Initial results on mathing with applications to integrated 5G antennas

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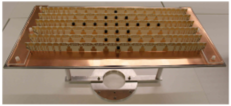
Table of Contents

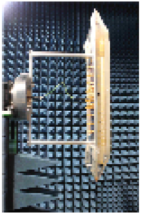
Introduction: Basestation antennas

Some properties of todays antenna base-stations

- Each cover a fixed sector the around the base station.
- The antenna has a fixed radiation pattern.
- The frequency is comparably low $\leq$5 GHz
- Fixed small frequency range

jazdcommunications.com, ,Ericsson





5G-base stations

- Beam steerability, massive MIMO
- Larger bandwidth, in a large frequency range
- Antenna adjustments (one type of base-stations)
- Both below 5GHz base stations and above 20GHz base stations

Antenna challenges:

For the <5 GHz:

- Each array are expected to work over a large bandwidth: 6:1
- Advantages: Same antenna in different regions (different center frequency bands) and for several frequency bands.
- Each frequency band is narrow, but the may occur at different center frequencies (due to provider and due to country).
- Disadvantages: Requires filters to remove radiation for unwanted frequencies. More complex antennas.

For $>$20GHz

- Narrow-band antennas e.g. about 5% BW
- Today, less efficient power amplifiers
- Higher losses, in the feeding systems, requirements on higher integration

Some challenges for base-station antennas:

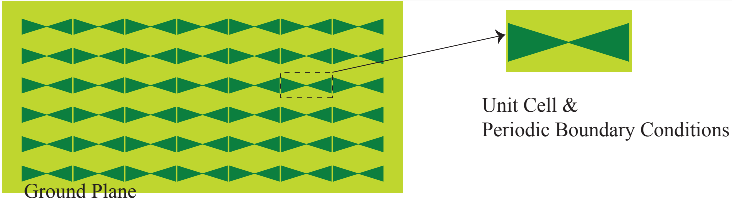
Three topics of today:

- Limits of the bandwidth in an wide-band antenna. [Sum-rule]
- Limits of the bandwidth in a narrow-band antenna system. [Q-factor and Current optimization]
- High integration and matching [Work in progress]

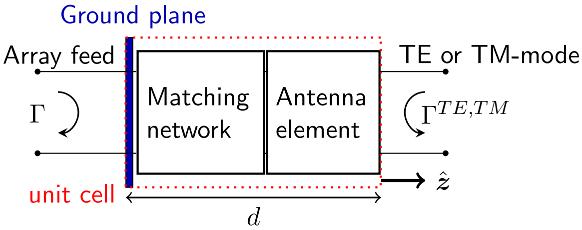
Table of Contents

Simplifications – assumptions

- A unit-cell model – the array is approximated as periodic.
- Each antenna element is build of passive linear and time-invariant materials.
- Impedance bandwidth model: One band or multi-band, with a given worst reflection coefficient as threshold.
- We consider here linear polarization, corresponding to the TE-mode (E-orthogonal to the surface normal)



We study the excitation and reception of the lowest TE-Floqquet mode.
A simplified antenna unit-cell system:



A sum-rule result

The refection coefficient Γ^{TE} is bounded and passive, with help of a Blaschke-product B we find that $-\mathrm{j}\ln(\Gamma^{TE}B)$ is a Herglotz-function, and sum-rules apply.

Bode-Fano type result for Γ^{TE}. (Rozanov 2000)

Passivity thus yields

$$I(\theta) := \int_0^\infty \omega^{-2} \ln(|\Gamma^{TE}(\omega,\theta)|^{-1}) \, \mathrm{d}\omega \leq q(\theta) \tag{1}$$

Sjöberg and Gustafsson, 2011 showed that

$$q(\theta) = \frac{\pi d}{c}(1 + \frac{\tilde{\gamma}}{2dA}) \cos\theta \leq \frac{\pi d \mu_s}{c} \cos\theta \tag{2}$$

d-thickness, A-unit cell area, $\tilde{\gamma}$-function of polarizability tensor, μ_s, maximum relative static permeability.

Array figure of merit

Limitations

- Loss-less system $|\Gamma| = |\Gamma^{TE}|$, see e.g. Doane et al 2013.
- Below grating lobe limit ω_G.
- The integrand is positive:

$$\eta_0 := \max_{\theta\in[\theta_0,\theta_1]} \frac{\int_0^{\omega_G} \omega^{-2} \ln(|\Gamma(\omega,\theta)|^{-1}|)\,\mathrm{d}\omega}{q(\theta)} \leq 1 \quad (3)$$

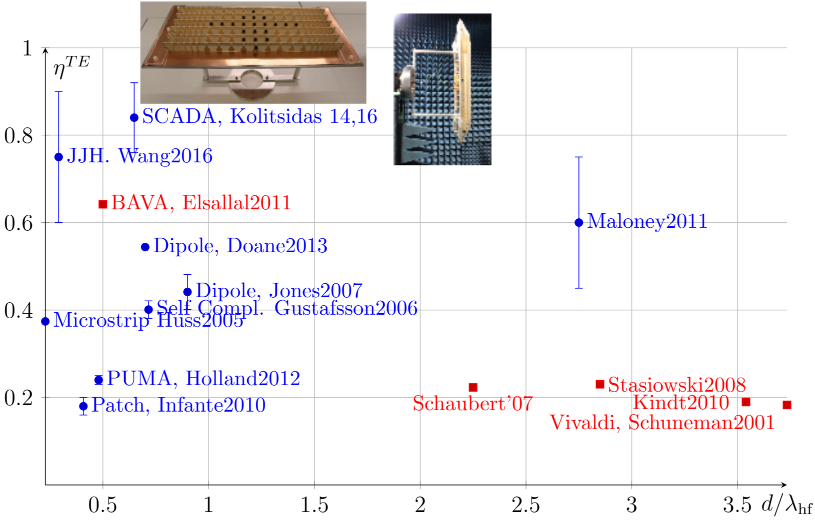
- Given M frequency bands $B_m := [\lambda_{-,m}, \lambda_{+,m}]$,
- Define $|\Gamma_m| := \max_{\lambda\in B_m,\theta\in[\theta_0,\theta_1]} |\Gamma(\lambda,\theta)|$.
- Clearly $\ln(|\Gamma(\lambda,\theta)|^{-1}) \geq \ln(|\Gamma_m|^{-1})$

Hence:

$$0 \leq \eta_M^{TE} := \frac{\sum_{m=1}^{M} \ln(|\Gamma_m|^{-1})(\lambda_{m,+} - \lambda_{m,-})}{2\pi^2 \mu_s d \cos\theta_1} \leq \eta_0 \leq 1 \quad (4)$$

Here η_M^{TE} is the *Array Figure of Merit* for a M-band antenna.

The figure of merit for some antennas



This resulted in two international patent applications for wide-band antennas and [Jonsson et al, Array antenna limitations, IEEE WPL, 2013].

TM-case

Follows same line of derivation as TE-case

$$\eta_M^{TM} := \frac{\sum_{m=1}^{M} \ln(|\Gamma_m|^{-1})(\lambda_{m,+} - \lambda_{m,+})}{2\pi^2 d\left[\frac{1}{n^2}\cos(\theta_*) + (1 - \frac{1}{n^2})\frac{1}{\cos\theta_*}\right]} \quad (5)$$

Here $n^2 = \varepsilon_s \mu_s$, where ε_s, μ_s maximal static relative values and θ_* defined as

$$\theta_* = \begin{cases} \theta_1, & \theta_1 < \theta_n,\ n \in [1, \sqrt{2}], \\ \theta_n, & \theta_n \in [\theta_0, \theta_1],\ n \in [1, \sqrt{2}], \\ \theta_0, & \theta_0 > \theta_n,\ \text{or } n > \sqrt{2}, \end{cases} \quad (6)$$

where $\theta_n = \arccos(\sqrt{n^2 - 1})$.

Table of Contents

Beyond sum-rules – stored energies

Herglotz-functions and sum-rules:

- A system perspective.
- Based on passivity, linearity and time-translation invariance
- The sum-rule based results describe performance for the entire bandwidth, with a few exceptions. [Shim etal 2019]
- Challenging to include additional constraints.

Q-factor based estimates

- Based on stored energies, in electromagnetic systems
- Tend to predict the bandwidth well for resonant systems
- The estimate utilize information from a single frequency
- Easy to include additional constraints, e.g. gain.

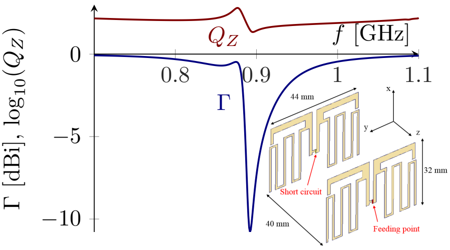
Resonant antennas have a fractional bandwidth

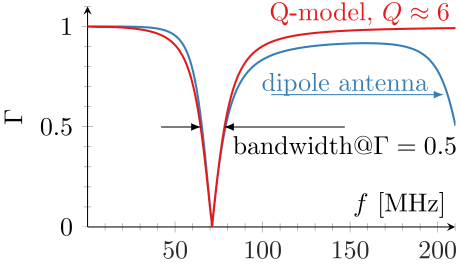
$$\frac{\omega_2-\omega_1}{\omega_0} = BW \approx \frac{2}{Q}\frac{|\Gamma_0|}{\sqrt{1-|\Gamma_0|^2}},$$

Given $Z_{in}(\omega) = R(\omega) + \mathrm{j}X(\omega)$.
We can use Yaghjian+Best '05:

$$Q_Z(\omega) = \frac{\sqrt{(\omega R')^2 + (\omega X' + |X|)^2}}{2R(\omega)}$$

Two different cases: 1) Given $Z(\omega)$, determine Q.
2) Given a region for the antenna determine the best possible Q.





Q-factor for antennas and the stored energy

We have that

$$FBW \approx \frac{2}{Q}\frac{|\Gamma_0|}{\sqrt{1-|\Gamma_0|^2}} \tag{7}$$

How can we determine the Q-factor for any antenna?

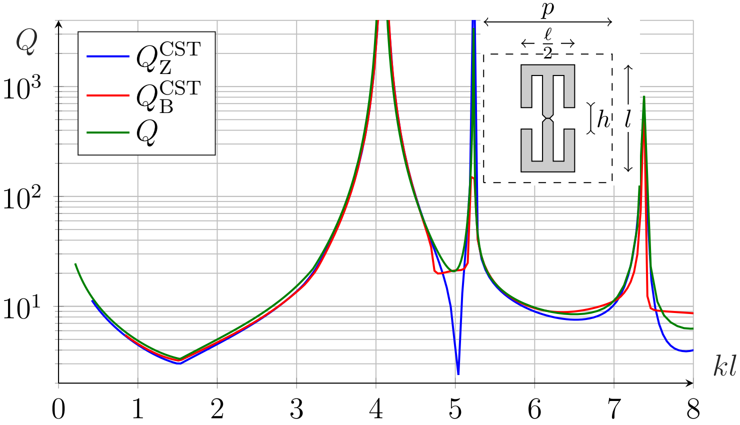
- Q-factor definition:

$$Q = \frac{2\omega\max(W_{\mathrm{e}}, W_{\mathrm{m}})}{P_{\mathrm{rad}} + P_{\Omega}} \tag{8}$$

  (stored energies and dissipated power)
- Key important fact: $W_{\mathrm{e}}, W_{\mathrm{m}}, P_{\mathrm{rad}}, P_{\Omega}$ are all expressed in terms of the antenna current.

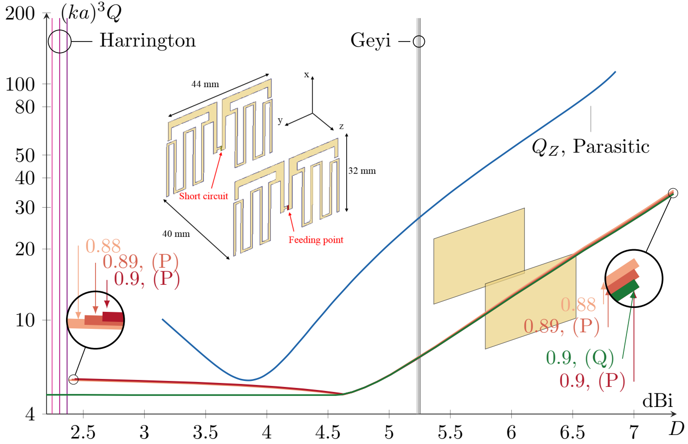
We have developed Q-factors, and bandwidth estimates for:

- small antennas, embedded antennas, periodic unit-cell antennas etc.



Ludvig-Osipov, Jonsson, *Stored energies and Q-factor of two-dimensionally periodic antenna arrays*, ArXiv 1903.01494, 2019

Q-factor vs Directivity



Trade-off between Q and Directivity of high-gain antenna

[Jonsson, Shi, Lei, Ferrero, Lizzi, IEEE Trans. Ant. Prop. 65(11) pp5686–5696, 2017]

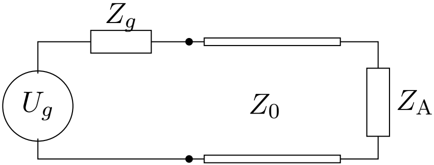
Table of Contents

For >20GHz 5G-base-stations we several challenges

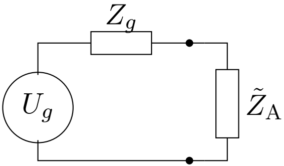
- Transmission line has higher radiation and substrate losses
- Lower power-amplifier efficiency
- Higher losses for the propagating waves

One suggested is to integrate the power-amplifier with the antenna.

-This requires a new antenna design.
-To maximize the radiated power, the optimal antenna needs to match the strong frequency dependence of the PA.



Traditional solution.



Suggested high integration solution

Finding the right question

Maximizing the radiated power

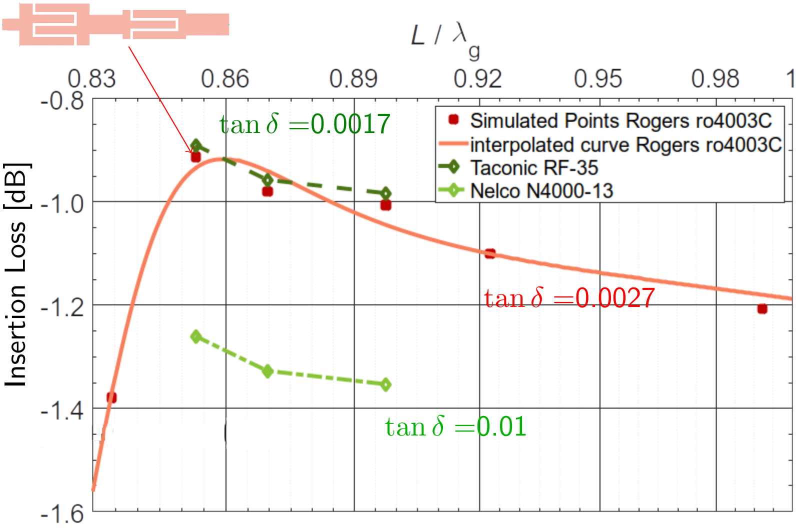
Can we use current optimization to include the matching in the antenna performance?

How do we formulate the question:

- Size and shape of a short balun/transmission line, maximize delivered power – what is the advantage of integration.
- Maximizing the power to the antenna.
- Radiated power, reciprocity.

There are well known techniques like Bode-Fano-type limitations [analyticity and sum-rules], H^∞ Helton-type bounds, Real-frequency technique of Carlin and Civalleri etc. Here we try to use a single-frequency optimization approach.

Empirical trade-off study for 5% BW vs size @22GHz



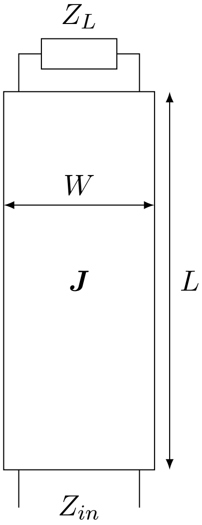
Ahmad Emadeddin, Jonsson, ICEAA 2019

Q-factor approach: a matching network

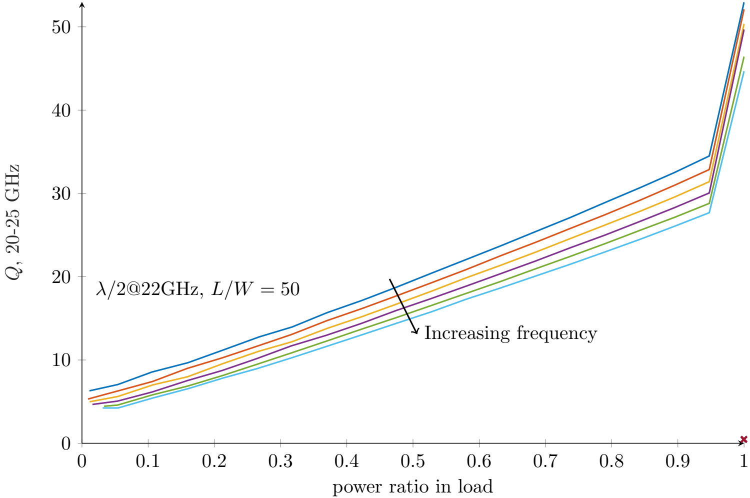
Consider the following optimization problem:

$$\max_{I} P_L$$
$$\text{s.t. } P_{\text{rad}} + P_L = 1$$
$$Q \leq q_0$$

+ Solveable
+ $Q = Q(P_L)$
- Connection to the generator.
- Can such a matching layer be realized.



Q-factor vs load power ratio

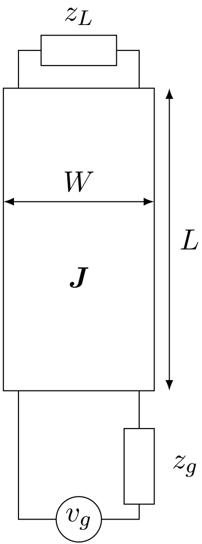


Comparable area give very low Q-factor. What guaranties the 'transmission' of power. A better model is needed.

$$\max_I P_L,$$
$$\text{s.t. } P_{\text{rad}} + P_L \le P_{in},$$
$$z_g|i_0|^2 + z_{in}|i_0|^2 = i_0^* v_g$$
$$\max(W_{\text{e}}, W_{\text{m}}) \le q$$

equivalently

$$\max_I R_L|I_n|^2,$$
$$\text{s.t. } I^H R_{\text{rad}} I + R_L|I_n|^2 \le \text{Re}(I_m^* V_g - Z_g|I_m|^2),$$
$$Z_g|I_m^2| + I_m^* Z_{in} I_m = I_m^* V_g$$
$$\max(I^H \mathbf{w}_{\text{e}} I, I^H \mathbf{w}_{\text{m}} I) \le q,$$



Assumption – increase power in load (antenna) increase power delivered.

MoM, impedance, and current optimization

Circuit theory yields:

$$z_g i_0 + z_{in} i_0 = v_g$$

The input impedance z_{in} satisfy Ohms law: $v_0 = z_{in} i_0$.

From an impedance matrix Z perspective, we find z_{in} from solving $ZI = V$, where $V = \hat{\boldsymbol{e}}_m v_0 \ell_m$, thus $I = Z^{-1}V$, and we find $i_0 = I_m \ell_m$, and

$$Y_{mm} = \frac{V_m}{I_m} = \frac{v_0 \ell_m^2}{i_0} = z_{in} \ell_m^2$$

Thus our model (N) has a fixed geometry dependent input impedance.

- Current optimization in the (N)-model **can not** account for impedance changes, associated with geometry changes through a current optimization.

Power reciprocity and scattering

The power reciprocity theorem [de Hoop etal 1974]:

$$\sigma_L(-\hat{\boldsymbol{k}}) = \frac{\lambda_0^2}{4\pi}\eta_L\eta_p(\hat{\boldsymbol{k}})G(\hat{\boldsymbol{k}}) \tag{9}$$

Here σ_L is the absorption crossection of the load: p_{in}/P_L, G is Gain, and η_p is the polarization missmatch. Furthermore

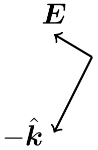
$$\eta_L = 1 - \frac{|Z_{\text{in}}^* - Z_L|^2}{|Z_{\text{in}} - Z_L|^2} \tag{10}$$

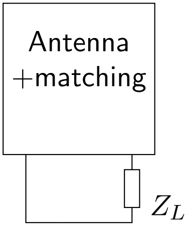
Thus if the load is conjugate-match, we have $\eta_L = 1$. Similarly a choice of polarization of the incomming wave can make $\eta_p = 1$;

Work in progress: A scattering approach

Given a plane wave $\boldsymbol{E} = \boldsymbol{E}_0 \mathrm{e}^{-\mathrm{j}\boldsymbol{k}\cdot\boldsymbol{r}}$, the recieved power in current optimization is

$$\begin{aligned} &\max_{I} P_L \\ \text{s.t. } &P_s + P_L \leq \frac{1}{2} \operatorname{Re} I^* V \\ &Q < q_0 \end{aligned}$$





where V is the MoM-coefficients associated with the plane wave.
$|\boldsymbol{E}_0|^2/(2\eta_0) = p_{\mathrm{in}}$,
We find $\sigma_L = P_L/p_{\mathrm{in}}$, with $p_{\mathrm{in}} = |\boldsymbol{E}_0|^2/(2\eta_0)$. Reciprocity gives $\eta_p \eta_L(\hat{\boldsymbol{k}}) G(\hat{\boldsymbol{k}})$.

Comparisons with a scattering sum-rule is interesting.

Conclusions

- Array antenna sum-rule, indicates a performance gap: improved arrays are possible.
  - Non-symmetric unit-cell shapes provided a method to increase the bandwidth
  - My student's work resulted in two patent on wide-band antennas
- A Q-factor representation for narrow-band arrays derived.
  - The method is validated against array performance for different elements
  - An optimization approach is ongoing.
- Different matching approaches has been considered – work in progress.