Electrically small antennas

A. Skrivervik
Laboratoire d’Electromagnétisme et d ’Acoustique
Ecole polytechnique Fédérale de Lausanne
Switzerland

Outline

• What is a small antenna ?
  • Introduction
• What is the problem ?
  • Physical limitations on small antennas
• How can we solve the problem ?
  • Design strategies and examples
• Why do some people not have the problem ?
  • The measurement of small antennas
What is a small antenna?

Wheeler:
• $\lambda/2\pi$ (radiansphere)

Usually:
• $\lambda/2$

Why small antennas?

<table>
<thead>
<tr>
<th>System</th>
<th>Frequency [MHz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>GSM (2G)</td>
<td>900 / 1800 / 1900</td>
</tr>
<tr>
<td>UMTS (3G)</td>
<td>1955 / 2155</td>
</tr>
<tr>
<td>DECT</td>
<td>1890</td>
</tr>
<tr>
<td>PHS</td>
<td>1900</td>
</tr>
<tr>
<td>CT 2</td>
<td>866</td>
</tr>
<tr>
<td>GPS</td>
<td>1575</td>
</tr>
<tr>
<td>Satellite link</td>
<td>1620 (up) / 2490 (down)</td>
</tr>
<tr>
<td>Pager</td>
<td>&lt; 900</td>
</tr>
</tbody>
</table>

The wavelength is between 14 cm and 35 cm.

On a portable device, the antenna is between a fifth and a tenth of a wavelength.
But the problem is old

f : 312 kHz => \( \lambda = 961m \)

Tip : for good ideas look in old radio-amateur publications

http://dspt.club.fr/Poldhu.htm

100 years later ...

- Size of the antenna : a tenth of wavelength
- Max. 3 dB bandwidth : 1.5%

The problem is the same !!!
Application example

The frog is the ground plane.

What is an antenna

guided wave

transition

wave in free space

$\lambda/2$
Where are the limitations?

- Limitations on the bandwidth
  - Theoretical minimum $Q$ as a function of the size (Chu, Harrington, McLean, Collin).

- Limitations on the efficiency
  - Theoretical maximum gain as a function of the size (Harrington).

for resonant antennas

![Circuit diagram with single resonance](image)
Bandwidth and quality factor

- No exact link between bandwidth and quality factor (antenna modelisation by lumped RLC circuit is approximate)
- For a second order lumped series RLC circuit, the half power bandwidth is given by:

\[
\frac{f_{\text{upper}} - f_{\text{lower}}}{f_{\text{centre}}} = \frac{1}{Q}
\]

for \( Q >> 1 \)

How can we get the Q of an antenna?
(linear polarization case)

- circuit approximation
- spherical wave expansion
Minimum quality factor

The antenna is approximated by a RLC circuit; and at resonance:

\[ Q = \frac{\omega W}{P} \]

mean stored energy

\[ W = W_e + W_m \]

radiated power

If the circuit is matched by a lossless network:

\[ Q = \frac{2\omega W_e}{P} \text{ for } W_e > W_m \]

\[ Q = \frac{2\omega W_m}{P} \text{ for } W_e < W_m \]

antenna reflection coefficient

\[ B_{3dB} = \frac{1}{Q} \]

\[ B = \frac{2}{Q} \frac{\Gamma}{\sqrt{1 - \Gamma^2}} \]

The antenna is enclosed in the smallest possible sphere.

The fields are represented by spherical waves functions.

Main problem: Evaluation of the energy stored in the reactive field.

- Chu: Equivalent ladder network (approximation).

- McLean: Directly from the fields.
Chu's method

- Enclose the antenna in the smallest possible sphere (radius $a$)
- The fields external to the sphere are represented by a weighted sum of spherical functions. These mode are orthogonal, and carry thus power independently from each other
- $Q$ is computed in terms of the time average non propagating energy external to the sphere, and of the radiated power. The energy stored inside the sphere would increase the $Q$
- The computation is difficult, because:
  - the total time-average stored energy outside the sphere is infinite, as for any propagating wave
  - A technique to separate the non propagation energy from the total energy is needed. We cannot simply use the near field components ($E$ and $H$) because the energy is non-linear

![Dipole antenna](image)
Chu's method (linear polarization)

- Compute the wave impedance of the modes

\[ Z_{TM}^r = -\frac{k}{j\omega} \frac{\hat{H}_n^{(2)}(kr)}{\hat{H}_n^{(2)}(kr)} = j\nu \frac{H_n^{(2)}(kr)}{H_n^{(2)}(kr)}, \]

- Use the modified Hankel functions

\[ Z_{TM}^r = j\nu \frac{\hat{H}_n^{(2)}(kr)}{\hat{H}_n^{(2)}(kr)} = j\nu \frac{(krh_n^{(2)})'}{krh_n^{(2)}}, \]

Chu's method (linear polarization)

- Use the recurrence formulas for modified Hankel functions

\[ h_{n-1}^{(2)}(kr) + h_{n+1}^{(2)}(kr) = \frac{2n+1}{kr} h_n^{(2)}(kr), \]

\[ \frac{n+1}{kr} h_n^{(2)}(kr) + h_{n}^{(2)}'(kr) = h_{n-1}^{(2)}(kr). \]

- Which can be re-arranged into

\[ \frac{h_{n-1}^{(2)}}{h_n^{(2)}} = \frac{1}{\frac{2n-1}{kr} - \frac{h_{n-2}^{(2)}}{h_{n-1}^{(2)}}}. \]
Chu's method (linear polarization)

- Finally

\[
Z_{TM}^{+r} = \nu \left\{ \frac{n}{jkr} + \frac{2n-1}{jkr} + \frac{1}{2n-3} \right\},
\]

where \( \nu = \sqrt{\frac{L}{C}}, c = \frac{1}{\sqrt{\mu}} \) and \( kr = \frac{2\pi}{X} = \frac{2\omega}{c} \).

Chu's method (linear polarization)

- Thus

\[
Z_{TM}^{+r} = \frac{n}{j\omega\epsilon} + \frac{2n-1}{j\omega\mu} + \frac{1}{2n-3} \left( \frac{3}{j\omega\epsilon} + \frac{1}{j\omega\mu + \nu} \right),
\]

- This is the transfer function of a Cauer network
The ladder network: $\text{TM}_n$ modes

\[ Q = \frac{2\omega W}{P}, \quad W_e = \frac{1}{2} \mathcal{C} V \mathcal{C}^* \quad \text{and} \quad P = R I_R I_R^* \]

Example: $\text{TM}_{01}$ mode only

\[ I = I_L + I_R, \quad V_R = V_L, \quad V_R = I_R Z_R, \quad V_L = I_L Z_L, \]

\[ I_R = \frac{Z_L}{Z_R + Z_L} I \quad \text{and} \quad I_R Z_R = I_L Z_L. \]
Example : TM\textsubscript{01} mode only

\[ V_e = V \frac{Z_C}{Z_C + Z_{LR}}. \]

Example : TM\textsubscript{01} mode only

\[ W_e > W_m \quad \text{(TM mode)} \quad \iff \quad Q_1 = \frac{2\omega W_e}{P}. \]

\[ W_e = \frac{1}{2} C V_C V_C^* = \frac{1}{2} C V V^* \frac{Z_C}{Z_C + Z_{LR}} \frac{Z_C^*}{(Z_C + Z_{LR})^*} \]

\[ = \frac{1}{2} C V V^* \frac{Z_C Z_C^*}{Z_{tot} Z_{tot}^*}. \]
Example : TM\textsubscript{01} mode only

\[ P = Z_R I_R I_R^* = Z_R \frac{Z_L}{Z_R + Z_L} \frac{Z_L^*}{(Z_R + Z_L)^*} I I^* \]

\[ Q_1 = \frac{\omega C V V^* Z_C Z_C^*(Z_R + Z_L)(Z_R + Z_L)^*}{Z_{tot} Z_{tot}^* Z_R Z_L Z_L^* I I^*} . \]

\[ Z_{tot} = \frac{V}{I} \]

Example : TM\textsubscript{01} mode only

\[ Q_1 = \frac{\omega C Z_C Z_C^*(Z_R + Z_L)(Z_R + Z_L)^*}{Z_R Z_L Z_L^*}, \]

with

\[ \omega = \frac{k}{\sqrt{\epsilon \mu}} , \quad Z_C = \frac{1}{j \omega C} = \frac{1}{j \omega \epsilon a} . \]

\[ Z_L = j \omega L = j \omega \mu a \text{ and } Z_R = \nu . \]
Example: $\text{TM}_{01}$ mode only

$$Q_1 = \frac{k\epsilon a (\nu + j\omega a)(\nu - j\omega a)}{\sqrt{\epsilon\mu}(j\omega a)(-j\omega a)\nu(j\omega a)(-j\omega a)}.$$  

Separation into 2 terms:

$$Q_1 = Q_{1.1} + Q_{1.2}$$

$$Q_{1.1} = \frac{k\epsilon a \nu^2}{\sqrt{\epsilon\mu}\omega^4 a^4 \mu^2 \epsilon^2 \nu} = \frac{k\nu \epsilon^2 \mu^2}{\sqrt{\epsilon\mu} k^4 a^4 \mu^2 \epsilon} \frac{\nu \epsilon}{\sqrt{\epsilon\mu} k^3 a^3} = \left(\frac{1}{ka}\right)^3$$

$$Q_{1.2} = \frac{k\epsilon a \omega^2 \mu^2 a^2}{\sqrt{\epsilon\mu}\omega^4 a^4 \mu^2 \epsilon^2 \nu} = \frac{1}{ka},$$
Example: TM\textsubscript{01} mode only

Finally:

\[ Q_1 = \frac{1}{ka} + \left(\frac{1}{ka}\right)^3 \]

Theoretical minimum \( Q \) for an antenna exciting the TM\textsubscript{01} mode only

Antenna exciting two modes

\[ Q_2 = \frac{3}{ka} + \frac{6}{(ka)^3} + \frac{18}{(ka)^5}. \]

\( Q_2 \) is always larger than \( Q_1 \)
Higher order mode approximation

- Too complicated for many modes (difficult to compute the energy stored in each capacitor and inductor of the ladder). We use an approximate equivalent circuit for each mode.

The ladder network: approximation for higher order modes

\[
Q_n = \frac{1}{2\eta} (kr)^2 \left| h_n^{(2)} \right|^2 \left[ \omega \frac{\partial X_n}{\partial \omega} - X_n \right]
\]

where

\[
X_n = \eta \frac{kr j_n (kr j_n) + kr n_n (kr n_n)}{(kr)^2 \left| h_n^{(2)} \right|^2}
\]
Higher order mode approximation

- Result for TM$_{01}$ mode

\[ Q_1 = \frac{1 + 2(ka)^2}{(ka)^3(1 + (ka)^2)} \]

- Exact solution

\[ Q_1 = \frac{1}{ka} + \left(\frac{1}{ka}\right)^3 \]

Lowest possible $Q$ for linearly polarized antennas

\[ Q_{\text{min}} = \frac{1 + 2(ka)^2}{(ka)^3(1 + (ka)^2)} \]
Chu’s method : TE modes

\[ Y_{+r}^{TE} = -\frac{1}{\nu} \left\{ \frac{n}{jk\rho} + \frac{1}{2n-1} \frac{1}{jk\rho} + \frac{1}{2n-3} \frac{1}{jk\rho} \right\} \]

\[ I_n \]
\[ Z_n^{TE} \]
\[ L = \frac{\mu_0}{(2n-1)} \]
\[ C = \frac{\epsilon_0}{2n-1} \]
\[ C = \frac{\epsilon_0}{(2n-5)} \]
\[ R = \nu \]
\[ \frac{3}{jk\rho} + \frac{1}{jk\rho+1} \]

Chu’s method

- Linear polarization antenna : Either TE or TM mode

\[ Q_{\text{min}} = \frac{1 + 2(ka)^2}{(ka)^3(1 + (ka)^2)} \]

- Circular polarization : combination of TE and TM modes with the proper phase shift

\[ Q_1 = \frac{1}{2} \frac{1 + 3(ka)^2}{(ka)^3(1 + (ka)^2)} \]
The field method: dipole

\[ \nabla^2 A_z + k^2 A_z = 0; \quad \frac{1}{r^2} \frac{\partial}{\partial r} \left( r^2 \frac{\partial A_z}{\partial r} \right) + k^2 A_z = 0 \]

\[ H_\phi = \frac{II}{4\pi} e^{-jkr} \left( \frac{jk}{r} + \frac{1}{r^2} \right) \sin \theta \]

\[ E_r = \frac{II}{2\pi} e^{-jkr} \left( \frac{\eta}{r^2} + \frac{1}{j\omega r^3} \right) \cos \theta \]

\[ E_\theta = \frac{II}{4\pi} e^{-jkr} \left( \frac{j\omega \mu}{r} + \frac{\eta}{r^2} + \frac{1}{j\omega r^3} \right) \sin \theta \]
The field method: dipole

\[ w_e = \frac{1}{2} \varepsilon EE^* \]

\[ w_e^{tot} = w_e^r + w_e^{np} \]

\[ w_e^{np} = w_e^{tot} - w_e^r \]

\[ W_e^{np} = \int_0^{2\pi} \int_0^\pi \int_0^\infty w_e^{np} r^2 \sin \theta \, dr \, d\theta \, d\varphi \]

\[ P_{rad} = \oint_S (E \times H) \, ds = \int_0^{2\pi} \int_0^\pi \int_0^\infty E_\theta H_\phi^* \sin \theta \, d\theta \, d\varphi \]

Linear antenna case: electric dipole in z direction

- The total non-propagating energy is obtained by integrating the non-propagating energy density over the entire space excepted for the sphere of radius enclosing the antenna.
Linear antenna case: electric dipole in z direction

- Computation of radiated power:

\[ P_{rad} = \oint_S (\mathbf{E} \times \mathbf{H}^*) \, ds. \]

\[ P_{rad} = \int_{0}^{2\pi} \int_{0}^{\pi} E_\theta H_\phi^* r^2 \sin \theta d\theta d\phi. \]

\[ P_{rad} = \frac{I^2 l^2 \omega \mu \kappa}{6\pi} \sin^3 \theta d\theta. \]

- Finally

\[ Q = \left( \frac{1}{ka} \right)^3 + \frac{1}{ka}, \]
Circular polarized antenna

\[ Q = \frac{1}{2} \left( \frac{1}{k^3 a^3} + \frac{2}{ka} \right). \]

Lowest possible Q for circularly polarized antennas

\[ Q_{\text{min}} = \frac{1}{2} \left( \frac{1}{k^3 a^3} + \frac{2}{ka} \right) \]

\[ Q_{\text{min}} = \frac{1}{2} \left( \frac{1 + 2(ka)^2}{(ka)^3 \left( 1 + (ka)^2 \right)} \right) \]
Q in presence of losses

Maximum gain of an antenna

- The gain is defined as

\[
G(\theta) = \frac{4\pi r^2 S_r(\theta)}{P_f},
\]

- Where \( S_r \) is the \( r \) component of the Poynting vector and \( P_f \) is the total radiated power, obtained integrating \( S_r \) over a large sphere
Maximum gain of an antenna: intuitive approach

- Spherical wave functions are obtained by solving Helmholtz’ equation in spherical coordinated. The fields radiated by an antenna oriented so that the maximum is at \( \theta = \pi/2 \) and \( \varphi = 0 \) is given by:

\[
E_\theta = -\frac{1}{\sin \theta} \sum_{m,n} A_{mn} \hat{H}_n^{(2)}(kr) P_n^m(\cos \theta) \sin(m\phi + \alpha_{mn}) \\
- \frac{\sin \theta}{j\omega \epsilon r} \sum_{m,n} B_{mn} \hat{H}_n^{(2)\prime}(kr) P_n^m(\cos \theta) \cos(m\phi + \beta_{mn}).
\]

\[
E_\phi = -\sin \theta \sum_{m,n} A_{mn} \hat{H}_n^{(2)}(kr) P_n^m(\cos \theta) \cos(m\phi + \alpha_{mn}) \\
- \frac{1}{j\omega \epsilon r \sin \theta} \sum_{m,n} m B_{mn} \hat{H}_n^{(2)\prime}(kr) P_n^m(\cos \theta) \sin(m\phi + B_{mn}).
\]

\( \theta = \pi/2 \).
The radiated field is obtained when $r$ is large, thus

$$H_n^{(2)}_{kr \to \infty}(kr) \sim \sqrt{\frac{2}{(\pi kr)}} e^{-jkr - \frac{1}{2} n \pi - \frac{\pi}{4}}.$$ 

Which is equivalent to

$$\hat{H}_n^{(2)}_{kr \to \infty}(kr) \sim j^{n+1} \frac{e^{-jkr}}{kr}.$$ 

We finally get

$$\frac{\partial}{\partial r} \left[ r \hat{H}_n^{(2)}_{kr \to \infty}(kr) \right] \sim j^n e^{-jkr}.$$ 

We finally get

$$G\left(\frac{\pi}{2}, 0\right) = \frac{\left| \sum_{m,n} j^{n-1} \left[ A_{mn} P_n^{m'}(0) + \sqrt{\frac{\mu}{\epsilon}} m B_{mn} P_n^m(0) \right] \right|^2}{\sum_{m,n} \frac{1}{\epsilon_m} \left[ |A_{mn}|^2 + \frac{\mu}{\epsilon} |B_{mn}|^2 \right]^{2} \frac{(n+1)(n+m)!}{(2n+1)(n-m)!}}.$$ 

Which we need to maximize
Maximum gain of an antenna

- After some cumbersome computations, and limiting the number of modes (wave functions) to \( N \), we finally get:

\[
G = N^2 + 2N
\]

Thus, if the number of modes can be increased, the gain has potentially no limit.

Maximum gain of an antenna

- What limits the gain:
  - Possibility to manufacture an antenna radiating many propagating modes
  - Losses (higher order modes have usually higher losses)
  - Bandwidth (the more modes, the smaller the bandwidth)
Practical gain limitation

\[ G(\theta) = \frac{4\pi r^2 S_e(\theta)}{P_f} \]

Wave impedance of a TM wave

\[ G = N^2 + 2N \]

\[ Z_{+r}^{TM} = \frac{j\eta}{kr} + \frac{\eta}{h_n^{(2)}} \left[ \frac{2}{\pi kr} + j(n_j n + n n') \right] \]

- The wave impedance is reactive when \( j n j_n' + n_n n_n' \) is large compared to \( 2/\pi kr \)
- \( n_n \) increases rapidly when \( kr < n \)
- The modes of order \( n > ka \) are rapidly cut off and are not naturally present in the field of an antenna of radius \( a \)
- Modes of order \( n > ka \) will increase heavily the stored reactive energy, but have no impact on radiated power
Maximum gain for a practical bandwidth: \( N = ka \)

\[
G_{\text{normal}} = (ka)^2 + 2ka
\]

Comparison with measured gains

Circular parabolic reflector antenna:
- Size 146 \( \lambda \), \( G_{\text{measured}} \): 50.4 dBi, \( G_{\text{max}} \): 53.3 dBi

Pyramidal horn antenna:
- Size 7.5 \( \lambda \), \( G_{\text{measured}} \): 24.5 dBi, \( G_{\text{max}} \): 27.7 dBi

Narda horn antenna:
- Size 2.5 \( \lambda \), \( G_{\text{measured}} \): 15-16 dBi, \( G_{\text{max}} \): 18.7 dBi

Rolled slot antenna:
- Size 0.2 \( \lambda \), \( G_{\text{measured}} \): -11.7 dBi, \( G_{\text{max}} \): 2.6 dBi

Slot-Dipole antenna:
- Size 0.2 \( \lambda \), \( G_{\text{measured}} \): 0 dBi, \( G_{\text{max}} \): 2.6 dBi
Loop antenna characteristics

small loop radiation resistance (single turn) \[ R_r = 20\pi^2 C_\lambda^4 \]

small loop radiation resistance (N turns) \[ R_r = 20\pi^2 C_\lambda^4 N^2 \]

small loop ohmic loss resistance (single turn) \[ R_{loss} = \frac{a}{b} \sqrt{\frac{\omega \mu_0}{2\sigma}} \]

small loop ohmic loss resistance (N turns) \[ R_{loss} = \frac{Na}{b} \sqrt{\frac{\omega \mu_0}{2\sigma}} \]

radiation efficiency \[ \eta = \frac{R_r}{R_r + R_{loss}} \]
single turn loop resistance

Loop made of 1 mm thick wire at 3 GHz
single turn loop efficiency

![Graph showing efficiency vs. circumference]

Small loop bandwidth

Single turn loop inductance: \[ L = \mu_0 a \left[ \ln \frac{8a}{b} - 1.75 \right] \]

N turn loop inductance: \[ L \approx \mu_0 a \left[ \ln \frac{8a}{b} - 1.75 \right] N^2 \]

Loop antenna relative bandwidth: \[ \frac{\Delta f}{f_0} = \frac{R_r + R_{loss}}{2 \pi f_0 L} \]
Small loop bandwidth

Loop made of 1 mm thick wire at 3 GHz

Example: microstrip patch antenna

\[ a \approx \frac{\lambda_0}{2\sqrt{\varepsilon_e}} \]

\[ 1 < \varepsilon_e < \varepsilon_r \]
Microstrip patch miniaturization

![Graph showing permittivity vs. relative patch size]

Miniature patch efficiency

![Graph showing efficiency vs. relative patch size for low and medium dielectric loss]
Miniature patch bandwidth

![Graph showing miniature patch bandwidth with low and medium dielectric loss at f=1 GHz]

Main miniaturization techniques. Effect on the performances

- **Antenna loading**
  - With lumped elements
  - With high permittivity or high permeability materials
- **Make some parts of the antenna virtual**
  - Using ground planes
  - Using short circuits
- **Optimizing the geometry**
- **Use the environment**
- **Multifrequency antennas**
Antenna loading (lumped elements)

- Antennas small compared to the wavelength are non-resonant (strong reactive part of the input impedance)
- Antennas small compared to the wavelength usually have a small radiation resistance
- => small antennas can be made resonant by reactively loading them
- => A matching network will usually be necessary to match the radiation resistance to the transmission line

- Two element top loaded monopole
- Four element top loaded monopole
- Coil loaded antenna
- Capacitor plate antenna
- Capacitively loaded loop
Antenna loading (lumped elements)

Effect on performances

- Lowers the antenna efficiency
  - If the added element has losses

- Enhances the antenna quality factor (lowers the bandwidth)
  - If the added element is lossless
Antenna loading (material)

- an antenna is resonant when at least one of its dimensions is of the size of half a wavelength
- the wavelength at a given frequency depends on the dielectric and magnetic properties of the material surrounding an antenna:
  \[ \lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r \mu_r}} \]
- the size of a resonant antenna can be decreased by increasing the dielectric or magnetic constant around the antenna
Antenna loading (material)

\[ a \approx \frac{\lambda_0}{2\sqrt{\varepsilon_e}} \]

\[ 1 < \varepsilon_e < \varepsilon_r \]

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Antenna loading : Effect on performances

- Concentration of electric (magnetic) fields in the high permittivity (permeability) regions

- Higher near fields

- Higher Q and lower bandwidth
Using ground planes and short circuits

- The image theory is used to simulate currents and charges
- Size reduction
Using ground planes and short circuits examples

The Planar Inverted F Antenna
Using ground planes and short circuits effects on performances

Difficult to predict in general!

Optimizing the geometry

- geometrical loading effects (notches, slots, ...)
- bends and curvature effects
Bend effect

\[ h \approx \frac{\lambda_0}{4} \]

\[ h + L \approx \frac{\lambda_0}{4} \]

monopole antenna  Inverted L Antenna (ILA)  Inverted F Antenna (IFA)

Curvature effect

\[ h \geq \frac{\lambda_0}{4} \]

\[ h < \frac{\lambda_0}{4} \]

monopole antenna  short helix antenna
Slot effect

\[ a \geq \frac{\lambda_g}{2} \]

microstrip antenna

\[ a < \frac{\lambda_g}{2} \]

microstrip antenna with notches

\[ a < \frac{\lambda_g}{2} \]

microstrip antennas with slots

Optimizing the geometry examples

The normal mode helical antenna

•

•
Optimizing the geometry
examples

Fractal antennas

Optimizing the geometry
effect on performances

- Loss of efficiency due to current concentration
- Loss of bandwidth due to frequency sensitivity of the technique itself (image theory)
- Alteration of polarization
Using the environment

- effect of the handset case
  - use the metallic parts of the case in a constructive way
  - use the maximum available volume
- effect of the human body
  - the human body acts as a ground plane

Using the environment example

Pager integrated in a wrist watch
Summary

- There are physical limitations on antenna performances related to size of the antenna.
- The limits are difficult to reach!
- Different miniaturization strategies have different impact on the performances. => use at least two strategies simultaneously.