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How Much Impedance Bandwidth Does an LTE or 5G NR Antenna Need? Closed-Form OFDM Link Penalties for Resonant Antennas

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Abstract: Antenna papers often declare a design fit for LTE or 5G NR when its −10 dB reflection bandwidth covers the band, but the cost to an orthogonal frequency-division multiplexing (OFDM) link is rarely quantified. This paper derives that cost for resonant antennas. Each subcarrier sees a power transfer of $1 - |\Gamma|^2$, and the link penalty depends only on the ratio r of occupied bandwidth to −10 dB impedance bandwidth and on the channel offset from resonance. Under Rayleigh fading, the required-SNR penalty is $10\log_{10}(1 + r^2/27)$ for one antenna and $10\log_{10}(1 + 2r^2/27 + r^4/405)$ for a transmit–receive pair, independent of the modulation order. Simulations of LTE 20 MHz and NR 100 MHz links in EPA, EVA and ETU channels with QPSK–64QAM and least-squares pilot estimation agree with the analysis within the 95% confidence interval in all 84 cases, and estimation changes it by at most 0.16 dB. A channel filling the −10 dB band costs 0.32 dB. A 0.5 dB budget allows $r \leq 1.25$, which for NR 100 MHz at 3.55 GHz means $Q \leq 30$ and a radius of at least 4.5 mm. Double tuning helps at equal antenna Q but is worse at equal −10 dB bandwidth when $r > 1.27$.

Keywords: OFDM; LTE; 5G NR; antenna bandwidth; quality factor; impedance mismatch; link-level simulation

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1. Introduction

Printed and microstrip antennas for cellular terminals, small cells and Internet-of-Things devices are usually judged by one number: the bandwidth over which the reflection coefficient stays below −10 dB, often stated as a voltage standing-wave ratio (VSWR) below about 2 [1], [2]. When that band covers an LTE or 5G New Radio (NR) operating band, the antenna is normally declared "suitable for LTE/5G". The −10 dB criterion is a convention, not a link-level requirement. At the band edge it still reflects 10% of the incident power. The criterion also says nothing about how the reflected power is distributed across the hundreds or thousands of subcarriers of an orthogonal frequency-division multiplexing (OFDM) signal.

The question matters because modern channel bandwidths are no longer small compared with antenna bandwidths. An LTE carrier occupies up to 18 MHz (100 resource blocks) [3], [4]. An NR carrier in band n78 at 3.3–3.8 GHz occupies up to 98.28 MHz (273 resource blocks at 30 kHz subcarrier spacing) [5], [6]. Compact antennas are pushed towards high quality factor Q by the fundamental size limits of Wheeler, Chu and McLean [7]–[9], and Q sets the impedance bandwidth [10]. Antenna designers therefore face a practical trade-off: how much −10 dB bandwidth, or how low a Q, does a given LTE or NR channel actually need? OFDM links of this kind also serve applications beyond cellular access, such as MIMO-OFDM data transfer in smart grids

[11], and LTE itself continues to be refined above the physical layer, for example in its security and handover procedures [12].

Two groups of existing results bracket this question without answering it. Antenna theory relates Q, size and matched bandwidth [7]–[10], [13], [14] but stops at the antenna port. Information-theoretic work has combined these limits with Shannon capacity. Shyianov et al. derived the achievable rate of size-constrained antennas using the Chu and Bode–Fano limits [15]. Deshpande et al. generalized the analysis to multiple-input single-output (MISO) systems with wideband matching constraints [16]. Circuit-aware estimators for multicarrier arrays [17] address channel estimation rather than link penalties. None of these studies models a standard-compliant OFDM receiver with pilot-based channel estimation, 3GPP fading channels and uncoded M-QAM error rates, or gives a design rule stated in the −10 dB bandwidth that antenna engineers actually measure. Other work treats mismatch from the transmitter side, such as the error vector magnitude (EVM) of power amplifiers under wideband load mismatch [18] or the radiated-power loss of mismatched IoT antennas [19]. Industry notes argue that linear RF filtering is largely removed by LTE channel estimation [20], but this has not been quantified for antenna mismatch.

This paper closes that gap for the most common antenna type in compact terminals, the single-resonance antenna, and for its double-tuned (broadbanded) version. Its contributions are as follows:

1. **A universal description of the per-subcarrier SNR profile.** For a lossless antenna, each subcarrier sees a power transfer of $1 - |\Gamma|^2$. The profile depends only on the ratio r of occupied bandwidth to −10 dB impedance bandwidth and on the channel's offset from resonance. It is the same for LTE and NR and for any carrier frequency.
2. **Closed-form link penalties.** Under Rayleigh fading with perfect channel knowledge, the required-SNR penalty at high SNR is $10\log_{10}(1 + r^2/27)$ for one antenna and $10\log_{10}(1 + 2r^2/27 + r^4/405)$ for a transmit–receive antenna pair. The penalty is independent of the modulation order. It generalises to off-centre channels and inverts to an explicit maximum r for a given penalty budget.
3. **Link-level validation with 3GPP numerologies and channels.** Monte Carlo simulation covers LTE 20 MHz and NR 100 MHz with EPA, EVA and ETU channels, QPSK to 64QAM, and least-squares (LS) pilot estimation. It shows that the closed form holds and that pilot-based estimation does not add a mismatch-specific penalty.
4. **A design table.** The table gives the maximum Q, minimum −10 dB bandwidth and minimum antenna size (via the Chu–McLean bound) for common LTE and NR channels. A reusable script computes the same penalty directly from a measured or simulated Touchstone S11 file.

The rest of the paper is organised as follows. Section 2 reviews related work. Section 3 presents the system and antenna models. Section 4 derives the closed-form penalties and the design rule. Section 5 describes the simulation method. Section 6 reports and discusses the results. Section 7 states the limitations, and Section 8 concludes.

2. Related Work

Antenna Q and bandwidth. Wheeler [8] and Chu [7] established that an electrically small antenna enclosed in a sphere of radius a has a minimum radiation Q that grows rapidly as ka decreases. McLean [9] gave the widely used form $Q \geq 1/(ka)^3 + 1/(ka)$ for a linearly polarised antenna. Yaghjian and Best [10] related Q to the matched VSWR bandwidth of a single-resonance antenna, $FBW_V \approx (s - 1)/(Q\sqrt{s})$ for a VSWR limit s. They also showed how Q can be obtained from the input impedance. Mavridis et al. [21] compared methods for evaluating the Q of small microstrip patches, the antenna class most used in LTE and 5G prototypes. The Bode–Fano limit [13], [14] bounds how much a lossless matching network can widen the bandwidth of a given load. All of these results end at the antenna port.

Antenna limits and communication rate. Shyianov et al. [15] combined the Chu and Bode–Fano limits with Shannon's formula. They derived the rate of a size-constrained single-antenna link and showed that an optimal signalling bandwidth exists. Deshpande et al. [16] extended this to MISO systems with Bode–Fano matching constraints. Tadele et al. [17] built circuit-aware LMMSE channel estimators for multicarrier systems with coupled broadband arrays. These works are information-theoretic or estimator-oriented. They do not report standard-compliant LTE/NR error rates or a rule stated in terms of the −10 dB bandwidth.

Mismatch in transmitters and devices. Gibiino et al. [18] measured the EVM of power amplifiers under wideband load mismatch. Ortiz et al. [19] quantified the radiated-power loss of mismatched IoT antennas. An industry white paper [20] reported that the amplitude and group-delay ripple of RF filters raises LTE EVM by only a few tenths of a percent, because reference-signal-based equalisation removes most linear distortion. That argument has not been tested for antenna mismatch across 3GPP channels.

Error rates and OFDM channel estimation. Error rates in Rayleigh fading follow by averaging the additive white Gaussian noise (AWGN) error rate over the fading distribution [22]. Exact bit-error-rate (BER) expressions for Gray-coded square M-QAM in AWGN are given by Cho and Yoon [23]. Comb-type pilots with LS estimation and interpolation are the textbook OFDM estimator [24], [25]. Link error performance under Rayleigh, Rician, Nakagami-m and two-wave fading has been compared for RIS-assisted single-antenna links [26], and space–time block coding over correlated and imperfectly estimated channels has been reviewed in [27]. Adaptive LMS equalisers, which have been implemented on FPGA [28], are an alternative to the pilot-based one-tap equaliser used in this work.

Table 1 positions this work against the closest studies.

Table 1. Positioning of this work with respect to the closest related studies.

Study	Antenna model	Link metric	OFDM numerology / 3GPP channels	Pilot-based estimation	Design rule in −10 dB bandwidth
[7]–[10], [21]	Q, size, VSWR bandwidth	None	No	No	No
[15]	Chu antenna + Bode–Fano matching	Shannon rate (SISO)	No	No	No
[16]	Chu antennas + Bode–Fano matching	Shannon rate (MISO)	No	No	No
[17]	Coupled broadband array	Estimation error	Generic multicarrier	LMMSE	No
[18], [19]	Load / antenna mismatch	PA EVM / radiated power	No	No	No
This work	Single- and double-tuned resonance, measured S11	Uncoded BER, required SNR, capacity	LTE 20 MHz, NR 100 MHz; EPA/EVA/ETU	LS on comb pilots	Yes (closed form)

3. System and Antenna Model

3.1 OFDM link

Fig. 1 shows the link. An OFDM transmitter maps Gray-coded M-QAM symbols X_k onto N subcarriers at frequencies $f_k = f_c + (k - (N - 1)/2)\Delta f$. The signal passes through the transmit antenna, a multipath channel and the receive antenna. The cyclic prefix (CP) is longer than the combined impulse response. This holds easily: the group delay of a resonant antenna at resonance is $Q/(2\pi f_0)$, for example 1.3 ns for Q = 30 at 3.55 GHz, whereas the normal CP lasts about 4.7 µs in LTE and about 2.3 µs in NR at 30 kHz spacing [3], [5]. Each subcarrier therefore sees a single complex gain:

$$Y_k = H_R(f_k)G_k H_T(f_k)X_k + Z_k, Z_k \sim CN(0, N_0), (1)$$

where G_k is the multipath channel with $E|G_k|^2 = 1$, and H_T and H_R are the antenna transfer functions defined below. The mean SNR of a flat, perfectly matched link is $\bar{\gamma} = E_s/N_0$.

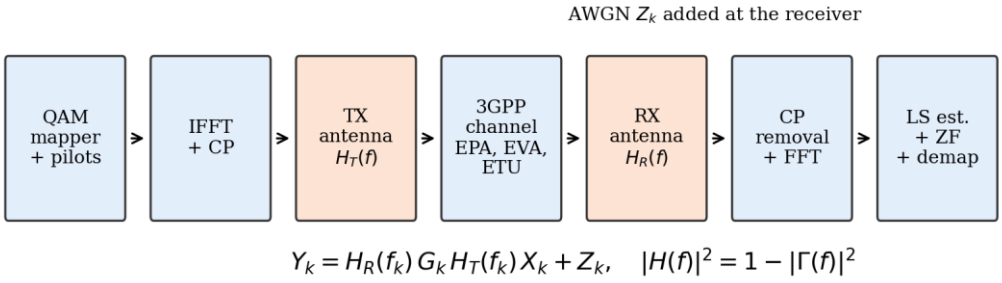


Figure 1. System model. The transmit and receive antennas act as frequency-selective power-transfer filters $|H(f)|^2 = 1 - |\Gamma(f)|^2$ in series with the 3GPP multipath channel.

3.2 Single-resonance antenna

Near its fundamental resonance, a single-resonance antenna (dipole, monopole, patch or slot) is well described by a series RLC circuit [1], [10]:

$$Z_A(f) = R[1 + jQv(f)], v(f) = \frac{f}{f_0} - \frac{f_0}{f}, (2)$$

where f_0 is the resonant frequency, Q is the antenna quality factor, and R is the input resistance at resonance. With the antenna matched at resonance ($R = Z_0$), the reflection coefficient is

$$\Gamma(f) = \frac{Z_A - R}{Z_A + R} = \frac{jQv}{2 + jQv}. (3)$$

The transfer function from the incident wave to the radiating current is $H = (1 + \Gamma)R/Z_A$. By reciprocity, the same response applies on receive. For a lossless antenna,

$$H(f) = \frac{2}{2 + jQv}, |H(f)|^2 = 1 - |\Gamma(f)|^2 = \frac{4}{4 + Q^2v^2}. (4)$$

The −10 dB edges satisfy $|\Gamma|^2 = 0.1$, that is, $Qv = \pm 2/3$. Because v(f) is antisymmetric in log-frequency, the fractional bandwidth follows exactly as

$$FBW_{-10} = \frac{BW_{-10}}{f_0} = \frac{2}{3Q}. (5)$$

This agrees with the VSWR relation of [10] for $s = 1.925$ ($|\Gamma| = 1/\sqrt{10}$).

3.3 Normalised subcarrier profile

Let $B = N\Delta f$ be the occupied bandwidth, centred at f_c. Define

$$r = \frac{B}{BW_{-10}}, d = \frac{f_c - f_0}{BW_{-10}}. (6)$$

Here r is the fraction of the −10 dB band that the channel fills, and d is its offset from resonance. For $FBW_{-10} \ll 1$, $v \approx 2(f - f_0)/f_0$. With $u_k = (f_k - f_c)/B \in [-½, ½]$, (4) and (5) give

$$|H(f_k)|^2 = \frac{1}{1 + x_k^2}, x_k = \frac{2}{3}(d + ru_k). (7)$$

The per-subcarrier SNR is $SNR_k = \bar{\gamma}\eta_k$, with

$$\eta_k = |H(f_k)|^{2n}, n = \{1, onemismatchedantenna(e.g., theterminal), 2, identicalantennasatbothends. (8)$$

Equation (7) contains no carrier frequency, subcarrier spacing or Q. Any LTE or NR channel with the same (r, d) therefore sees the same profile. This universality is confirmed numerically in Section 6.1.

3.4 Double-tuned antenna

A common way to widen the impedance bandwidth is to add a lossless shunt resonator, tuned to f_0, at the antenna port. The input admittance becomes

$$Y\ (f) = \frac{1}{Z_A(f)} + \frac{jQ_p v(f)}{R}, (9)$$

where Q_p is the slope parameter of the resonator. For each antenna Q (denoted Q_a in this case), Q_p was chosen numerically to maximise the −10 dB bandwidth. The optimum widens BW_{-10} by a factor of 1.44 relative to the single-

tuned antenna with the same Q_a (Section 6.1). The response is flat in the band centre but falls more steeply at the band edges, which matters below. For the double-tuned antenna H and $|H|^2$ were computed from (9) without approximation.

4. Closed-Form Link Penalties

4.1 SNR loss

For d = 0 the worst subcarriers lie at the band edges, where x = r/3. Their loss is

$$L_{edge} = 10nlog_{10}\left(1+\frac{r^2}{9}\right)dB. (10)$$

At r = 1, the edge loss is 0.46 dB per antenna, the familiar $10\log_{10}(10/9)$. Averaging (7) over u gives the mean loss. With a = r/3,

$$L_{mean,1} = 10log_{10}\frac{a}{arctana}, L_{mean,2} = -10log_{10}\left[\frac{1}{2a}\left(\frac{a}{1+a^2}+arctana\right)\right]. (11)$$

The mean loss describes the received power but not the error rate. Error rate and capacity depend on how the SNR is distributed across subcarriers. The capacity-equivalent loss at mean SNR $\bar{\gamma}$ is

$$\Delta_C = 10log_{10}\gamma^- - 10log_{10}(2^{C^-}-1), C^- = \frac{1}{N}\sum_k log_2(1+\gamma^-\eta_k). (12)$$

4.2 BER penalty under Rayleigh fading

In EPA, EVA and ETU, every subcarrier gain G_k is complex Gaussian with unit power. The individual subcarriers are correlated, but each is marginally Rayleigh. With perfect channel knowledge and zero-forcing (ZF) equalisation, the average BER is therefore exactly

$$P^-_b(\gamma^-) = \frac{1}{N}\sum_k P_b^R(\gamma^-\eta_k), P_b^R(\gamma^-) = \int_0^\infty P_b^{AWGN}(\gamma^- x)e^{-x}dx, (13)$$

where P_b^{AWGN} is the exact Gray-coded M-QAM BER [23]. The expression holds whatever the frequency correlation of the channel.

At high SNR, every square M-QAM constellation has diversity order one in Rayleigh fading [22]:

$$P_b^R(\gamma^-) \simeq \frac{c_M}{\gamma^-}, (14)$$

where c_M depends only on M. Substituting (14) into (13) makes the average BER proportional to the mean of $1/\eta_k$. The flat link needs $\bar{\gamma}_0 = c_M/P_b$, so the required-SNR penalty is

$$\Delta = 10log_{10}\left(\frac{1}{N}\sum_k \eta_k^{-1} = 10log_{10}E[(1+x^2)^n]. (15)\right)$$

The constant c_M cancels. The penalty is therefore the same for QPSK, 16QAM and 64QAM: it is set by the harmonic mean of the subcarrier SNRs, not by the arithmetic mean. With u uniform on [−½, ½], the moments of x in (7) are $\mu_2 = E[x^2] = (4/9)(d^2 + r^2/12)$ and $\mu_4 = E[x^4] = (16/81)(d^4 + d^2r^2/2 + r^4/80)$. Hence

$$\Delta_1 = 10log_{10}(1+\mu_2), \Delta_2 = 10log_{10}(1+2\mu_2+\mu_4). (16)$$

For a channel centred on resonance (d = 0), (16) reduces to

$$\Delta_1 = 10log_{10}\left(1+\frac{r^2}{27}\right), \Delta_2 = 10log_{10}\left(1+\frac{2r^2}{27}+\frac{r^4}{405}\right). (17)$$

Two practical bounds follow from (15) and (10). A channel lying entirely inside the −10 dB band has $|x| \le 1/3$ on every subcarrier, so its penalty can never exceed 0.46n dB, i.e. 0.92 dB for two antennas. A channel of width r placed off resonance costs more than a centred one of the same width. For example, (16) gives the same 0.32 dB (n = 2) for a centred channel with r = 1 as for a half-width channel (r = 0.5) placed at d = 0.25, i.e. touching one −10 dB edge.

4.3 Design rule

Inverting (17) for a penalty budget Δ (dB) gives the largest admissible ratio:

$$r_{max,1} = \sqrt{}$$

The antenna must therefore provide $BW_{-10} \ge B/r_max$. By (5), its quality factor is limited to

$$Q_{max} = \frac{2r_{max}f_0}{3B}. (19)$$

Finally, the Chu–McLean bound [7], [9]

$$Q \geq \frac{1}{(ka)^3} + \frac{1}{ka}, k = \frac{2\pi f_0}{c}, (20)$$

turns Q_{max} into the smallest radius of a sphere that can enclose a lossless single-resonance antenna meeting the budget.

5. Simulation Method

5.1 Link-level simulator

A frequency-domain OFDM simulator was written in Python 3 (NumPy, SciPy); its parameters are listed in Table 2. Each slot contains one pilot OFDM symbol followed by 12 data symbols. The channel is block-static within a slot and independent between slots; Doppler is not modelled. The multipath gain G_k is generated from the 3GPP EPA, EVA and ETU tapped-delay-line profiles [4]. Two numerologies are simulated:

- **LTE 20 MHz** [3], [4]: 1200 subcarriers at 15 kHz (B = 18 MHz), f_0 = 1.8425 GHz. Pilots are placed on every third subcarrier. This is the union of the two cell-specific reference-signal (CRS) combs of one antenna port [29].
- **NR 100 MHz** [5], [6]: 3276 subcarriers at 30 kHz (B = 98.28 MHz), f_0 = 3.55 GHz in band n78. Pilots use a comb-2 pattern, as in demodulation reference signal (DMRS) configuration type 1 [30].

The receiver uses either perfect channel state information (CSI) or LS estimates on the unit-modulus pilots. Linear interpolation of the real and imaginary parts fills the data subcarriers [24], [25], and ZF equalisation and hard Gray demapping follow. The antenna response of (4), (8) or (9) multiplies the channel on every subcarrier, and the estimator sees the same product through the pilots, so no knowledge of the antenna is assumed.

Table 2. Simulation parameters.

Parameter	Value
Numerologies	LTE 20 MHz: N = 1200, Δf = 15 kHz, f_0 = 1.8425 GHz; NR 100 MHz: N = 3276, Δf = 30 kHz, f_0 = 3.55 GHz
Channels	3GPP EPA, EVA, ETU [4]; AWGN for verification
Modulation	QPSK, 16QAM, 64QAM (Gray)
Antenna	Single-tuned (4); double-tuned (9) with optimal Q_p; TX + RX (n = 2) or RX only (n = 1)
r = B/BW_{-10}	0, 0.5, 1, 1.5, 2, 3 (Monte Carlo); 0–4 in steps of 0.1 (analytic)
Channel estimation	Perfect CSI; LS on comb pilots (spacing 3 for LTE, 2 for NR) + linear interpolation
Equalisation / detection	One-tap ZF; hard-decision Gray demapping
Slots per point	400 (LTE), 150 (NR); 12 data symbols per slot
Target BER	10^{-3} (uncoded)
Random numbers	Seed 20261002; common random numbers per slot index

5.2 Estimating the penalty

The penalty is a small difference between two required SNRs, each read from a BER curve with Monte Carlo noise. Two steps keep this difference precise.

Common random numbers. Slot s of every configuration uses the same random stream. The channel realisation, the data bits and the unit-variance noise samples are therefore identical across r, SNR and CSI mode. Only the antenna response and the noise scale change. The penalty is a paired difference, so the fading-induced variance largely cancels.

Bootstrap confidence intervals. For each configuration, BER was evaluated on a 1 dB grid from 2 dB below to 14 dB above the flat-antenna requirement. The required SNR at BER 10^{-3} was found by log-linear interpolation. The 95% confidence interval (CI) of the penalty was then obtained with a percentile bootstrap [31] with 400 replicates. Slots were resampled jointly for the flat and mismatched configurations so that the pairing is preserved.

5.3 Experiments

Five experiment groups were run:

- **E0 (verification):**
- the exact −10 dB bandwidth (5) against a numerical search of (3);
- power conservation $|H|^2 = 1 - |\Gamma|^2$ for both antenna types;
- the Rayleigh integral in (13) against the closed-form QPSK result;
- Monte Carlo BER with a flat antenna against theory in AWGN and EVA;
- the universality of (7) between LTE and NR.
- **E1 (analytic penalties):** (10)–(13) evaluated exactly for r = 0–4, for single- and double-tuned antennas at one or both ends. The BER penalty is solved from (13) without the high-SNR approximation.
- **E2/E3 (link level):** Monte Carlo penalties for all numerology–channel pairs with 16QAM, and for QPSK and 64QAM in LTE EVA. E3 compares perfect CSI with LS estimation.
- **E4 (design):** (18)–(20) applied to LTE 5 and 20 MHz and NR 40 and 100 MHz channels.
- **E6 (sensitivity):** double tuning, RX-only mismatch, and off-centre channels.

A case study with a measured antenna (E5) is left for future work (Section 7).

All results are produced by one script from a fixed seed, and every figure is generated from the saved result files.

6. Results and Discussion

6.1 Verification

Table 3 summarises the verification. All checks passed. In AWGN and EVA, all 18 flat-antenna Monte Carlo BER points contain the exact value in their 95% CI. The mean subcarrier loss for LTE 20 MHz and NR 100 MHz agrees to within 10^{-4} dB at the same r, which confirms that r is the only parameter that matters.

Table 3. Verification results (E0).

Check	Result
FBW_{-10} from (5) vs numerical search, Q = 3–100	Max. relative error 2.5×10^{-5} (grid resolution)
$\|H\|^2 = 1 - \|\Gamma\|^2$, single / double tuned	Max. error 6.7×10^{-16} / 8.9×10^{-16}
Rayleigh integral (13) vs closed-form QPSK, −20 to 80 dB	Max. relative error 4.0e-09
Flat-antenna BER, Monte Carlo vs exact (AWGN 8–16 dB, EVA 18–30 dB; QPSK–64QAM)	18/18 points inside the 95% CI
Mean loss at r = 0.5 / 1 / 2, LTE vs NR	0.0792 / 0.3029 / 1.0394 dB for both
Double-tuned BW_{-10} gain at equal Q_a	1.44

6.2 Analytic penalty curves

Fig. 2 shows the antenna responses. Fig. 3 and Table 4 show how the penalties grow with r.

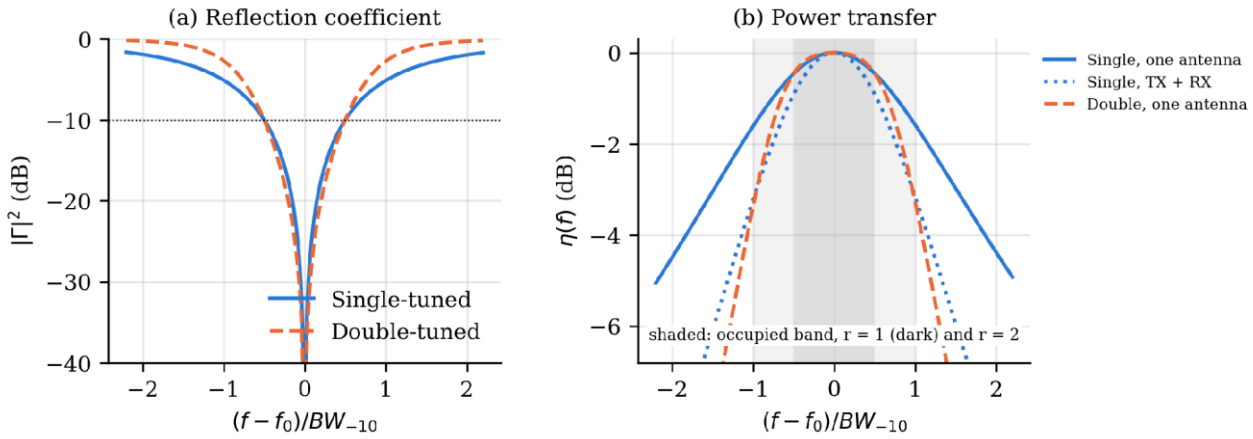


Figure 2. (a) Reflection coefficient and (b) power transfer η(f) of single- and double-tuned antennas with the same −10 dB bandwidth. Shaded regions show the occupied band for r = 1 and r = 2.

Three observations follow.

The penalty is small inside the −10 dB band. For a channel that exactly fills the −10 dB band (r = 1), the Rayleigh BER penalty is 0.16 dB with one mismatched antenna and 0.32 dB with two. The band-edge subcarriers lose 0.46 dB and 0.92 dB respectively. The link penalty is much smaller than the edge loss, because only a few subcarriers are near the edges.

Beyond the −10 dB band, the penalty grows quickly. At r = 2 the two-antenna penalty reaches 1.26 dB, and at r = 3 it reaches 2.71 dB. The closed form (17) matches the exact solution of (13) within 0.008 dB for all r ≤ 4 and all three modulations (Fig. 3b). This confirms that the BER penalty does not depend on the modulation order.

AWGN is harsher than Rayleigh fading. In AWGN, the BER is dominated by the weakest subcarriers, whereas in Rayleigh fading it is dominated by deep fades that occur on every subcarrier alike. At r = 2, the 64QAM penalty is 1.58 dB in AWGN against 1.26 dB in Rayleigh fading (two antennas). A fixed-wireless or line-of-sight link therefore needs a slightly wider antenna for the same budget: r ≤ 1.19 instead of 1.25 for 0.5 dB.

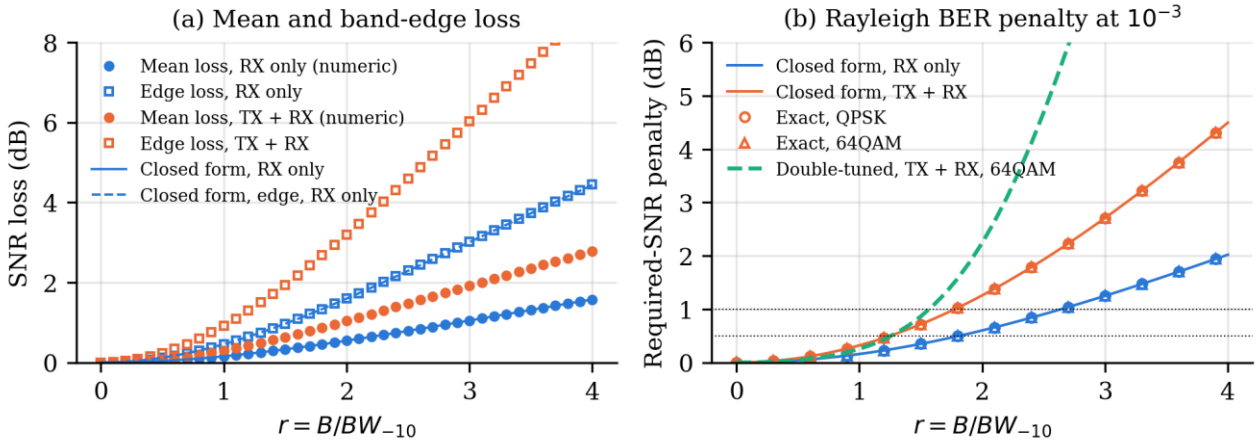


Figure 3. (a) Mean and band-edge SNR loss versus r. (b) Required-SNR penalty at BER 10^{-3} in Rayleigh fading: closed forms (17) (lines) and exact solutions of (13) (markers) for QPSK and 64QAM, plus the double-tuned antenna.

Table 4. Analytic penalties (dB) of the single-tuned antenna versus r = B/BW$_{-10}$, channel centred on resonance.

r	Edge loss, 1 ant.	Mean loss, 2 ant.	Edge loss, 2 ant.	Capacity loss (12) at 20 dB, 2 ant.	BER penalty, Rayleigh, 1 ant.	BER penalty, Rayleigh, 2 ant.	BER penalty, AWGN 64QAM, 2 ant.
0.5	0.12	0.08	0.24	0.08	0.04	0.08	0.08
1.0	0.46	0.30	0.92	0.31	0.16	0.32	0.35
1.5	0.97	0.64	1.94	0.68	0.35	0.72	0.83
2.0	1.60	1.04	3.20	1.14	0.60	1.26	1.58
3.0	3.02	1.92	6.04	2.29	1.25	2.71	3.73
4.0	4.45	2.78	8.89	3.56	2.02	4.49	6.23

6.3 Link-level validation

Fig. 4 compares Monte Carlo BER curves with the analytic prediction (13) for LTE 20 MHz in EVA. With perfect CSI, the simulated points follow the prediction at every r. Across all 969 perfect-CSI points with an expected BER above 10^{-5}, the median ratio of simulated to predicted BER is 1.004 (interquartile range 0.962–1.064). With LS estimation the curves shift right by about 2.0–2.7 dB, but they shift by the same amount at every r.

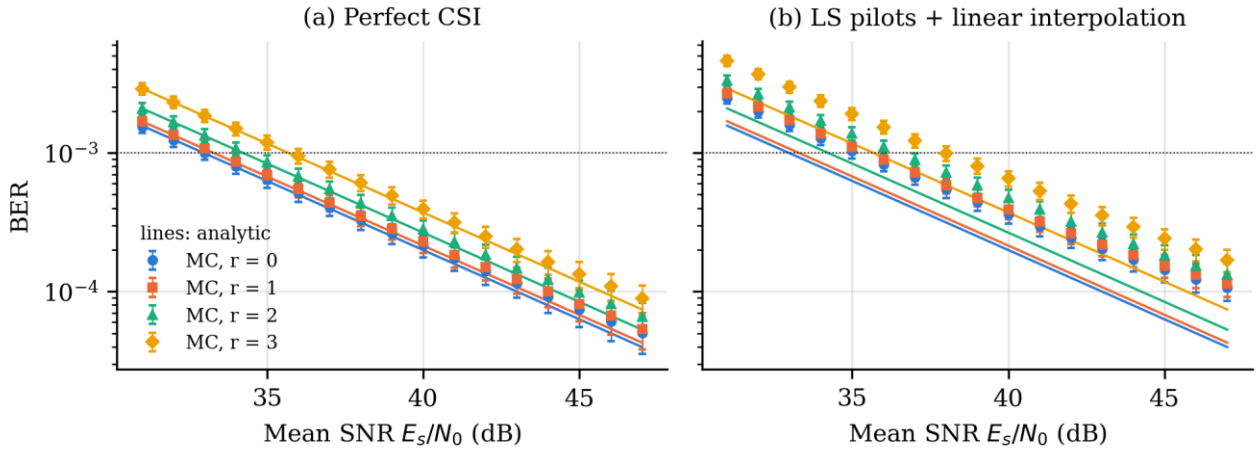


Figure 4. BER of 16QAM, LTE 20 MHz, EVA channel, identical antennas at both ends. Markers: Monte Carlo with 95% CIs; lines: analytic prediction (13) for perfect CSI.

Fig. 5 and Table 5 report the required-SNR penalties. For the single-tuned antenna, the analytic penalty lies inside the 95% bootstrap CI of the Monte Carlo penalty in 84 of the 84 cases with a defined penalty. These cases cover both numerologies, all three channels, QPSK to 64QAM, both CSI modes, and single- and double-tuned antennas at one or both ends. The median CI half-width is 0.06 dB. The penalty depends neither on the numerology nor on the delay spread (EPA, EVA and ETU agree), as (13) predicts.

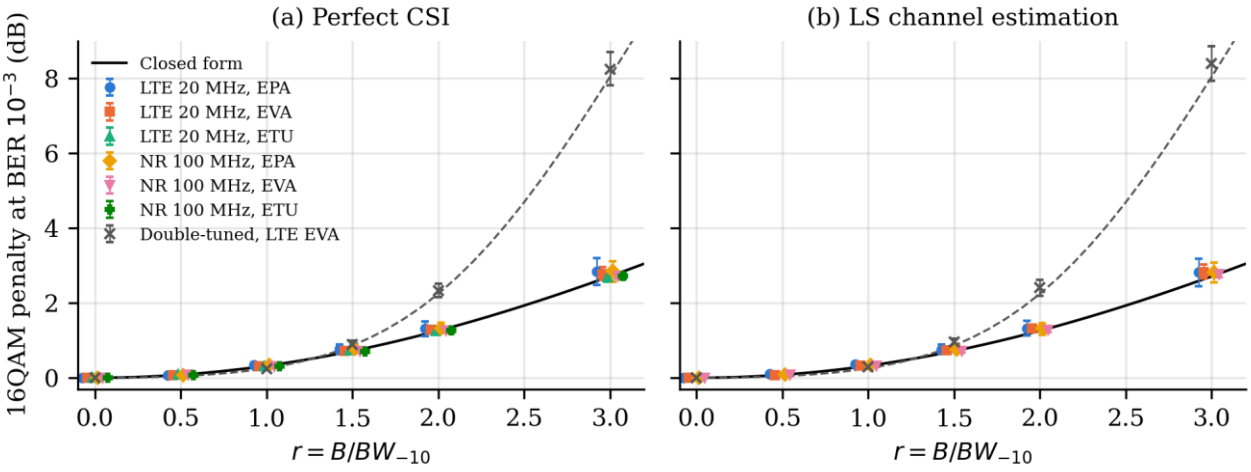


Figure 5. Required-SNR penalty at BER 10^{-3} for 16QAM versus r. Markers: Monte Carlo with 95% bootstrap CIs for LTE 20 MHz and NR 100 MHz in EPA, EVA and ETU; black line: closed form (17); grey: double-tuned antenna (markers: Monte Carlo, dashed line: exact analysis).

Table 5. Monte Carlo required-SNR penalty (dB) at BER 10^{-3} with 95% bootstrap CI in brackets, compared with the exact analysis (13) in the last three rows.

Numerology, channel, modulation	Antenna	CSI	r = 1	r = 2	r = 3
LTE 20 MHz, EPA, 16QAM	single-tuned, TX + RX	perfect	0.35 [0.28, 0.40]	1.32 [1.12, 1.51]	2.84 [2.48, 3.20]
LTE 20 MHz, EPA, 16QAM	single-tuned, TX + RX	LS	0.36 [0.29, 0.42]	1.32 [1.12, 1.52]	2.82 [2.45, 3.17]
LTE 20 MHz, EVA, 16QAM	single-tuned, TX + RX	perfect	0.31 [0.28, 0.36]	1.28 [1.19, 1.38]	2.77 [2.62, 2.95]
LTE 20 MHz, EVA, 16QAM	single-tuned, TX + RX	LS	0.32 [0.27, 0.36]	1.32 [1.21, 1.43]	2.83 [2.65, 3.03]
LTE 20 MHz, ETU, 16QAM	single-tuned, TX + RX	perfect	0.31 [0.28, 0.34]	1.24 [1.18, 1.32]	2.68 [2.57, 2.79]
LTE 20 MHz, ETU, 16QAM	single-tuned, TX + RX	LS	n/a (floor)	n/a (floor)	n/a (floor)
NR 100 MHz, EPA, 16QAM	single-tuned, TX + RX	perfect	0.33 [0.29, 0.38]	1.32 [1.20, 1.47]	2.85 [2.56, 3.11]
NR 100 MHz, EPA, 16QAM	single-tuned, TX + RX	LS	0.34 [0.29, 0.39]	1.31 [1.15, 1.45]	2.81 [2.55, 3.08]
NR 100 MHz, EVA, 16QAM	single-tuned, TX + RX	perfect	0.31 [0.28, 0.34]	1.27 [1.21, 1.32]	2.73 [2.65, 2.81]
NR 100 MHz, EVA, 16QAM	single-tuned, TX + RX	LS	0.33 [0.29, 0.36]	1.28 [1.21, 1.34]	2.78 [2.68, 2.87]

NR 100 MHz, ETU, 16QAM	single-tuned, TX + RX	perfect	0.31 [0.28, 0.34]	1.26 [1.20, 1.31]	2.71 [2.64, 2.79]
NR 100 MHz, ETU, 16QAM	single-tuned, TX + RX	LS	n/a (floor)	n/a (floor)	n/a (floor)
LTE 20 MHz, EVA, QPSK	single-tuned, TX + RX	perfect	0.35 [0.30, 0.41]	1.35 [1.23, 1.46]	2.88 [2.70, 3.06]
LTE 20 MHz, EVA, QPSK	single-tuned, TX + RX	LS	0.32 [0.25, 0.39]	1.33 [1.20, 1.45]	2.85 [2.63, 3.04]
LTE 20 MHz, EVA, 64QAM	single-tuned, TX + RX	perfect	0.34 [0.31, 0.37]	1.31 [1.24, 1.40]	2.77 [2.64, 2.89]
LTE 20 MHz, EVA, 64QAM	single-tuned, TX + RX	LS	0.35 [0.31, 0.39]	1.31 [1.22, 1.39]	2.79 [2.64, 2.95]
LTE 20 MHz, EVA, 16QAM	double-tuned, TX + RX	perfect	0.25 [0.21, 0.30]	2.33 [2.15, 2.52]	8.24 [7.82, 8.70]
LTE 20 MHz, EVA, 16QAM	double-tuned, TX + RX	LS	0.30 [0.24, 0.35]	2.41 [2.19, 2.61]	8.40 [7.93, 8.86]
LTE 20 MHz, EVA, 16QAM	single-tuned, RX only	perfect	0.16 [0.13, 0.18]	0.61 [0.56, 0.67]	1.27 [1.19, 1.36]
LTE 20 MHz, EVA, 16QAM	single-tuned, RX only	LS	0.15 [0.13, 0.18]	0.60 [0.54, 0.66]	1.29 [1.19, 1.39]
Analysis (13), any channel and M	single-tuned, TX + RX	perfect	0.32	1.26	2.71
Analysis (13)	double-tuned, TX + RX	perfect	0.25	2.24	8.04
Analysis (13)	single-tuned, RX only	perfect	0.16	0.60	1.25

6.4 Effect of channel estimation

A natural concern is that antenna mismatch might also degrade channel estimation, because the edge subcarriers carry weaker pilots. The results do not support this. Across all single-tuned configurations with a defined penalty, the LS and perfect-CSI penalties differ by at most 0.06 dB, with a mean difference of +0.004 dB. Every difference lies within the Monte Carlo uncertainty. The largest difference, 0.16 dB, occurs for the double-tuned antenna at r = 3, where its penalty exceeds 8 dB.

The reason is that the antenna response is extremely smooth on the pilot grid. It varies over a scale of BW_{-10}, which spans hundreds of subcarriers, so linear interpolation reproduces it almost perfectly. The weaker pilots at the band edges raise the estimation noise there in exact proportion to the weaker data. The equaliser therefore absorbs the antenna, as has been argued for RF filters [20], and the antenna acts purely as an SNR profile.

LS estimation does have one limitation that is independent of the antenna. In ETU, whose delay spread reaches 5 µs, linear interpolation across a 45 kHz (LTE) or 60 kHz (NR) pilot gap leaves an interpolation-error floor. The flat-antenna BER stops falling at 1.8e-03 (LTE) and 3.9e-03 (NR). Because both floors lie above 10^{-3}, the LS penalty in ETU is undefined at this target. The floor itself increases only marginally with r (for LTE, from 1.8e-03 at r = 0 to 1.9e-03 at r = 3). A practical receiver would use a smoother estimator, such as MMSE or DFT-based estimation [25].

6.5 Double tuning, one-sided mismatch and channel placement

Double tuning. Double tuning should be judged at equal antenna Q, since the radiator and its size are what is fixed. Table 6 makes that comparison. For the same radiator, the added resonator lowers the two-antenna penalty from 0.32 to 0.10 dB at r = 1 and from 1.26 to 0.67 dB at r = 2. Equivalently, it raises the admissible Q_a by a factor of 1.45 at a 0.5 dB budget. At equal −10 dB bandwidth, however, the double-tuned antenna is worse once r exceeds 1.27. At r = 2 its penalty is 2.24 dB against 1.26 dB, because its response falls more steeply outside the band (Fig. 2). The Monte Carlo results agree with the analysis in both regimes (Table 5).

The −10 dB bandwidth alone is therefore not a sufficient specification. Two antennas with the same S11 bandwidth can differ by 1 dB in link penalty, depending on the shape of their band edges.

Table 6. Two-antenna Rayleigh penalty (dB) of single- and double-tuned antennas built on the same radiator (same Q_a). Double tuning widens BW_{-10} by a factor of 1.44, so $r_{double} = r_{single}/1.44$.

r (single-tuned)	Penalty, single-tuned	r (double-tuned, same Q_a)	Penalty, double-tuned
1.0	0.32	0.69	0.10
1.5	0.72	1.04	0.29
2.0	1.26	1.39	0.67
2.5	1.93	1.73	1.39
3.0	2.71	2.08	2.56

One-sided mismatch. When only one end has a compact antenna, n = 1 and the budget is much looser. A 0.5 dB budget allows r ≤ 1.82 instead of 1.25, and 1 dB allows r ≤ 2.64 instead of 1.78. The Monte Carlo RX-only penalties (0.16, 0.61 and 1.27 dB at r = 1, 2 and 3, perfect CSI) agree with (17).

Channel placement. In frequency-division duplex bands, one antenna usually covers the uplink and downlink sub-bands, so a given channel rarely sits at resonance. Equation (16) quantifies the cost. A channel with r = 0.5 costs 0.08 dB at resonance, 0.32 dB at d = 0.25 (touching the −10 dB edge), and 1.00 dB at d = 0.5 (centred on the edge). The exact solution of (13) agrees with (16) to within 0.002 dB in all tested cases (results file E6_offset_channel.csv). For multi-channel operation, designers should therefore evaluate (16) at the worst channel position rather than at the band centre.

6.6 Design table and use with measured antennas

Table 7 and Fig. 6 apply (18)–(20) to common channels, for identical antennas at both ends and a 64QAM Rayleigh budget. For LTE 20 MHz at 1.8425 GHz, a 0.5 dB budget requires $BW_{-10} \geq 14.4$ MHz (FBW 0.78%), i.e. Q ≤ 85. This is easily met by most printed antennas. NR 100 MHz at 3.55 GHz is far more demanding. It requires $BW_{-10} \geq 78.6$ MHz (FBW 2.21%) and Q ≤ 30. By the Chu–McLean bound this implies ka ≥ 0.33, i.e. an enclosing radius of at least 4.5 mm for a lossless antenna. In practice the radius must be larger, since real antennas exceed the bound [10], [21]. Relaxing the budget to 1 dB raises the Q ceiling to 43 for NR 100 MHz.

By (16), a link whose antennas at both ends are designed for “band n78” but meet −10 dB over only a 60 MHz sub-band (r = 1.64) loses 0.85 dB with a 100 MHz carrier centred on resonance, and 1.10 dB when the carrier is offset by a quarter of the −10 dB band (d = 0.25).

Table 7. Antenna requirements for a two-antenna Rayleigh penalty ≤ 0.5 dB and ≤ 1 dB (r_{max} = 1.25 and 1.78). The minimum radius is the Chu–McLean lower bound for a lossless single-resonance antenna.

Channel	f_0 (GHz)	B (MHz)	Budget (dB)	Min. BW_{-10} (MHz)	Min. FBW_{-10} (%)	Max. Q	Min. ka	Min. radius (mm)
LTE 5 MHz	1.843	4.50	0.5	3.6	0.20	341	0.144	3.73

LTE 20 MHz	1.843	18.00	0.5	14.4	0.78	85	0.231	5.98
LTE 20 MHz	2.655	18.00	0.5	14.4	0.54	123	0.204	3.66
NR 40 MHz	3.55	38.16	0.5	30.5	0.86	78	0.239	3.21
NR 100 MHz	3.55	98.28	0.5	78.6	2.21	30	0.333	4.47
LTE 5 MHz	1.843	4.50	1.0	2.5	0.14	485	0.128	3.31
LTE 20 MHz	1.843	18.00	1.0	10.1	0.55	121	0.205	5.30
LTE 20 MHz	2.655	18.00	1.0	10.1	0.38	175	0.181	3.25
NR 40 MHz	3.55	38.16	1.0	21.5	0.60	110	0.212	2.84
NR 100 MHz	3.55	98.28	1.0	55.3	1.56	43	0.294	3.95

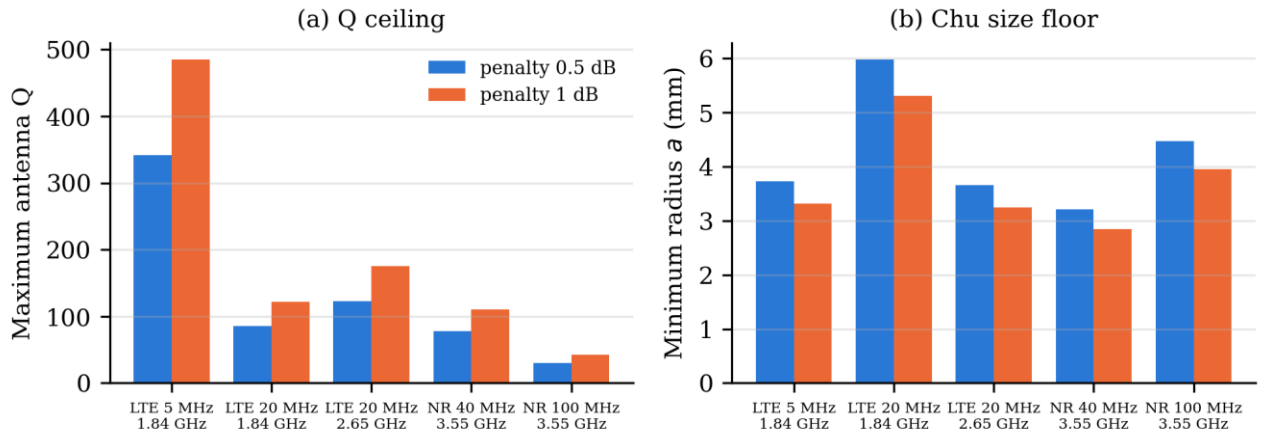


Figure 6. (a) Maximum antenna Q and (b) minimum enclosing-sphere radius for a two-antenna penalty of 0.5 and 1 dB.

Use with measured or simulated antennas. Real antennas are rarely a perfect single resonance. The accompanying script s1p_penalty.py therefore reads a Touchstone S11 file from a vector network analyser or an electromagnetic solver. It computes $\eta_k = [(1 - |S11(f_k)|^2)e_{rad}]^n$ on the actual subcarrier grid and solves (13) for the exact penalty. It also reports the prediction of the single-resonance fit, (16), based on the measured −10 dB band. A self-test on a synthetic single-resonance file reproduces (16) to within 0.001 dB.

6.7 Implications

The results lead to four practical conclusions for antenna and system designers:

1. **The −10 dB criterion is conservative, not adequate by itself.** A channel inside the −10 dB band never loses more than 0.46 dB per antenna. Where the budget allows it, r slightly above 1 is acceptable.
2. **Report r and channel position, not only BW_{-10}.** With two compact antennas the penalty factor in (17) grows as r^4 for large r, and the penalty also grows with the offset from resonance. Claims of "5G suitability" should state the channel bandwidth and position that were assumed.
3. **Band-edge steepness matters.** Broadbanding helps at a fixed antenna size, but it trades gentle edges for steep ones. A double-tuned antenna specified only by its −10 dB bandwidth can perform worse than a simple one.

4. **Compact NR terminals are close to the limit.** At 3.5 GHz with 100 MHz carriers, a radiator smaller than about 4.5 mm in radius cannot meet a 0.5 dB budget even at the Chu limit.

7. Limitations

The study has the following limitations:

- **Antenna model.** The model assumes a single- or double-tuned lossless antenna near resonance. Radiation efficiency, which scales η by e_{rad}, and multi-resonance behaviour are handled only through the S11-based script. Antenna-to-antenna coupling and pattern variation with frequency are ignored.
- **Uncoded metrics.** Only uncoded BER, required SNR and a capacity measure are reported. Channel coding, HARQ and link adaptation convert the SNR penalty into throughput in implementation-specific ways, although for a well-designed link the SNR penalty remains the relevant input. Error-correcting codes and their decoders [32] would change the absolute SNR requirement; how they translate the mismatch penalty into throughput is left for future work.
- **Channel model.** The channel is block-static. Doppler, inter-carrier interference and time interpolation of pilots are not modelled.
- **Single-antenna links.** The results apply to SISO links. MIMO and diversity, which reduce the fading penalty, would change the weighting in (15). With L independent diversity branches sharing the same antenna profile, the penalty becomes $(10/L)\log_{10}E[\eta^{(-L)}]$. Space–time block coding with two transmit antennas [27] is one such case.
- **Receiver.** The LS estimator is the simplest practical one. MMSE estimation would remove the ETU floor but is not expected to change the antenna penalty.
- **Validation.** The case study with the author's own antenna (E5) and over-the-air measurements remain to be carried out.

8. Conclusions

This paper translated the standard antenna specification, the −10 dB impedance bandwidth, into LTE and 5G NR OFDM link penalties. For a single-resonance antenna, the penalty depends only on the ratio r of occupied bandwidth to −10 dB bandwidth and on the channel's offset from resonance. Under Rayleigh fading it takes the closed form $10\log_{10}(1 + r^2/27)$ for one antenna and $10\log_{10}(1 + 2r^2/27 + r^4/405)$ for two, for any M-QAM order.

Monte Carlo simulation of LTE 20 MHz and NR 100 MHz links in EPA, EVA and ETU channels, with pilot-based LS estimation, confirmed the closed forms within the 95% confidence interval in every case with a defined penalty. It also showed that channel estimation does not add a mismatch-specific penalty. A channel that fills the −10 dB band costs 0.32 dB with two antennas, and a 0.5 dB budget allows $r \leq 1.25$. For NR 100 MHz at 3.55 GHz this implies $Q \leq 30$ and an enclosing radius of at least 4.5 mm. Double tuning helps at a fixed antenna size but is worse at equal −10 dB bandwidth for $r > 1.27$, so the band-edge shape must be specified as well as the bandwidth.

Future work will apply the method to measured antennas, add channel coding and MIMO, and extend the analysis to millimetre-wave NR bands. Energy-aware operation of beyond-5G links, surveyed in [33], is a further setting in which the closed-form penalties can be used to budget the antenna.

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