

What Is the Maximum Density of Microwave Control Lines in a Superconducting Quantum Computer?

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Scaling up superconducting quantum computers requires denser control wiring, introducing microwave crosstalk along the entire cryogenic signal path rather than only near the chip. Here, we introduce a framework that combines microwave transmission-line theory with a quantum Hamiltonian description to quantify the wiring-induced microwave crosstalk and the resulting gate infidelity for an all-microwave universal gate set. Using this approach, we determine the equal-level far-end crosstalk (ELFEXT) required to operate fixed-frequency transmons at arbitrary target fidelities and find that achieving 99.99% fidelity requires ELFEXT levels of approximately -43 dB for the R_X gate and -73 dB for the cross-resonance gate. Finally, we use these results to bound the number of physical qubits on a chip.

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Similar to classical computers, the interconnects that route input and output information to a quantum processor will govern the computing performance [1–4]. In cryogenic quantum computing, interconnects themselves influence multiple criteria at the system level [5], including (1) scalability [2,6–9], (2) the capability of qubit-specific measurement [10–12], and even (3) the rates of decoherence arising from external noise [13–16]. The standard approach toward fulfilling criteria (1) and (2) has been to increase the number of physical control cables within a dilution refrigerator system (DR). The total volume of a DR is governed by its cooling capacity and therefore poses a constraint on the total number of cables that can be installed. Given these constraints, denser interconnects are necessary, which inevitably increase microwave crosstalk.

In current designs, semirigid coaxial cables transmit signals and provide shielding against electromagnetic interference. Consequently, most of the crosstalk occurs near the quantum processing unit (QPU), where conductors are crowded and shielding is partially removed, or directly on-chip. These conditions have been addressed extensively in the literature from multiple perspectives, including flux-mediated effects [17–20], microwave interactions

across different superconducting qubit modalities and operations [21–25], and dynamical decoupling for crosstalk suppression during readout operations [26,27]. Recent works have also shed light on on-chip crosstalk and how it affects a multitude of qubit operations [28–33]. Finally, the topic of benchmarking quantum devices that suffer from correlated errors has also been explored, and correlated randomized benchmarking schemes have been implemented [34,35].

In this Letter, we analyze a different regime where signal density in the DR increases up to the limit set by the pitch of the QPU. Recent efforts have been made in this direction by introducing flexible printed circuit (FPC) wiring that accommodates multiple conductors in a single laminar structure, such as a stripline [36–39]. Because of close proximity, these conductors are less efficiently shielded from one another. Hence, contrary to the current coaxial-based approach, significant crosstalk might already develop far from the quantum processor and affect drastically the relevant fidelities. Motivated by these possible limitations, we seek to (1) determine the overall crosstalk from FPC-based control lines, (2) evaluate its impact on qubit control, and (3) estimate how it may limit the qubit number. To connect seamlessly all these aspects we introduce a general theoretical framework that combines microwave theory with Hamiltonian models. This enables the derivation of analytical equations for single- and two-qubit gate fidelities as functions of design parameters of the control wiring. We will use the trade-off between benchmark fidelities and microwave crosstalk to determine an upper bound to the number of qubits that can be efficiently manipulated.

In Fig. 1 we illustrate a comprehensive framework to investigate the limit where the line-to-line pitch (d_l)

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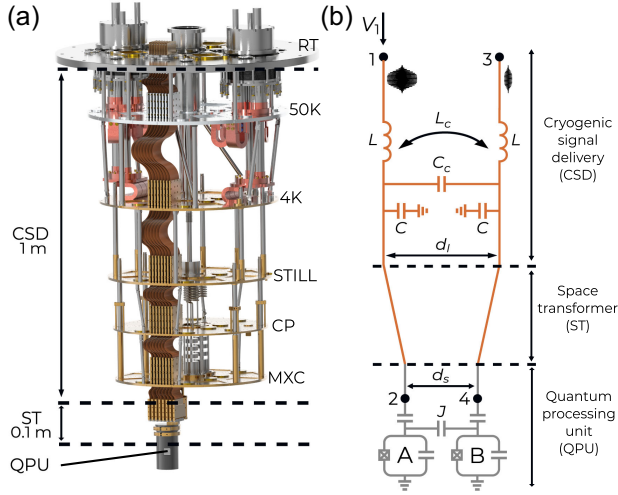


FIG. 1. (a) Illustration of a dilution refrigerator depicting an implementation of flexible printed circuit wiring. A dense multi-conductor cable traverses the CSD region, while the ST section provides the interface between the cable and the QPU. (b) Equivalent circuit diagram of a pair of transmission lines spanning the CSD, ST, and QPU regions.

approaches the on-chip pitch (d_s). Figure 1 presents both the FPC integration in a DR and a circuit schematic. The signal transmission lines fill a dedicated volume that runs through a line-of-sight port from the room-temperature (RT) feedthrough to the base-temperature stage of the dilution refrigerator. We name this volume the cryogenic signal delivery (CSD) region, which spans four decades of temperature and approximately 1 m in length. In contrast to the more traditional coaxial approach with typical $d_1 = 5$ mm (and space transformation ratio of $d_1/d_s = 50$) in a typical high-density configuration [40], FPCs enable space transformation ratios of order unity. In Fig. 1, below the mixing chamber (MXC) flange, a space transformer (ST) (e.g., a routing interposer) is usually employed to reduce further the line pitch down to on-chip spacings, which can be as small as $d_s = 0.1$ mm [32]. The assembly in Fig. 1 shows three distinct regions where microwave crosstalk can be present: the CSD region, the ST region, and the QPU region. The latter two have smaller spatial extent and are positioned below the MXC flange.

We model the cable in the CSD region as a transmission line (TL), whose equivalent circuit diagram is depicted in Fig. 1(b). Here, L and C denote the self-inductance and self-capacitance per unit length of the TL, respectively. The signal conductors in the TL will experience a mutual capacitance per unit length C_c and a mutual inductance per unit length L_c [41], both dependent on d_1 . This microwave structure is known as a coupled transmission line and, under single-ended excitation, the interline couplings will induce signal leakage between conductors. While the scenario generalizes to a coupled TL with N signal conductors, the dominant interaction occurs between

adjacent traces (see Sec. I in Ref. [42]); thus, we focus on the $N = 2$ case to capture the essential physics. We describe the coupled TL as a 4-port network and use standard scattering parameters to characterize its properties [41]. As shown in Fig. 1(b) we define input ports as 1 and 3 at the RT feedthrough, and the output ports 2 and 4 at the QPU reference plane. We write the corresponding output voltage waves as a function of the input voltage wave V_1 for an impedance matched system in phasor representation as

$$V_{2/4}(\ell) = \frac{V_1}{2} \left(e^{-\gamma_e \ell} \pm e^{-\gamma_o \ell} \right), \quad (1)$$

where ℓ is the length of the transmission line and $\gamma_{e/o}$ are the complex propagation constants of the even and odd modes, which depend on the distributed parameters of the coupled TL. In this setting, the relevant S parameters are $S_{21} \equiv V_2(\ell)/V_1$ and $S_{41} \equiv V_4(\ell)/V_1$, which represent the transmission of the intended signal, and the far-end crosstalk (FEXT) between ports 1 and 4, respectively. To quantify the effect of crosstalk on qubit control operations, we use the equal-level far-end crosstalk (ELFEXT), introduced in Refs. [46,47], which reads

$$\xi = 20 \log_{10} \left| \frac{S_{41}}{S_{21}} \right| \text{dB} \simeq 20 \log_{10} \left[\frac{\ell \omega \sqrt{LC}}{2} \left(\frac{L_c}{L} + \frac{C_c}{C} \right) \right] \text{dB}, \quad (2)$$

where the approximation in the second line holds in the weak coupling limit, that is $L_c/L, C_c/C \ll 1$. We choose this quantity because it normalizes the total amount of signal leakage to the amount of intended, attenuated signal delivered [2].

Following this analysis, we measured the frequency-dependent behavior of ELFEXT as a function of normalized pitch (d/W) for 12 different samples. The samples were 100-mm-long coupled stripline pairs on FPC with a cross section illustrated in the inset of Fig. 2(a) (see Sec. II A in Ref. [42]). In Fig. 2(a), we plot the measurement results as cross markers at the frequencies 1, 4, and 8 GHz, with d/W ratios spanning 1 to 4. Next, we describe the measured data using Eq. (2). First, we convert the even- and odd-mode impedances described by Cohn [48] into the mutual inductance and capacitance values L_c and C_c that serve as inputs to Eq. (2). Both mutual inductance and mutual capacitance decrease exponentially in d/W within the model. However, the analytically derived results from Cohn overestimate the measured ELFEXT. We therefore introduce two global fit parameters that rescale both the overall magnitude and the exponential decay length of L_c and C_c as functions of d/W , while preserving the functional form of the model. After optimizing those global parameters via numerical minimization (see Sec. II B in Ref. [42]), we found that Eq. (2) provides an excellent fit to the data, plotted as the solid lines. Importantly, the solid

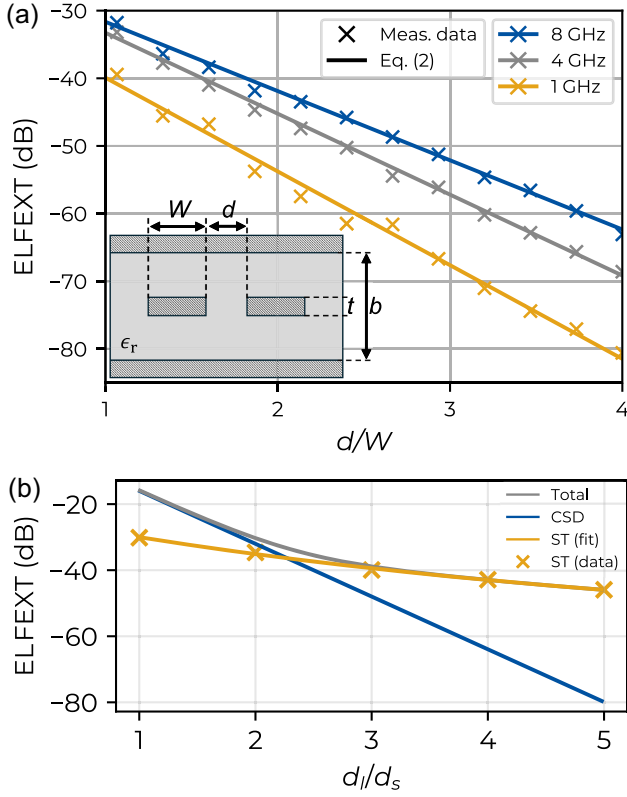


FIG. 2. (a) Stripline ELFEXT vs d/W . The solid lines show Eq. (2) with ξ evaluated using fitted coupling parameters L_c and C_c ; crosses denote measured data. Inset: cross section of a coupled stripline with two conductors of thickness t embedded in a dielectric filling of thickness b characterized by an effective permittivity ϵ_r . (b) The CSD, ST, and combined ELFEXT expressed as a function of the d_l/d_s ratio.

lines in Fig. 2(a) represent an experimentally verified, closed-form expression for $\xi(d/W)$ that serves as a direct input to the fidelity analysis presented in the following section. Finally, as a consistency check we performed a finite element method (FEM) simulation and achieved fair agreement with both the measured data and theory (see Sec. II C in Ref. [42]).

Recall from Fig. 1 that the interface between the CSD region and the QPU is a space transformer, which is required in the regime where $d_l > d_s$ and is physically realized by a pitch reduction component (e.g., an interposer). Here, we model the ST as a tapered coupled stripline with a line pitch decreasing linearly from an initial d_l to $d_s = 0.1$ mm and compute its contribution to the total ELFEXT using FEM simulations. Figure 2(b) compares the total ELFEXT (CSD + ST, gray) calculated using a cascaded model (see Sec. III in Ref. [42]), with the individual contributions from the CSD (blue) and ST (orange) regions, as a function of the pitch reduction ratio d_l/d_s for 1-m-long CSD stripline and a 100-mm-long ST, with $W = 75$ μm . Below the crossover point at $d_l/d_s \approx 2.3$, the total ELFEXT is dominated by the CSD stripline, with a

negligible ST contribution. On the other hand, at values greater than the crossover point, where the CSD lines are sparse, the total ELFEXT is limited by the ST contribution, rendering any further improvement on the CSD stripline crosstalk ineffective. Thus, for a fixed on-chip signal pitch of order 0.1 mm, Fig. 2(b) reveals the critical pitch in the CSD region of $d_l = 0.23$ mm, below which, a novel regime occurs where the cryogenic wiring will limit the total crosstalk. We note that for a different type of ST, the value of the crossover point will change as described in Sec. III in Ref. [42].

The outputs of the TL model can be incorporated into arbitrary Hamiltonian models via the derived voltage waves V_2 and V_4 . We consider the following Hamiltonian [cf. Fig. 1(b)]:

$$\hat{H} = \frac{\omega_A}{2} \hat{\sigma}_A^z + \frac{\omega_B}{2} \hat{\sigma}_B^z + gV_2 \hat{\sigma}_A^x + gV_4 \hat{\sigma}_B^x + J \left(\hat{\sigma}_A^+ \hat{\sigma}_B^- + \hat{\sigma}_A^- \hat{\sigma}_B^+ \right), \quad (3)$$

where we assumed $\hbar = 1$, $\omega_{A(B)}$ is the resonance frequency of qubit $A(B)$, g is the qubit-line coupling constant, J the qubit-qubit coupling constant and $\hat{\sigma}^z, \hat{\sigma}^x, \hat{\sigma}^\pm$ are z and x Pauli and raising and lowering operators, respectively. Note that even in the absence of direct interaction ($J = 0$), the qubits are dynamically correlated due to the coupling between the lines. Equation (3) enables both single- and two-qubit gates. In circuit implementations, single-qubit operations generally require individual addressability via control lines transversely coupled to the qubit. Two-qubit gates can be implemented using fixed capacitive coupling [49,50], tunable couplers [51,52], or microwave resonators [53,54]. The coupled TL model can be used to quantify the effect of microwave crosstalk during gate operations in all of these architectures. As an extension, we also derive the circuit Hamiltonian including crosstalk terms between qubits and a tunable coupler drive [43] (see Sec. VI in Ref. [42]).

Now, let us consider fixed-frequency transmons, treated as two-level systems (see Sec. VII in Ref. [42] for details on anharmonic contribution to the dynamics), coupled via a fixed capacitor and focus on arbitrary single-qubit rotations and the cross-resonance (CR) gate [55,56], both implemented using Gaussian pulses. Owing to its operating principle, the strong sensitivity of the CR gate to resonant and quasiresonant microwave crosstalk makes it the perfect candidate for our analysis. To evaluate the impact of microwave crosstalk on a target qubit gate U we compute the average gate fidelity defined as

$$\bar{F}_U = \frac{|\text{Tr}(U^\dagger V)|^2 + D}{D(D+1)}, \quad (4)$$

where V is the unitary operation implemented via the corresponding control Hamiltonian and D is the dimension

of the relevant Hilbert space. Any general single-qubit rotation can be implemented using $\hat{\sigma}_z$ and $\hat{\sigma}_x$. While $\hat{\sigma}_z$ can be realized virtually through a qubit phase modulation and, as such, does not incur additional errors [57], realizing $\hat{\sigma}_x$ actively changes the qubit state. Thus, any single-qubit input voltage at port 1 (or 3) will excite both the even and odd modes of the coupled stripline, affecting the spectator qubit at port 4 (or 2), depending on the magnitude of L_c and C_c . To illustrate this effect best, we set $J = 0$ and $\omega_A = \omega_B = 2\pi \times 6$ GHz, assume $L_c/L, C_c/C \ll 1$, and impose the boundary conditions $V_1 = V(t), V_3 = 0$ [Fig. 1(b)]. The detailed derivation of a single-qubit rotation Hamiltonian is given in Sec. IV in Ref. [42]. In the case of an X drive this reads

$$\hat{H}_{\text{rf}}^d = \frac{gV(t)e^{-\alpha\ell}}{2} \left(-\hat{\sigma}_A^x Q_+ + \hat{\sigma}_B^y I_- \right), \quad (5)$$

where $Q_+ = \cos(\beta_e \ell/2 - \beta_o \ell/2)$, $I_- = \sin(\beta_e \ell/2 - \beta_o \ell/2)$, α is the attenuation constant of the line, and β_e, β_o are the imaginary parts of the propagation constant of the even and odd mode, respectively. The second term generates an unintended rotation of the spectator qubit state at a detuned Rabi frequency. Using Eqs. (4) and (5) the average gate fidelity of a single-qubit X rotation by an angle θ , affected by ELFEXT can be calculated analytically using the coupled stripline model, leading to

$$\bar{\mathcal{F}}_{R_X}(\theta) = \frac{4}{5} \left[\cos^2 \left(\frac{\theta}{2} \frac{I_-}{Q_+} \right) + \frac{1}{4} \right] \simeq 1 - \frac{\theta^2}{5} 10^{\frac{\epsilon}{10 \text{ dB}}}. \quad (6)$$

For the CR gate, we must assume that the qubits are nearly resonant and set $\omega_A = 2\pi \times 6$ GHz, $\omega_B = 2\pi \times 5.8$ GHz, and $J = 2\pi \times 3.8$ MHz. Following the standard derivation reported in Ref. [44], we can recast the CR Hamiltonian in the following form (see Sec. V in Ref. [42] for full derivation):

$$\hat{H}_{\text{rf}}^d = -\frac{gV(t)e^{-\alpha\ell}}{2} \left(\hat{\sigma}_A^z \hat{\sigma}_B^x \frac{J}{\Delta} Q_+ + \hat{\sigma}_B^y I_- \right), \quad (7)$$

where $\Delta = \omega_A - \omega_B$ is the qubit-qubit frequency detuning. The last term in Eq. (7) originates from the coupling of conductors in the coupled TL and, similarly to the single-qubit case, leads to unintended detuned Rabi oscillations. We can again derive an analytical expression for the average gate fidelity of the CR gate using Eq. (4), which takes the following simplified form when phase errors from the Y rotation are neglected:

$$\bar{\mathcal{F}}_{CR} \simeq \frac{4}{5} \left[\cos^2 \left(\frac{\pi \Delta}{4J} \frac{I_-}{Q_+} \right) + \frac{1}{4} \right] \simeq 1 - \frac{\Delta^2 \pi^2}{20J^2} 10^{\frac{\epsilon}{10 \text{ dB}}}. \quad (8)$$

We emphasize that our approach ties together the transmission line model of the coupled stripline, the ELFEXT, calculated using the 4-port approach, and the average

fidelity of single- and two-qubit gates in a single analytical framework.

The infidelities of Eqs. (6) and (8) are displayed in Fig. 3(a), as a function of d_1/W for a stripline with $l = 1$ m, assuming no pitch reduction. Both infidelities decrease as the distance between the conductors in the TL increases, with the CR gate being more strongly affected for a given d_1/W . This can be attributed to the damping of the intended ZX interaction in Eq. (7) by the J/Δ ratio. For instance, in order to achieve a CR gate fidelity of 99.9%, the spacing between the signal conductors must satisfy the condition $d_1/W > 6$, which imposes a practical limit on how densely signal conductors can be arranged. While this requirement might still be achievable without sacrificing signal density, crosstalk originating in the ST region—such as from a connector or an interposer—can remain a limiting factor.

When cascading the CSD with the ST, the total ELFEXT can be expressed as a function of the individual component contributions (see Sec. VIII in Ref. [42]). Combined with Eqs. (6) and (8), this enables determination of the total infidelity. Using this approach, Fig. 3(b) displays the trade-off between the ST and the CSD ELFEXT as contour lines

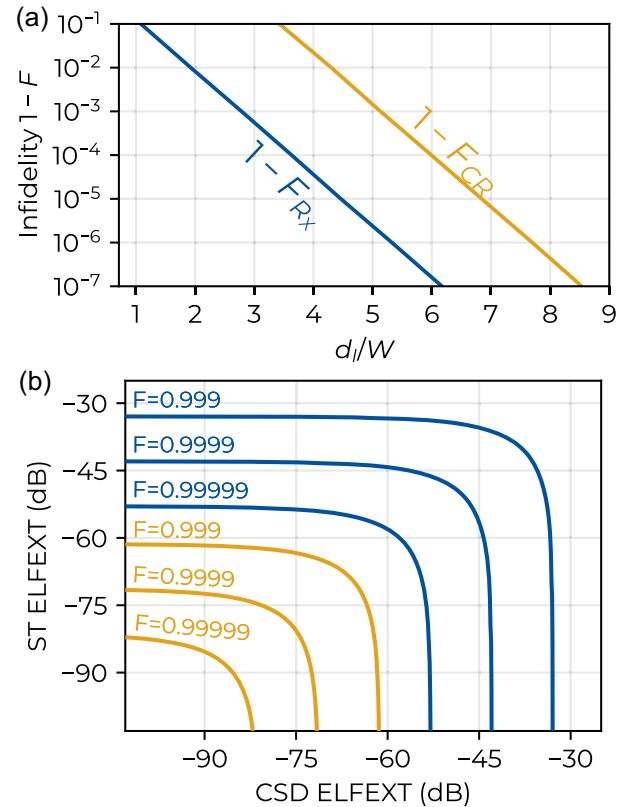


FIG. 3. (a) Average gate infidelities of $R_X(\pi)$ and CR vs the rescaled line spacing d_1/W assuming a 1 m-long stripline. (b) Fidelity contour lines for the $R_X(\pi)$ gate (blue) and the CR gate (orange) as functions of both the CSD (x axis) and the ST (y axis) ELFEXT. The qubit frequencies are set to $\omega_A = 2\pi \times 6$ GHz, $\omega_B = 2\pi \times 5.8$ GHz, while $J = 2\pi \times 3.8$ MHz.

of average gate fidelities, both for the $R_X(\pi)$ (blue) and the CR (orange) gate. Each additional order of magnitude improvement in fidelity (an extra nine) requires approximately a 10 dB reduction in the total ELFEXT. For example, in the limit where only one component, either the CSD or the ST, is limiting the performance, benchmark fidelities of 99.99% for both the $R_X(\pi)$ and CR gates require an ELFEXT of approximately -43 and -73 dB, respectively. These results are pivotal, as they directly connect the control system crosstalk budget to the fidelity of its quantum operations.

Most importantly, as we now show, Fig. 3 enables us to estimate an upper bound n_q to the number of physical qubits that can be controlled while achieving those fidelities. We consider the scenario where the system performance is limited by the CSD lines. In Fig. 4, we show the bound n_q obtained for both the $R_X(\pi)$ (dashed) and the CR (solid) gate, assuming a 2D square configuration with side length S , where the CSD wiring connects only to the sides of the QPU sample holder, e.g., via bond wires. When computing n_q , we assume that the chip interconnects are distributed uniformly on all sides and impose that the outer perimeter of the chip divided by the total number of connections per side cannot exceed a minimum distance $d_l^{\min}$ set by Eqs. (6) and (8). Moreover, consistent with current developments [9,58], we assume that every physical qubit requires at least one XY line, leading to the constraint $n_q = 4S/d_l^{\min}$. This approach to deriving scaling laws for bounds in quantum computing is broadly applicable whether the chip is a standalone planar QPU, a module within a 3D-integrated stack [59], or a quantum chiplet [60,61], and it can be generalized to include other types of qubit lines, qubit designs and circuit Hamiltonians. As shown in Fig. 4, when different chip sizes are considered, the above condition will provide an estimate for the

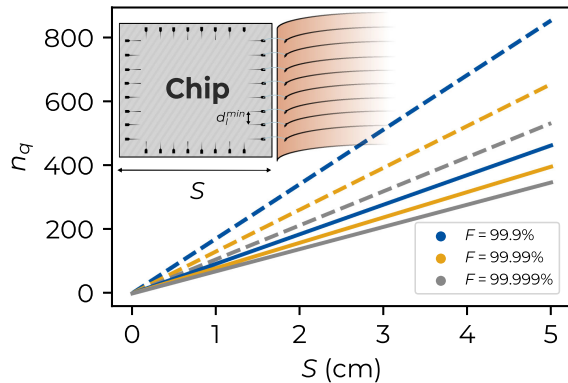


FIG. 4. Upper bound n_q to the number of physical qubits, as determined by target gate fidelities. The dashed lines correspond to $\tilde{F}_{R_X(\pi)}$ and the solid lines correspond to $\tilde{F}_{CR}$. Inset: direct cable-to-chip configuration, where the minimum line pitch $d_l^{\min}$ defines the maximum achievable interconnect density for a square chip with uniformly distributed connections.

maximum number of qubits on a square chip for which a specific target fidelity can be achieved using a multi-conductor transmission line. For example, requiring a CR gate fidelity of 99.9% in a square chip with $S = 30$ mm yields an upper bound of about 270 physical qubits. In passing, we note that by employing vectoring techniques [62] on the input signal, all fidelities may be improved, resulting in larger n_q .

In conclusion, by integrating transmission line theory with a two-qubit Hamiltonian model, we introduced a methodology to connect analytically microwave crosstalk and quantum operations. We illustrated our approach by focusing on the average fidelity of some universal all-microwave gates. Furthermore, we used this connection to make quantitative predictions of chip controllability as a function of qubit number and determined bounds for a planar geometry. At the system-architecture level, our analysis provides a quantitative guidance for crosstalk allocation.

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Data availability—The data that support the findings of this article are not publicly available because they contain commercially sensitive information. The data are available from the authors upon reasonable request.

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