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Antennas for EAWS

Anja Skrivervik
anja.skrivervik@epfl.ch, 021 693 4635

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Outline

- What is the problem
- Electrically Small Antennas
- Antennas in Lossy media (implants and wearables)
- Some regulations
- Some simulation issues

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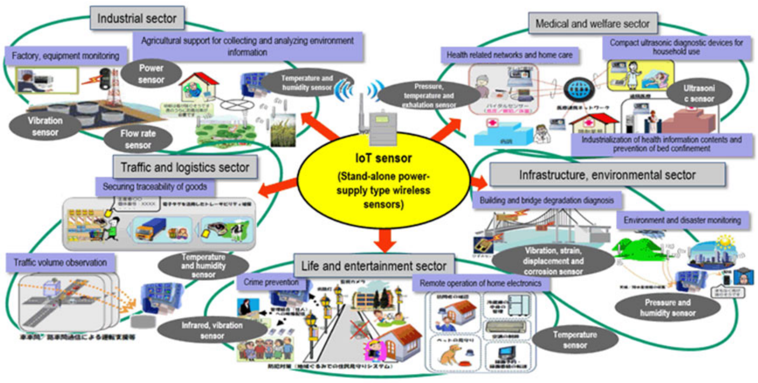
Global wireless network



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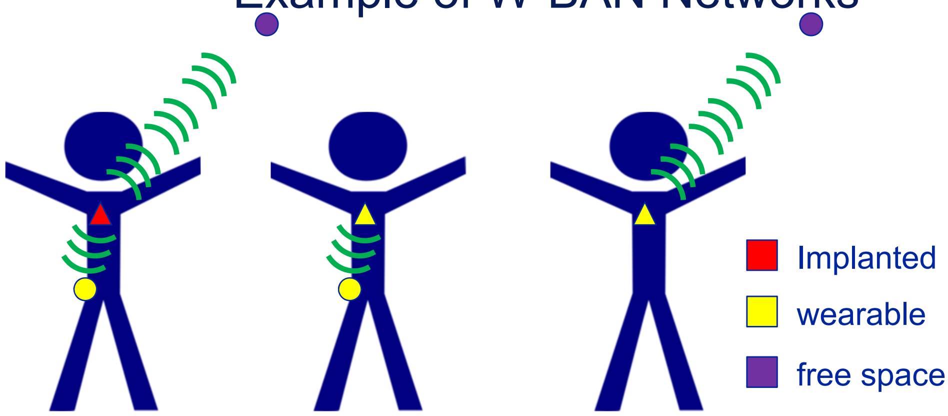
Example of W-BAN Networks

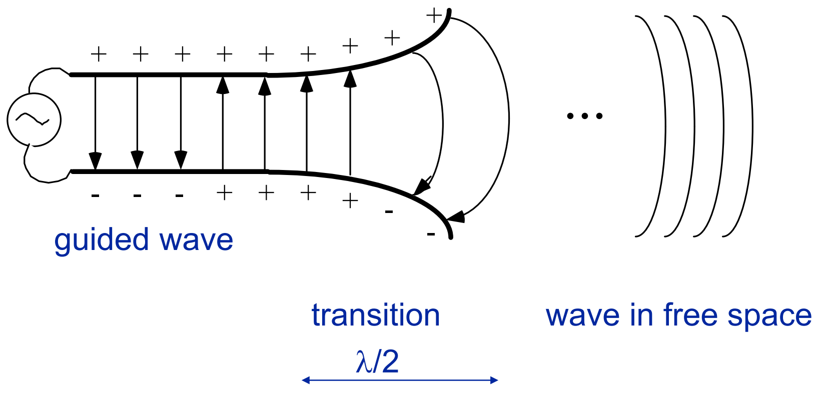


In Body On Body Off Body

Why are antennas important for wireless powering

- Saving power is crucial
- Parts of systems (typically sensors) can be small
- Wireless powering can take place in or on a lossy medium (body)
- The difference between a good or bad antenna design can make differences of up to 10 dB in power efficiency of the system

What is an antenna



What is an antenna ?

- An antenna transforms a guided wave to a radiated wave
- An antenna is a link between Krichhoff's world (circuit) and Maxwell's world (fields)
- An antenna is a spatial filter and a spectral filter
- An antenna is a one port device
- An antenna is a two port device
- An antenna is part of a system

Circuit characteristics

- Frequency response of the antenna (antenna bandwidth)
- Input impedance of the antenna
- Reflection coefficient of the antenna

Can be measured using S parameter measurements or Power measurements

Field characteristics

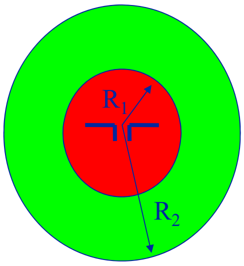
- Radiation pattern
- Directivity
- Gain
- Polarization
- Radiation efficiency

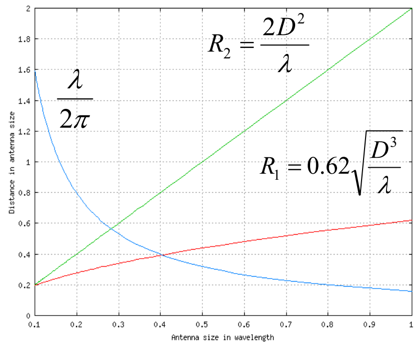
Characterizing those features imply the measurement of an attenuation coefficient (transfer function). They need to be characterized in an anechoic chamber or an outdoor range

Cases where antenna size is important

- Sensor telemetry (electrically small)
- Implants and wearables

What is an electrically small antenna ?

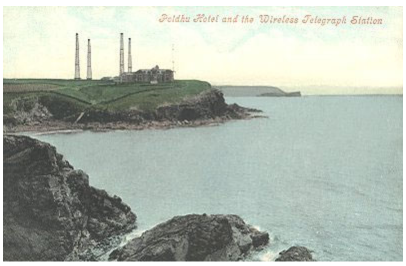


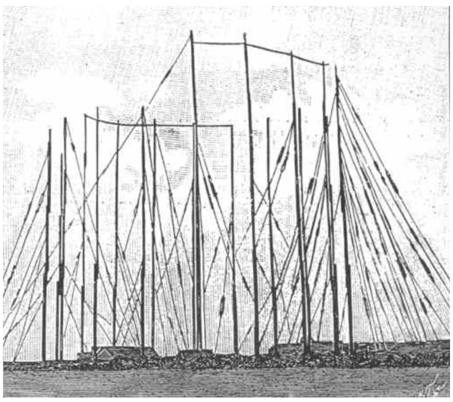


Wheeler:
- $\lambda/2\pi$ (radiansphere)

Usually:
- $\lambda/2$

1st boom of ESAs: the old radio days





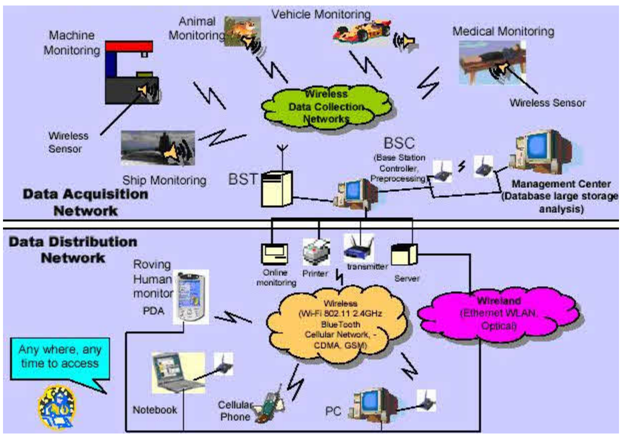
f : 312 kHz => λ = 961m

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

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

The crazy1990s

The new millenium: IoT and IoE



https://www.elprocus.com/introduction-to-wireless-sensor-networks-types-and-applications/

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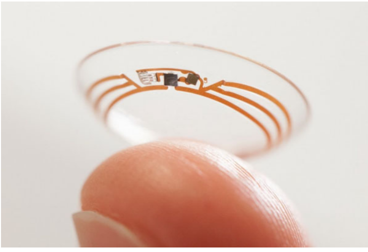
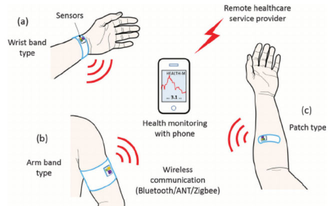
Today: worn and implanted sensors

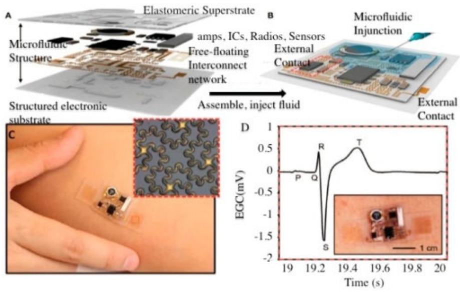


http://orthogonal.io/medical-softtware/wearable-sensors-are-the-future-of-personalized-medicine-html/



https://www.researchgate.net/figure/Three-types-of-wearable-sensor-nodes-powered-by-thermoelectric-energy-harvesters-The_fig1

https://medicalxpress.com
Implantable sensor



https://openi.nlm.nih.gov/detailedresult.php?img=PMC4367357_sensors-15-03236f4&req=4

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Application example

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Physical bounds

Why do we need them ?

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Physical bounds of antennas

- Provide a limitation to the antenna characteristics
- Are usually hard to reach
- Can be used as benchmarks to assess designs
- Are extremely useful to the system and design engineers to assess feasibility
- Bounds are existing for classic ESAs
- Classic bounds for ESAs were introduced 70 years ago and were based on Spherical Wave Expansions
- Research on bounds for implantable antennas has just started. Can we use SWE to gain insight in implantable antenna characteristics ?

Two classic uses of spherical wave expansions: Limits on quality factor and gain

- L.J. Chu, "Physical limitations on omni-directional antennas", *Journal of Applied Physics*,. 19, 1948, pp. 1163-1175.
- R.F. Harrington, "On the Gain and Beamwidth of Directional Antennas", *IRE Transactions on Antennas and Propagation*, vol. AP-6, 1958, pp. 219-225.

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 :

$$\boxed{\frac{f_{\text{upper}} - f_{\text{lower}}}{f_{\text{centre}}} = \frac{1}{Q}}$$

valid for Q >>1

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:

$$Q = \frac{\omega W}{P}$$

If the circuit is matched by a lossless network:

$$Q = \frac{2\omega W_e}{P} \; for\, W_e > W_m$$

$$Q = \frac{2\omega W_m}{P} \; for\, W_e < W_m$$

and $$B_{3dB} = \frac{1}{Q}$$

Minimum quality factor

- The antenna is enclosed in the smallest possible sphere.
- The fields are represented by spherical waves functions.

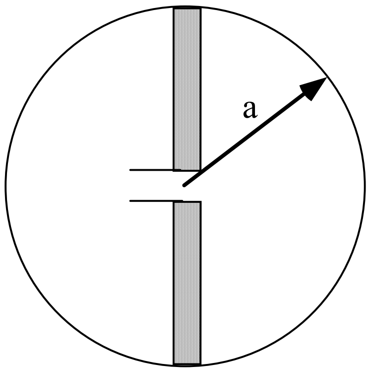
Main problem: Evaluation of the energy stored in the reactive field.

- Chu: Equivalent ladder network (approximation).
  - L.J. Chu, Journal of Appl. Physics, vol. 19, pp. 1163-1175, 1948
- McLean: Directly from the fields.
  - J.S. Mc Lean, IEEE Trans on AP, vol. AP-44, pp. 672-675, 1996

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

Chu's method



dipole antenna

Chu's method (linear polarization)

- Compute the wave impedance of the modes

$$Z_{+r}^{TM} = -\frac{k}{j\omega\epsilon}\,\frac{\hat{H}_n^{'\,(2)}(kr)}{\hat{H}_n^{(2)}(kr)} = j\nu\frac{\hat{H}_n^{'\,(2)}(kr)}{\hat{H}_n^{(2)}(kr)},$$

- Use the modified Hankel functions

$$Z_{+r}^{TM} = j\nu\frac{\hat{H}_n^{'\,(2)}(kr)}{\hat{H}_n^{(2)}(kr)} = j\nu\frac{\left(krh_n^{(2)}\right)'}{krh_n^{(2)}},$$

Chu's method (linear polarization)

- Use the recurrence formulas for modified Hankel functions

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

$$\frac{n+1}{kr} h_n^{(2)}(kr) + h_n^{(2)\,\prime}(kr) = h_{n-1}^{(2)}(kr).$$

- 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

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

where $\nu = \sqrt{\frac{\mu}{\epsilon}}$, $c = \frac{1}{\sqrt{\epsilon\mu}}$ and $kr = \frac{2\pi r}{\lambda} = \frac{2r\omega}{c}$,

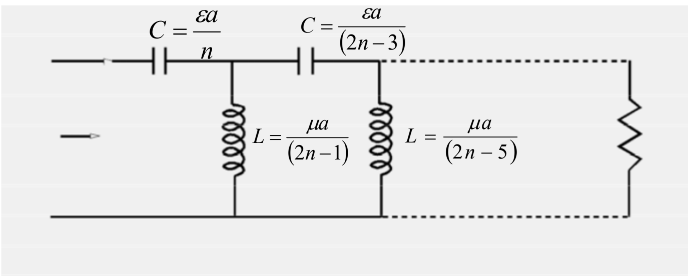
Chu's method (linear polarization)

- Thus

$$Z_{+r}^{TM} = \frac{n}{j\omega r\epsilon} + \cfrac{1}{\cfrac{2n-1}{j\omega r\mu} + \cfrac{1}{\cfrac{2n-3}{j\omega r\epsilon} + \cfrac{\ddots}{\cfrac{1}{\cfrac{3}{j\omega r\epsilon} + \cfrac{1}{\cfrac{1}{j\omega r\mu} + \nu}}}}}.$$

- This is the transfer function of a Cauer network

The ladder network: TM_n modes



$$Q = \frac{2\omega W}{\overline{P}} \qquad W_e = \frac{1}{2} C V_c V_c^* \qquad P = R I_R I_R^*$$

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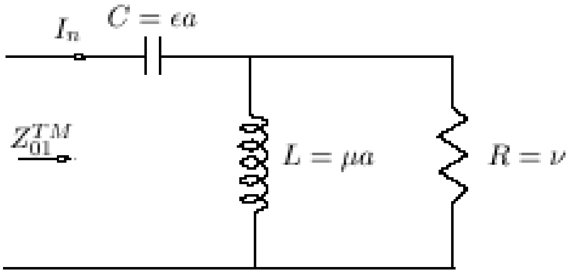
Example : TM_{01} mode only



$$I = I_L + I_R,$$

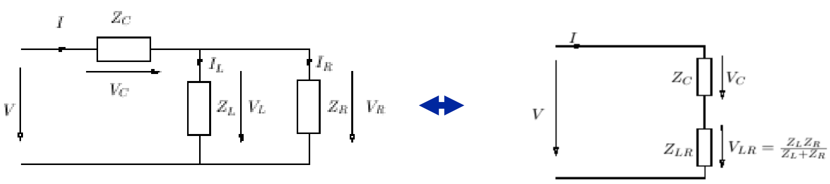
$$V_R = V_L,$$

$$V_R = I_R Z_R,$$

$$V_L = I_L Z_L,$$

$$I_R = \frac{Z_L}{Z_R + Z_L} I. \quad \longleftarrow \quad I_R Z_R = I_L Z_L.$$

Example : TM_{01} mode only



$$V_c = V \frac{Z_C}{Z_C + Z_{LR}}.$$

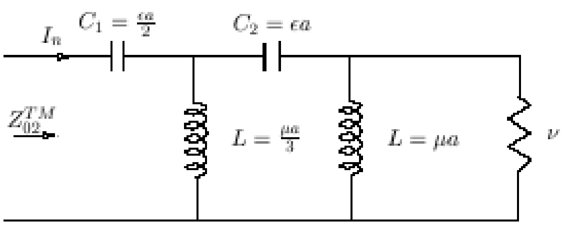
Example : TM_{01} mode only

Finally :

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

Theoretical minmum Q for an antenna exciting the TM_{01} mode only

Antenna exciting two modes

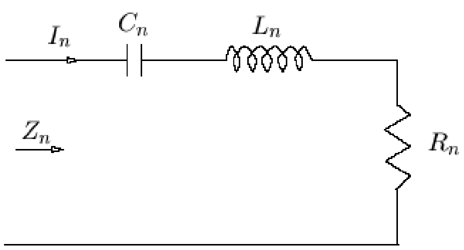


$$Q_2 = \frac{3}{ka} + \frac{6}{(ka)^3} + \frac{18}{(ka)^5}.$$

Q_2 is always larger than Q_1

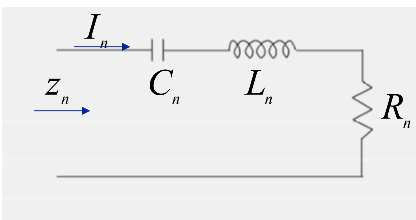
Higher order mode approximation

- 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

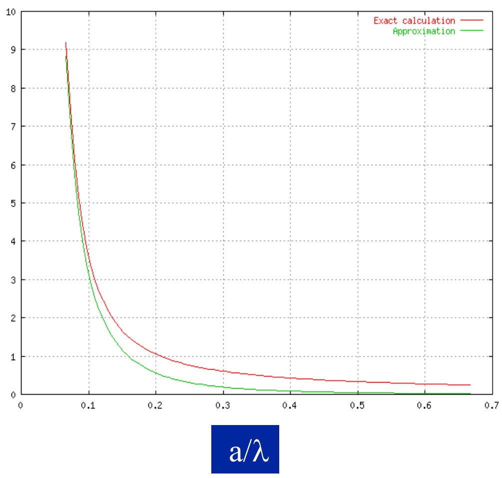
- Result for TM_{01} mode

$$Q_1 = \frac{1+2(ka)^2}{(ka)^3(1+(ka)^2)}.$$

- Exact solution

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

Lowest possible Q for linearly polarized antennas

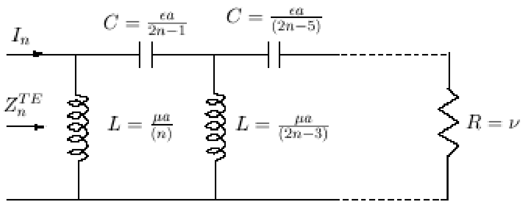


$$Q_{\min} = \frac{1}{ka} + \frac{1}{(ka)^3}$$

$$Q_{\min} = \frac{1+2(ka)^2}{(ka)^3\left(1+(ka)^2\right)}$$

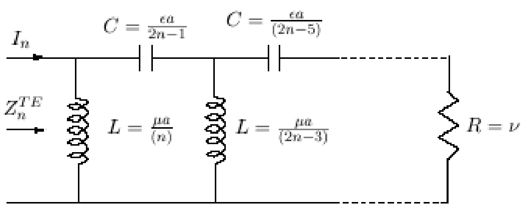
Chu's method : TE modes

$$Y_{+r}^{TE} = -\frac{1}{\nu}\left\{\frac{n}{jkr} + \cfrac{1}{\cfrac{2n-1}{jkr} + \cfrac{1}{\cfrac{2n-3}{jkr} + \ddots \cfrac{1}{\cfrac{3}{jkr} + \cfrac{1}{\cfrac{1}{jkr} + 1}}}}\right\}$$



Chu's method : TE modes

$$Y_{+r}^{TE} = -\frac{1}{\nu}\left\{\frac{n}{jkr} + \cfrac{1}{\cfrac{2n-1}{jkr} + \cfrac{1}{\cfrac{2n-3}{jkr} + \ddots \cfrac{1}{\cfrac{3}{jkr} + \cfrac{1}{\cfrac{1}{jkr} + 1}}}}\right\}$$



Chu's method

- Linear polarization antenna : Either TE or TM mode

$$Q_{\min} = \frac{1}{ka} + \frac{1}{(ka)^3}$$

- 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: TM case (e.g. dipole)

$$w_e = \frac{1}{2}\varepsilon \mathbf{E}\mathbf{E}^* \Rightarrow w_e^{rad},\; w_e^{tot}$$

$$w_e^{np} = w_e^{tot} - w_e^{rad}$$

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

$$P_{rad} = \oint_S \left(\mathbf{E}\times\mathbf{H}\right) ds = \int_0^{2\pi}\int_0^{\pi} E_\theta H_\phi^* \sin\theta \, d\theta \, d\phi$$

Example : TM_{01} mode only

Finally :

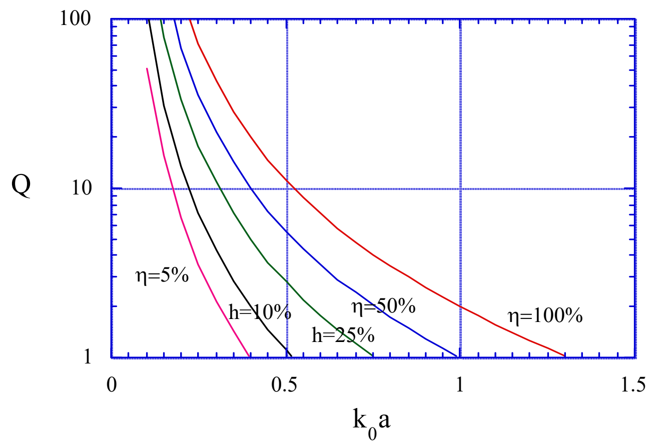
$$Q = \frac{1}{ka} + \left(\frac{1}{ka}\right)^3$$

Theoretical minmum Q for an antenna exciting the TM_{01} mode Only

k is the wavenumber

A the radius of the sphere enclosing the antenna

Q in presence of losses



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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

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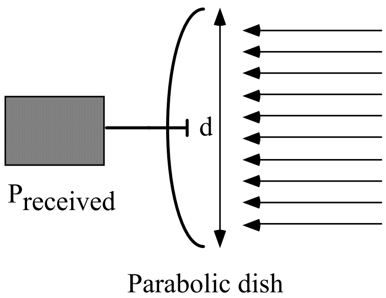
Maximum gain of an antenna : intuitive approach



$P_{received} = p \; A_e(\theta,\varphi)$

$A_e(\theta,\varphi)$: *effective aperture or effective surface*

$$P_{received}^{\max} = p \; \pi \; \frac{d^2}{4} \; \rho_A \quad , \quad \rho_A \le 0.82$$

$$G(\theta,\varphi) = \frac{4\pi}{\lambda^2} A_e(\theta,\varphi) \quad , \quad G^{\max} = \left(\frac{\pi d}{\lambda}\right)^2 \rho_A$$

Maximum gain of an antenna

Harrington, IRE Trans on AP, vol. AP-6, pp. 219-225, 1958

- Spherical wave functions are obtained by solving Helmoltz' equation in spherical coordinated. The fields radiated by an antenna oriented so that the maximum is at $\theta=\pi/2$ and φ=0 is given by :

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

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

Maximum gain of an antenna

- 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

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

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

Maximum gain of an antenna

- The radiated fields are given by

$$E_\theta = \frac{e^{-jkr}}{kr} \sum_{m,n} j^{n+1} \left[\frac{m A_{mn}}{\sin\theta} P_n^m(\cos\theta) \sin(m\phi + \alpha_{mn}) + \sqrt{\frac{\mu}{\epsilon}} \sin\theta B_{mn} P_n^m{}'(\cos\theta) \cos(m\phi + \beta_{mn}) \right],$$

$$E_\phi = \frac{e^{-jkr}}{kr} \sum_{m,n} j^{n+1} \left[-\sin\theta A_{mn} P_n^m{}'(\cos\theta) \cos(m\phi + \alpha_{mn}) - \sqrt{\frac{\mu}{\epsilon}} \frac{m B_{mn}}{\sin\theta} P_n^m(\cos\theta) \sin(m\phi + \beta_{mn}) \right].$$

Maximum gain of an antenna

- In the far field, the radiated power density is given by :

$$S_r = \sqrt{\frac{\epsilon}{\mu}} \left(|E_\theta|^2 + |E_\phi|^2 \right).$$

- And the radiated power by

$$P = \int_0^{2\pi} d\phi \int_0^{\pi} d\theta \sin\theta \left(|E_\theta|^2 + |E_\phi|^2 \right) = \frac{4\pi}{k^2} \sum_{m,n} \frac{1}{\epsilon_m} \left[\sqrt{\frac{\epsilon}{\mu}} |A_{mn}|^2 + \sqrt{\frac{\mu}{\epsilon}} |B_{mn}|^2 \right] \frac{n(n+1)(n+m!)}{(2n+1)(n-m)!},$$

$$\epsilon_m = \begin{cases} 1 & m = 0, \\ 2 & m > 0. \end{cases}$$

Maximum gain of an antenna

- We finally get

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

- Which we need to maximize

Maximum gain of an antenna (linear polarization)

- After some cumbersome computations and using the spherical wave expansions, 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

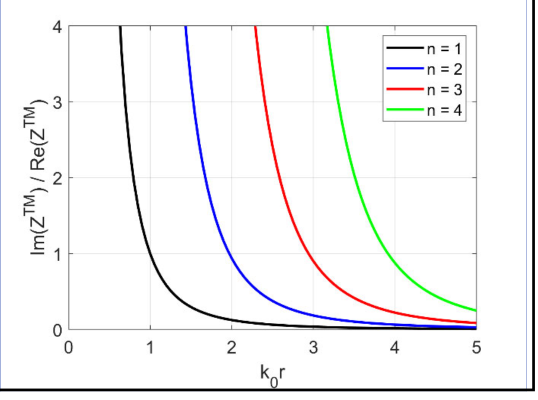
Practical gain limitation

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

$$G = N^2 + 2N$$

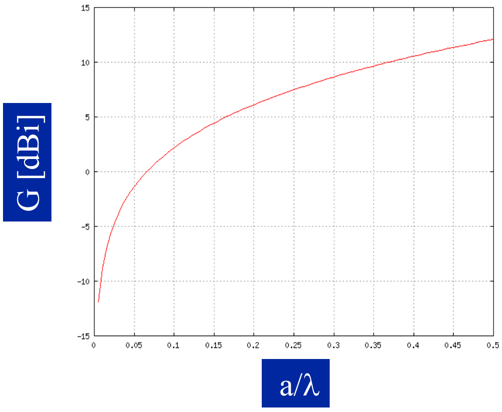
Wave impedance of a TM wave

$$Z_{+r}^{TM} = \frac{j\eta}{kr} + \frac{\eta}{\left|h_n^{(2)}\right|^2}\left[\frac{2}{\pi kr} + j\left(j_n j_n^{'} + n_n n_n^{'}\right)\right]$$

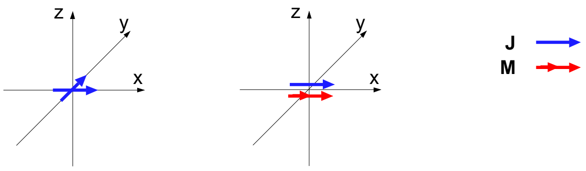


Maximum gain for a practical bandwidth : N = ka

$$G_{normal} = (ka)^2 + 2ka$$



Combination of elementary sources



Depending on combination od type of source, orientation and phases, we obtain

- LP or CP
- Q_0 or $Q_0/2$
- G=1.5 or G=3

See: D. Pozar, New Results for Minimum Q, Maximum Gain, and Polarization Properties of Electrically Small Arbitrary Antennas, EuCAP 2009

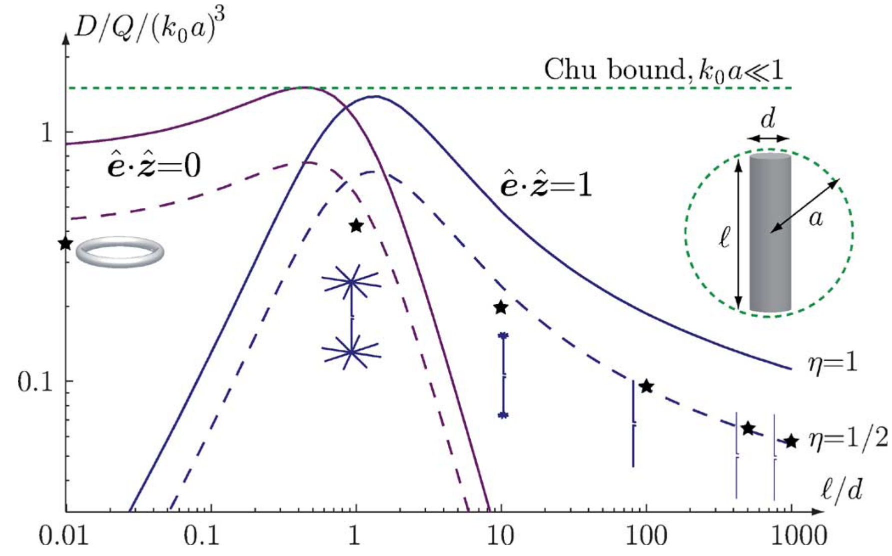
Comparison with measured gains

Circular parabolic reflector antenna:
Size 146 λ, $G_{measured}$: 50.4 dBi, G_{max}: 53.3 dBi
Pyramidal horn antenna:
Size 7.5 λ, $G_{measured}$: 24.5 dBi, G_{max}: 27.7 dBi
Narda horn antenna:
Size 2.5 λ, $G_{measured}$: 15-16 dBi, G_{max}: 18.7 dBi
Rolled slot antenna:
Size 0.2 λ, $G_{measured}$: -11.7 dBi, G_{max}: 2.6 dBi
Slot-Dipole antenna:
Size 0.2 λ, $G_{measured}$: 0 dBi, G_{max}: 2.6 dBi

limitations including antenna form factor

- works by several authors, the most interesting by Gustafsson et al.
- Mats Gustafsson, Christian Sohl, Gerhard Kristensson, Physical limitations on antennas of arbitrary shape, Proceedings of the Royal Society of London A: Mathematical, Physical and Engineering Sciences, vol 463, pp. 2589-2607, 2007
- Mats Gustafsson, Christian Sohl, Gerhard Kristensson, Illustrations of new physical bounds on linearly polarized antennas, IEEE transactions on antennas and propagation, vol. 59, 2009, pp. 1319-1327

example



from IEEE Trans. on AP, VOL. 57, NO. 5, MAY 2009, Gustafsson et al.

IMPLANTBLE ANTENNAS

EPFL

System Requirements

- Data transmission
- Wearer comfort
  - large autonomy => Low power consumption
  - small volume, conformable, flexible
  - sufficient reading distance
- Wearer health
  - avoid battery if possible (implants)
  - biocompatible encapsulation(implants)
  - emission values have to be respected
  - max SAR has to be respected
  - high reliability (implants)

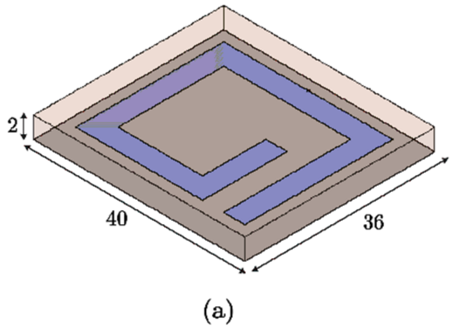
EPFL

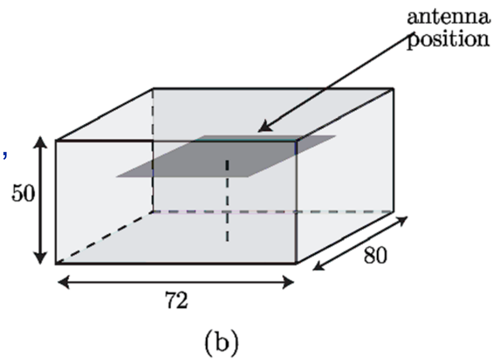
Antenna requirements

- Physically small => electrically very small @ MedRad, TETRA or ISM bands
- Enough bandwidth for the required data transmission
- Good radiation efficiency

We want to maximize the power radiated out/away of the body

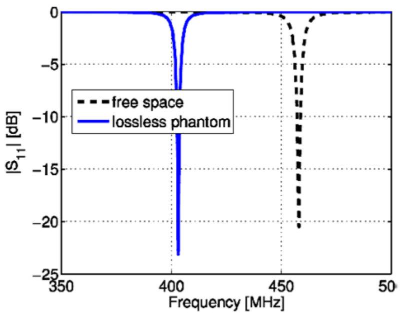
Antennas radiating into a lossy medium

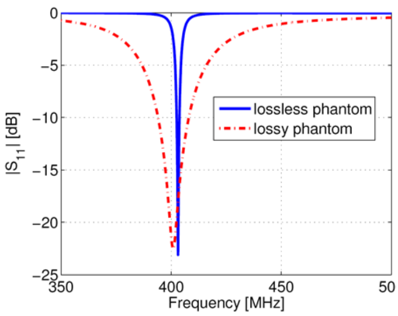




generic antenna for MedRadio, derived from design in:
J.Kim and Y. Rhamat-Samii, IEEE Trans. MTT, vol. 52. pp. 1934-1943, 2004

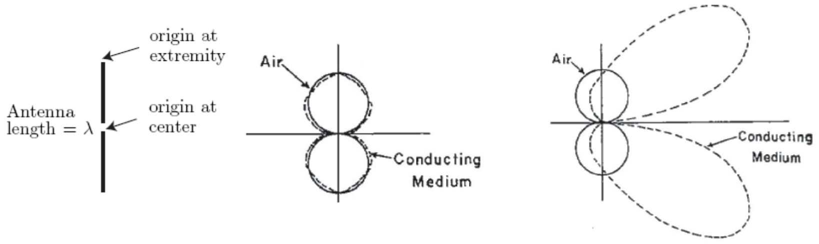
Antennas radiating into a lossy medium : effect on bandwidth





What is the meaning of the bandwidth ?

Antennas radiating into a lossy medium : effect on the pattern



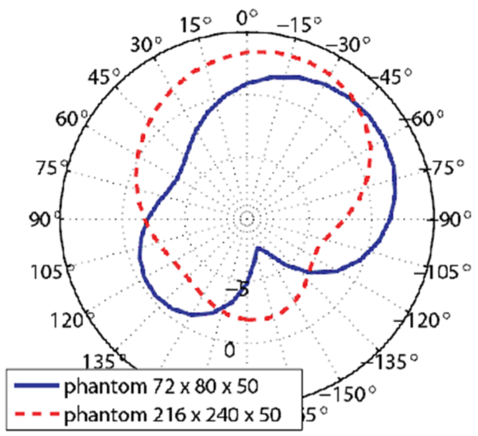
Antenna with uniform curent

Origin at center

Origin at top

R. Moore, “Effects of a surrounding conducting medium on antenna analysis,” IEEE Trans. AP., vol. 11, no. 3, pp. 216–225, May 1963

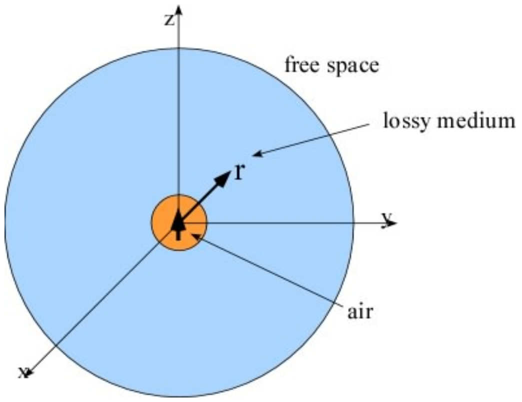
Antennas radiating into a lossy medium : effect on the pattern



In the case of our implantable antenna

What is the meaning of the radiation pattern ?

Antennas radiating into a lossy medium : Definition of efficiency ?



$$P_{Rad}^{TE} \sim \frac{1}{r^3}$$

$$P_{Rad}^{TM} \sim \frac{1}{r}$$

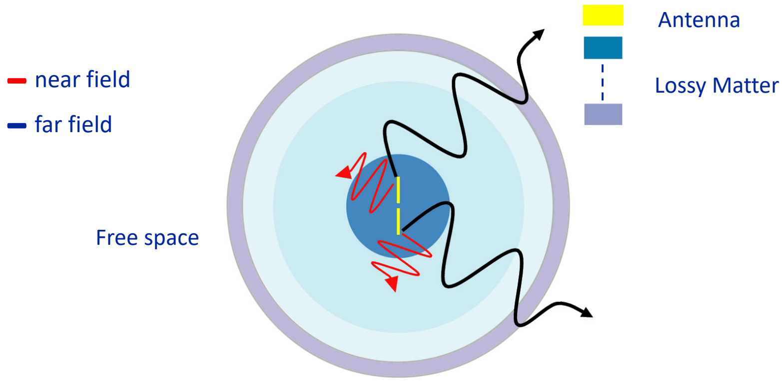
In the case of an implanted antenna : $\eta_{Rad} = \frac{P_{Rad}|_{\text{free space}}}{P_{Source}}$

Depends on the host body !!!

Main issues

- We have an electrically small antenna problem
- But : the antenna radiates into a lossy medium first, then into free space
- An insulation layer is required between the antenna and the lossy medium
- How does this modify our design strategy from a classical electrically small antenna design ?
- How does this affect the antenna characterization ?
- What is an adequate model of the host body ?
- What implications do the safety issues have ?

Antennas in Lossy Matter



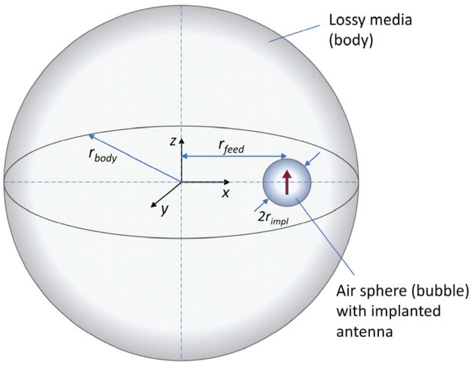
- Strong coupling between the near-field and the surrounding materials
- Attenuation of the far field propagating in the lossy dielectrics

A better understanding is needed

THE SPHERICAL WAVE CANONICAL MODEL

Canonical Implant: the elementary dipole

- Electrically small source is located inside the surface of the body (spherical medium which can be single- or multi-layered)



***Single-layer* @ 403.5 MHz**
IEEE Head model… ε_r = 43.5 - j34.75

***Multi-layer* @ 403.5 MHz**
Muscle… 82mm ε_r = 57.10 – j35.51
Fat… 86 mm ε_r = 5.58 – j1.83
Dry skin… 90 mm ε_r = 46.7 –j30.72

Air sphere r_{impl} = 1 mm (unless written otherwise)

M. Bosiljevac, Z. Sipus, and A. K. Skrivervik, "Propagation in Finite Lossy Media: An Application to WBAN"

Spherical mode expansion

- We can express the EM field via spherical modes:

$$\mathbf{E} = -\sum_n \sum_m a_{mn}\mathbf{M}_{mn} + b_{mn}\mathbf{N}_{mn} \qquad \mathbf{H} = -\frac{j}{\eta}\sum_n \sum_m b_{mn}\mathbf{M}_{mn} + a_{mn}\mathbf{N}_{mn}$$

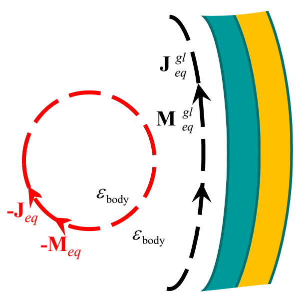
$$\mathbf{M}_{mn} = \nabla \times \mathbf{r}\psi_{mn} \qquad \mathbf{N}_{mn} = \frac{1}{k}\nabla \times \mathbf{M}_{mn} \qquad \psi_{mn}(r,\theta,\phi) = z_n(kr)P_n^m(\cos\theta)e^{jm\phi}$$

- The solution scheme:

$$\begin{Bmatrix} \tilde{J}_{eq}(r'_{impl},\nu,\mu) \\ \tilde{M}_{eq}(r'_{impl},\nu,\mu) \end{Bmatrix} \Rightarrow \begin{Bmatrix} \tilde{E}_{eq}(r',\nu,\mu) \\ \tilde{H}_{eq}(r',\nu,\mu) \end{Bmatrix}$$

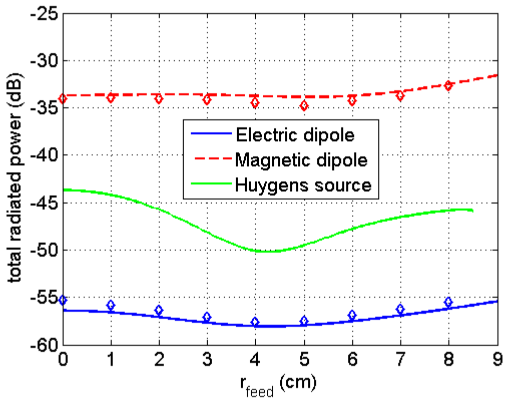
$$\Rightarrow \begin{Bmatrix} \sum_{n,m} \tilde{E}_{eq}(r,n,m) \\ \sum_{n,m} \tilde{H}_{eq}(r,n,m) \end{Bmatrix} \Rightarrow \begin{Bmatrix} \sum_{n,m} \tilde{J}_{eq}(r_{eq},n,m) \\ \sum_{n,m} \tilde{M}_{eq}(r_{eq},n,m) \end{Bmatrix}$$

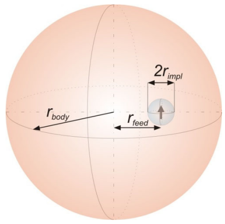
$$\Rightarrow \begin{Bmatrix} \sum_{n,m} \tilde{E}_{scat}(r,n,m) \\ \sum_{n,m} \tilde{H}_{scat}(r,n,m) \end{Bmatrix} \Rightarrow \begin{Bmatrix} \sum_{\nu,\mu} \tilde{E}_{scat}(r'_{impl},\nu,\mu) \\ \sum_{\nu,\mu} \tilde{H}_{scat}(r'_{impl},\nu,\mu) \end{Bmatrix}$$



Implanted case – Influence of the position of the antenna @ 403.5 MHz
- dipole is moved parallel to the boundary

- Total radiated power as a function of the source position
  → comparison with Huygens source

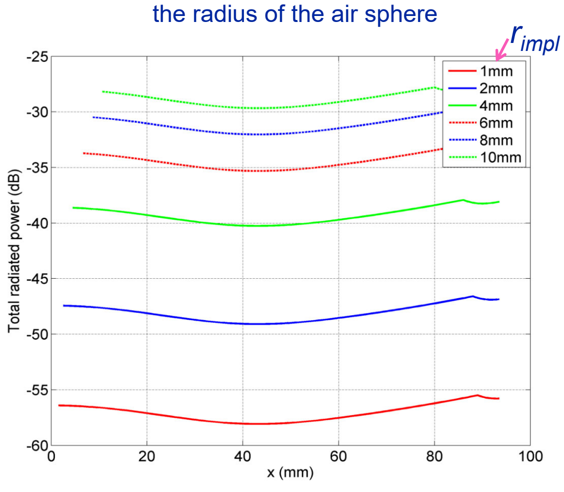


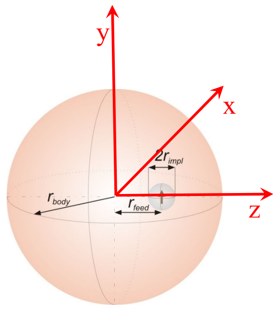


r_{impl} = 1 mm
r_{body} = 90 mm
$\varepsilon_{r,body}$ = 43.5-j34.75

Implanted case – Influence of the position and the size of the air sphere @ 403.5 MHz

- Dipole is moved parallel to the boundary (y-polarization, moved in x or z direction)
- Total radiated power as a function of the source position and the radius of the air sphere

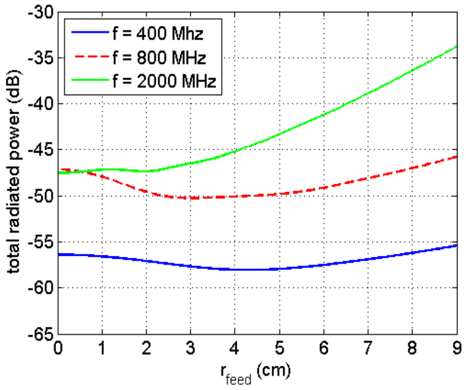


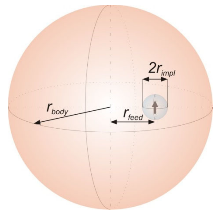


r_{body} = 90 mm
$\varepsilon_{r,body}$ = 43.5-j34.75
f = 403.5 MHz

Implanted case – Influence of the position and the size of the air sphere

- Dipole is moved parallel to the boundary (y-polarization, moved in x or z)
- Total radiated power as a function of the frequency and of the source position





f = 403.5 MHz

ε_{body} = 43.50 – j34.75

f = 800 MHz

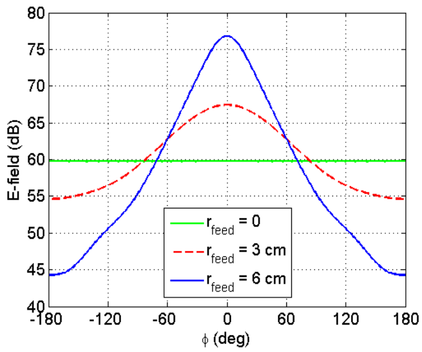
ε_{body} = 41.5 - j20.22

f = 2000 MHz

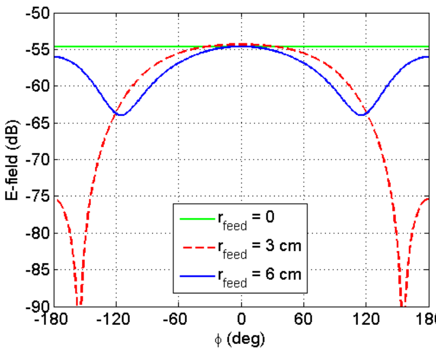
ε_{body} = 40.0 - j12.58

Implanted case – Near and far field outside the phantom

- Dipole is in parallel to the interface
- f=403 MHz



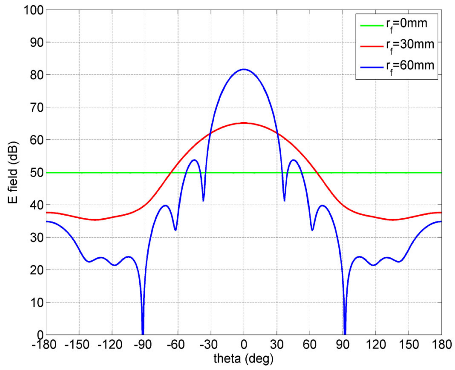
Electric field at a distance of 10 cm of the centre of the phantom versus the φ angle.



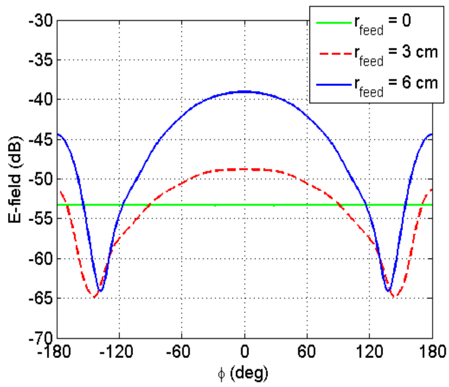
Electric far field versus the φ angle, for an electric source at 403 MHz

Implanted case – Near and far field outside the phantom

- Dipole is parallel to the interface
- f=2.45 GHz

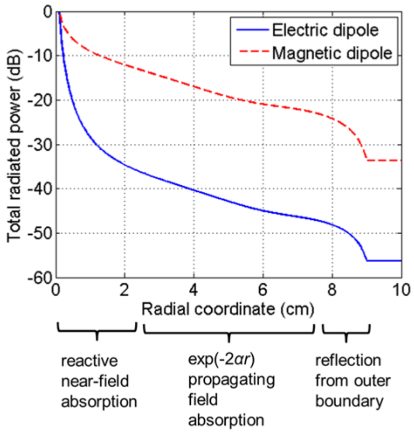


Electric field at a distance of 10 cm of the centre of the phatonm versus the φ angle.



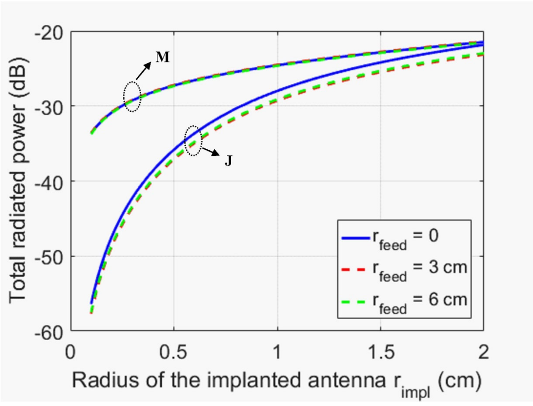
Electric far field versus the φ angle, for an electric source at 403 MHz

Decrease of power due to the lossy medium



f: 403 MHz
r_{body}: 9cm
r_{impl}: 1mm
$\varepsilon_{r,body}$ = 43.5-j34.75

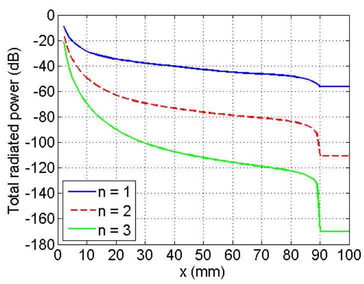
Effect of near field coupling to the lossy medium



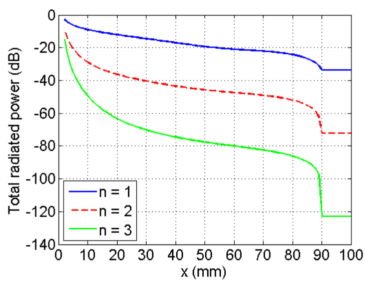
f: 403 MHz
r_{body}: 9cm
$\varepsilon_{r,body}$ = 43.5-j34.75

Effect of spherical modes

f: 403 MHz
r_{body}: 9cm
r_{impl}: 1mm
$\varepsilon_{r,body}$ = 43.5-j34.75



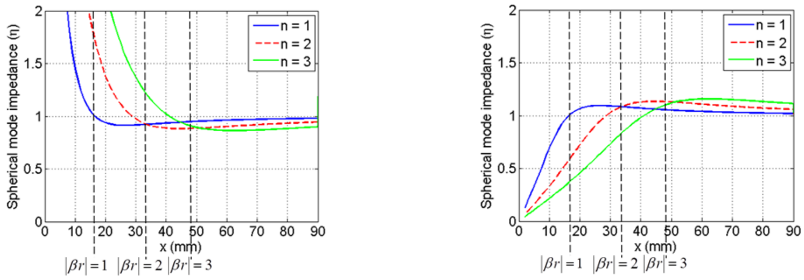
Electric source



Magnetic source

Impedance of spherical modes

f: 403 MHz
r_{body}: 9cm
r_{impl}: 1mm
$\varepsilon_{r,body}$ = 43.5-j34.75



Electric source

Magnetic source

The outer radius of the reactive near-field region of the *n*th spherical harmonic is defined with:

$$|\beta r| \approx n$$

β – propagation constant in lossy media

Desing rules

- The losses of an implantable antenna will depend on three main contributions: the near field losses, the propagating field losses, and the reflection at the body/air interface.
- The near field losses depend on the type of the radiating source (electric or magnetic) and the electric size of the lossless encapsulation around the antenna
- For an electrically small encapsulation, a magnetic type of antenna is more favorable than an electric one
- In order to avoid unnecessary losses, the antenna should only excite modes with order n < kr_{impl}, to keep the near fields as much as possible in the lossless encapsulation of the antenna. Indeed, as seen in Figure 11, the distance of influence of the near fields increases with the order of the mode.

First approximate bounds obtained using SWE

$$W_{\substack{\text{reaching}\\ \text{free}\\ \text{space}}} = W_{\substack{\text{entering}\\ \text{the}\\ \text{body}}} \cdot \frac{r_{impl}^2}{\Delta^2} \cdot e_{\substack{\text{losses}\\ \text{in the}\\ \text{reactive}\\ \text{near-field}}} \cdot e_{\substack{\text{propagating}\\ \text{field}\\ \text{absorption}\\ \text{losses}}} \cdot e_{\substack{\text{losses}\\ \text{due to}\\ \text{reflections}}}$$

$$e_{\substack{\text{propagating}\\ \text{field}\\ \text{absorption}\\ \text{losses}}} = \exp(-2\alpha(\Delta - r_{impl}))$$

$$e_{\substack{\text{losses}\\ \text{due to}\\ \text{reflections}}} = \frac{\mathrm{Re}\left\{|T|^2 / Z_{air}\right\}}{\mathrm{Re}\left\{1/Z_{body}\right\}}, \qquad T = \frac{2Z_{air}}{Z_{air}+Z_{body}}$$

$$e_{\substack{\text{losses}\\ \text{in the}\\ \text{reactive}\\ \text{near-field}}} = \frac{f_1(\Delta)}{f_1(r_{impl})},$$

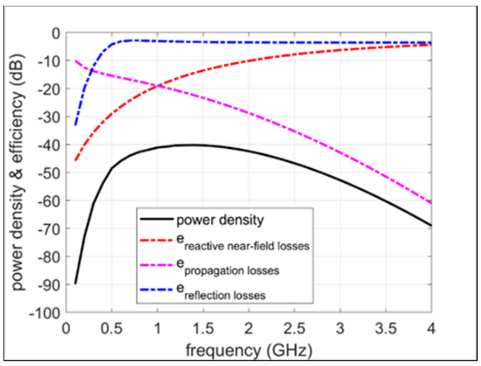
$$f_1(r) = \mathrm{Re}\left\{\eta \cdot \left(|k|^2 + \frac{2\alpha}{r} + \left(1 - \frac{k^*}{k}\right)\frac{1}{r^2} - j\frac{1}{kr^3}\right)\right\}$$

Electric sources

$$e_{\substack{\text{losses}\\ \text{in the}\\ \text{reactive}\\ \text{near-field}}} = \frac{f_2(\Delta)}{f_2(r_{impl})}, \qquad f_2(r) = |k|^2 + 2\alpha / r$$

Magnetic sources

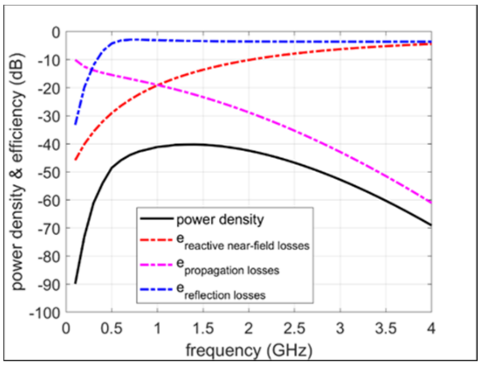
Application: homogeneous muscle phantom

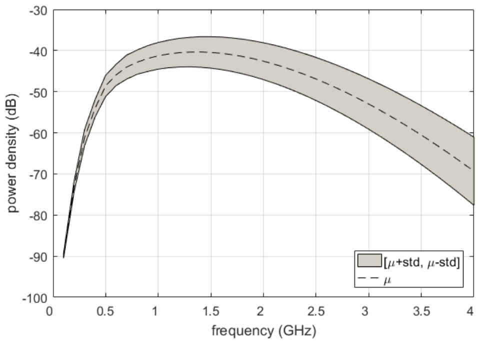


- At low frequencies near field losses dominate
- At high frequencies propagating losses dominate

R_{impl}= 1 mm
R_{body}= 90 mm

Comparison to uncertainties due to tissue parameters





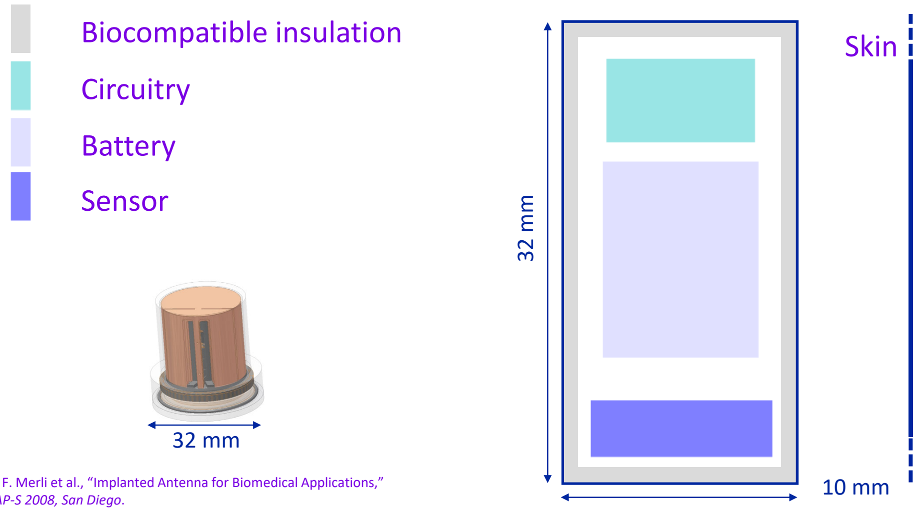
Permittivity is varied by 20 %

dual band antenna : data transmission @ 401-406 MHz, wake up signal @ 2.45 GHz

Figure of merit: maximize reading distance

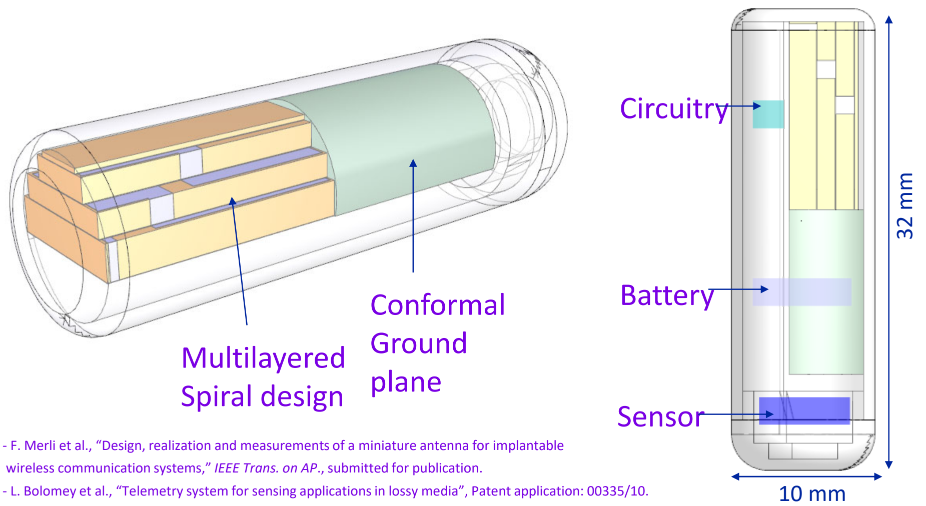
EXAMPLE1 : THE DESIGN OF ANTENNA FOR AN IMPLANTABLE GENERIC BODY MONITORING MODULE

Antenna conception



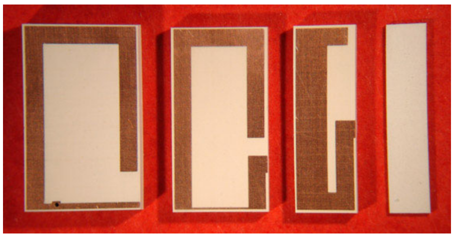
- F. Merli et al., "Implanted Antenna for Biomedical Applications," *AP-S 2008, San Diego*.

Antenna Structure



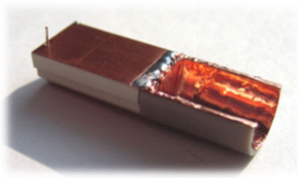
- F. Merli et al., "Design, realization and measurements of a miniature antenna for implantable wireless communication systems," *IEEE Trans. on AP.*, submitted for publication.
- L. Bolomey et al., "Telemetry system for sensing applications in lossy media", Patent application: 00335/10.

Antenna Realization



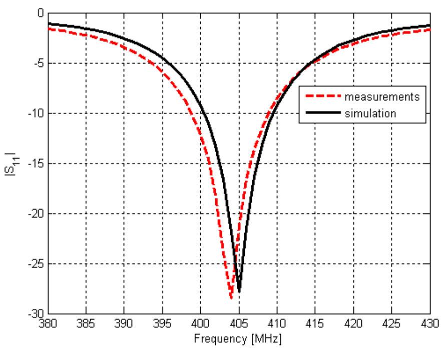
Substrate: Roget TMM (ε_r = 9.2)

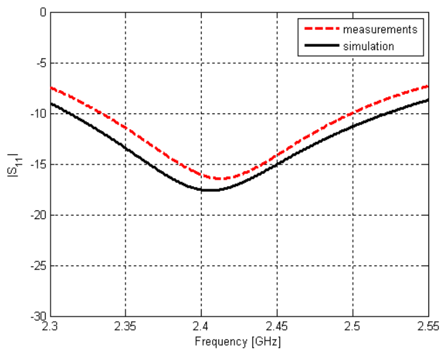
Insulation: PEEK (ε_r = 3.2)



Antenna Matching measurement (*in-vitro*)

EM performances of the antenna alone have been checked with a feeding coaxial cable (present only for testing



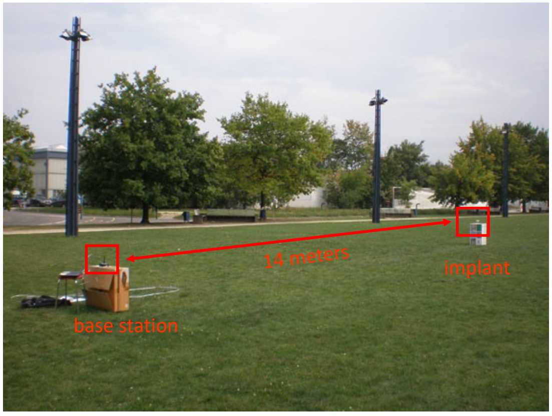


System Measurements *In-vitro*

Zarlink BS is always considered
System controlled via a laptop (*Labview*)

Outdoor MedRadio Tests:
- TX power -3 dBm

channel	max range [m]
0	7
4	14
9	14



System *In-vivo*

Implantation (*in collaboration with* the Stem Cell Dynamics Laboratory, LDCS):

Two devices have been implanted at different locations,

- subcutaneous (5 mm)

Target: - in muscle tissue (30 mm)

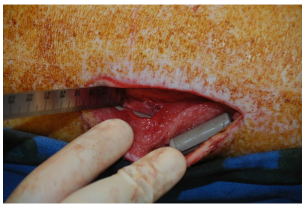
Continuous monitoring of subcutaneous temperature of a porcine animal (= pig)

Characteristics:

- Measurement during the implantation procedure
- Temperature check every 5 min. Complete working cycle (*wake-up, measurement, transmission…)* for 15 days

EPFL

System *In-vivo*



Implantation in accordance to all ethical considerations and the regulatory issues related to animal experiments.

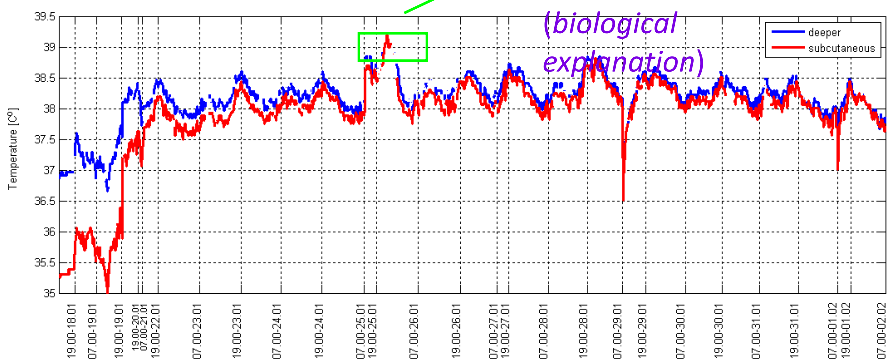
In vitro sensor for room temperature comparison

EPFL

System Measurements *In-vivo*

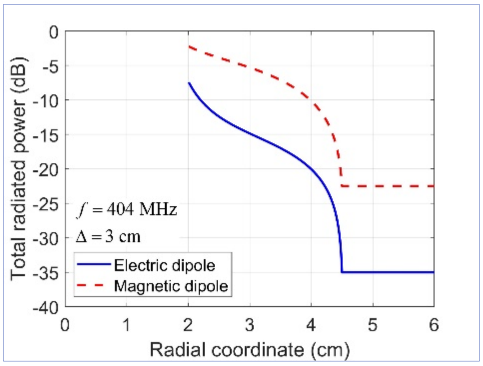
Highest relative error

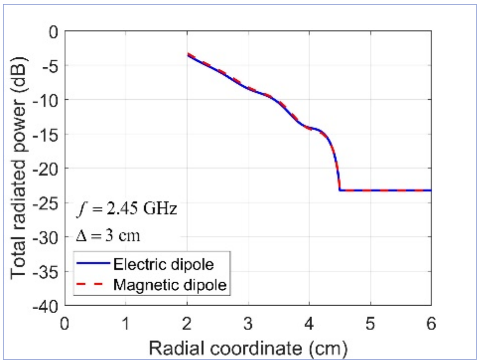
(*biological explanation*)



Comparison SWE with real antenna results

Considering r_{impl} = 5mm and r_{body} = 45 mm





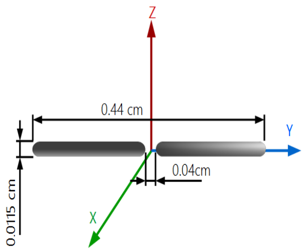
Full wave simulation
Antenna efficiency: -32.6 dB

In vitro measurements:
Antenna efficiency: -33 dB

Full wave simulation
Antenna efficiency: -20 dB

In vitro measurements:
Antenna efficiency: -21 dB

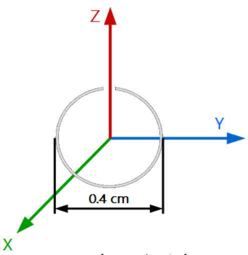
Examples: Real (but ideal) matched antennas



Electrical dimensions

ISM band:
λ/30

MedRadio band:
λ/185



$$X_{in} \approx -120 \frac{\lambda}{\pi D}\left(\ln\left(\frac{D}{2a}\right) - 1\right)$$

$$X_{in} \approx 120 \frac{2r\pi^2}{\lambda}\left(\ln\left(\frac{r}{a}\right)\right)$$

Matched with **inductance** lumped element:

$$L = -\frac{X_{in}}{2\pi f}$$

Matched with **capacitance** lumped element:

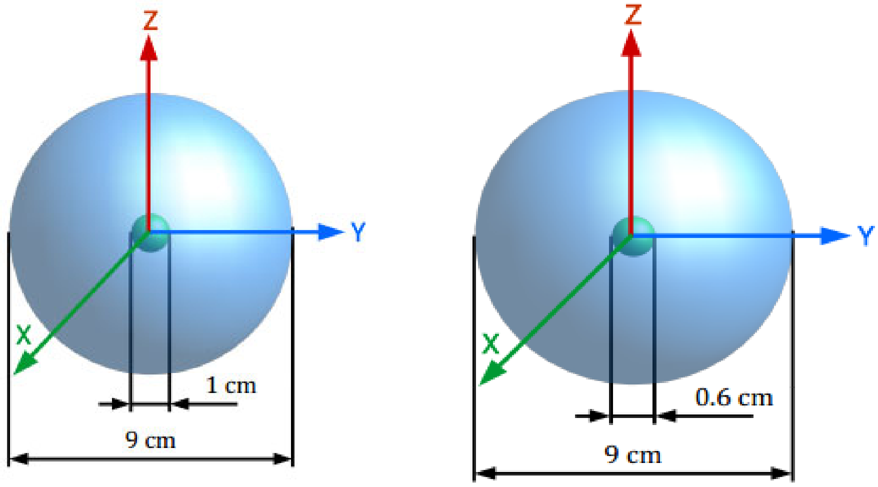
$$C = \frac{1}{X_{in} 2\pi f}$$

K. T. McDonald, "Reactance of Small Antennas," 2012

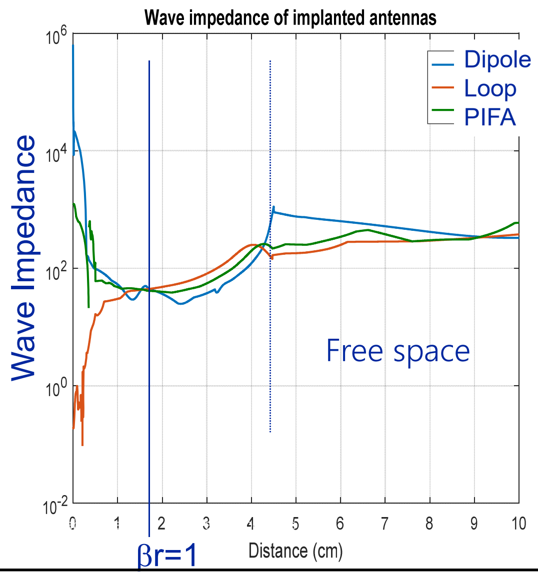
Compared antennas: the dipole and loop

Dipole

Loop

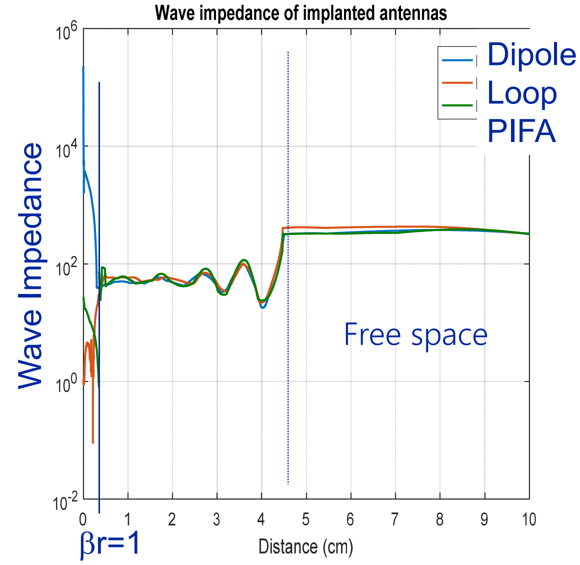


Analysis of wave impedance



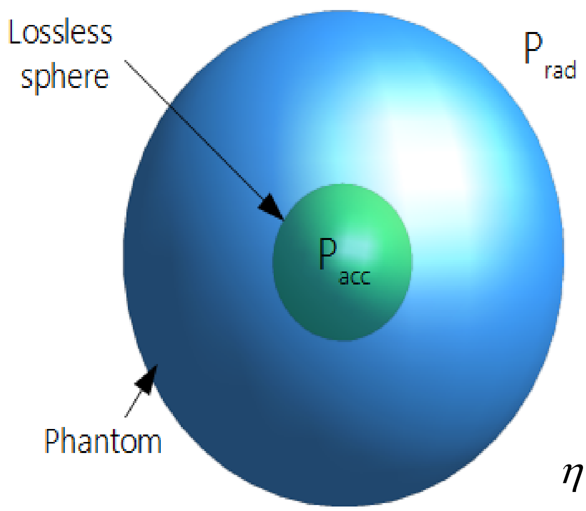
@MedRadio
(403MHz)

Analysis of wave impedance



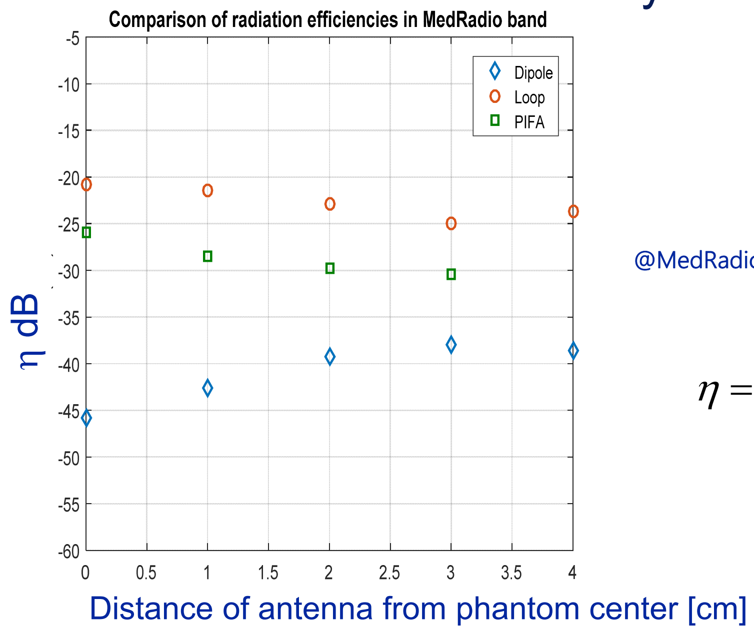
@ISM (2.45GHz)

Radiation efficiency



$$\eta = \frac{P_{rad}}{P_{acc}}$$

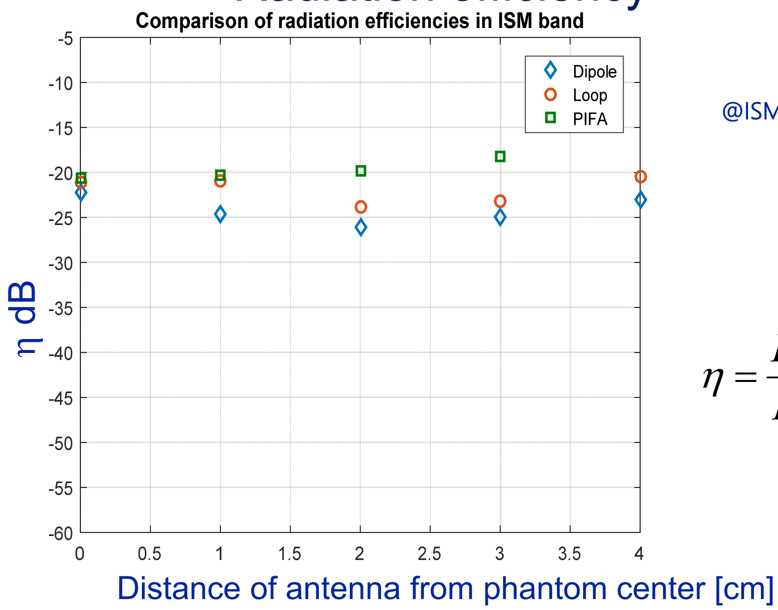
Radiation efficiency



@MedRadio (403 MHz)

$$\eta = \frac{P_{rad}}{P_{acc}}$$

Radiation efficiency



@ISM (2.45GHz)

$$\eta = \frac{P_{rad}}{P_{acc}}$$

EPFL

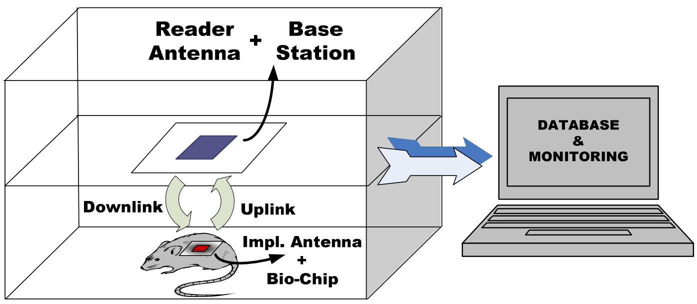
Remote powered antenna @ ISM 2,45 GHz Band

Figure of Merit: comply to reulatory safety issues

EXAMPLE 2: ANTENNA FOR A GENERIC IMPLANTS FOR RODENTS (MICE)

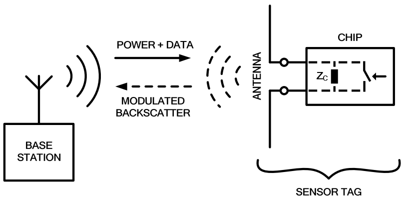
EPFL

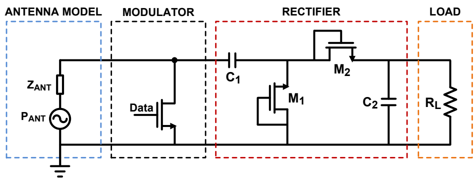
The problem



- Remote powering
- Short reading distance (5-10 cm)
- Small volume

The system





- Regulation issues, Base station :
  - Max EIRP (EU RFID regulation): 27dBm
  - Max Re[S] at mouse position: 10 (50) W/m^2
  - Max field level at mouse position: 87 (193) V/m
- From this we obtain the reading distance:

$$r = \sqrt{\frac{EIRP}{4\pi\,\mathrm{Re}[S]}} = 6cm$$

The system

We would like a re-emmited power of Pout=0.8 dBm

$$P_{\text{out}} = \text{Real}\{S_{\text{in}}\}\ \frac{\lambda^2}{4\pi}\ G_{\text{rx}}\ \chi\ \tau\ \eta_{\text{rect}}$$

P_{out}= 1.2 mW (+0.8 dBm)
S_{in}= Poynting vector at the mouse surface
λ = 12.24 cm at 2.45 GHz
G_{rx} = -1.5 dBi gain of the implanted antenna LP (incl.body losses)
χ = 0.5 pol. mismatch (impl. antenna is LP)
τ= 0.7 power transmission coefficient (antenna to implant rectifier)
η_{rect}=0.2 rectifier efficiency

This implies $\boxed{\text{Real}\{S_{\text{in}}\} = 20.36\ \text{W}/m^2}$

$\boxed{\text{Regulatory compliant level } (\text{Real}\{S_{\text{in}}\} = 10\ \text{W}/m^2) \ \text{->}\ P_{\text{out}} = 0.588\ \text{mW (-2.3 dBm)}}$

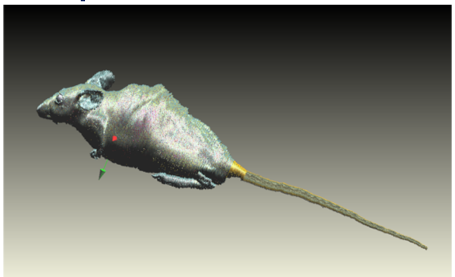
SAR and Temperature increase

B6F3Cl female Mouse, pregnant
28.7 g developed by IT'IS [5]
voxeled at 0.5 mm

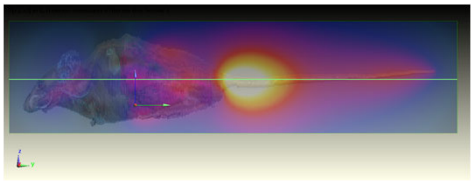
Whole Body SAR = 0.66 W/kg
(lim. 0.4 / 0.08)

10-g av. SAR = 0.78 W/kg
(lim. 10 / 2)

@ 120 s = max increase 0.151º



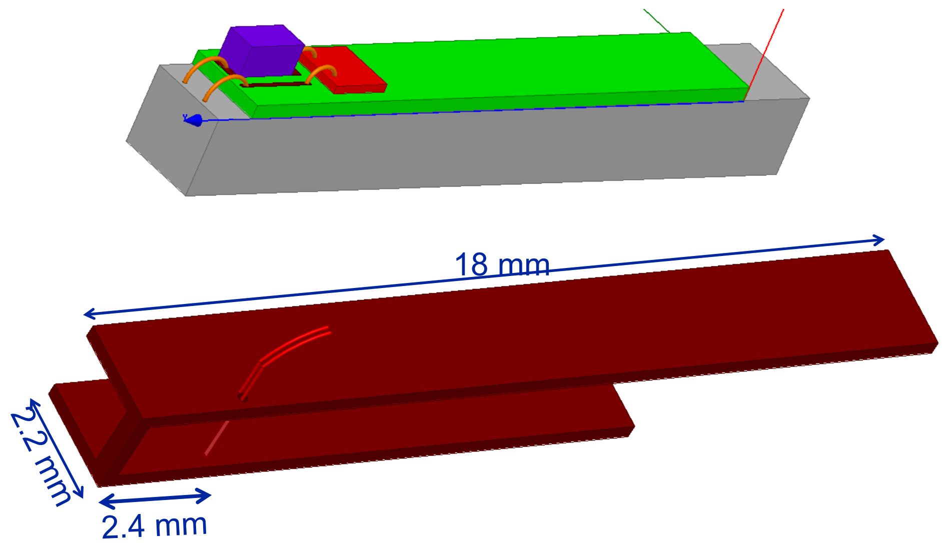
$\mathrm{Real}\{S_{\mathrm{in}}\} = 10\ \mathrm{W}/m^2$



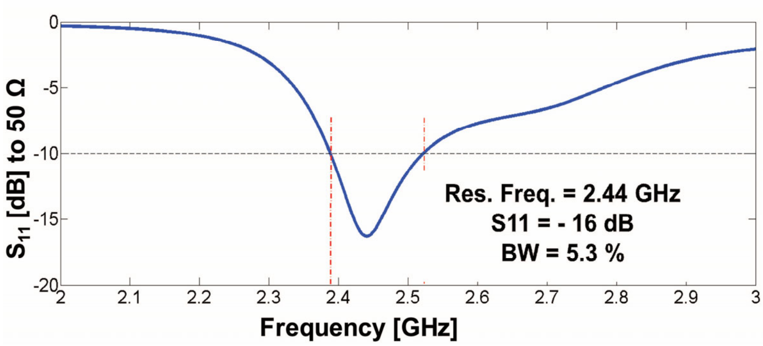
To ensure SAR limits, we have
P_{out}=-11.5 (-4.5) dBm

[5] ITIS FOUNDATION http://www.itis.ethz.ch/services/anatomical-models/animal-models/

The Implant and the antenna



Antenna characteristics



Efficiency including the mouse : -3 dB
Matched to 30-j250 Ω

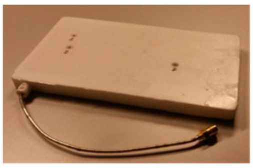
Wearable antennas

wearable antenna at 380 MHz for security communications

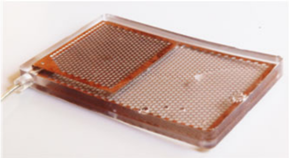
Figure of merit: maximize distance and robstness

EXAMPLE : THE SAR AND EFFICIENCY OF DIFFERENT WEARABLE TETRAPOL ANTENNAS

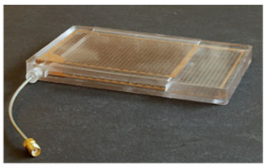
Antennas compared



a)



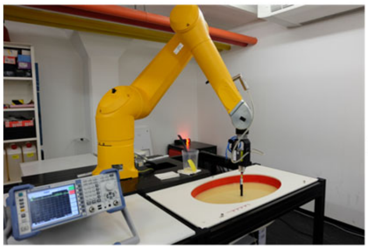
b)



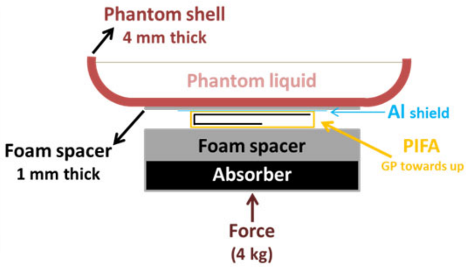
c)

d)

SAR measurement setup at IT'IS foundation, Switzerland



a)

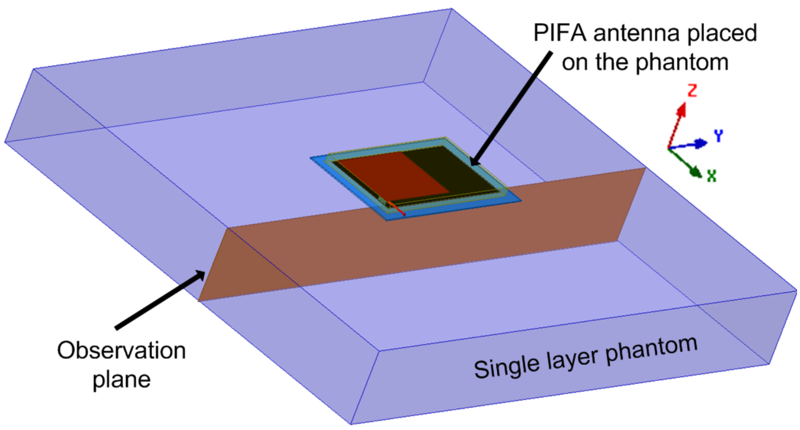


b)

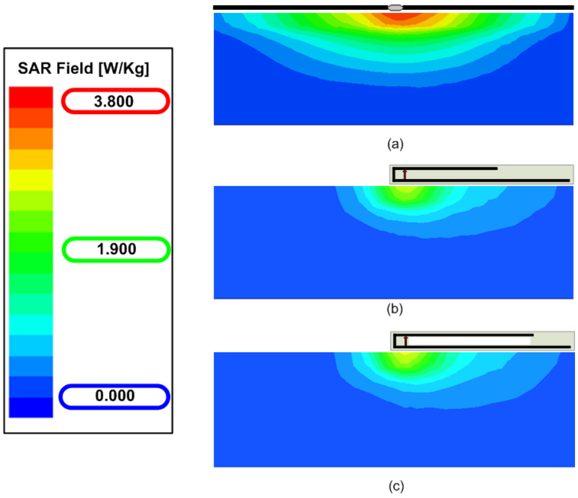
Limits

SAR [W/kg]	Averaging mass	Occupational exposure	
		Whole body	**Head/trunk**
IEEE/ICNIRP	10 g	0.4	**10**
FCC	1 g	0.4	**8**

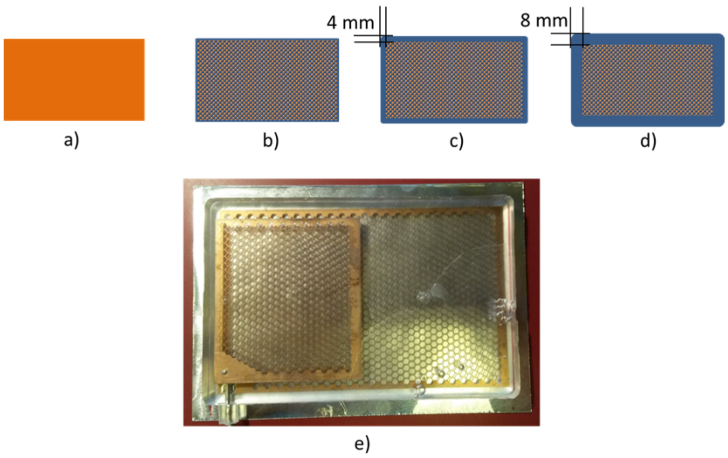
Simulated results



Simulated results



Different ground planes for the PIFA

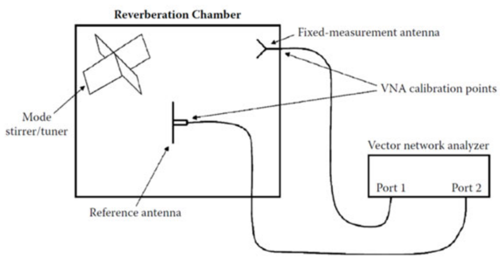


Measured results

	PIFA Hex-solid		PIFA Hex-Air Gap		PIFA Copper-mesh		Dipole Panorama	
Averag.	1g	10g	1g	10g	1g	10g	1g	10g
Normal	9.12	4.36	9.2	4.56	4.41	2.31	NA	7.82
Double GP	6.26	3.17	7.32	3.72	2.71	1.42	NA	NA
Ext. 4 mm	3.87	1.96	4.77	2.57	1.6	0.90	NA	NA
Ext. 8 mm	3.18	1.75	3.42	1.96	1.09	0.61	NA	NA

Measured efficiency

a)



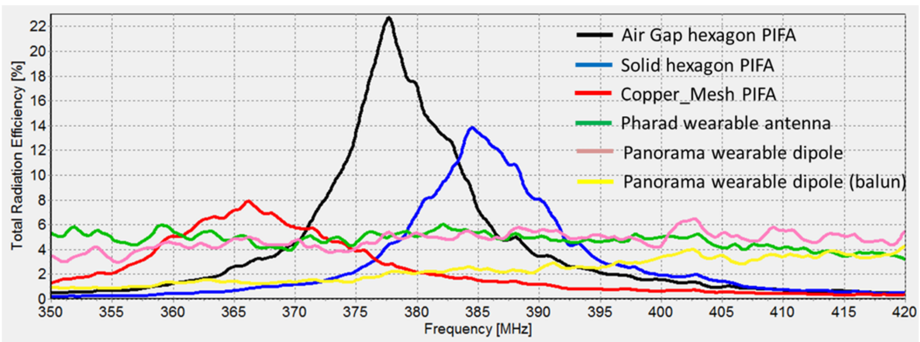
b)

done in a reverberation chamber at Bluetest

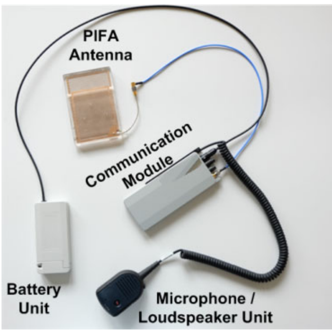
Measured efficiency

a) b) c)

Measured efficiency

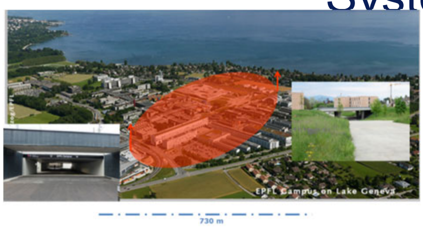


System measurements



Communication system used
(Courtesy of RUAG SA)

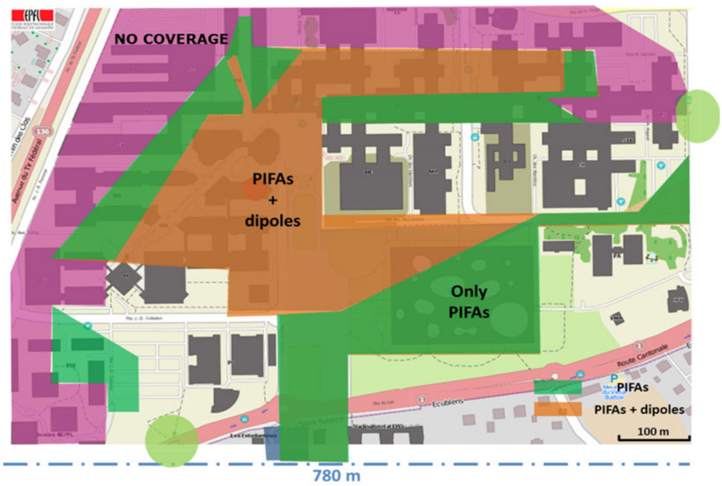
System measurement



An urban environment : EPFL campus

![EPFL]

System measurements



UWB,

Figure of merit: low profile, minimize coupling into the body

EXAMPLE 2 : THE DESIGN OF WEARABLE UWB ANTENAS FOR WBAN

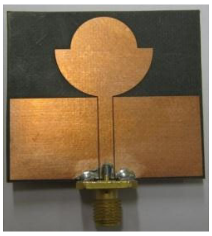
EPFL

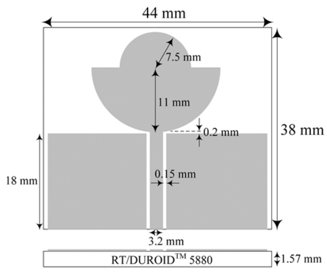
How to decouple an UWB antenna from the Body ?

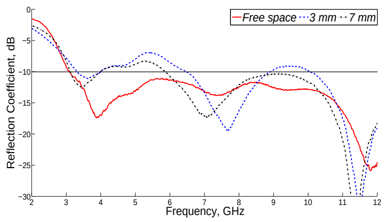
- Use ground planes
  - possible only for polarization orthogonal to the body.
  - for flat (printed) antennas, the ground plane is used to achieve the band width => cannot be beneath the antenna
- Use dielectrics to control the near field
- Select the right feed structure

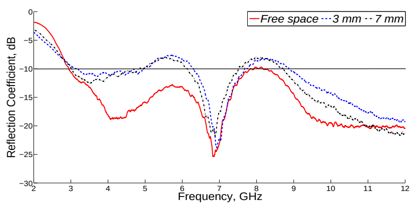
EPFL

Printed UWB antenna







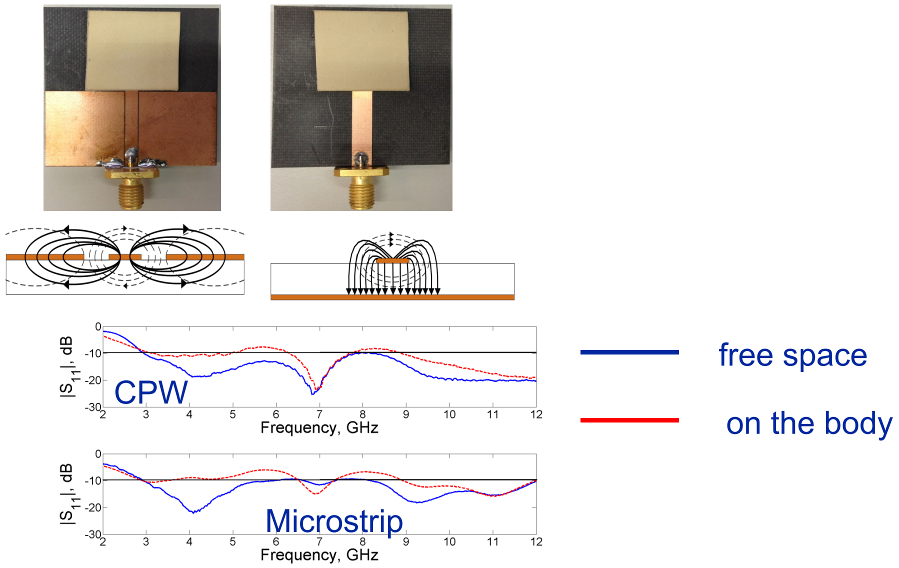


unloaded antenna

loaded antenna

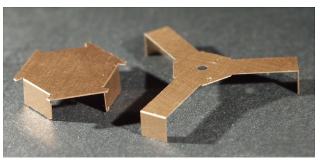
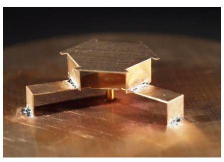
EPFL

Printed UWB antenna: different feed models



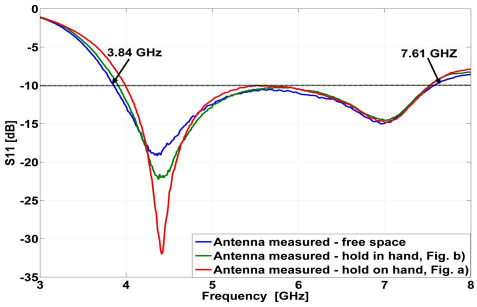
EPFL

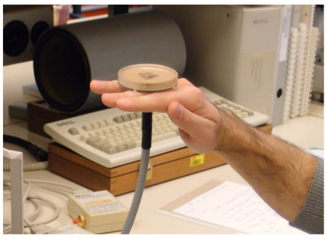
Tri-Pod Kettle Antenna

Etched and folded antenna parts	Assembled antenna with the mounted connector	TKA encapsulated in PDMS
		

h=8 mm
ø=12 mm

Tri-Pod Kettle Antenna





a)

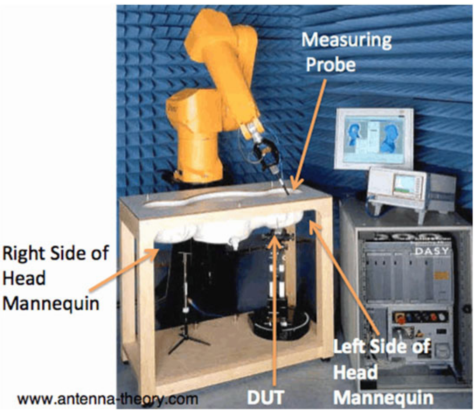
b)

Some legislation and rules

Standards and regulations

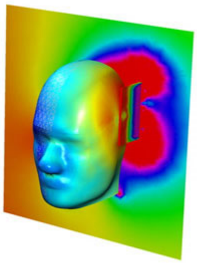
- Standarts are set to protect our health, e.g in
  - food additives
  - air or water pollutants
  - EM field levels
  - SAR levels
  - temperature increase
- Each country regulates its standarts
- Standarts are based on the latest sceintific knowledge: they may change

SAR measurement



DASY measurement system from SPEAG

Examples



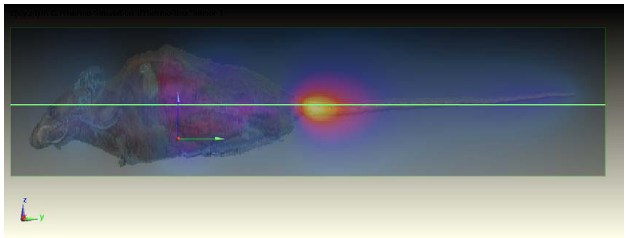


Table 1: Exposure limits for the general public for electromagnetic fields in inhabited areas in member states of the European Union and selected industrial nations outside the European Union (situation April 2011)

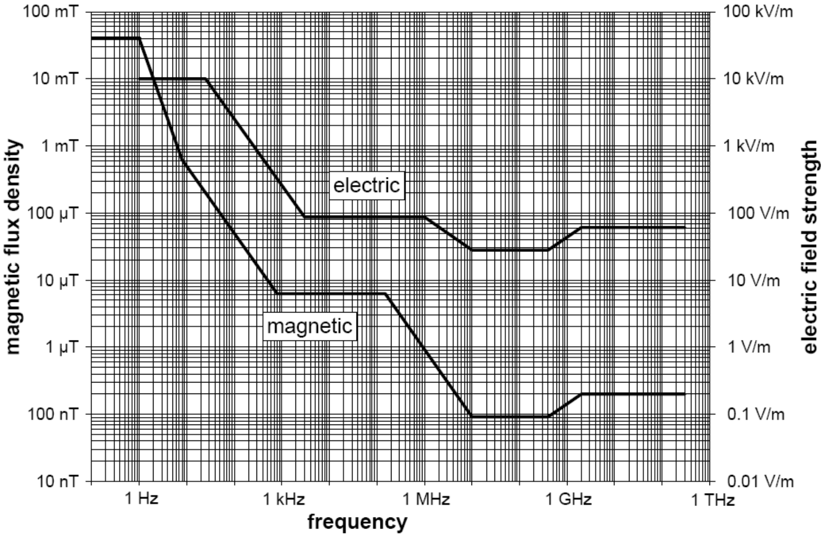
Country:	50 Hz (ELF) electric field strength (V/m)	50 Hz (ELF) magnetic flux density (µT)	900 MHz (GSM) electric field strength (V/m)	900 MHz (GSM) magnetic flux density (µT)	900 MHz (GSM) equivalent plain wave power density (W/m²)	1800 MHz (GSM) electric field strength (V/m)	1800 MHz (GSM) magnetic flux density (µT)	1800 MHz (GSM) equivalent plain wave power density (W/m²)	2100 MHz (UMTS) electric field strength (V/m)	2100 MHz (UMTS) magnetic flux density (µT)	2100 MHz (UMTS) equivalent plain wave power density (W/m²)
Recommendation 1999/519/EC	**5000**	**100**	**41**	**0.14**	**4.5**	**58**	**0.20**	**9**	**61**	**0.20**	**10**
Austria	[5000]	[100]	[41]	[0.14]	[4.5]	[58]	[0.20]	[9]	[61]	[0.20]	[10]
Belgium (Flanders)	—	10	21 (1	—	—	29 (1	—	—	31 (1	—	—
Bulgaria	– (2	– (2	—	—	0.1	—	—	0.1	—	—	0.1
Cyprus	[5000]	[100]	41	0.14	4.5	58	0.20	9	61	0.20	10
Czech Republic	5000	100	41	0.14	4.5	58	0.20	9	61	0.20	10
Denmark	– (3	– (3	—	—	—	—	—	—	—	—	—
Estonia	5000	100	41	0.14	4.5	58	0.20	9	61	0.20	10
Finland	[5000]	[100]	41	0.14	4.5	58	0.20	9	61	0.20	10
France	5000 (4	100 (4	41	0.14	4.5	58	0.20	9	61	0.20	10
Germany	5000	100	41	0.14	4.5	58	0.20	9	61	0.20	10
Greece	5000	100	32 (5	0.11 (5	2.7 (5	45 (5	0.15 (5	5.4 (5	47 (5	0.16 (5	6 (5
Hungary	5000	100	41	0.14	4.5	58	0.20	9	61	0.20	10
Ireland	[5000]	[100]	41	0.14	4.5	58	0.20	9	61	0.20	10
Italy	– (6	3 (6	6 (7	0.02 (7	0.1 (7	6 (7	0.02 (7	0.1 (7	6 (7	0.02 (7	0.1 (7
Latvia	—	—	—	—	—	—	—	—	—	—	—
Lithuania	500 (8	—	—	—	0.1	—	—	0.1	—	—	0.1
Luxembourg	5000 (9	100 (9	41 (10	0.14	4.5	58 (10	0.2	9	61 (10	0.20	10
Malta	[5000]	[100]	41	0.14	4.5	58	0.20	9	61	0.20	10
Netherlands	– (11	– (11	—	—	—	—	—	—	—	—	—
Poland	1000	75	7	—	0.1	7	—	0.1	7	—	0.1
Portugal	5000	100	41	0.14	4.5	58	0.20	9	61	0.20	10
Romania	5000	100	41	0.14	4.5	58	0.20	9	61	0.20	10
Slovakia	5000	100	41	0.14	4.5	58	0.20	9	61	0.20	10

http://ec.europa.eu/health/electromagnetic_fields/docs/emf_comparision_policies_en.pdf

Main regulators

- USA:
  - FCC (Federal communication commission)
  - FDA (Food and Drug Administration)
- EU:
  - ICNIRP (International Commission on Non-Ionizing Radiation Protection)
- Switzerland:
  - BAFU (Bundesamt für Umweltschutz), published the Ordinance on Non-Ionizing Radiation (ONIR)

Reference levels for the general population recommended by ICNIRP [1]



[1] Guidelines for limiting exposure to time-varying electric, magnetic and electromagnetic fields (up to 300 GHz) International Commission on Non-Ionizing Radiation Protection. Health Phys. 1998, 54, 115-123

Regulations USA

Limits depend on frequency allocation:
electrical field strength is evaluated at 3 meters
Equivalent Isotropic Radiated Power

$$\frac{|E_{3m}|^2}{Z_0} = \frac{|E_{3m}|^2}{120\pi} = \frac{P_{tx}G_{tx}}{4\pi 3^2}$$

$$EIRP = P_{tx}G_{tx}$$

Frequency	Electrical Field Strength	Corresponding EIRP
30 ... 88 MHz	100 μV/m	– 55.2 dBm
88 ... 216 MHz	150 μV/m	– 51.7 dBm
216 ... 960 MHz	200 μV/m	–49.2 dBm
> 960 MHz	500 μV/m	–41.2 dBm

Almost nothing could work with that! So there are exceptions…

[1] FCC Title 47, Part 15 (47 CFR 15) Rules and regulations regarding unlicensed transmissions
[2] Texas instruments Application Report SWRA048–May 2005: ISM-Band and Short Range Device Regulatory Compliance Overview

Regulations USA: the 2.4 -2.4835 GHz ISM band

Transmission Type	Fundamental		Harmonics	
	E at 3m	EIRP	E at 3m	EIRP
Frequency hopping		≥ 75 channels: +36 dBm < 75 channels: +30 dBm	20 dB below the peak in-band emission in any 100-kHz bandwidth	
Digitally spread		+36 dBm		
Other	50 mV/m	–0.23 dBm	500 μV/m	–41.23 dBm

max transmitted power 1 W (+30 dBm)

[1] FCC Title 47, Part 15 (47 CFR 15) Rules and regulations regarding unlicensed transmissions
[2] Texas instruments Application Report SWRA048–May 2005: ISM-Band and Short Range Device Regulatory Compliance Overview

Regulations EU

Limits depend on:
- *Frequency allocations (Non-Specific Short Range Devices SRD)*
- *Applications*

Frequency Band	ERP	Duty Cycle	Channel Bandwidth	Remarks
433.05 – 434.79 MHz	+10 dBm	<10%	No limits	No audio and voice
433.05 – 434.79 MHz	0 dBm	No limits	No limits	≤– 13 dBm/10 kHz, no audio and voice
433.05 – 434.79 MHz	+10 dBm	No limits	<25 kHz	No audio and voice
868 – 868.6 MHz	+14 dBm	< 1%	No limits	
868.7 – 869.2 MHz	+14 dBm	< 0.1%	No limits	
869.3 – 869.4 MHz	+10 dBm	No limits	< 25 kHz	Appropriate access protocol required
869.4 – 869.65 MHz	+27 dBm	< 10%	< 25 kHz	Channels may be combined to one high speed channel
869.7 -870 MHz	+7 dBm	No limits	No limits	
2400 – 2483.5 MHz	+7.85 dBm	No limits	No limits	Transmit power limit is 10-dBm EIRP

[2] Texas instruments Application Report SWRA048–May 2005: ISM-Band and Short Range Device Regulatory Compliance Overview
[3] ERC RECOMMENDATION 70-03 RELATING TO THE USE OF SHORT RANGE DEVICES (SRD) 2011

Regulations EU (RFID)

Frequency Band	Power	Spectrum access and mitigation requirement	Channel spacing	ECC/ERC Decision	Notes
a1 2446-2454 MHz	≤500 mW e.i.r.p.	No requirement	No spacing		
a2 2446-2454 MHz	>500 mW-4 W e.i.r.p	≤ 15% duty cycle FHSS techniques should be used	No spacing		Power levels above 500 mW are restricted to be used inside the boundaries of a building and the duty cycle of all transmissions shall in this case be ≤15 % in any 200 ms period (30 ms on /170 ms off).
b1 865.0-865.6 MHz	100 mW e.r.p.	No requirement	200 kHz		
b2 865.6-867.6 MHz	2 W e.r.p.	No requirement	200 kHz		
b3 867.6-868.0 MHz	500 mW e.r.p.	No requirement	200 kHz		

duty cycle? Spread spectrum? Indoor/outdoor?
Because otherwise we should not transmit more than 500 mW (+27 dBm)…

[2] Texas instruments Application Report SWRA048–May 2005: ISM-Band and Short Range Device Regulatory Compliance Overview
[3] ERC RECOMMENDATION 70-03 RELATING TO THE USE OF SHORT RANGE DEVICES (SRD) 2011

Regulation ISM band summary

- **We can transmit up to EIRP = 4 W (+36 dBm) in both the USA and EU if**
  - *Tx < 15% Duty cycle*
  - *Spread spectrum (more than 75 channels or digital)*
  - *indoor*
- **We can transmit up to EIRP = 1 W (+30 dBm) in the USA if**
  - *Spread spectrum (less than 75 channels or digital)*
- **We can always transmit up to EIRP = 0.5 W (+27 dBm) in the EU**
- **We can always transmit up to EIRP = 0.00095 W (-0.23 dBm, |E|=50 mV/m) in the USA**

Regulations on SAR

Table 3: ICNIRP SAR limits (100 kHz – 10 GHz)

	Whole-body average SAR (W/kg)	Localized SAR (Head and Trunk) (W/kg)	Localized SAR (Limbs) (W/kg)
Occupational Exposure	0.4	10	20
General Public Exposure	0.08	2	4

1. All SAR limits are to be averaged over any six-minute period.
2. Localized SAR averaging mass is any 10 g of contiguous tissues; the maximum SAR so obtained should be the value used for the estimation of exposure.

Table 4: ANSI/IEEE SAR limits (100 kHz – 6 GHz)

	Whole-body average SAR (W/kg)	Localized SAR (Head and Trunk) (W/kg)	Localized SAR (Limbs) (W/kg)
Occupational Exposure	0.4	8	20
General Public Exposure	0.08	1.6	4

1. For occupational exposure, the SAR limits are averaged over any six-minute interval.
2. For general public exposure, the averaging time for SAR limits varies from six minutes to 30 minutes.
1. Whole-body SAR is averaged over the entire body, partial-body SAR is averaged over any 1 g of tissue defined as a tissue volume in the shape of a cube. SAR for hands, wrists, feet and ankles is averaged over any 10 g of tissue defined as a tissue volume in the shape of a cube.

 https://arquivo.splat-n.com/engineering/comparison-of-radio-frequency/

Measurement

Outline

- What is the problem
- Baluns
- Wheeler cap method
- System measurements
  - Reverberation chamber
  - Anechoic chamber

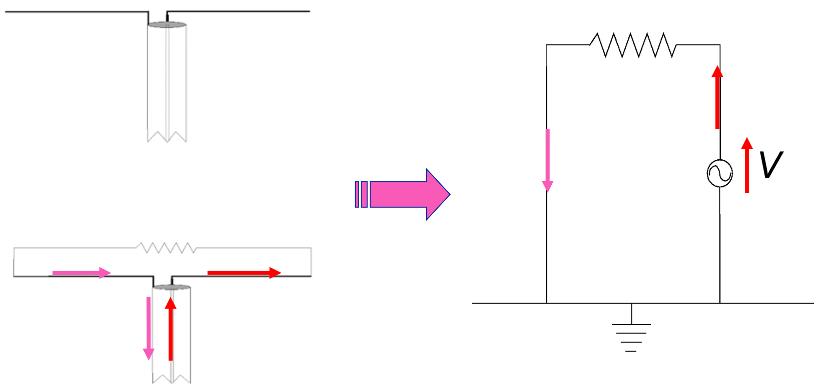
relevant characteristics

- bandwidth
- max gain
- radiation efficiency

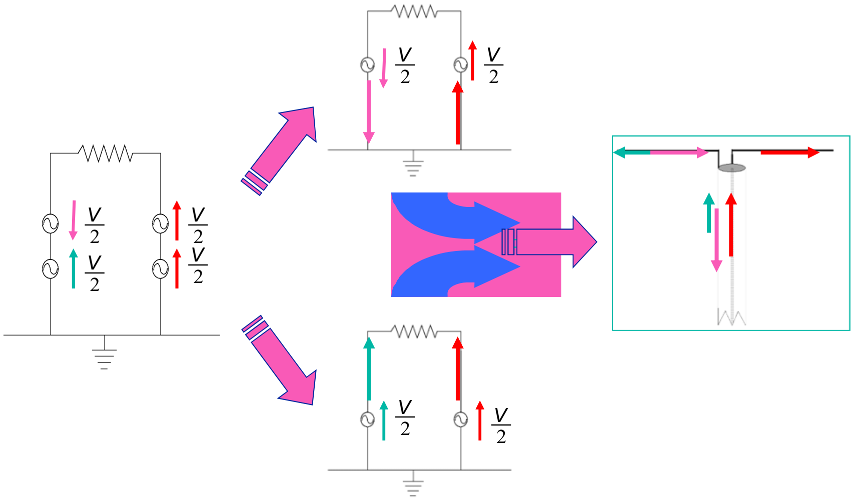
What is the problem ?

- Commercial DECT antenna :
  - Ceramic chip, 6 x 9 x 1.8 mm
  - Gain of 2.2dBi at 1.89 GHz
  - Max Gain after Harrington : -3.3 dBi !!
  - Gain measured at LEMA : -8 ± 2 dBi
- The discrepancy is due to measurement errors

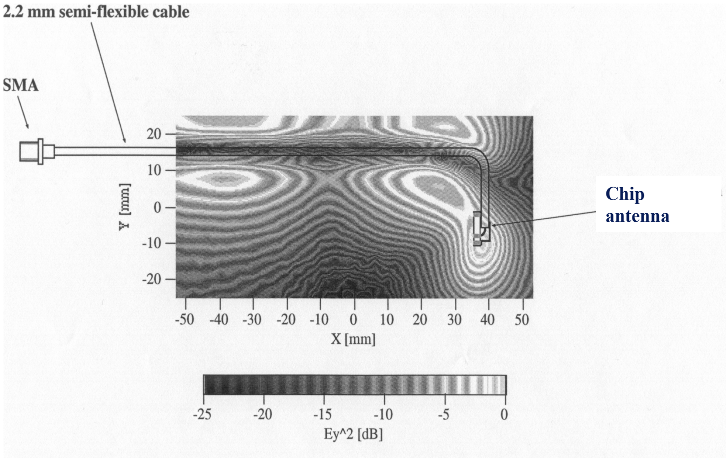
Spurious radiation from cables



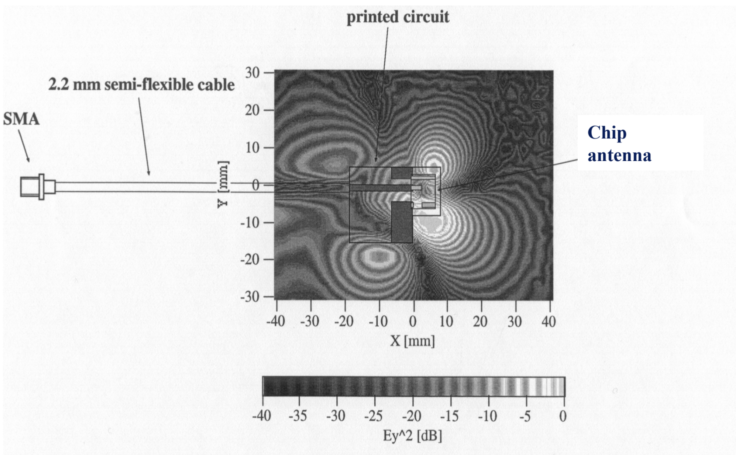
Spurious radiation from cables



Effect of Spurious radiation in the case of the chip antenna



Effect of Spurious radiation in the case of the chip antenna



Effects on radiation characteristics

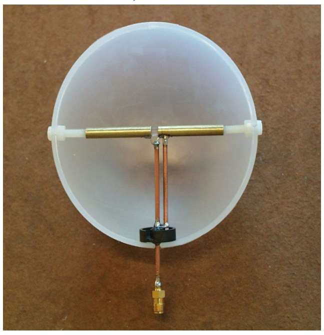
- Unwanted radiation in unwanted directions
- Increase of measured gain up to 10 dB
- Destruction of both polarization and radiation pattern

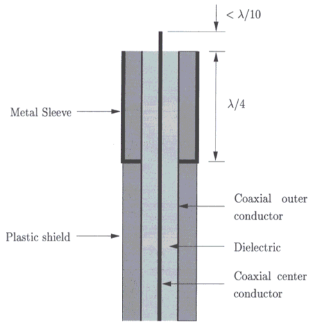
Measurement solutions

- Baluns
- Wheeler cap method
- System measurement methods
  - reverberation chamber
  - anechoic chamber

Baluns

The spurious radiation can be attenuated using for instance ferrite cores, chockes or baluns.





Baluns

- Advantage :
  - good for both circuit and radiation measurements
- Disadvantages :
  - mostly narrow-band
  - cumbersome for the characterization of multi-band antennas

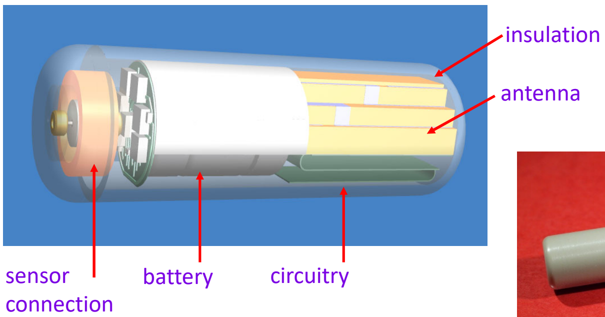
Measurement issues

- ESA are difficult to characterize as they do not have a proper port
- In certain cases, they cannot be measured as the connection to the cable modifies severely their characteristics
- In case of implantable antennas, the problem is worse due to the lossy environment. This is an old problem known from microwave hyperthermia.

- But it is important to characterize the antenna before implanting the system !!!

- F. Merli and A.K.Skrivervik, Design and Measurement Considerations for Implantable Antennas for Telemetry Applications, EUCAP 2010, Barcelona

Illustration : a real case [1-2]



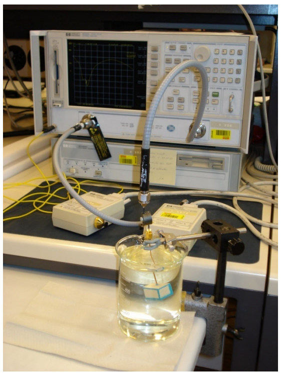
size 10x32mm

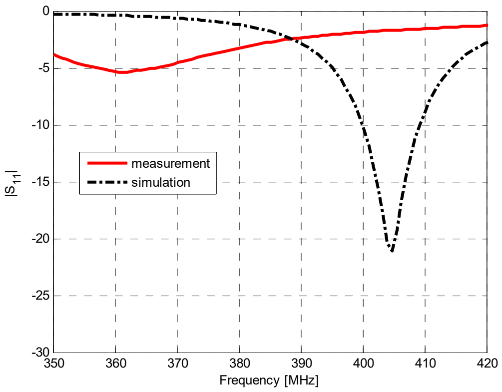
[1] F. Merli, L. Bolomey, J.-F. Zürcher, G. Corradini, E. Meurville, and A. K. Skrivervik, "Design, realization and measurements of a miniature antenna for implantable wireless communication systems," *IEEE Trans. Antennas Propagat.*, submitted for publication.

[2] "Telemetry system for sensing applications in lossy media", L. Bolomey; F. Merli; E. Meurville; J-F. Zürcher and A.K. Skrivervik, Patent number: 00335/10

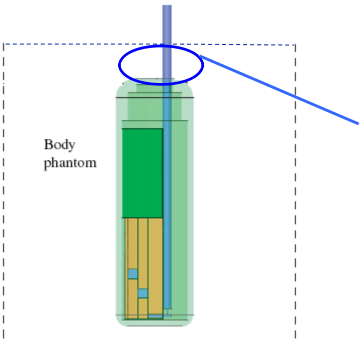
LEMA Miniature Antenna: First simulations and measurements

Electromagnetic performances of the antenna alone have been checked with a feeding coaxial cable (present only for testing purposes).





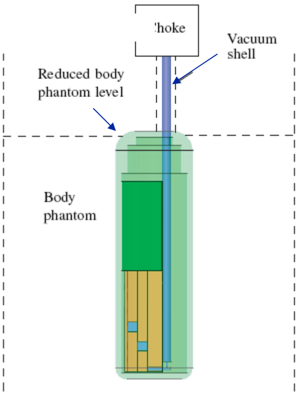
Main difference between simulation and measurement?

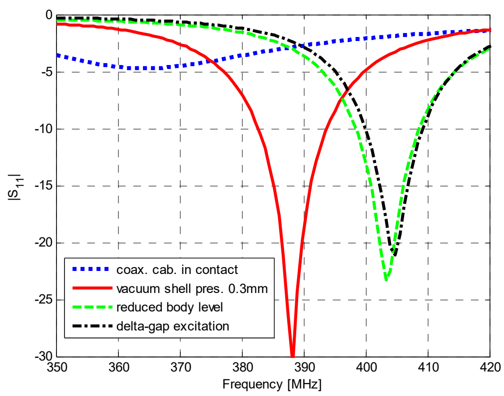


As is well known, the feeding coaxial cable affects the performances of electrically small antennas.

In the measurement setup, the coaxial cable is in direct contact with the biological liquid!!

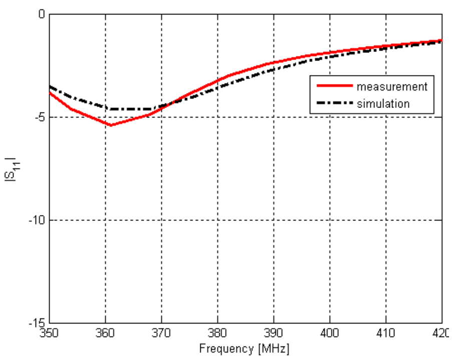
LEMA Miniature Antenna: how to mitigate the cable + body effect?

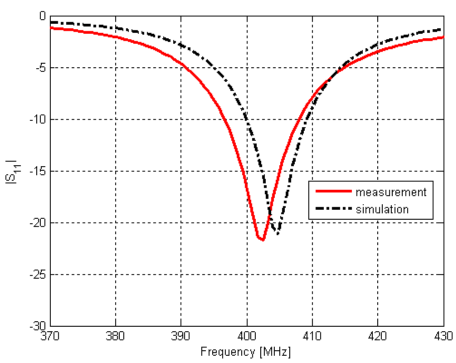




simulations

LEMA Miniature Antenna: how to mitigate the cable + body effect?





Coaxial cable in direct contact

Reduced body phantom level and ideal excitation

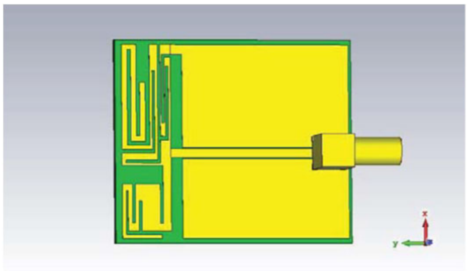
Some issues with the correct definition of the excitations and loads in full-wave simulations

Anja Skrivervik, Microwave and Antenna Group, Ecole Polytechnique Fédérale de Lausanne, Switzerland

Outline

- An example
- Some definitions
- Classification of Full Wave simulation methods
- Numerical excitation of a simulated problem versus physical excitation of a component or device
  - FDTD (FEM)
  - IE + MoM
- Examples

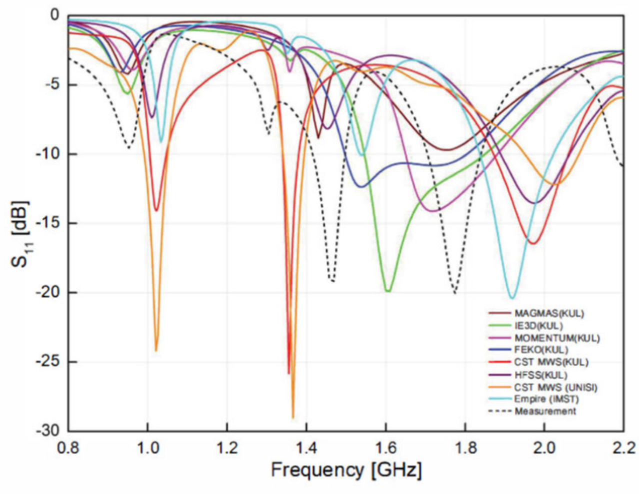
What is the problem: an old benchmark



G. A. E. Vandenbosch, "State-of-the-art in antenna software benchmarking—'Are we there, yet?'" *IEEE Antennas Propag. Mag.*, vol. 56, no. 4, pp. 300–308, Aug. 2014.

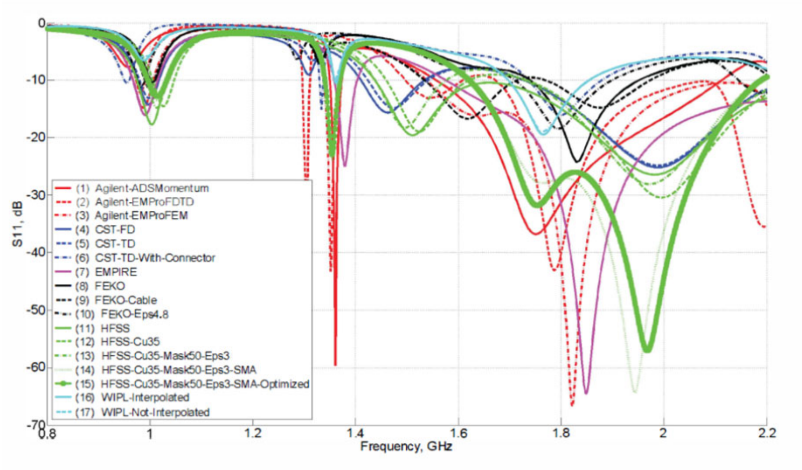
Benchmark first proposed in the ACE Network of excellence WG
On numerical simulation, continued by the EurAAP WG on numerical simulation

Results in 2009

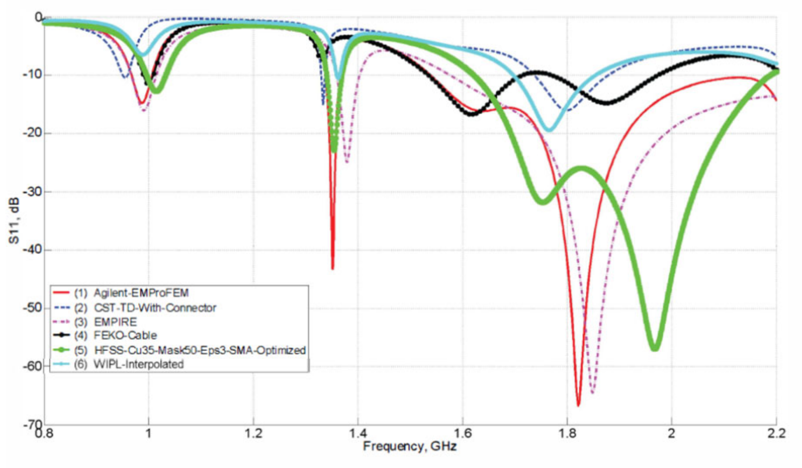


G. A. E. Vandenbosch, "State-of-the-art in antenna software benchmarking—'Are we there, yet?'" *IEEE Antennas Propag. Mag.*, vol. 56, no. 4, pp. 300–308, Aug. 2014.

Results in 2013



Best results in 2013



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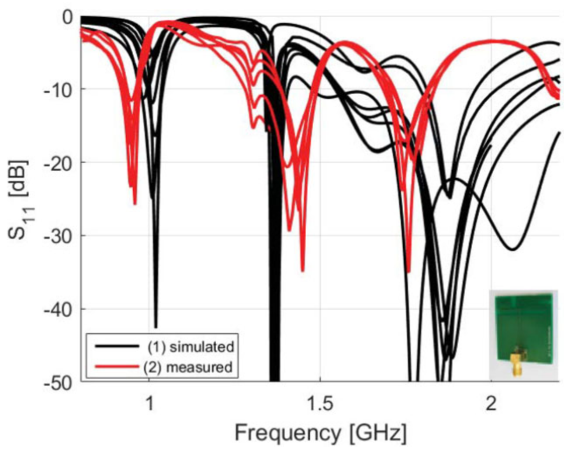
Results in 2018



Guy A.E. Vandenbosch, Modeling and Design Tools for Small Antennas: State of the Art and Future Perspectives
IEEE Antennas and Propagation Magazine
Year: 2018 , Volume: 60 , Issue: 4
Pages: 18 - 20

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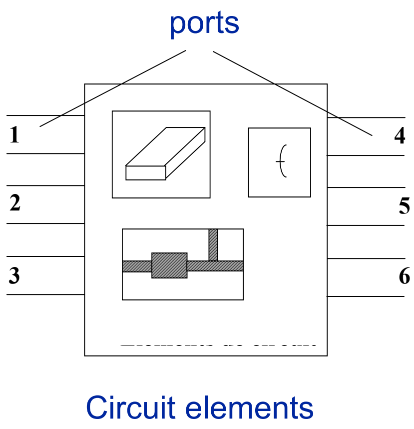
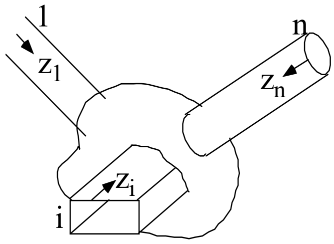
Some definitions

Microwave component



Reference planes



On each access line i of a component, a coordinate axis z_i is defined. The origin of this axis lies in the reference plane of the port i.

Assumptions :

- The transmission lines are lossless
- They support only the dominant mode or any other single mode. If several modes can propagate, we will need to define one port per mode
- The reference plane is distant enough from discontinuities to ensure that the none relevant modes are attenuated

Intermediate findings

- A port is a well defined circuit element
- In microwaves, circuits are defined between ports
- However, elements can be interconnected by just using wires or soldering them together
- Micorwave measurements can be done at ports only
- What about numarical simulations ???

An overview on methods

Selection of solution domain

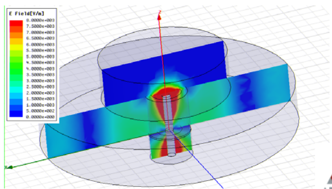
- Time domain
- Frequency domain

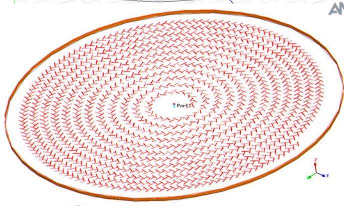
Selection of the Field Propagator

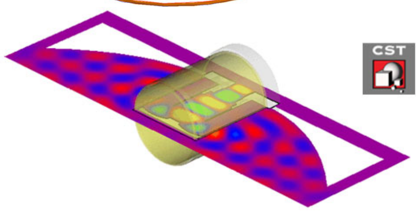
- Integral Equation (IE) model
  - Global field propagator
- Differential Equation (DE) model
  - Local field propagator

Selection of the Sampling Functions

- entire-domain functions
- sub-domain functions







Solution domain

- Time domain

$$\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t}$$
$$\nabla \times \mathbf{H} = \frac{\partial \mathbf{D}}{\partial t} + \mathbf{J}$$
$$\nabla \cdot \mathbf{D} = \rho$$
$$\nabla \cdot \mathbf{B} = \mathbf{0}$$

- Frequency domain

$$\nabla \times \mathbf{E} = -j\omega \mathbf{B}$$
$$\nabla \times \mathbf{H} = j\omega \mathbf{D} + \mathbf{J}$$
$$\nabla \cdot \mathbf{D} = \rho$$
$$\nabla \cdot \mathbf{B} = \mathbf{0}$$

Field Propagator

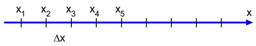
- Local
- An unknown (usually E and H) interacts only with its closest neighbours

- Global
- All unknowns interact with each other

Example of local field propagator: 1-D FDTD

Consider the 1-d wave equation $$\frac{\partial^2 u(x,t)}{\partial x^2} = \frac{1}{c^2}\frac{\partial^2 u(x,t)}{\partial t^2}$$

For the simulation domain $0 \le x \le d$

And the following grid 

with

$$x_m = (m-1)\Delta x$$
$$\Delta x = \frac{d}{M-1}$$
$$t_n = (n-1)\Delta t$$
$$u_m^n = u(x_m, t_n) = u\left[(m-1)\Delta x, (n-t)\Delta t\right]$$

Example of local field propagator: 1-D FDTD

We need to find the numerical expression for the derivatives:

$$\frac{\partial^2 u(x,t)}{\partial x^2} \simeq \frac{\partial}{x}\left[\frac{u(x+\Delta x/2,t)-u(x-\Delta x/2,t)}{\Delta x}\right]$$
$$\simeq \frac{u(x+\Delta x,t)-2u(x,t)+u(x-\Delta x,t)}{(\Delta x)^2}$$

$$\frac{\partial^2 u(x,t)}{\partial t^2} \simeq \frac{\partial}{t}\left[\frac{u(x,t+\Delta t/2)-u(x,t-\Delta t/2)}{\Delta t}\right]$$
$$\simeq \frac{u(x,t+\Delta t)-2u(x,t)+u(x,t-\Delta t)}{(\Delta t)^2}$$

Example of local field propagator: 1-D FDTD

With: $r = \frac{c\Delta t}{\Delta x}$ we write

$$r^2\left[u(x+\Delta x,t)-2u(x,t)+u(x-\Delta x,t)\right]=$$
$$\left[u(x,t+\Delta t)-2u(x,t)+u(x,t-\Delta t)\right]$$

Which can be written as

$$r^2\left(u_{m+1}^n-2u_m^n+u_{m-1}^n\right)=u_m^{n+1}-2u_m^n+u_m^{n-1}$$
$$u_m^{n+1}=r^2\left(u_{m+1}^n-2u_m^n+u_{m-1}^n\right)+2u_m^n-u_m^{n-1}$$

Each unknown interacts only
With its neighbours

Initial and boundary conditions

Initial condition in time: required for two time steps u_m^1 and u_m^2

Often, we take $u_m^1=0,\ u_m^2=0$

x_1 and x_M

Boundary conditions in space, required at

Dirichlet BC: $u(0,t)=u_1^n=0$

$u(d,t)=u_M^n=0$

Neumann BC: $u_1^n=u_2^n$

$u_M^n=u_{M-1}^n$

Sources: hard sources

A hard source sets the value of a field at one or more grid points equal to a specific function of time and is thus a type of Dirichlet BC.

An issue with hard sources is that wave propagating towards them are reflected by them, which can cause modeling errors. A solution is to remove the source after launching the incident wave but before reflections arrive at the source location.

Sources: soft sources

A soft source corresponds to a forcing solution added to the wave equations, for EM problems an impressed electric current. The equation is thus modified as follows:

$$\nabla^2 \mathbf{E} - \frac{1}{c^2}\frac{\partial^2 \mathbf{E}}{\partial t^2} = \mu \frac{\partial \mathbf{J}}{\partial t}$$

In 1D it becomes

$$\frac{\partial^2 u(x,t)}{\partial x^2} - \frac{1}{c^2}\frac{\partial^2 u(x,t)}{\partial t^2} = \mu \frac{\partial J(x,t)}{\partial t}$$

Which can be written as

$$u_m^{n+1} = r^2\left(u_{m+1}^n - 2u_m^n + u_{m-1}^n\right) + 2u_m^n - u_m^{n-1} - c^2(\Delta t)^2 \mu \left.\frac{\partial J(x_m,t)}{\partial t}\right|_{t=t_n}$$

Finite element methods

- Very rigorous, as based on function and functional analysis
- More rigorous than FDTD
- Can be used to solve any Partial differential equations
- Can take many forms
- Can be applied in time or frequency domain
- The mathematical source in the simulator is far from a physical source at a port!!!
- This problem is common to all simulators with local propagators

Example of global field propagator: Electric Field Integral equation + Method of Moments

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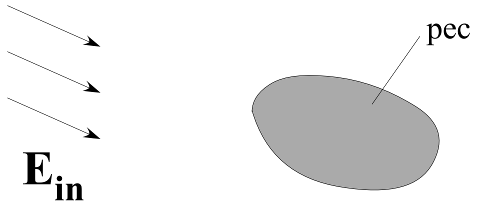
Electric field integral equation : principle



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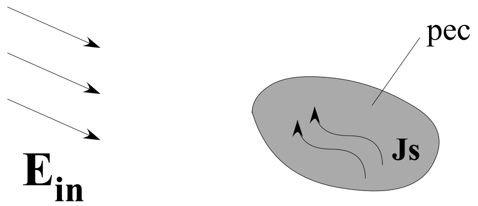
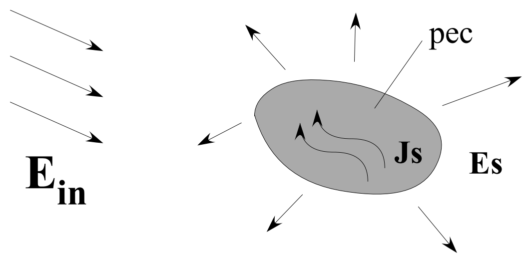
Electric field integral equation : principle



Electric field integral equation : principle



The electric field has to be normal to the body in pec:

$$\boldsymbol{n} \times \boldsymbol{E}_{in} + \boldsymbol{n} \times \boldsymbol{E}_s = 0 \big|_{\text{on the pec}}$$

Electric field integral equation

$$\boldsymbol{E}_s = \overline{\overline{\boldsymbol{G}}}_{EJ} \otimes \boldsymbol{J}_s$$

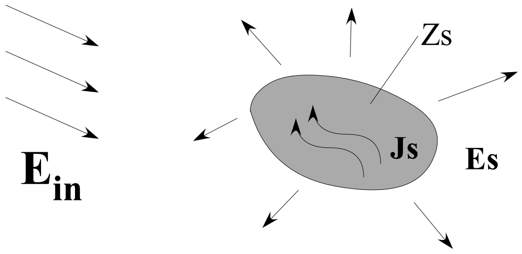
where $\overline{\overline{\boldsymbol{G}}}_{EJ}$ is the Green's function (field of point sources) for the electric field and

$$G \otimes f = \int_{sources} G(\mathbf{r}|\mathbf{r}') f(\mathbf{r}') dv'$$

and finally :

$$\boldsymbol{n} \times \boldsymbol{E}_{in} - \boldsymbol{n} \times \overline{\overline{\boldsymbol{G}}}_{EJ} \otimes \boldsymbol{J}_s = \boldsymbol{0}$$ **EFIE**

Electric field integral equation

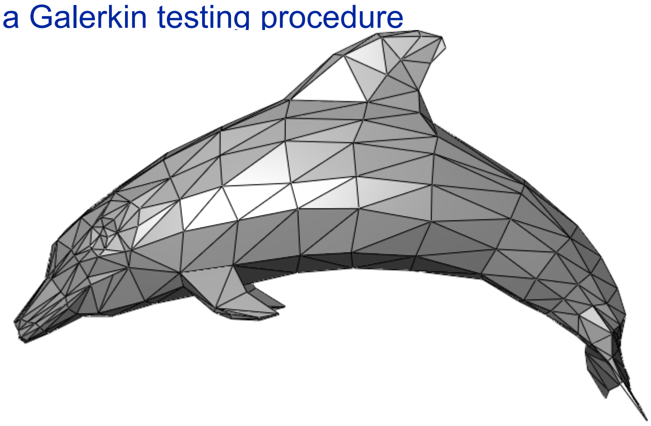


$$n \times E_{in} - n \times \overline{\overline{G}}_{EJ} \otimes J_s = Zs J_s$$

Leontovich impedance condition

Method of Moments

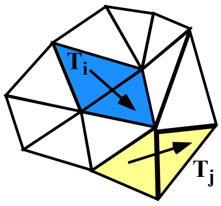
Let us consider a MoM using subsectionnal basis functions and a Galerkin testing procedure



The unknown current is expressed as a sum of basis functions

Method of Moments

The unknown current is expressed as a sum of basis functions



$$\mathbf{J_s} = \sum_{i=1}^{N} \alpha_i \mathbf{T}_i$$

$$\boldsymbol{n} \times \boldsymbol{E}_{in} - \boldsymbol{n} \times \overline{\overline{\boldsymbol{G}}}_{EJ} \otimes \boldsymbol{J}_s = 0$$

$$\boldsymbol{n} \times \boldsymbol{E}_{in} = \boldsymbol{n} \times \overline{\overline{\boldsymbol{G}}}_{EJ} \otimes \sum_{i=1}^{N} i_i \boldsymbol{J}_s$$

Galerkin testing procedure

$$Z_{ij} = \int_{s_i} \boldsymbol{T}_i(\boldsymbol{\rho}) ds_i \int_{s'_j} \overline{\overline{\boldsymbol{G}}}_{EJ}(\boldsymbol{\rho} \mid \boldsymbol{\rho}') \cdot \boldsymbol{T}_j(\boldsymbol{\rho}') ds'_i \qquad [i] = [Z]^{-1}[U]$$

$$u_i = \int_{s_i} \boldsymbol{T}_i(\boldsymbol{\rho})(\boldsymbol{n} \times \boldsymbol{E}_{in}) ds_i$$

Mathematical source,
physically a voltage!!

Popular Commercial softwares

- Time domain
  - Finite Difference Time Domain (FDTD)
    - Discretization of space (3-D) and time
    - Requests absorbing boundary conditions
    - Local field propagator
    - Unknowns E and H fields
    - Mathematical excitation is a time pulse in a specific cell.
    - Physical feeds are defined as special functions (for instance transmission line modes in waveguides or cables)

Popular Commercial softwares

- Frequency domain
  - Finite Element Method (FEM)
    - Volume discretization of space (3-D)
    - Requests absorbing boundary conditions
    - Local field propagator
    - Unknowns E and H fields
    - Mathematical excitation is a field in a cell.
    - Physical feeds are defined as special functions (for instance as transmission line modes)

Popular Commercial softwares

- Integral Equation + Method of Moment (MoM)
  - Surface discretization of conducting surfaces (2-D)
  - Global field propagator
  - Unknowns are surface currents
  - Mathematical excitation is a current or voltage in a cell
  - Great flexibility in the feed definition
  - Good compatibility with circuit simulators
  - Limited treatment of inhomogeneous problems

How are excitations and lumped loads specified in commercial tools ?

- FEM or FDTD
- IE+MoM

Modeling issues : Example of FDTD cell

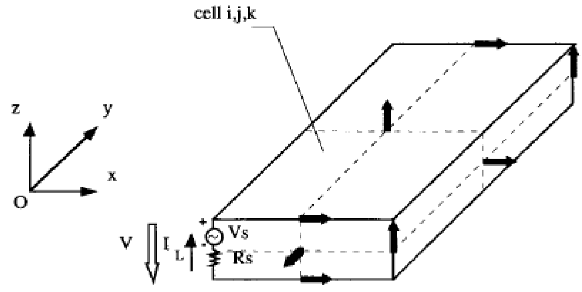


Fig. 1. Standard Yee cell with a lumped resistive voltage generator.

Mathematical excitation in FEM (similar but

The port excitation is pre-computed : Wave ports or lumped ports

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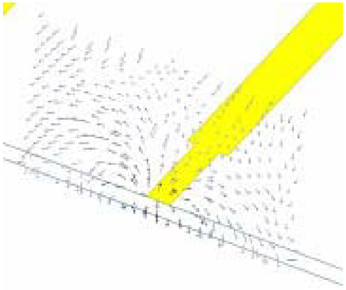
Wave ports example: a microstrip line



Wave port: the modes are solved
In the plane transverse to the port
Wave ports solve for characteristic
impedance and propagation constants
at the port cross-section

An infinite long transmission line is assumed
At the port. It is assumed to support only one
Mode. This mode is resolved, and used as
excitation f or the initial problem

We have a port in the circuit sense

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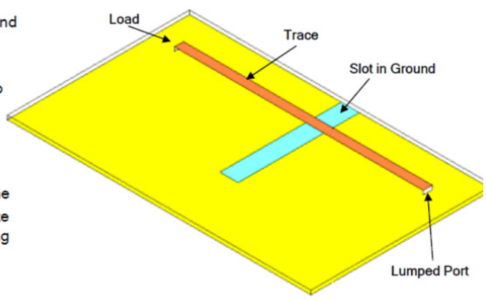
Lumped ports

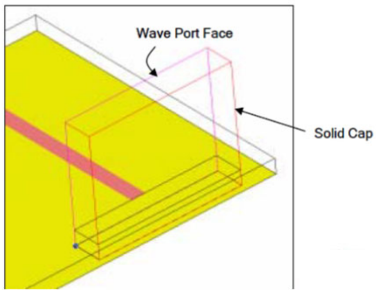
- Lumped ports excite a simplified, single-mode field excitation assuming a user-supplied Zo for S-parameter referencing
- A Terminal line may still be defined, but only one per port.
- Impedance and Propagation constants are not computed
- Port boundaries are simplified to support simple uniform field distributions.
- Edges touching perfect_E or finite conductivity faces, such as ground planes and traces, take on the same definition for the port computation
- Edges not touching conductors become perfect_H edges for the port computation
- This is different than the assumption made by Wave ports!!
- Impedance and Calibration line assignments are required for Lumped port assignments

Wave ports versus lumped ports

- Wave Ports are more Rigorous
  - True modal field distribution solution
  - Multiple mode, multiple terminal support
  - Use Wave ports by preference if there are no specific reasons their usage would be discouraged
- Port Spacing may force Selection
  - Widely spaced individual excitations usually permit Wave ports
  - Closer-spaced, yet still individual excitations may require Lumped ports
  - Closely-spaced, coupled excitations require Wave ports
    - Only Wave Ports handle multiple modes, multiple terminals.
- Port Location may force Selection
  - Wave ports are best on model exterior surface; interior use requires cap
  - Lumped ports are best for internal excitations, where caps would provide undue disruption to modeled geometry and fields
  - Wave Ports permit de-embedding to remove excess uniform input transmission lengths
  - Lumped Ports cannot be de-embedded to remove or add uniform input transmission lengths
- Transmission Line and Solution Frequency may force Selection
  - Lumped Ports support only uniform field distributions
  - Only Wave Ports solve for TE mode distributions, TM mode distributions, or multiple modes in same location
  - Most non-TEM excitations will require Wave Ports

- Microstrip Port on RF Board
  - Circuit board modeled inside air volume for ground slot excitation and EMI analysis
  - Trace does not extend to end of board
    - For above reasons, port must be interior to modeled volume
  - Wave port would require cap embedded in substrate [see bottom]
    - Port face extends from ground surface beneath substrate to well above trace plane
    - Cannot have intersecting cap and substrate solids, therefore Boolean subtraction during model construction is required
  - Use Lumped Port for simplicity
    - Easier to draw
    - Sufficiently accurate solution for isolated line input (no coupled behavior to be neglected)
    - No large metal cap object present to perturb solution of ground plane resonance or radiation effects



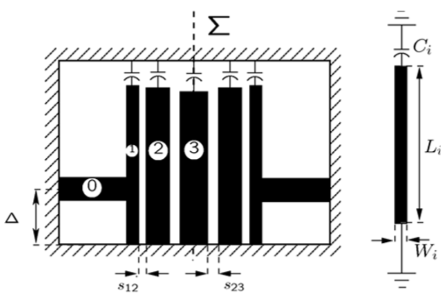


Excitations and Ports in IE+MoM

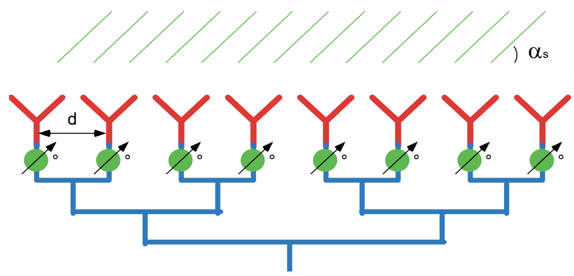
- The unknown is the surface or volume current
- - The mathematical source is a volage
- - thus no problem

Lumped loads

Loads are not on TLines

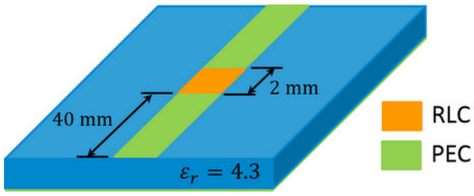


Loads are on TLs



Lumped loads not on TLs (not linked to ports)

- For FDTD and FEM, they are represented by surfaces, and a new complete simulations will have to be made for each change of load



https://apps.lumerical.com/rf_pcb_microstrip_rlc.html

Lumped loads are not on TLs (not linked to ports)

After discretization of the integral equation and projection on the test function, we obtain :

- For IE+MoM, some time can be saved (in theory)

$$[Z][I]=[V_{ex}]$$

$$[Z_{mom}][I]+[Z_{load}][I]=[V_{ex}]$$

[Zload] is a diagonal matrix containing the load impedances

$[Z_{MoM}]$ Needs to be computed only once

Summary

- The largest source of mistakes or discrepancy in EM numerical modelling is due to the handling of the ports.
- A port is defined on a guiding structure
- It is difficult to handle lumped elements in EM simulators
- The user needs to be careful !