW<sup>+</sup>10] J. M. Amini, H. Uys, J. H. Wesenberg, S. Seidelin, J. Britton, J. J. Bollinger, D. Leibfried, C. Ospelkaus, A. P. VanDevender, and D. J. Wineland. Toward scalable ion traps for quantum information processing. *New Journal of Physics*, 12(3):033031, 2010. [cited at p. 16, 20]
- [Bal14] C. J. Ballance. *High-fidelity quantum logic in  $\text{Ca}^+$* . PhD thesis, University of Oxford, 2014. <http://www.physics.ox.ac.uk/research/ion-trap-quantum-computing-group/publications>. [cited at p. 70]

- [BHL<sup>+</sup>16] C. J. Ballance, T. P. Harty, N. M. Linke, M. A. Sepiol, and D. M. Lucas. High-fidelity quantum logic gates using trapped-ion hyperfine qubits. *Phys. Rev. Lett.*, 117:060504, 2016. [cited at p. 21, 35]
- [BK98] S. B. Bravyi and A. Y. Kitaev. Quantum codes on a lattice with boundary. *ArXiv e-prints, quant-ph/9811052*, 1998. [cited at p. 7]
- [Blo08] I. Bloch. Quantum coherence and entanglement with ultracold atoms in optical lattices. *Nature*, 453:1016–1022, 2008. [cited at p. 3]
- [BMJS16] C. D. Bruzewicz, R. McConnell, Chiaverini J., and J. M. Sage. Scalable loading of a two-dimensional trapped-ion array. *Nature Commun.*, 7:13005, 2016. [cited at p. 16]
- [BN09] I. Buluta and F. Nori. Quantum simulators. *Science*, 326(5949):108–111, 2009. [cited at p. 3]
- [BSH<sup>+</sup>15] C. J. Ballance, V. M. Schäfer, J. P. Home, D. J. Szwer, S. C. Webster, D. T. C. Allcock, N. M. Linke, T. P. Harty, D. P. L. Aude Craik, D. N. Stacey, A. M. Steane, and D. M. Lucas. Hybrid quantum logic and a test of Bell’s inequality using two different atomic isotopes. *Nature*, 528:384–386, 2015. [cited at p. 70]
- [BSWL10] A. H. Burrell, D. J. Szwer, S. C. Webster, and D. M. Lucas. Scalable simultaneous multiqubit readout with 99.99% single-shot fidelity. *Phys. Rev. A*, 81:040302, 2010. [cited at p. 33]
- [Bur10] A. H. Burrell. *High fidelity readout of trapped ion qubits*. PhD thesis, University of Oxford, 2010. <http://www.physics.ox.ac.uk/research/ion-trap-quantum-computing-group/publications>. [cited at p. 50]
- [BWC<sup>+</sup>11] K. R. Brown, A. C. Wilson, Y. Colombe, C. Ospelkaus, A. M. Meier, E. Knill, D. Leibfried, and D. J. Wineland. Single-qubit-gate error below  $10^{-4}$  in a trapped ion. *Phys. Rev. A*, 84:030303, 2011. [cited at p. 20]
- [CBB<sup>+</sup>05] J. Chiaverini, R. B. Blakestad, J. Britton, J. D. Jost, C. Langer, D. Leibfried, R. Ozeri, and D. J. Wineland. Surface-electrode architecture for ion-trap quantum information processing. *Quantum Information and Computation*, 5:419, 2005. [cited at p. 15]
- [CMBK14] S. Crain, E. Mount, S. Baek, and J. Kim. Individual addressing of trapped  $^{171}\text{Yb}^+$  ion qubits using a microelectromechanical systems-based beam steering system. *Applied Physics Letters*, 105(18), 2014. [cited at p. 24]
- [CS96] A. R. Calderbank and P. W. Shor. Good quantum error-correcting codes exist. *Phys. Rev. A*, 54:1098–1105, 1996. [cited at p. 5]
- [CZ12] J. I. Cirac and P. Zoller. Goals and opportunities in quantum simulation. *Nature Phys.*, 8:264–266, 2012. [cited at p. 3]
- [Deu85] D. Deutsch. Quantum theory, the Church-Turing principle and the universal quantum computer. *Proceedings of the Royal Society of London Series A*, 400:97–117, 1985. [cited at p. 3]

- [DGK<sup>+</sup>14] E. A. Dauler, M. E. Grein, A. J. Kerman, F. Marsili, S. Miki, S. W. Nam, M. D. Shaw, H. Terai, V. B. Verma, and T. Yamashita. Review of superconducting nanowire single-photon detector system design options and demonstrated performance. *Opt. Eng.*, 53:081907, 2014. [cited at p. 33]
- [DHK<sup>+</sup>83] R. W. P. Drever, J. L. Hall, F. V. Kowalski, J. Hough, G. M. Ford, A. J. Munley, and H. Ward. Laser phase and frequency stabilization using an optical resonator. *Applied Physics B: Lasers and Optics*, 31(2):97, 1983. [cited at p. 43]
- [Div00] D. P. Divincenzo. The physical implementation of quantum computation. *Fortschritte der Physik*, 48:771–783, 2000. [cited at p. 3]
- [EWA<sup>+</sup>13] A. M. Eltony, S. X. Wang, G. M. Akselrod, P. F. Herskind, and I. L. Chuang. Transparent ion trap with integrated photodetector. *Appl. Phys. Lett.*, 102:054106, 2013. [cited at p. 33]
- [FTS11] R. Folman, P. Treutlein, and J. Schmiedmayer. Atom chip fabrication. In J. Reichel and V. Vuletić, editors, *Atom chips*, chapter 3. Wiley-VCH Verlag GmbH & Co. KGaA, Weinheim, Germany, 2011. [cited at p. 118, 119]
- [Gho95] P. K. Ghosh. *Ion Traps*. International series of monographs on physics. Clarendon Press, 1995. [cited at p. 12]
- [GRS93] M. Gillick, I. D. Robertson, and Joshi J. S. An Analytical Method for Direct Calculation of E & H-Field Patterns of Conductor-Backed Coplanar Waveguides. *IEEE Transactions on Microwave Theory and Techniques*, 41(9):1606–1610, 1993. [cited at p. 121]
- [GTL<sup>+</sup>16] J. P. Gaebler, T. R. Tan, Y. Lin, Y. Wan, R. Bowler, A. C. Keith, S. Glancy, K. Coakley, E. Knill, D. Leibfried, and D. J. Wineland. High-fidelity universal gate set for  ${}^9\text{Be}^+$  ion qubits. *Phys. Rev. Lett.*, 117:060505, 2016. [cited at p. 9]
- [HAB<sup>+</sup>14] T. P. Harty, D. T. C. Allcock, C. J. Ballance, L. Guidoni, H. A. Janacek, N. M. Linke, D. N. Stacey, and D. M. Lucas. High-fidelity preparation, gates, memory, and readout of a trapped-ion quantum bit. *Phys. Rev. Lett.*, 113:220501, 2014. [cited at p. 9, 10, 21, 22, 38, 39, 111, 112]
- [HCW<sup>+</sup>12] D. A. Hite, Y. Colombe, A. C. Wilson, K. R. Brown, U. Warring, R. Jördens, J. D. Jost, K. S. McKay, D. P. Pappas, D. Leibfried, and D. J. Wineland. 100-fold reduction of electric-field noise in an ion trap cleaned with *in situ* argon-ion-beam bombardment. *Phys. Rev. Lett.*, 109:103001, 2012. [cited at p. 16]
- [HDT15] A. Horsley, G.-X. Du, and P. Treutlein. Widefield microwave imaging in alkali vapor cells with sub-100  $\mu\text{m}$  resolution. *New Journal of Physics*, 17(11):112002, 2015. [cited at p. 133, 134]
- [Hou08] M. G. House. Analytic model for electrostatic fields in surface-electrode ion traps. *Physical Review A*, 78:033402, 2008. [cited at p. 75, 77]
- [HSA<sup>+</sup>16] T. P. Harty, M. A. Sepiol, D. T. C. Allcock, C. J. Ballance, J. E. Tarlton, and D. M. Lucas. High-fidelity trapped-ion quantum logic using

- near-field microwaves. *ArXiv e-prints, quant-ph/1606.08409*, 2016. [cited at p. 9, 10, 21]
- [KMK<sup>+</sup>04] E. Kukhareuka, Z. Moktadir, M. Kraft, M. E. Abdelsalam, D. Bagnall, C. Vale, M. P. A. Jones, and E. A. Hinds. Microfabrication of gold wires for atom guides. *Sensors and Actuator, A*, 115:600–607, 2004. [cited at p. 53]
- [KMW02] D. Kielpinski, C. Monroe, and D. J. Wineland. Architecture for a large-scale ion-trap quantum computer. *Nature*, 417(6890):709–711, 2002. [cited at p. 19, 20, 23]
- [KWPO<sup>+</sup>11] I. Kassal, J. D. Whitfield, A. Perdomo-Ortiz, M.-H. Yung, and Aspuru-Guzik A. Simulating chemistry using quantum computers. *Annu. Rev. Phys. Chem.*, 62:185–207, 2011. [cited at p. 3]
- [KWS<sup>+</sup>12] S. Kuehn, M. Wild, P. Sepan, E. Grobelaar, and N. Kuster. Automated near-field EMC/EMI scanning system with active electro-optical field probes. In *IEEE Electrical Design of Advanced Packaging and Systems Symposium (EDAPS)*, pages 109–112, 2012. [cited at p. 135]
- [Lab08] J. Labaziewicz. *High fidelity quantum gates with ions in cryogenic microfabricated ion traps*. PhD thesis, Massachusetts Institute of Technology, 2008. [cited at p. 52]
- [Lei99] D. Leibfried. Individual addressing and state readout of trapped ions utilizing RF micromotion. *Phys. Rev. A*, 60:R3335–R3338, 1999. [cited at p. 14]
- [LGA<sup>+</sup>08] J. Labaziewicz, Y. Ge, P. Antohi, D. Leibbrandt, K. R. Brown, and I. L. Chuang. Suppression of heating rates in cryogenic surface-electrode ion traps. *Phys. Rev. Lett.*, 100:013001, 2008. [cited at p. 17]
- [Lin12] N. M. Linke. *Background-free detection and mixed-species crystals in micro- and macroscopic ion-traps for scalable QIP*. PhD thesis, University of Oxford, 2012. <http://www.physics.ox.ac.uk/research/ion-trap-quantum-computing-group/publications>. [cited at p. 38, 50]
- [LOJ<sup>+</sup>05] C. Langer, R. Ozeri, J. D. Jost, J. Chiaverini, B. DeMarco, A. Ben-Kish, R. B. Blakestad, J. Britton, D. B. Hume, W. M. Itano, D. Leibfried, R. Reichle, T. Rosenband, T. Schaetz, P. O. Schmidt, and D. J. Wineland. Long-lived qubit memory using atomic ions. *Phys. Rev. Lett.*, 95:060502, 2005. [cited at p. 10]
- [LWG<sup>+</sup>10] B. P. Lanyon, J. D. Whitfield, G. G. Gillett, M. E. Goggin, M. P. Almeida, I. Kassal, J. D. Biamonte, M. Mohseni, B. J. Powell, M. Barbieri, Aspuru-Guzik A., and A. G. White. Towards quantum chemistry on a quantum computer. *Nat. Chem.*, 2:106–111, 2010. [cited at p. 3]
- [MBG<sup>+</sup>12] L. Mazza, A. Bermudez, N. Goldman, M. Rizzi, Martin-Delgado M. A., and M. Lewenstein. An optical-lattice-based quantum simulator for relativistic field theories and topological insulators. *New Journal of Physics*, 14(1):015007, 2012. [cited at p. 3]
- [MBM<sup>+</sup>16] K. K. Mehta, D. Bruzewicz, R. McConnell, R. J. Ram, J. M. Sage, and J. Chiaverini. Integrated optical addressing of an ion qubit. *Nat. Nanotechnol.*, 2016:139, 2016. [cited at p. 21]

- [MSW<sup>+</sup>08] A. H. Myerson, D. J. Szwer, S. C. Webster, D. T. C. Allcock, M. J. Curtis, G. Imreh, J. A. Sherman, D. N. Stacey, A. M. Steane, and D. M. Lucas. High-fidelity readout of trapped-ion qubits. *Physical Review Letters*, 100:200502, 2008. [cited at p. 9, 33, 39]
- [OIB<sup>+</sup>07] R. Ozeri, W. M. Itano, R. B. Blakestad, J. Britton, J. Chiaverini, J. D. Jost, C. Langer, D. Leibfried, R. Reichle, S. Seidelin, J. H. Wesenberg, and D. J. Wineland. Errors in trapped-ion quantum gates due to spontaneous photon scattering. *Phys. Rev. A*, 75:042329, 2007. [cited at p. 22]
- [OWC<sup>+</sup>11] C. Ospelkaus, U. Warring, Y. Colombe, K. R. Brown, J. M. Amini, D. Leibfried, and D. J. Wineland. Microwave quantum logic gates for trapped ions. *Nature*, 476:181–184, 2011. [cited at p. 21]
- [Pou46] R. V. Pound. Electronic frequency stabilization of microwave oscillators. *Review of Scientific Instruments*, 17(11):490–505, 1946. [cited at p. 43]
- [Poz05] D. M. Pozar. *Microwave Engineering*. Wiley, 3rd edition, 2005. [cited at p. 143]
- [PSVW14] C. Piltz, T. Sriarunothai, A. F. Varón, and C. Wunderlich. A trapped-ion-based quantum byte with  $10^{-5}$  next-neighbour cross-talk. *Nature Communications*, 5:4679, 2014. [cited at p. 25]
- [RH07] R. Raussendorf and J. Harrington. Fault-tolerant quantum computation with high threshold in two dimensions. *Physical Review Letters*, 98(19):190504, 2007. [cited at p. 7]
- [Sep16] M. A. Sepiol. *A high-fidelity microwave driven two-qubit quantum logic gate in  $^{43}\text{Ca}^+$* . PhD thesis, University of Oxford, 2016. Submitted Trinity term 2016. [cited at p. 41]
- [Sho97] P. W. Shor. Polynomial-time algorithms for prime factorization and discrete logarithms on a quantum computer. *J. Sci. Statist. Comput.* 26 (1997) 1484, 26:1484, 1997. [cited at p. 2]
- [SPM<sup>+</sup>10] K. Singer, U. Poschinger, M. Murphy, P. Ivanov, F. Ziesel, T. Calarco, and F. Schmidt-Kaler. Colloquium: Trapped ions as quantum bits: Essential numerical tools. *Reviews of Modern Physics*, 82:2609–2632, 2010. [cited at p. 78]
- [Ste96] A. Steane. Multiple-particle interference and quantum error correction. *Proceedings of the Royal Society of London Series A*, 452:2551–2577, 1996. [cited at p. 5]
- [Ste03a] A. M. Steane. A quantum computer only needs one universe. *Studies in History and Philosophy of Science B - Studies in History and Philosophy of Modern Physics*, 34(3):469–478, 2003. [cited at p. 2]
- [Ste03b] A. M. Steane. Overhead and noise threshold of fault-tolerant quantum error correction. *Phys. Rev. A*, 68(4):042322, 2003. [cited at p. 8]
- [Ste06] A. M. Steane. A tutorial on quantum error correction. In G. Casati, D. L. Shepelyansky, and P. Zoller, editors, *Proceedings of the International School of Physics “Enrico Fermi”, course CLXII, “Quantum Computers, Algorithms and Chaos”*, pages 1–32, 2006. [cited at p. 6]

- [Sti07] D. L. Stick. *Fabrication and characterization of semiconductor ion traps for quantum information processing*. PhD thesis, University of Michigan, 2007. [cited at p. 81]
- [Szw09] D. Szwed. *High fidelity readout and protection of a  $^{43}\text{Ca}^+$  trapped ion qubit*. PhD thesis, University of Oxford, 2009. <http://www.physics.ox.ac.uk/research/ion-trap-quantum-computing-group/publications>. [cited at p. 39, 41]
- [TGH<sup>+</sup>08] M. Trinker, S. Groth, S. Haslinger, S. Manz, T. Betz, S. Schneider, I. Bar-Joseph, T. Schumm, and J. Schmiedmayer. Multilayer atom chips for versatile atom micromanipulation. *Applied Physics Letters*, 92(25):254102, 2008. [cited at p. 118]
- [TKK<sup>+</sup>00] Q. A. Turchette, D. Kielpinski, B. E. King, D. Leibfried, D. M. Meekhof, C. J. Myatt, M. A. Rowe, C. A. Sackett, C. S. Wood, W. M. Itano, C. Monroe, and D. J. Wineland. Heating of trapped ions from the quantum ground state. *Phys. Rev. A*, 61:063418, 2000. [cited at p. 16]
- [VCA<sup>+</sup>10] A. P. VanDevender, Y. Colombe, J. Amini, D. Leibfried, and D. J. Wineland. Efficient fiber optic detection of trapped ion fluorescence. *Phys. Rev. Lett.*, 105:023001, 2010. [cited at p. 33]
- [WBB<sup>+</sup>02] D. J. Wineland, J. C. Bergquist, J. J. Bollinger, R. E. Drullinger, and W. M. Itano. Quantum computers and atomic clocks. In P. Gill, editor, *Proceedings of the 6th Symposium Frequency Standards and Metrology*, pages 361–368, 2002. [cited at p. 3]
- [WEL<sup>+</sup>07] J. H. Wesenberg, R. J. Epstein, D. Leibfried, R. B. Blakestad, J. Britton, J. P. Home, W. M. Itano, J. D. Jost, E. Knill, C. Langer, R. Ozeri, S. Seidelin, and D. J. Wineland. Fluorescence during doppler cooling of a single trapped atom. *Phys. Rev. A*, 76:053416, Nov 2007. [cited at p. 95]
- [WFSH09] D. S. Wang, A. G. Fowler, A. M. Stephens, and L. C. L. Hollenberg. Threshold error rates for the toric and surface codes. *ArXiv e-prints quant-ph/0905.0531*, 2009. [cited at p. 7]
- [WMI<sup>+</sup>98] D. J. Wineland, C. Monroe, W. M. Itano, D. Leibfried, B. E. King, and D. M. Meekhof. Experimental issues in coherent quantum-state manipulation of trapped atomic ions. *J. Res. Natl. Inst. Stand. Tech.*, 103:259, 1998. [cited at p. 14]
- [WOC<sup>+</sup>13] U. Warring, C. Ospelkaus, Y. Colombe, R. Jördens, D. Leibfried, and D. J. Wineland. Individual-Ion Addressing with Microwave Field Gradients. *Physical Review Letters*, 110(17):173002, 2013. [cited at p. 153]
- [WZR<sup>+</sup>12] A. Walther, F. Ziesel, T. Ruster, S. T. Dawkins, K. Ott, M. Hettrich, K. Singer, F. Schmidt-Kaler, and U. Poschinger. Controlling fast transport of cold trapped ions. *Phys. Rev. Lett.*, 109:080501, 2012. [cited at p. 20]
- [Yag86] A. D. Yaghjian. An overview of near-field antenna measurements. *IEEE Trans. Antennas Propag.*, 34(1):30–45, 1986. [cited at p. 135]

# List of abbreviations

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| Symbol | Description                                    |
|--------|------------------------------------------------|
| AM     | amplitude modulation                           |
| AOM    | acousto-optic modulator                        |
| AR     | anti-reflection                                |
| ASE    | amplified spontaneous emission                 |
| CCD    | charge-coupled device                          |
| CF     | ConFlat                                        |
| CPGA   | ceramic pin grid array                         |
| CPW    | coplanar waveguide                             |
| DAC    | digital-to-analogue convertor                  |
| DDS    | direct digital synthesis                       |
| DI     | de-ionised                                     |
| EOM    | electro-optic modulator                        |
| FPGA   | field-programmable gate array                  |
| GPIO   | general purpose interface bus                  |
| HFSS   | high-frequency structural simulator            |
| IPA    | isopropyl alcohol                              |
| LCU    | laser control unit                             |
| LO     | local oscillator                               |
| MAT    | Microwave addressing trap                      |
| MiLoT  | Microwave loop trap                            |
| NIST   | National Institute of Standards and Technology |
| NPL    | UK National Physical Laboratory                |
| PCB    | printed circuit board                          |
| PBS    | polarising beamsplitter                        |
| PCI    | peripheral component interconnect              |
| PDH    | Pound Drever Hall                              |
| PEEK   | polyetheretherketone                           |
| PI     | photoionisation                                |
| PMT    | photomultiplier tube                           |
| PZT    | piezoelectric transducer                       |
| QC     | quantum computer                               |
| QCCD   | quantum charge-coupled device                  |
| SMA    | subminiature version A                         |
| TTL    | transistor-transistor logic                    |
| UHV    | ultra-high vacuum                              |
| VNA    | vector network analyser                        |