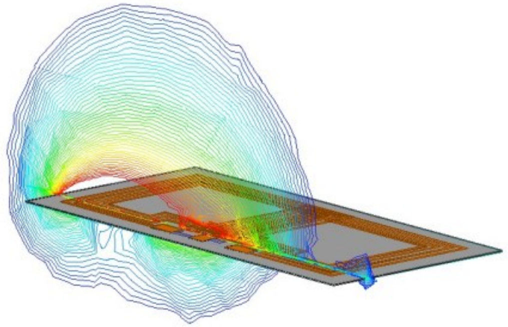

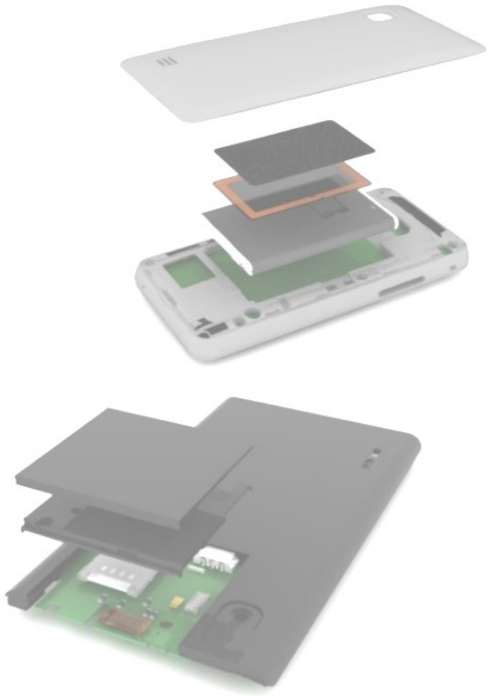

03 Contactless Readers

3rd unit in course 451.417, RFID Systems, TU Graz

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RFID Systems, Graz University of Technology

SS 2023, March 06st

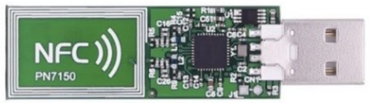
rfid-systems.at
Advancing Technology

Some NFC Reader appearences – form factors

ATM Terminal Reader

Access Reader (HID)

POS Terminal Reader (Verifone)



USB Reader (Identiv, MikroElektronika)

Pen Reader (MicroSensys)

Selection criteria for NFC reader IC

- Type of IC / intended concept for electronics
- Package
- Interface options (supply, digital interface, clock,…)
- NFC protocol capabilities
- Contactless communication capabilities (output power / sensitivity / filters)
- Ambient conditions (temperature, humidity,..)
- Lifetime / aging (data retention, pin protection, …)
- Additional (product lifetime, documentation / Appnotes, Eval-Kits, SW-libraries, design ressources)

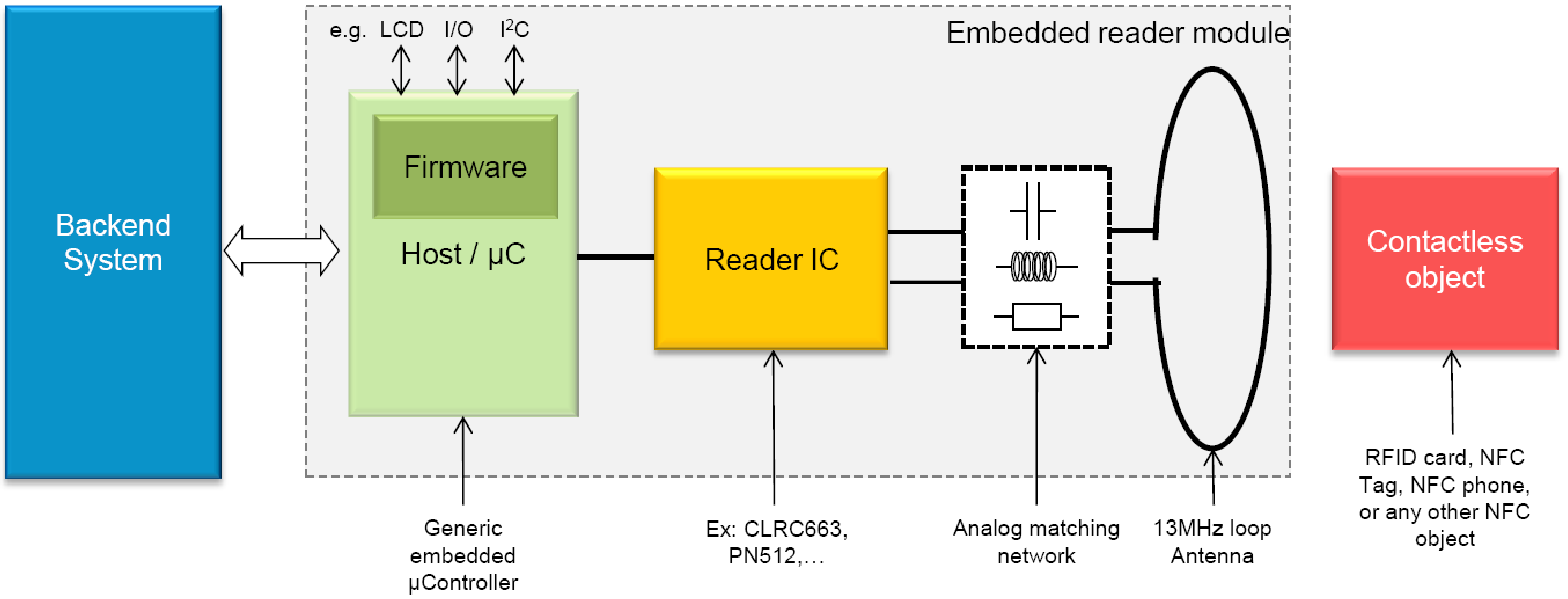
Electronics concepts

- **NFC Reader Frontend IC + Microcontroller** – **classical design**
  - Typical „infrastructure“-reader design.
  - Firmware on µC configures e.g. the protocol capabilities of the reader.
  - E.g. USB reader, Payment Terminal NFC reader
  - IC´s: CLRC663xx + LPCxx (NXP), SE610, ST25R95 (ST), TRF79xxA + MSP430 (TI), PTX100R (Panthronix)

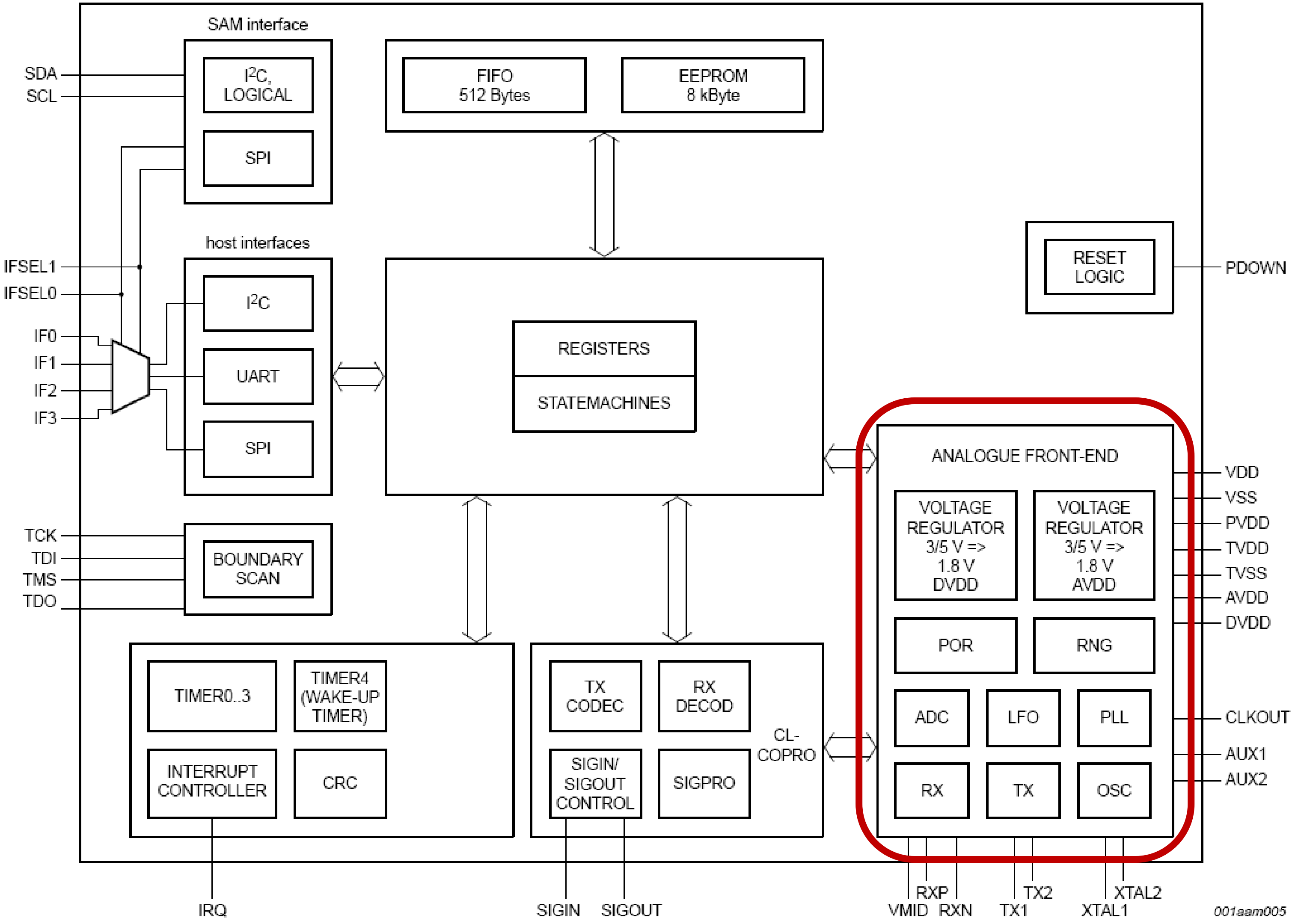
- **Fully integrated NFC IC** – **cost and space saving**
  - Typical for mobile phone NFC interface
  - IC´s: PN548, PN7462 (NXP)

- **Multi function radio IC featuring NFC option** – **NFC als add on**
  - Typical for
  - E.g. car key

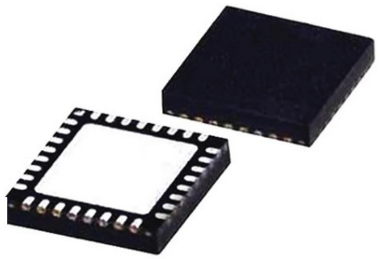
Contactless reader architecture – classical design



Functional blocks in Reader IC

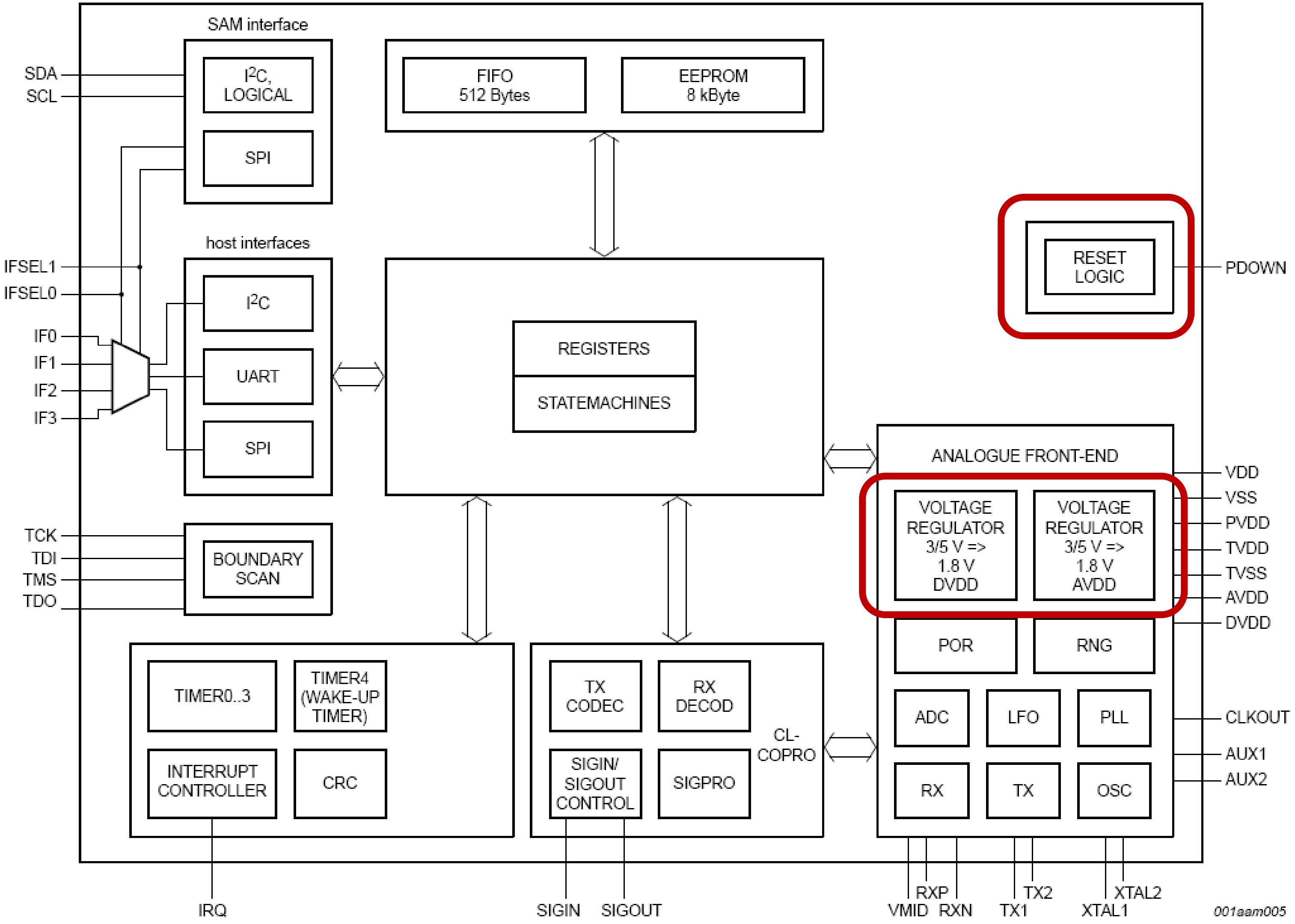


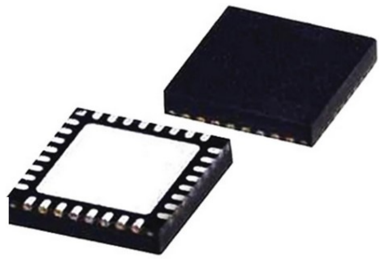
- **Analog Frontend**



  - Does the modulation and demodulation of signals for the contactless interface
  - Connection for clock (XTAL1, 2), typically to a quartz crystal. This is the time reference for the carrier signal, as well as encoder and decoder
  - PLL clock line (CLKOUT) can serve as clock source to MCU µC
  - Analog test signals (AUX1, 2) to debug or optimize the reader HW

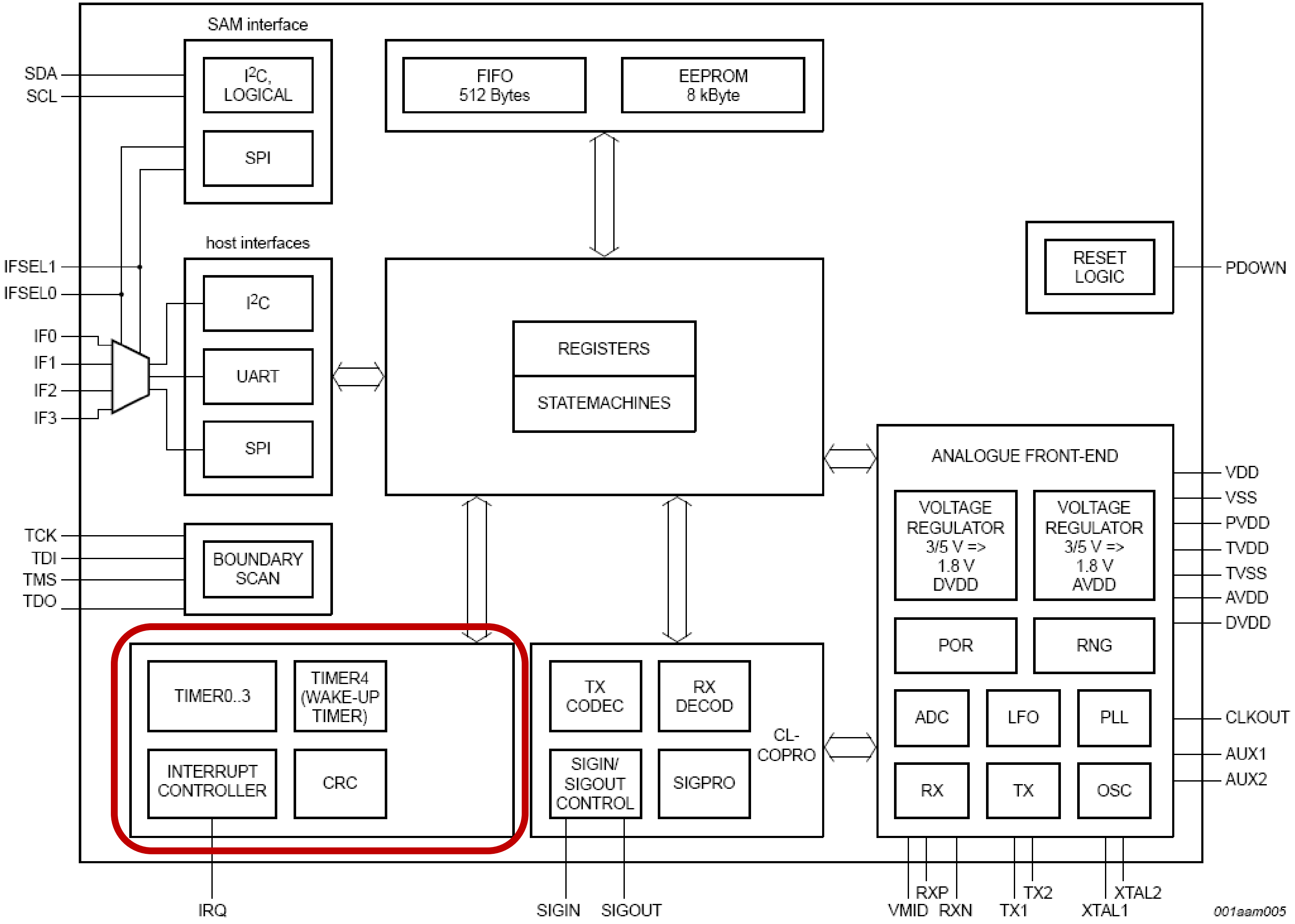
Functional blocks in Reader IC

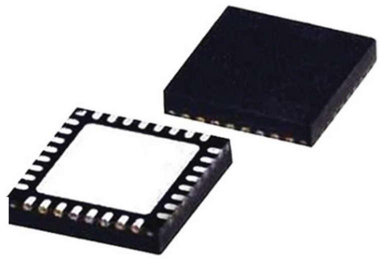




- **Power management**
  - Here we find 3 individual voltage supplies
    - VDD… general chip supply
    - PVDD…pad logic level supply
    - TVDD…TX supply (can be higher)
    - AVDD…analogue supply
  - All should be buffered with capacitors
  - For energy saving there are power-down options…
  - There is a specific low power card detection option, in which the reader IC is not fully powered all the time

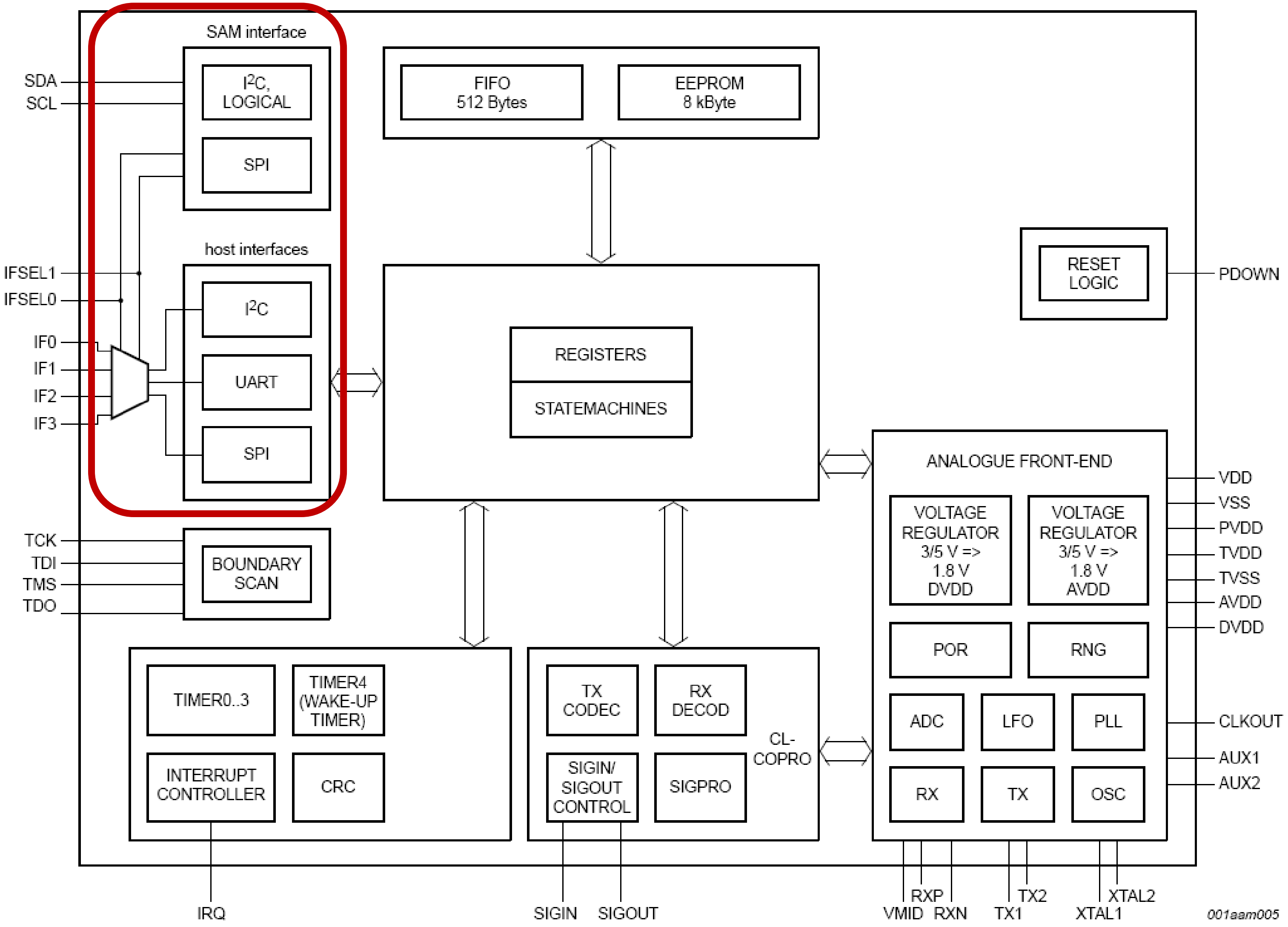
Functional blocks in Reader IC



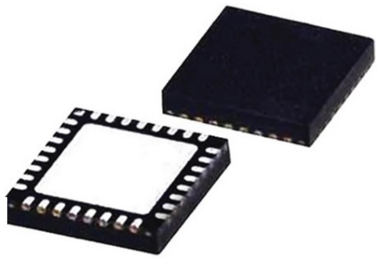


- **Interrupt & Timer**
  - All interrupts can be configured by FW
  - The interrupt controller handles the enabling / disabling of interrupt requests
  - It indicates certain events by setting bit IRQ and pin IRQ to high level – this may be used to interrupt the host µC
- The external host may use the timers to manage time-relevant tasks (watchdog counters, periodical triggers,…)

Functional blocks in Reader IC

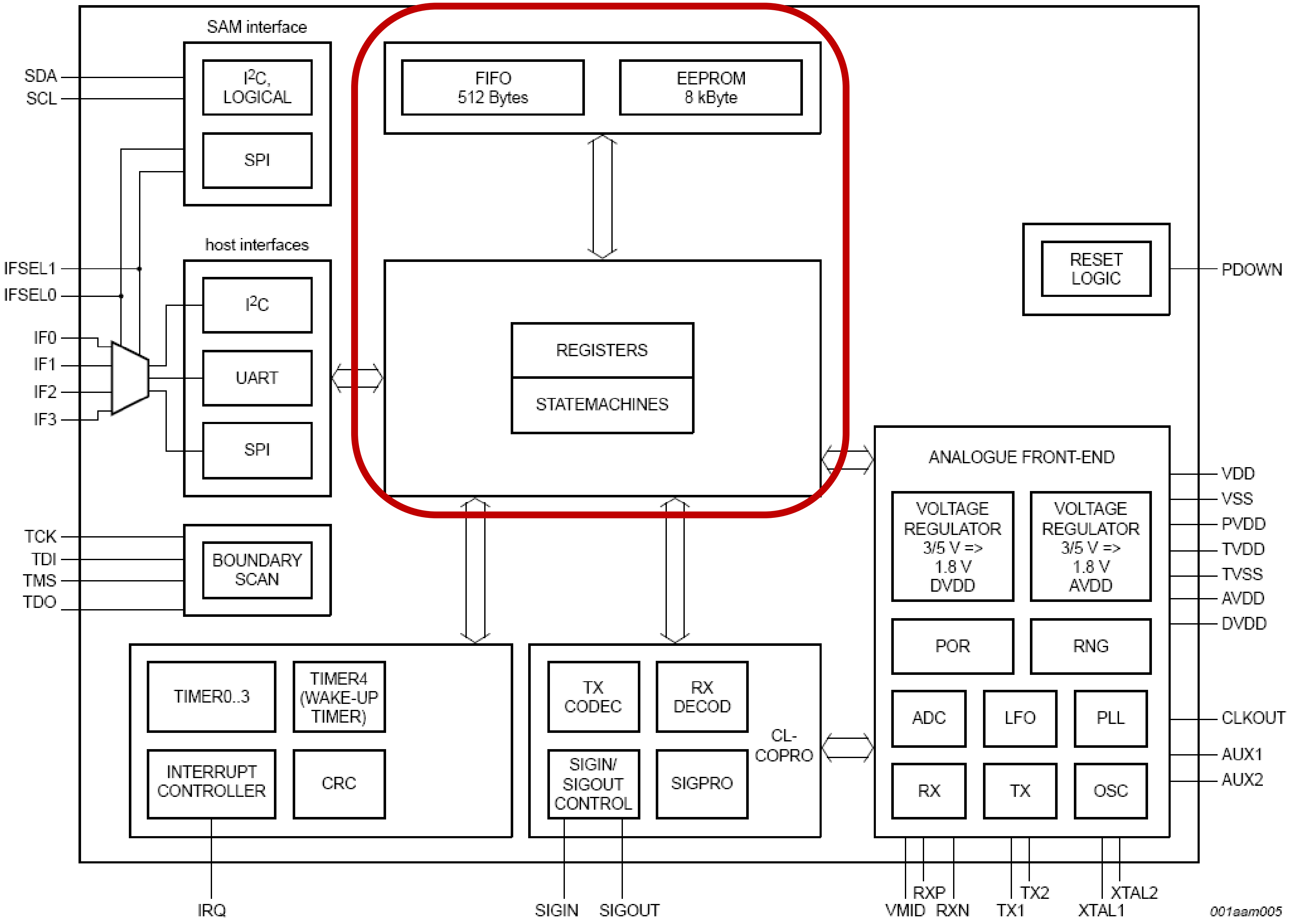


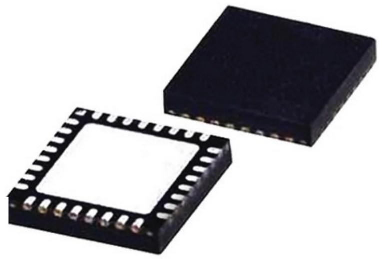
- **Dig. Control Interfaces**



  - Direct control by host µCs is supported
  - Commonly used digital interfaces (I²C, SPI, UART) are available
  - The host interface is selected by means of the logic levels on control pins (IFSEL0, IFSEL1) after the reset phase

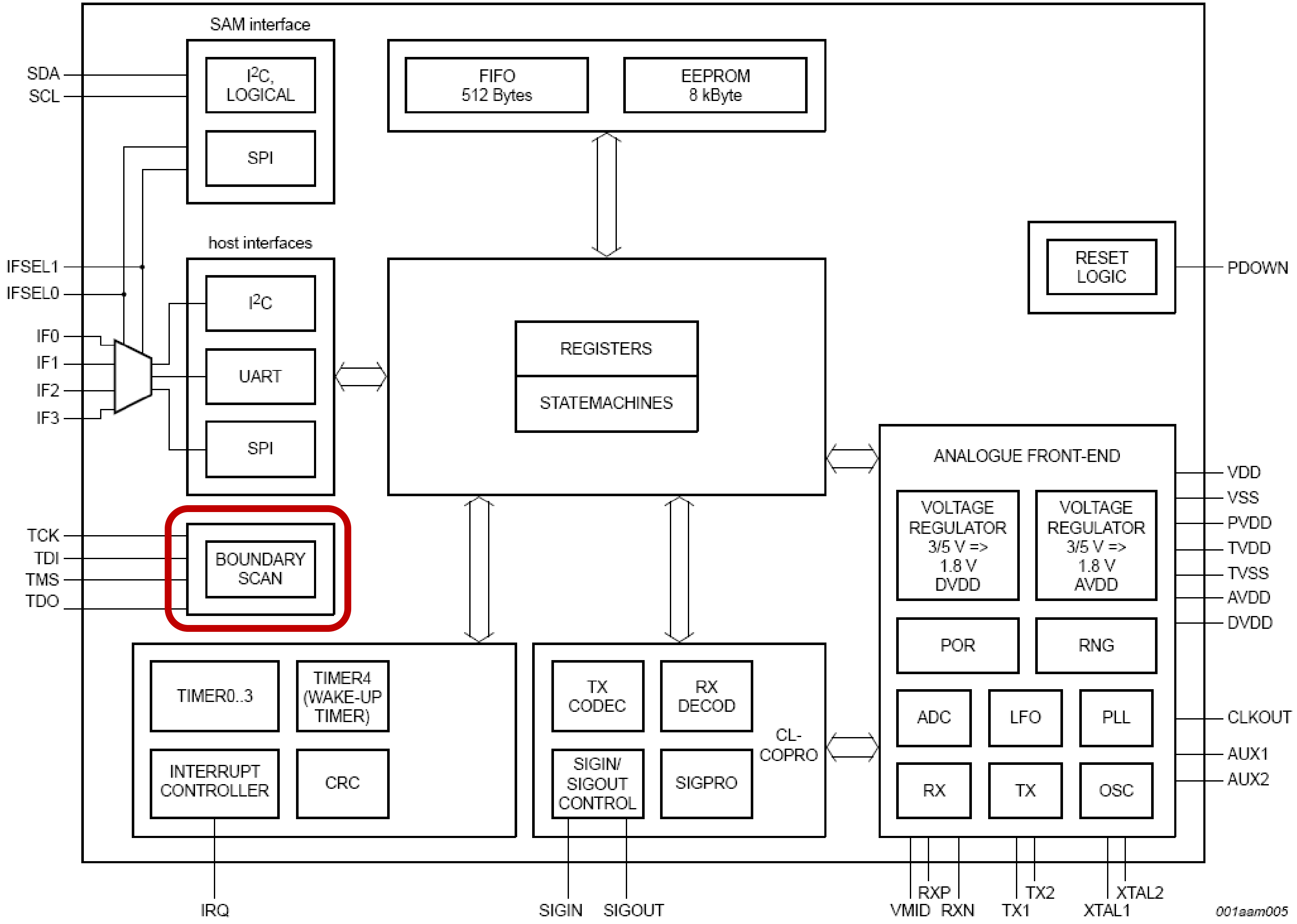
Functional blocks in Reader IC

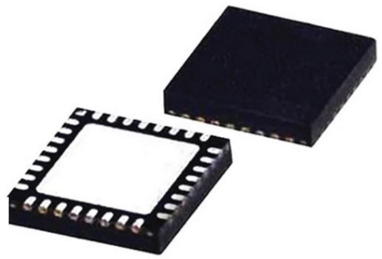




- **Memory**
  - The register bank contains settings for analog and digital functionality
  - EEPROM allows to store alternative, user defined register settings. Settings for another protocol can be loaded by a command from host µC
  - FIFO: This buffer handles a receive and send data frame size (here, 512 bytes)

Functional blocks in Reader IC

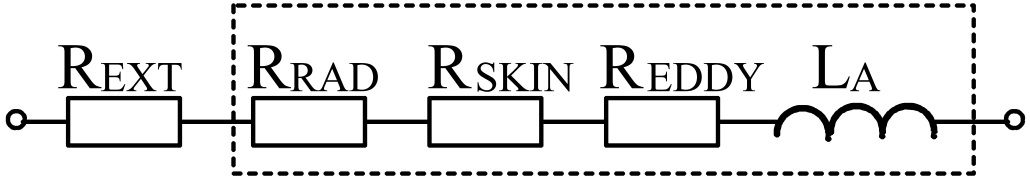




- **Boundary Scan**
  - Allows to test interconnections without having to use physical probes
  - Usually, a test socket on the reader PCB is foreseen to allow function testing
  - An interface (according to IEEE1149,1) between IC and environment is implemented via this interface
  - It uses its own description language (Boundary Scan Description Language)

Reader loop antenna – RF power losses

- We differentiate the „directly matched“ antenna, very close to and typical for an NFC reader, and (coaxial) cable connected loop antennas, as used for long-range readers and vicinity applications.
- In any case, it is instructive to consider which power losses apply:



R_{EXT} *Power loss in the external damping resistor*

R_{RAD} *Radiation as electromagnetic wave (typ. P ~ 20 µW, negligible)*

$$R_{RAD} = 31000 \cdot N^2 \cdot \frac{A^2}{\lambda^4}$$

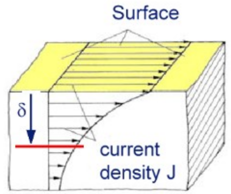
N.......number of turns

A.......antenna area

λ.......wavelength

R_{SKIN} *Losses caused by skin effect (AC current is forced from center to “skin” of the conductor). For copper at 13.56 MHz the skin depth δ (decrease to 1/e) is determined by*

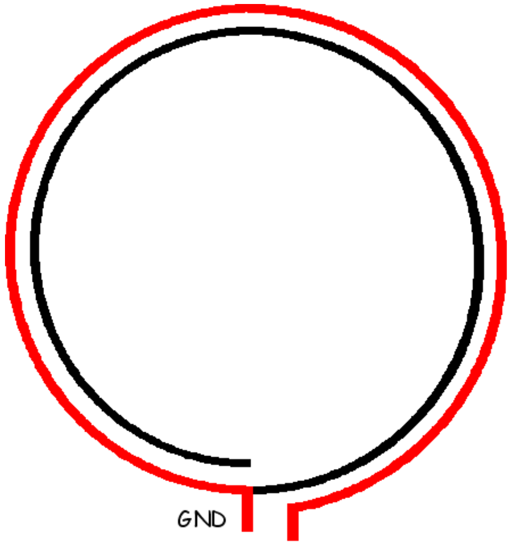
$$\delta = \frac{1}{\sqrt{f_C \cdot \pi \cdot 4\pi \cdot 10^{-7} \cdot 5.8 \cdot 10^7}} = \frac{0{,}0661}{\sqrt{f_C}} \cong 18\mu m$$

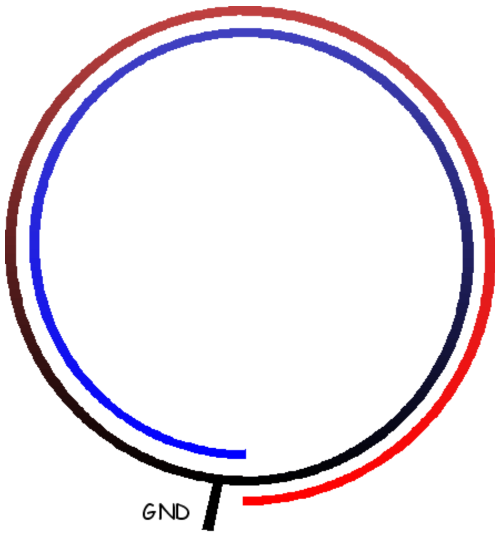


R_{EDDY} *Losses caused by eddy currents in surrounding metal plates. Consequently, avoid closed metal areas (or split them up) or use ferrite material for magnetic isolation.*

Electrical compensation

- It is possible to reduce electromagnetic wave propagation (in far field) to a minimum (and so the *H*-field amplitude in near-field is allowed to be even higher), if an electrically compensated loop antenna is used.
- Especially for cable-connected loop antennas (e.g. ISO/IEC10373-6 test antenna), such electrical compensation is applied.





- This also helps, to confine radiation to the antenna and to prevent common mode RF currents (allowing to connect the antenna to coaxial cables).

- AC current, the root cause for *H*-field emission, flows from one end of the loop to a center tap (virtual Ground).
- AC voltage, root cause for the *E*-field, ranges from one polarity at one end, to the inverse polarity at the opposite (open) end, and cancels out in distance.

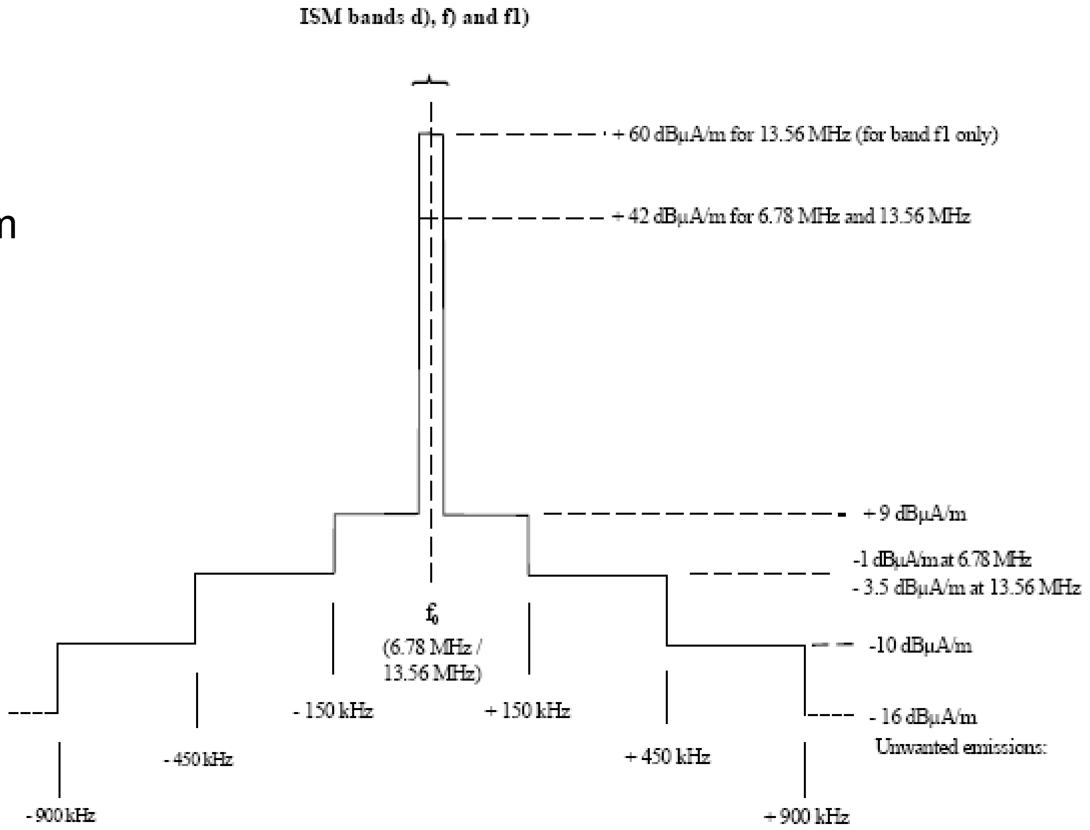
Emission limits

- Especially for Vicinity (long range) reader systems, a starting point are normally geometrical dimensions of the loop antenna and applicable emission limits for the carrier frequency. (For Proximity / NFC readers, the modulation bandwidth is more interesting.)
- In a 1st step the maximum antenna current is considered, which - for the antenna geometry - generates the maximum allowable *H*-field (or in the US the *E*-field) limit in the specified measurement distance. This can be estimated using the **Magnetic Momentum method**, or the extended **Biot-Savart law**.
- To anticipate, in which orientation and direction the maximum field strength can be expected, it is necessary to consider for the wavelength (of the carrier) and the measurement distance, whether it is in near-field or far-field.
- In near-field, *H*-field emission is stronger for coaxial orientation, in far-field the free em. wave propagation in coplanar orientation.
- Usually you start with the carrier frequency limit (Vicinity systems). If the modulation is more critical (Proximity systems), it is simulated relative to the carrier, taking all aspects (duty cycle…) into account.

Emission limits

- Especially for Vicinity (long range) reader systems, a starting point are normally geometrical dimensions of the loop antenna, and applicable emission limits for the carrier frequency. (For Proximity / NFC readers, the limits for modulation bandwidth are more interesting.)

- Relevant documents for Europe
  - CEPT ERC/REC 70-03 (2005) Annex 9 gives H-field limits in A/m or dB(µA/m)
  - EN 300 330 – 1 (2002) 7.2.1.2. gives measurement method and measurement distance (10 m typ.)



Re-calculation between absolute and decibel values

- Limits are given (e.g.in CEPT ERC/REC) in decibels (dBµA/m). A re-calculation to absolute values (A/m) can be done according to

$$H_{dB} = 20\log\left(\frac{H_{ABS}}{H_{REF}}\right) \qquad H_{ABS} = H_{REF} \cdot 10^{H_{dB}/20}$$

For example $$H_{LIMIT} = H_{REF} \cdot 10^{H_{dB}/20} = 1\mu A/m \cdot 10^{60dB(\mu A/m)/20} = 1000\ \mu A/m = 1\ mA/m$$

- Similarly it is possible to re-calculate (for far-field) *E*-field limit values to *H*-field values, if required (e.g. FCC Spec is for *E*-field). This is done as follows:

$$Z_{0,dB} \cong 20\log(377\ \Omega) \cong 51.5\ dB(\Omega)$$

$$E_{LIMIT} = H_{LIMIT} \cdot Z_0 = 1000\ \mu A/m \cdot 377\ \Omega = 377000\ \mu V/m$$

$$E_{dB} = H_{dB} + Z_{0,dB} = H_{dB} + 51.5dB(\Omega) = 60\ dB(\mu A/m) + 51.5\ dB(\Omega) = 111.5\ dB(\mu V/m)$$

Maximum allowable loop antenna current

- The maximum current (for N antenna turns) follows from Magnetic Momentum as…

$$\left|M_D\right| = N \cdot I \cdot A \quad \rightarrow \quad I_{LIMIT} = \frac{\left|M_D\right|_{LIMIT}}{N \cdot A}$$

total current in antenna geometry: $$I_{LIMIT} = \frac{\left|M_D\right|_{LIMIT}}{A}$$

- ….and we calculate $\left|M_D\right|_{LIMIT}$ as follows:
  - ¬ If applies $$d_{MESS} \leq \frac{\lambda}{2\,\pi} \cdot 2.354$$

…we measure in the transition region to near field, where emission in coaxial direction generates more *H*-field. Using the Magnetic Momentum method, we can estimate…

$$\left|M_D\right|_{LIMIT} = 2\pi \cdot \frac{\bar{\lambda}\, d_{MESS}}{\sqrt{\bar{\lambda}^2 + d_{MESS}^{\;2}}} \cdot H_{LIMIT} \qquad \text{with} \quad \bar{\lambda} = \frac{\lambda}{2\,\pi}$$

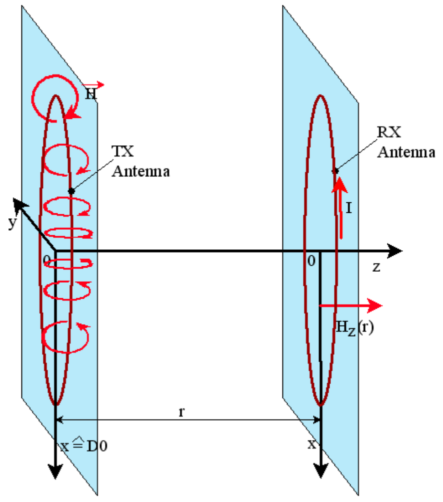
  - ¬ If applies $$d_{MESS} > \frac{\lambda}{2\,\pi} \cdot 2.354$$

…we measure in the transition region to far field, where wave propagation in coplanar direction generates more *H*-field. Using Magnetic Momentum, we can estimate…

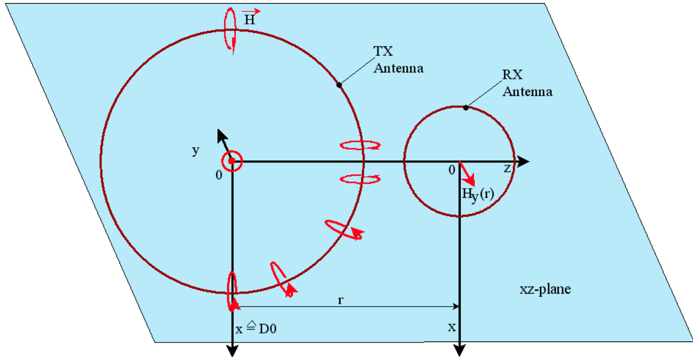
$$\left|M_D\right|_{LIMIT} = 4\pi \cdot \frac{\bar{\lambda}^2\, d_{MESS}^{\;3}}{\sqrt{\bar{\lambda}^4 + \bar{\lambda}^2 d_{MESS}^{\;2} + d_{MESS}^{\;4}}} \cdot H_{LIMIT}$$

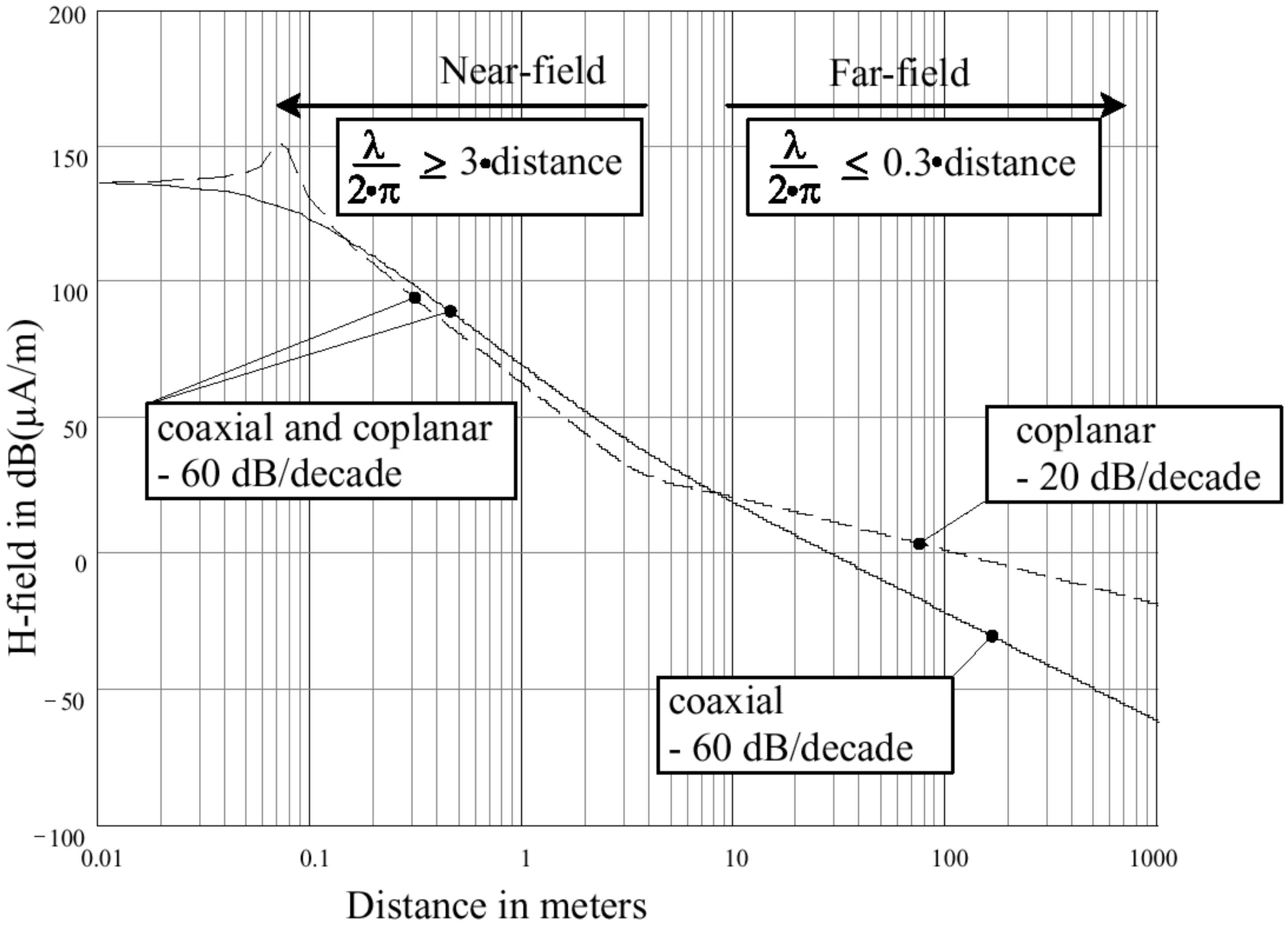
Maximum allowable loop antenna current

Coaxial antenna arrangement



Coplanar antenna arrangement





Determining the Q-factor

- The loop antenna current is the root cause for the emitted alternating *H*-field, and in fact, this involves effective and reactive current. For an antenna resonant at 13,56 MHz, on one hand it seems reasonable to have a Q-factor as high as possible, to **emit high *H*-field at carrier frequency** (still below emission limit) with a minimum required amplifier power.
- On the other hand, **communication requires a certain bandwidth**, defined by the intended protocol (highest data rate, modulation type, duty cycle, etc.). This requirement limits the maximum allowable Q-factor (involving the complete path of amplifier, matching network, and loop antenna).
- A trade-off between these two contradicting requirements is required to specify a good value for the Q-factor:
  - Communication theory allows to determine the required bandwidth
  - This bandwidth allows to determine a maximum Q-factor

Determining the Q-factor

- The rule for time and bandwidth of communication engineering determines the required bandwidth.
- From bandwidth results a maximum Q-factor

- To note: This may be a current gain in a more complex filter function, in real life!

e.g. ISO/IEC14443 (3 µs @106kbit/s)

typ. Q < 35

ISO/IEC15693 (9,44 µs)

typ. Q < 100

Time – Bandwidth Product

$$B \cdot T \geq 1$$

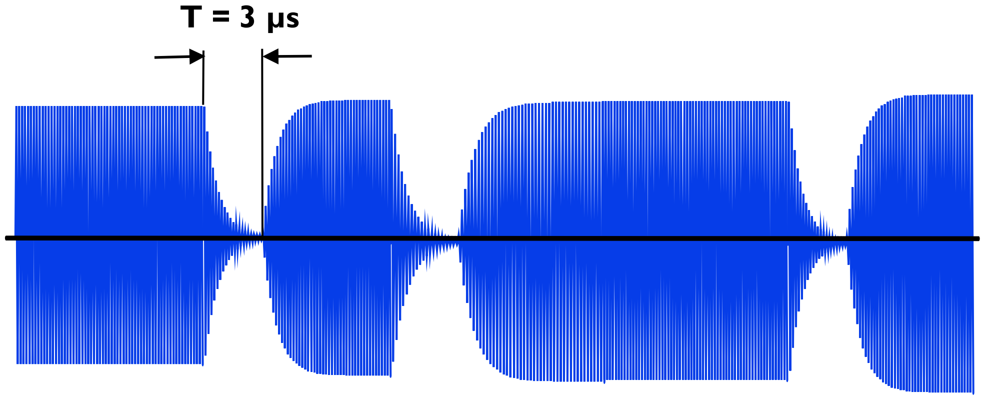
Q-factor definition (freq.)

$$Q = \frac{f_{RES}}{B}$$

$$Q \leq f_{RES} \cdot T$$

$$Q \leq 13{,}56 MHz \cdot 3 \mu s$$

$$Q \leq 40{,}68$$



Power and Q requirement – for single resonance

- The unloaded *H*-field strength emitted by an antenna with resonance at carrier frequency can be approximated by...

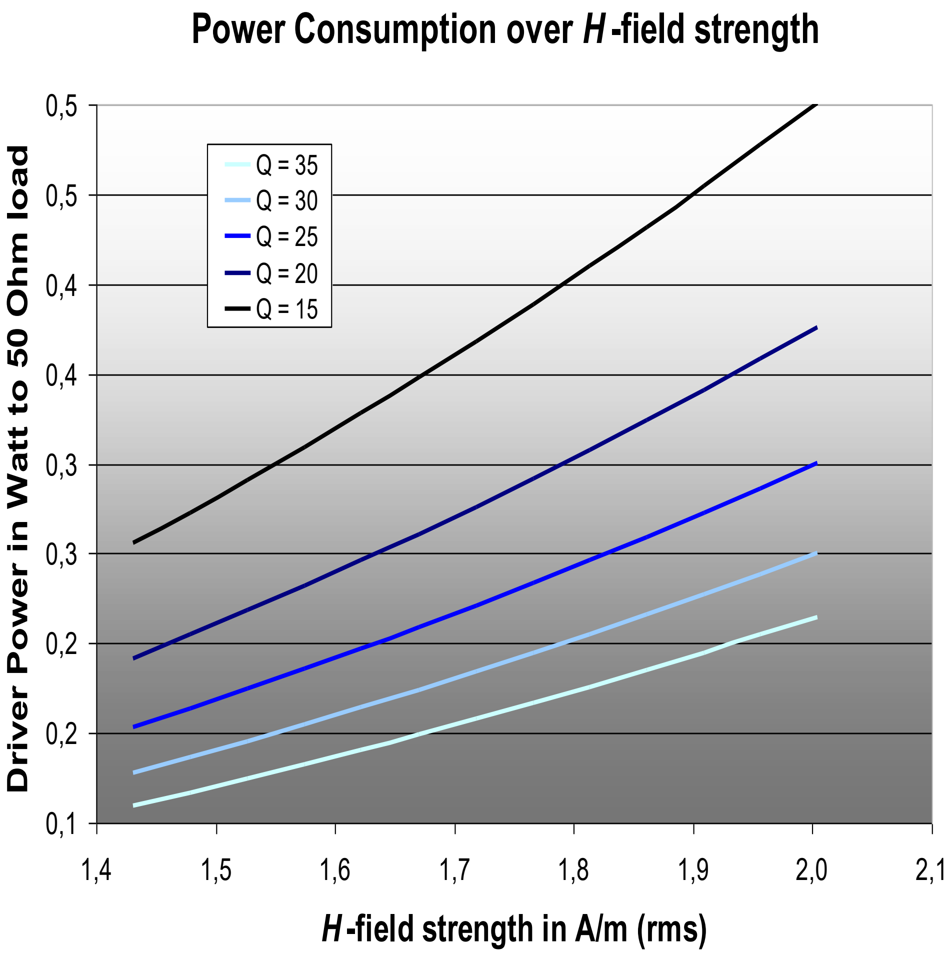
$$P = U \cdot I = I^2 R = {U^2}/{R}$$

$$Z = R + jX \quad where \quad X = \omega L$$

$$Q = \frac{X}{R} \quad \rightarrow \quad R = \frac{\omega L}{Q}$$

$$\Rightarrow P = \frac{I^2 \omega L}{Q}$$ *…required driver power*

$$\Rightarrow \quad H \sim I \approx \sqrt{\frac{P \cdot Q_A}{\omega \cdot L_A}}$$



Determining the required RF power

- So, starting from our previous consideration (using e.g. the extended Biot-Savart law) it is possible to determine the required antenna current to achieve an intended *H*-field strength for the given antenna geometry. (Note, for a loop antenna consisting of multiple (N) turns, the current for one turn is I/N).
- Knowing Q (or current gain at carrier frequency for the antenna network filter function), knowing the antenna current and the operating carrier frequency allows to estimate the required amplifier RF output power:

$$P_D \cong I^2 \cdot R_S \quad with \quad Q_A = \frac{\omega_C \cdot L_A}{R_S}$$

$$P_D \cong \frac{I^2 \cdot \omega_C \cdot L_{ANT}}{Q_{ANT}}$$

I...............total current in the antenna geometry

ω_C...........angular carrier frequency

L_A............antenna inductance

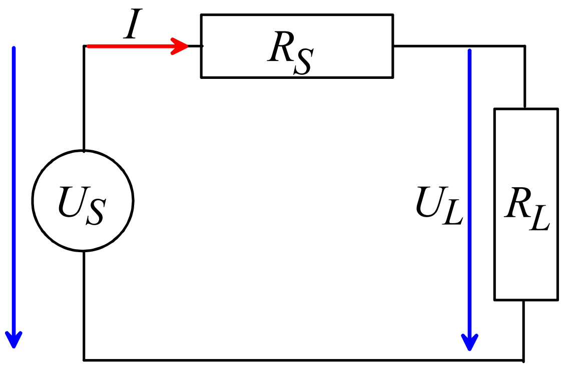
N..............number of turns

Q_A............antenna Q-factor

P_D............driver output power (at carrier)

RF power – maximum power and efficiency

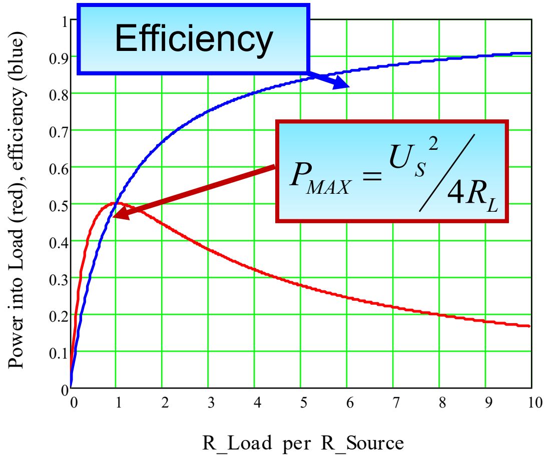
- Assuming conjugate reactance matching, the remaining problem is …



- Maximum power transfer into load… (assuming constant source resistance > 0)

$$P_L = I^2 R_L = \left(\frac{U_S}{R_S + R_L}\right)^2 R_L = \frac{U_S{}^2}{\frac{R_S{}^2}{R_L} + 2R_S + R_L}$$

$$\frac{d}{dR_L}\left(\frac{R_S{}^2}{R_L} + 2R_S + R_L\right) \equiv 0 = -\frac{R_S{}^2}{R_L{}^2} + 1 \Rightarrow R_L = \pm R_S$$



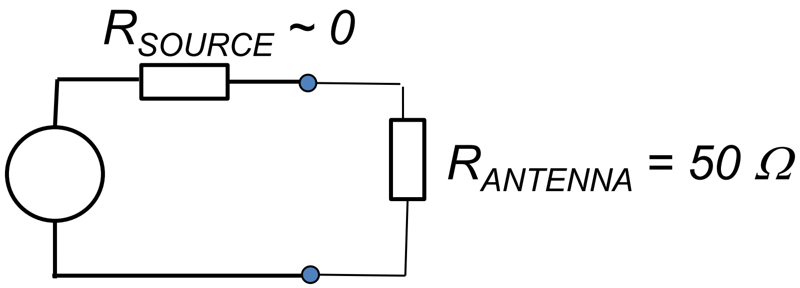
- Maximum efficiency for power transfer…

$$P_L = I^2 \cdot R_L = \left(U_S \cdot \frac{R_L}{R_S + R_L}\right)^2 \quad and \quad P_S = \frac{U_S{}^2}{R_S + R_L}$$

$$\eta = \frac{P_L}{P_S} = \frac{R_L}{R_S + R_L}, \quad for\, \eta \to 1 (= \max) \Rightarrow R_S << R_L$$

Driver concepts – Power and Q requirement

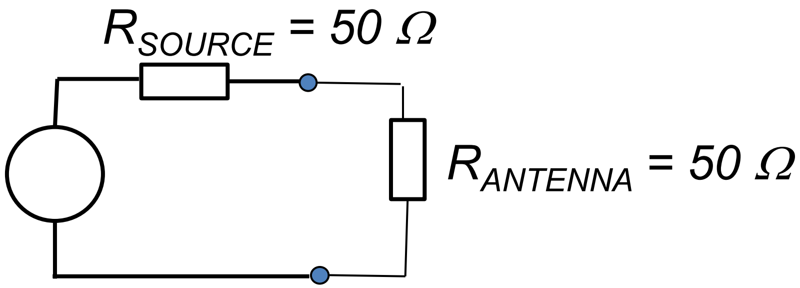
- If we consider the quality factor, it depends how the load / antenna is matched to the source / driver:



Low output impedance

$Q_{ANT} \sim Q_{SYS}$, *e.g. ~ 12,5*

$P_{SYS} \sim P_{ANT} \sim 1/Q_{ANT}$, *e.g. ~ 400 mW*



Load Matching

$Q_{ANT} \sim 2Q_{SYS}$, *e.g. ~ 25*

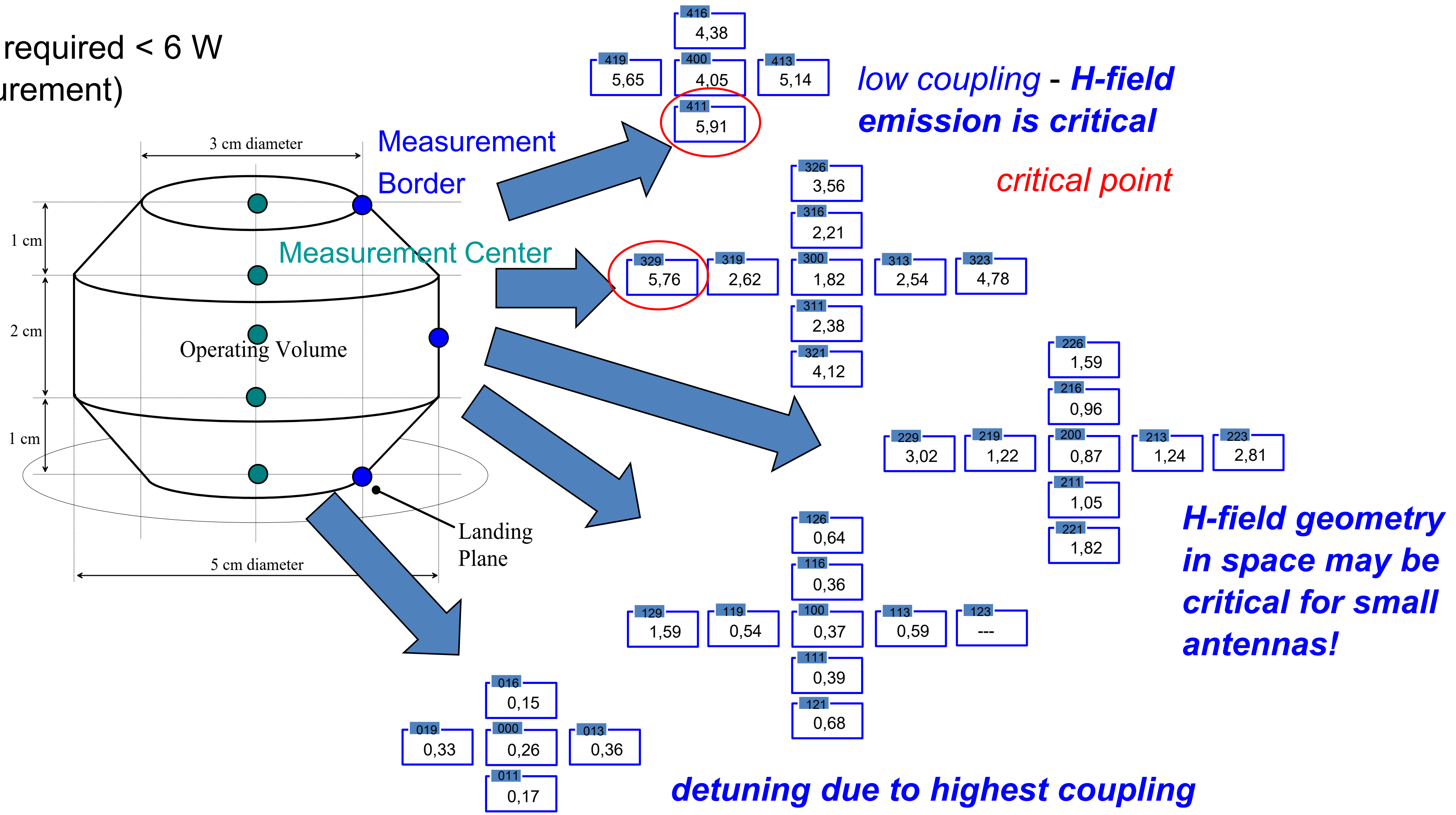
$P_{SYS} \sim P_{SOURCE} + P_{ANT} \sim 2P_{ANT} \sim 2/2Q_{ANT}$, *e.g. ~ 2 * 200 mW*

- We need a certain operational system Q to achieve time constants for modulation (e.g. Q_{SYS} ~ 12.5).
- In Load Matching, Q_{SYS} is half the value of the open antenna: Q_{ANT} can be doubled.
- The power consumed in the antenna is related to 1/Q.
- For Load Matching, the required total power is the same, as for low output impedance.
- But for low output impedance no power is dissipated in the amplifier, all in the antenna network (advantage for chip).

Test / Measure RF power requirement for antenna & MN

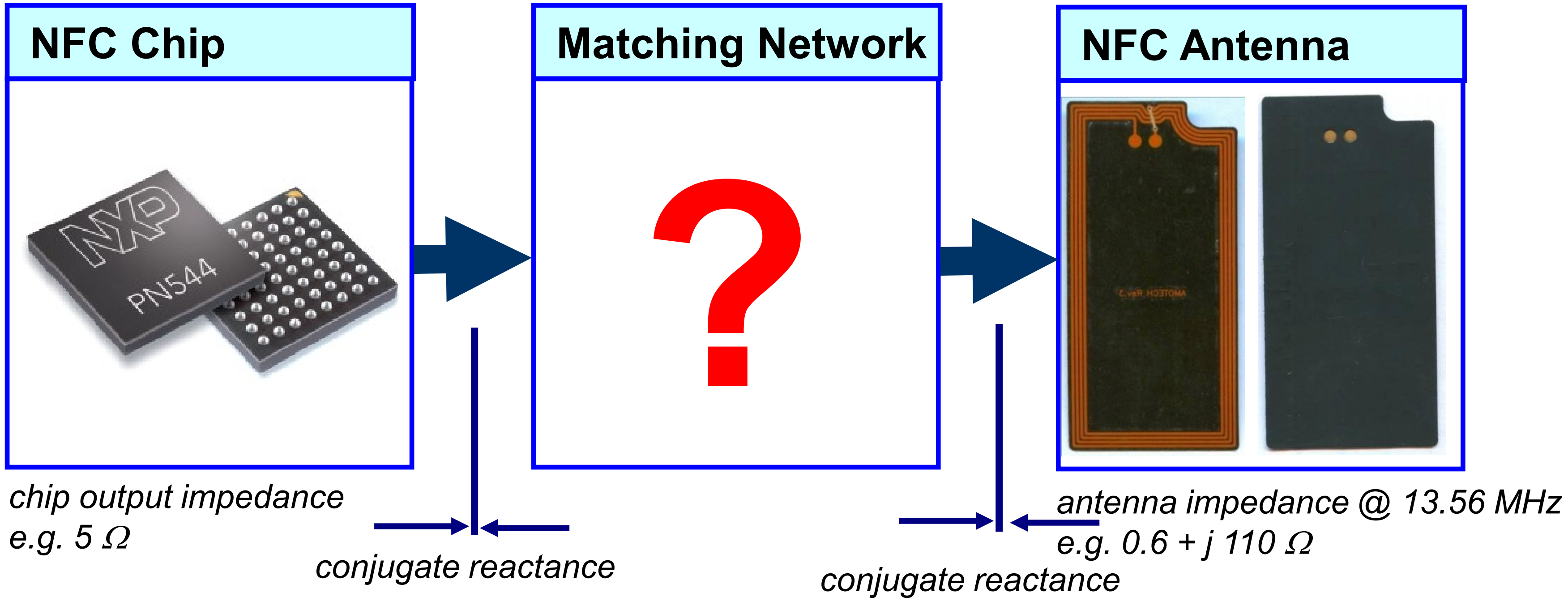
e.g. for Proximity, EMVCo Reader Hmin condition

Lab Amp. feed power required < 6 W
(no Modulation measurement)



How to match the Reader TX to the antenna?

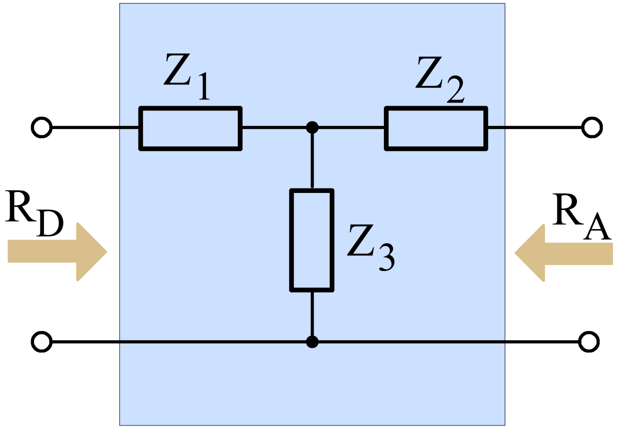
- Reader / NFC device in Reader Mode, a network adjusts (transforms) the antenna impedance to a desired value for the chip driver output. This allows optimum power transfer and to meet other contactless property requirements:

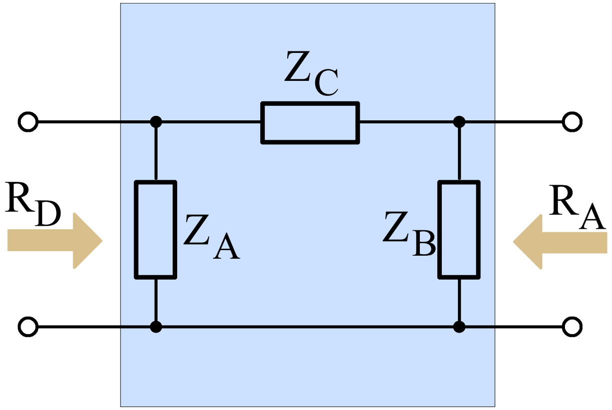


***Narrow-band** "matching" for the 13,56 MHz carrier frequency in reader mode*

Topology and design of Matching Networks

- In general, T- or π-structures of networks allow to match any arbitrary impedances to desired values (power matching of conjugate complex impedances). One network can be re-calculated (for one frequency!) to an equivalent network in the other topology.





$$Z_1 = -j\frac{R_D\cos(\varphi)-\sqrt{R_D R_A}}{\sin(\varphi)} = -j\frac{\sqrt{R_D R_A}}{\sin(\varphi)}\left(\sqrt{\frac{R_D}{R_A}}\cos(\varphi)-1\right)$$

$$Z_A = j\frac{R_D R_A\sin(\varphi)}{R_A\cos(\varphi)-\sqrt{R_D R_A}} = j\sqrt{R_D R_A}\sin(\varphi)\left(\sqrt{\frac{R_A}{R_D}}\cos(\varphi)-1\right)^{-1}$$

$$Z_2 = -j\frac{R_A\cos(\varphi)-\sqrt{R_D R_A}}{\sin(\varphi)} = -j\frac{\sqrt{R_D R_A}}{\sin(\varphi)}\left(\sqrt{\frac{R_A}{R_D}}\cos(\varphi)-1\right)$$

$$Z_B = j\frac{R_D R_A\sin(\varphi)}{R_D\cos(\varphi)-\sqrt{R_D R_A}} = j\sqrt{R_D R_A}\sin(\varphi)\left(\sqrt{\frac{R_D}{R_A}}\cos(\varphi)-1\right)^{-1}$$

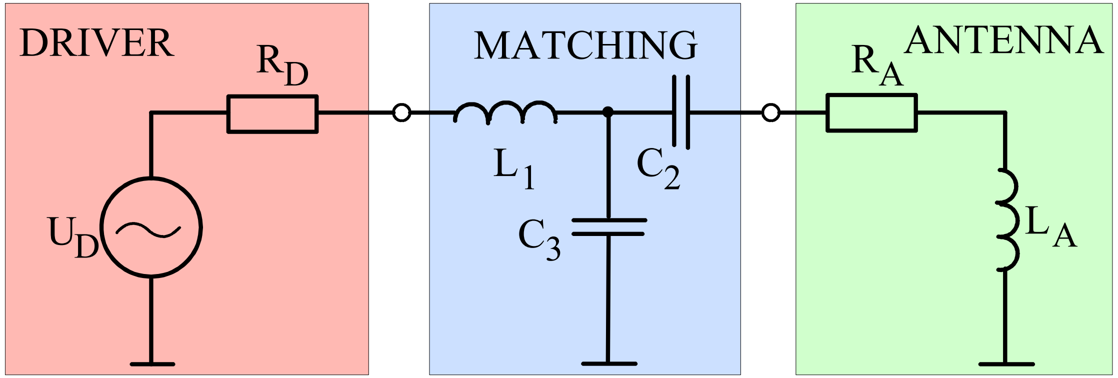
$$Z_3 = -j\frac{\sqrt{R_D R_A}}{\sin(\varphi)}$$

φ....Phasenshift output to input

$$Z_C = j\sqrt{R_D R_A}\sin(\varphi)$$

Example: Matching network in T – topology

- The following example shows a potential practical approach to this general method:

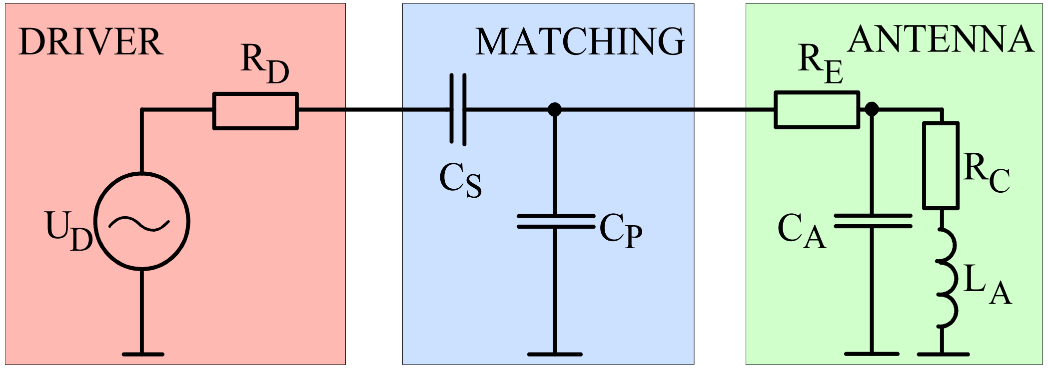


- Matching network element values can be calculated according to following method:

1. Dimension C_2 to compensate inductive antenna load, making it real $\quad \frac{1}{\omega C_2} = \omega L_A \rightarrow C_2 = \frac{1}{\omega^2 L_A}$
2. For the remaining ohmic load R_A dimension C_3 (if $R_A > R_D$) $\quad Z_3 = -j\frac{\sqrt{R_D R_A}}{\sqrt{1-R_D/R_A}} \equiv -j\frac{1}{\omega C_3} \rightarrow C_3 = \frac{\sqrt{1-R_D/R_A}}{\omega\sqrt{R_D R_A}}$
3. Dimension L_1 as follows (if $R_A > R_D$) $\quad Z_1 = j\sqrt{R_D R_A\left(1-R_D/R_A\right)} \equiv j\omega L_1 \rightarrow L_1 = \frac{1}{\omega}\sqrt{R_D R_A\left(1-R_D/R_A\right)}$

Matching network in L – topology

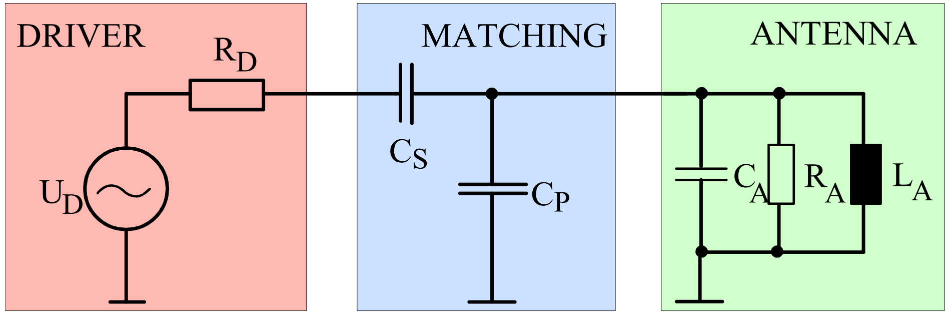
- For practical Reader applications, especially Proximity or NFC Readers, some requirements can be omitted and the matching network can be simplified to an L-structure.
  - Typical loop antennas in this context are ohmic-inductive loads (not arbitrary loads)
  - The phase relation between amplifier output current and antenna *H*-field is irrelevant
  - Power losses in capacitors are negligibly small, while losses in inductors can be relevant



- To dimension the matching capacitors, convert the antenna EQC in a parallel resonance circuit structure
- Check, if the antenna can be matched (is the square-root expression real?)
- Calculate capacitances from the conditions for the impedance (according the method shown on following pages)

Matching network in L – topology

- Calculate the electrical load connected to the driver output and set it equal to the driver source impedance (R_D):



$$R_D \equiv \frac{s^2\left(R_A L_A C_S + R_A L_A \left(C_A + C_P\right)\right) + sL_A + R_A}{s^3 R_A L_A C_S \left(C_A + C_P\right) + s^2 L_A C_S + sR_A C_S}$$

- Split the equation up in real and imaginary part (s ➔ jω). Solve for C_S. From this solve C_P as quadratic equation:

$$p = \frac{2\omega^2 L_A C_A - 2}{\omega^2 L_A} \qquad q = \frac{R_D R_A\left(\omega^4 R_A L_A{}^2 C_A{}^2 - 2\omega^2 R_A L_A C_A + 1\right) - \omega^2\left(R_A - R_D\right)}{\omega^4 R_G R_A{}^2 L_A{}^2}$$

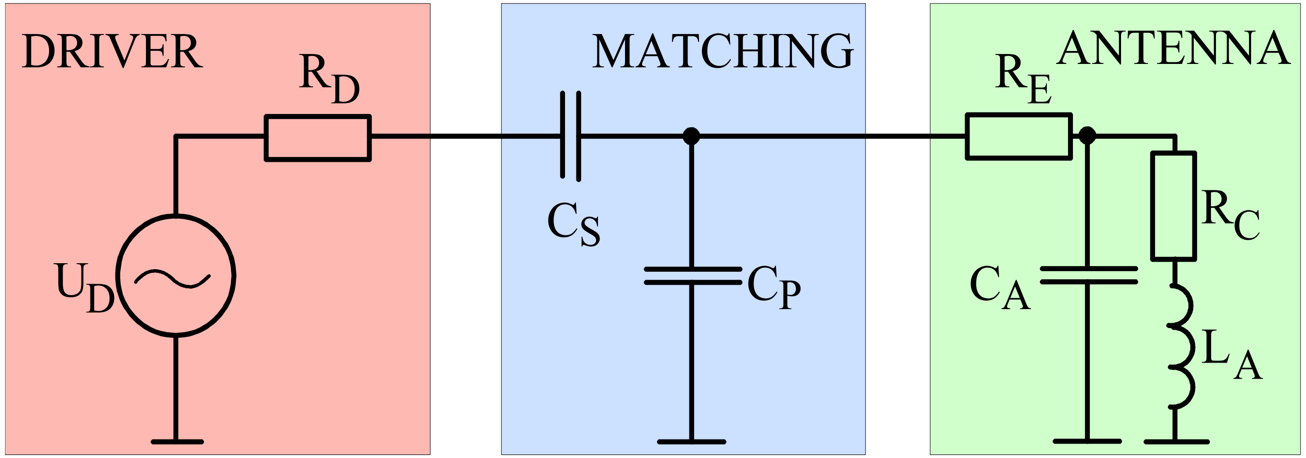
- C_P can be solved, if $\frac{p^2}{4} - q \geq 0$

$$C_{P\,a,b} = -\frac{p}{2} \pm \sqrt{\frac{p^2}{4} - q}$$

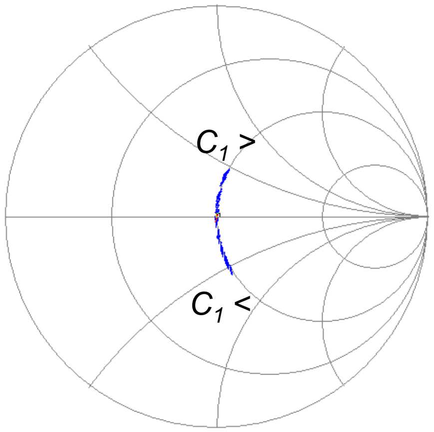
- C_S can be solved from C_P (only positive, real caps are physically meaningful)

Practical RF system work with matching network in L – topology

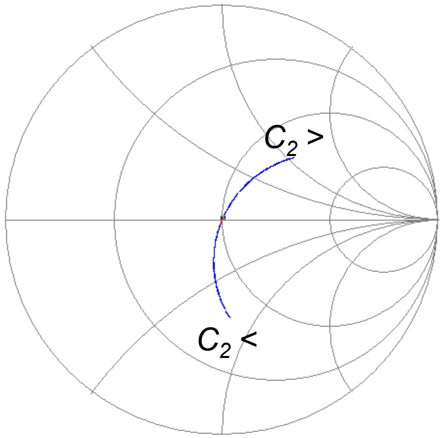
- The exact formulas to calculate C_P and C_S can also be implemented in an Excel sheet, or in mathematical software.
- For practical work, also circuit simulator software / freeware, such as LTSpice or RFSim99 can be helpful.
- The result, in practice, is only as accurate, as the extracted antenna EQC data allows. So, an approximation can be a good starting point, and the RF System engineer should also investigate tolerances (e.g. multi-sweep…). Everything should be confirmed by impedance measurement…



Serial capacitance



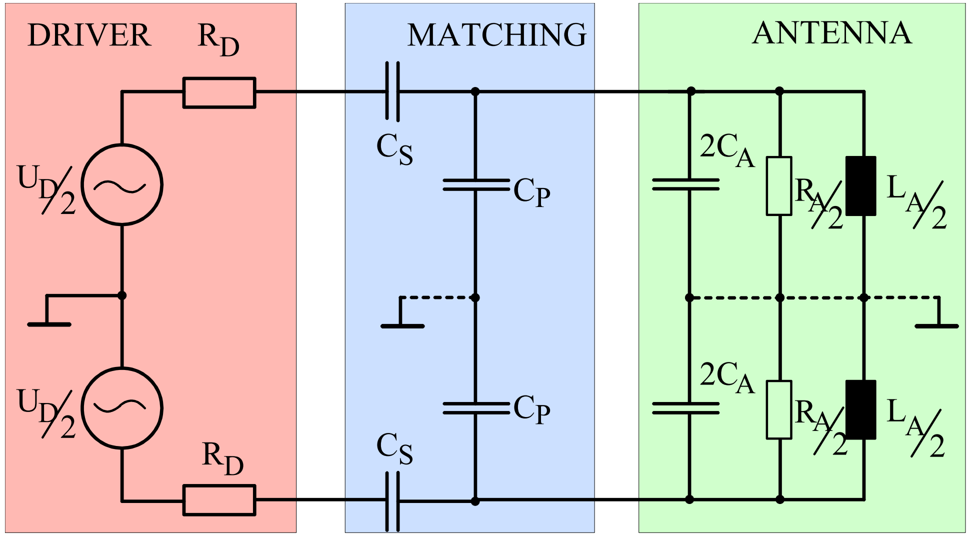
Parallel capacitance



Differential matching network for L – topology

- In practice, integrated Reader ICs typically have a differential output.
- From a limited supply voltage, and a not negligible output resistance (e.g. 5…2 Ohms) this allows higher efficiency.
- Same considerations as before can be used to calculate matching capacitors, if we split up the antenna in 2 half parts:

- For the driver output impedance, we consider the resistance against GND.

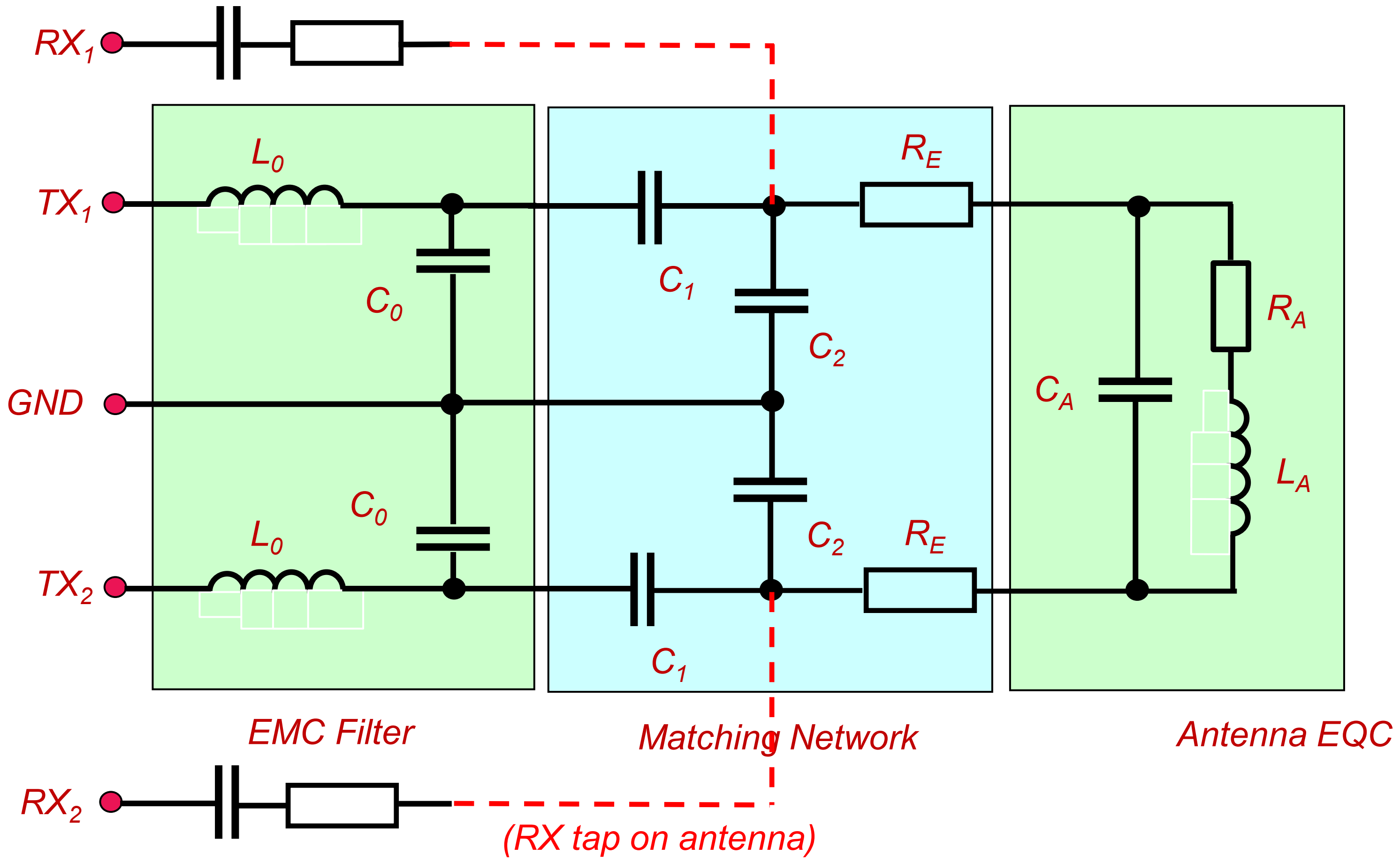


- We introduce a virtual GND (0 V_{AC}) in the antenna middle. For one branch of the antenna, the EQC values are modified as shown in the schematic:
  - Resistance and inductance get the half value
  - Capacitance gets double value

- One branch of the MN can then be calculated as before for the asymmetric circuit.
- Typically GND of the driver is connected to the MN. It can be better, not to connect the virtual GND of the antenna though, as any asymmetry in design or by capacitive coupling else may cause unwanted compensation currents.

Matching network in L – topology with EMC filter

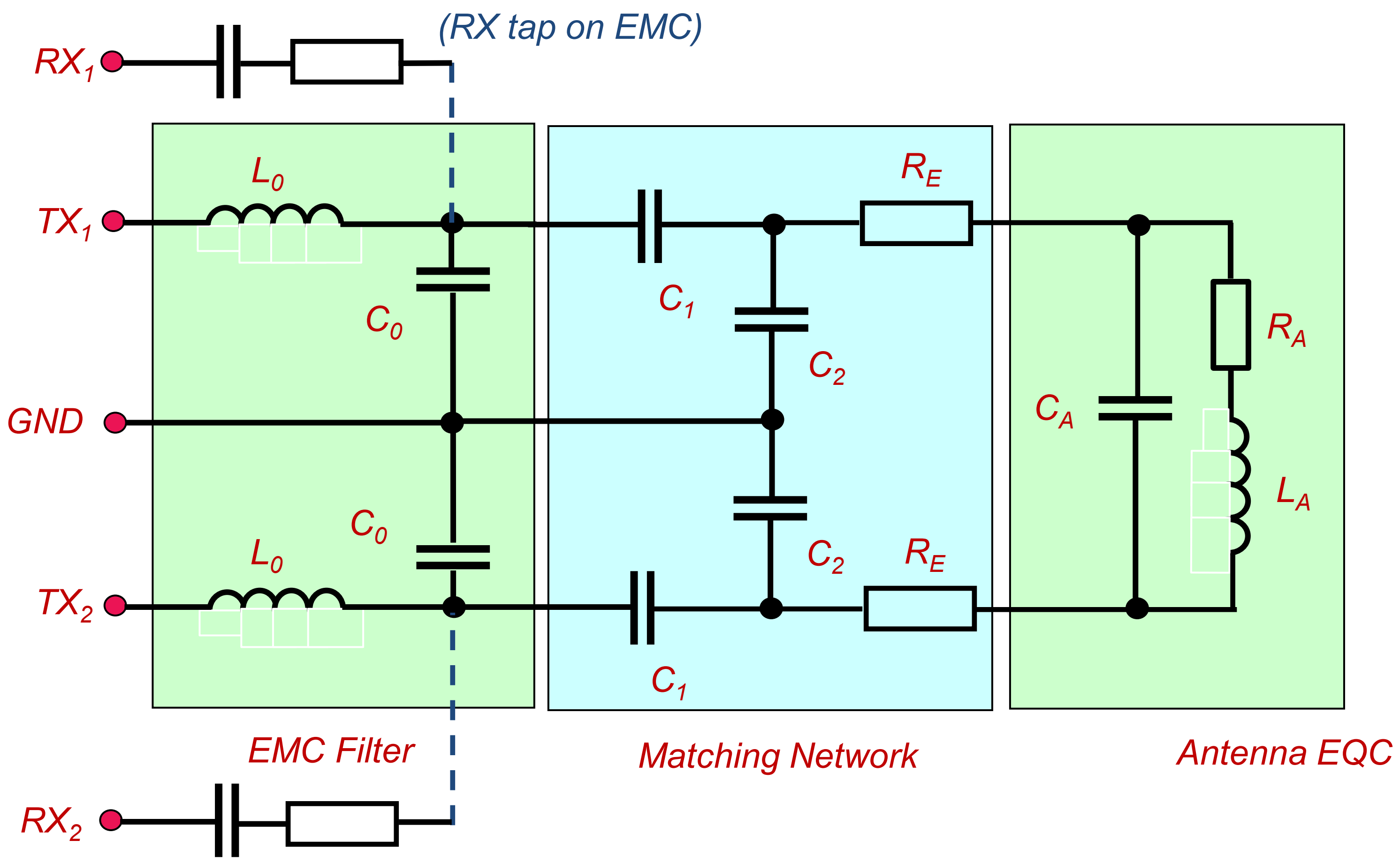
- Very commonly used, the antenna matching network for integrated NFC Reader ICs have this structure:



- Also the RX input can be differential.

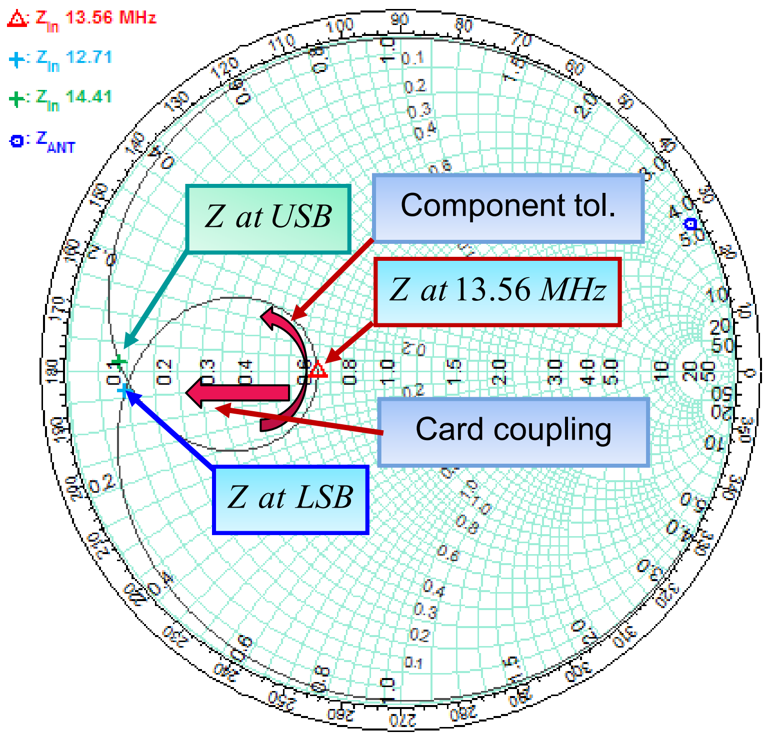
Matching network in L – topology with EMC filter

- Very commonly used, the antenna matching network for integrated NFC Reader ICs have this structure:



- Also the RX input can be differential.
- E.g. for eval-boards, the antenna can be separated from the reader board.
- Typically the separation is done between the EMC filter, the rest of the matching remains with the antenna.

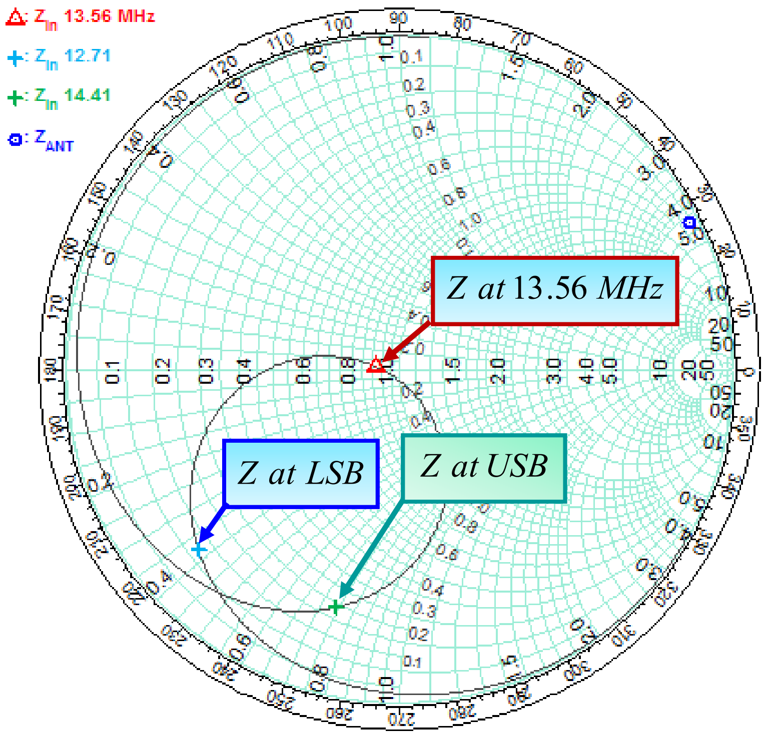
Matching: Symmetric parallel resonance



- ❑ *Note: Preferred option*

- ❑ **EMC filter design requirement:**
  - – resonance close to carrier (~ 14.5 MHz)
- ❑ **Component tolerance**
  - – less critical, Load Z deviation is mainly imaginary
- ❑ **Coupling to a Card**
- o RF current (for Card Loading)
  - – strong increase (e.g. x 5)
- o Reader Modulation Envelope
  - – steep edges, tends to overshoot
- o Card Sideband Amplitudes
  - – USB and LSB are symmetric, increase with coupling

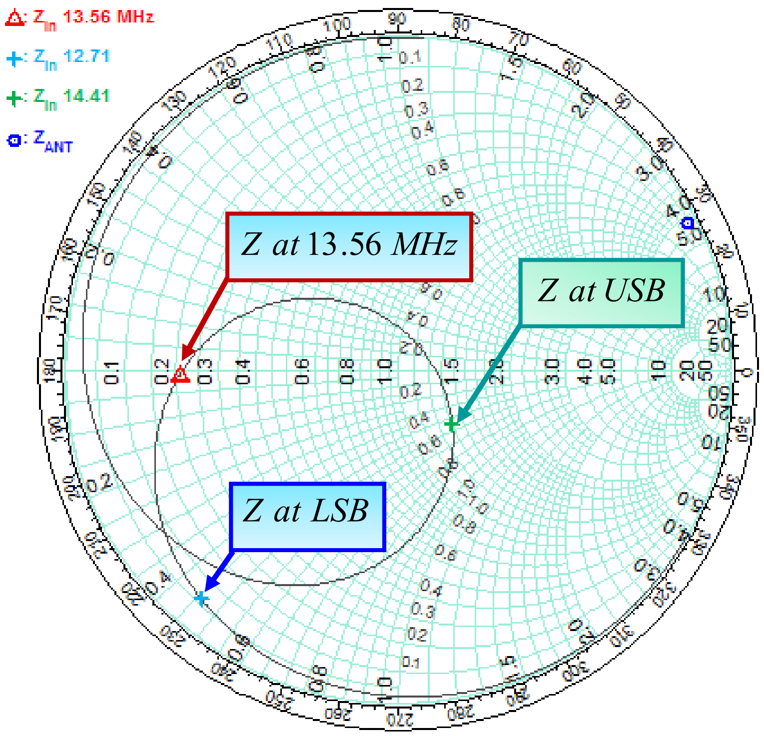
Matching: Asymmetric parallel resonance



- *Note: Can adjust only to higher matching impedances!*

- **EMC filter design requirement:**
  - higher resonance (e.g. 17 MHz)
- **Component tolerance**
  - more critical, Load Z deviation is mainly resistive (e.g. 2 pF => 30 Ω ➔ 70 Ω)
- **Coupling to a Card**
- RF current (for Card Loading)
  - May stay *relatively* constant (e.g. < x 2)
- Reader Modulation Envelope
  - medium / slow edges, low overshoots
- Card Sideband Amplitudes
  - USB and LSB are asymmetric, any variation with coupling

Matching: Asymmetric parallel resonance



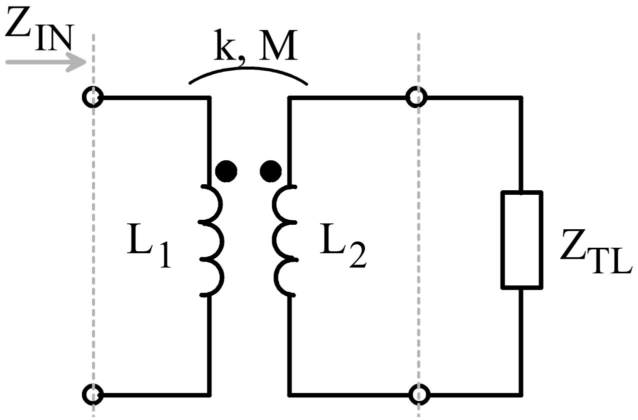
- ❑ *Note: Can adjust to low matching resistance!*

- ❑ **EMC filter requirement:**
  - – higher resonance (e.g. 17 MHz)
- ❑ **Component tolerance**
  - – less critical, Z deviation is mainly imaginary
- ❑ **Coupling to a Card**
- ○ RF current (for Card Loading)
  - – no experience
- ○ Reader Modulation Envelope
  - – very slow edges expected
- ○ Card Sideband Amplitudes
  - – USB and LSB very asymmetric, operates with USB only

Matched Reader loop antenna in coupling scenario

- So, by dimensioning the EMC filter, we can determine the impedance trace around resonance, the way, how the reader should perform, especially in a coupling with a Transponder Card, when it gets de-tuned (antenna Z is changed!).
- The typical approach is, in a first step to dimension the EMC filter. Then, the remaining MN is adjusted to the impedance at the EMC filter output.
- Coupled impedances can be represented by an equivalent T-network:

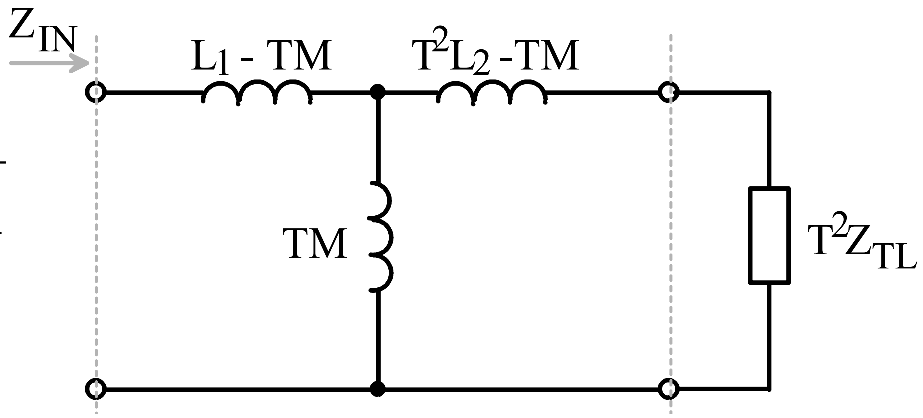
$$k = \frac{M}{\sqrt{L_1 L_2}}$$



k.....coupling

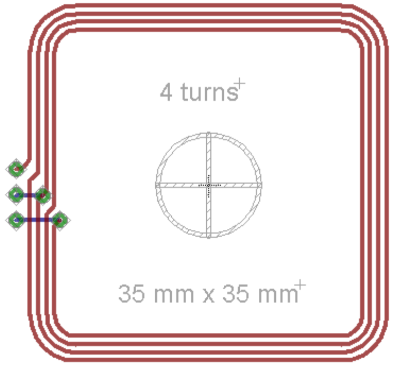
M....mutual inductance

$$T = \sqrt{\frac{L_1}{L_2}}$$



T.....Transformation

Matched Reader loop antenna in coupling scenario



- Step by step, let us consider the impedance of loop antenna and matching structure:

Z_A

$$f_{RES} >> f_{CAR}$$

$$Q_A >> \quad (> Q_{SYS})$$

$$\boxed{Z_A = \frac{R_A + j\omega L_A}{1 + j\omega R_A C_A - \omega^2 L_A C_A}}$$

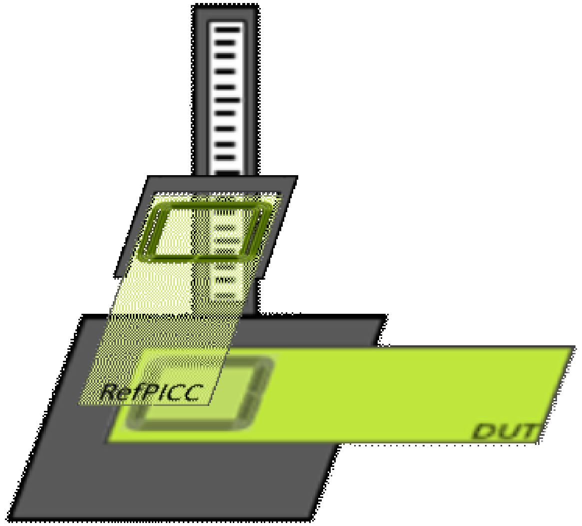
- An equivalent circuit of the loop antenna may have the above structure.
- It can be extracted from measurement.
- Complex antenna impedance Z_A can be calculated (over angular frequency ω).

$$Z_A = \frac{0.58\Omega + j\left(2\cdot\pi\cdot 13.56\cdot 10^6\,Hz\cdot 1.314\cdot 10^{-6}\,H\right)}{1 + j\left(2\cdot\pi\cdot 13.56\cdot 10^6\,Hz\cdot 0.58\Omega\cdot 2.35\cdot 10^{-12}\,F\right) - \left[\left(2\cdot\pi\cdot 13.56\cdot 10^6\,Hz\right)^2 1.314\cdot 10^{-6}\,H\cdot 2.35\cdot 10^{-12}\,F\right]}$$

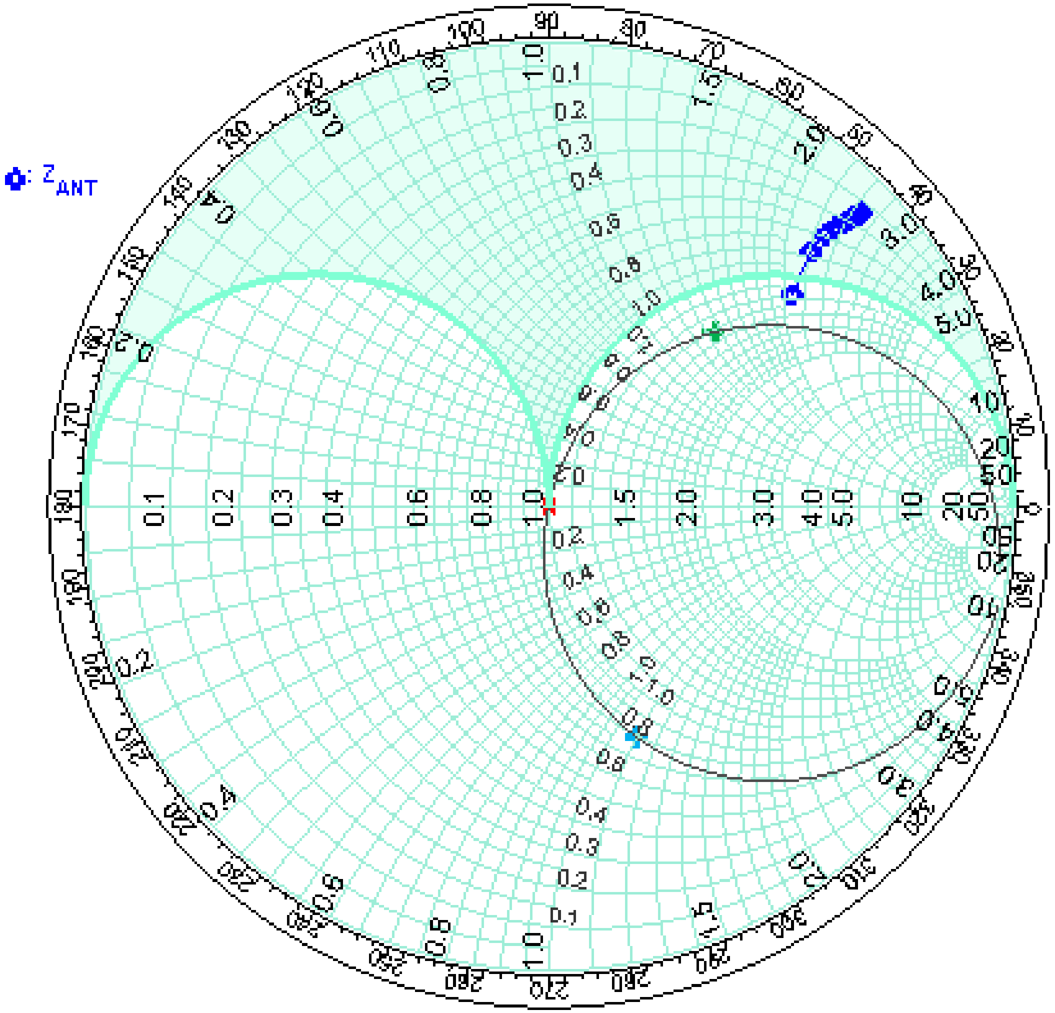
$$Z_A(@13.56\ MHz) = 0.607 + j114.52$$

Antenna coupling – antenna only

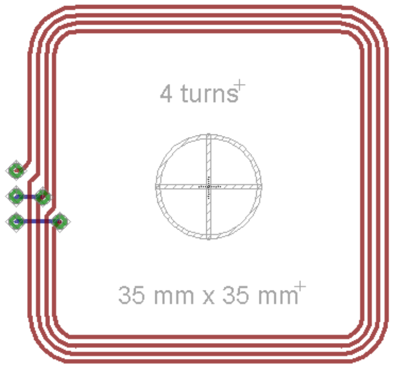
- Coupling affects the antenna impedance significantly
  - mutual inductance ➔ L_A changes
  - „Loading“ ➔ Q changes
- Impedance trace is shown in Smith Chart



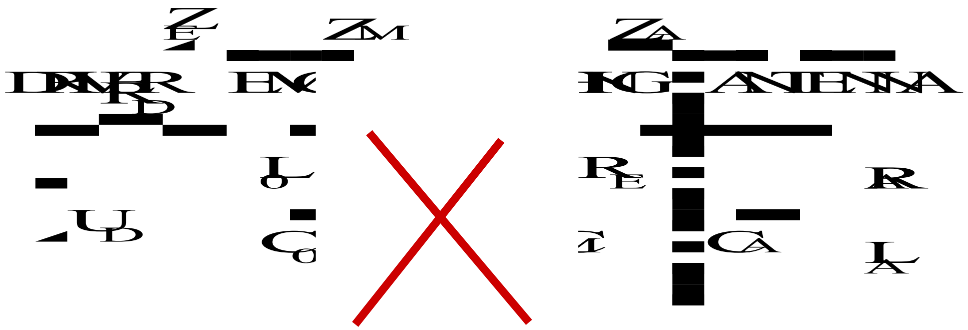
distance variation to “loading” antenna



Impedance adjustment using L-topology



- Step by step, let us consider the impedance of loop antenna and matching structure:



No EMC

$$f_{RES} \equiv f_{CAR}$$

$$\mathrm{Im}\{\underline{Z}\} \equiv 0$$

$$\mathrm{Re}\{\underline{Z}\} \equiv R_{DESIRED}$$

$$Z_A = \frac{R_A + j\omega L_A}{1 + j\omega R_A C_A - \omega^2 L_A C_A}$$

$$C_{SM}(C_{PM}) = \frac{R_A L_A}{(\omega L_A)^2 R_D \left[1 - \omega^2 L_A (C_A + C_{PM})\right]}$$

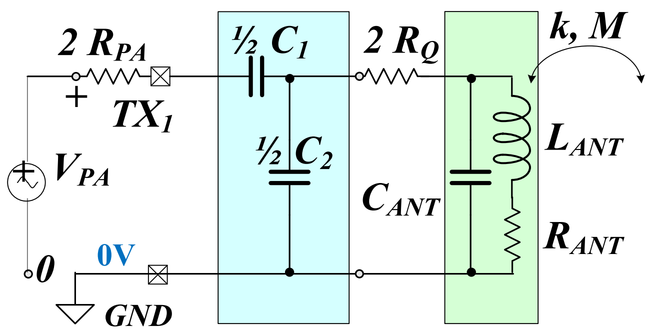
$$C_{PM} = -\frac{p}{2} \pm \sqrt{\frac{p^2}{4} - q}$$

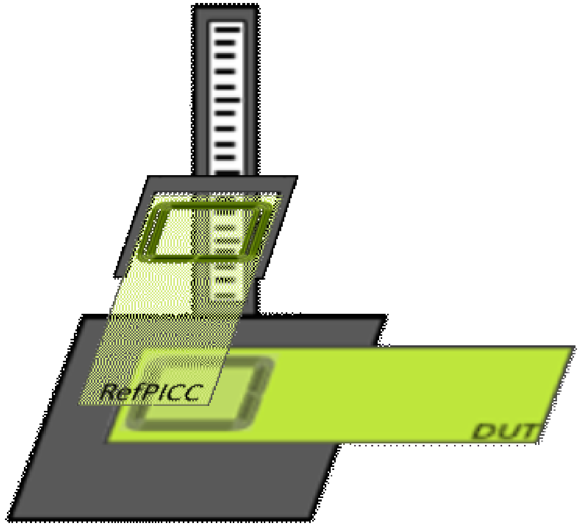
$$Z_M = Z_E = \frac{1 + j\omega Z_A (C_{SM} + C_{PM})}{\omega \left(jC_{SM} - \omega C_{SM} C_{PM} Z_A\right)}$$

$$p = \frac{2\omega^2 L_A C_A - 2}{\omega^2 L_A}$$

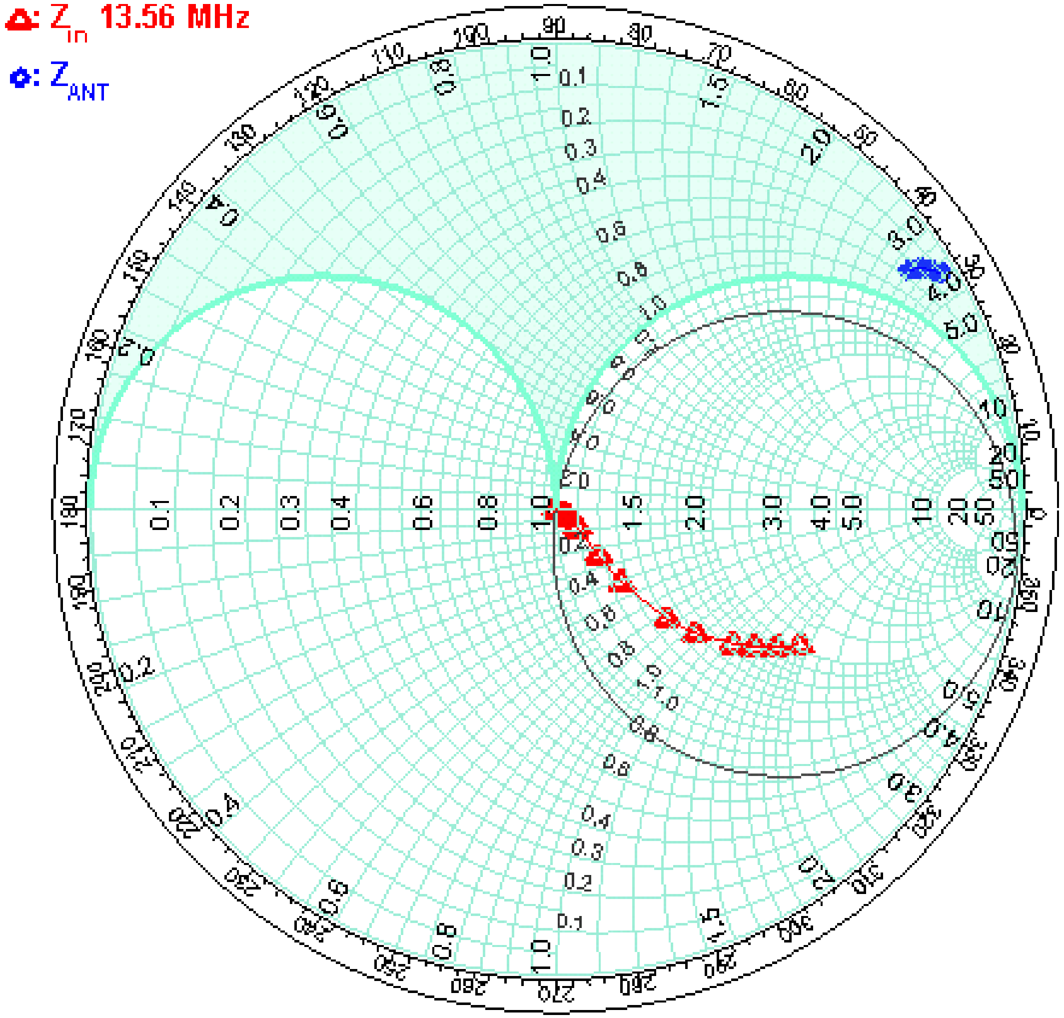
$$q = \frac{R_A (1 + R_A)}{(\omega L_A)^6} - \frac{R_A}{\omega^4 R_D L_A{}^4} - \frac{2 C_A}{\omega^2 L_A} + C_A{}^2$$

Antenna coupling – no EMC filter



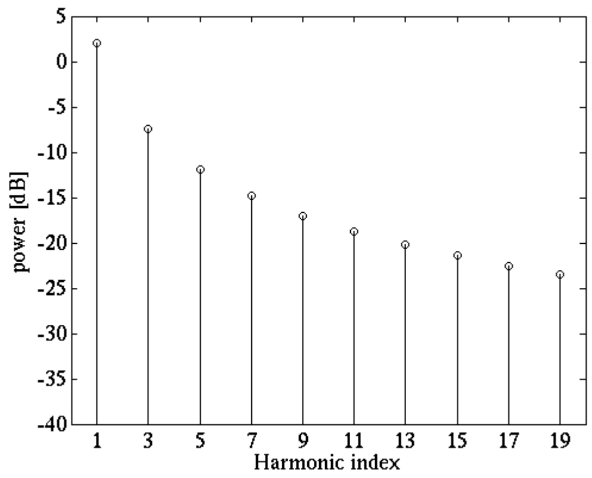


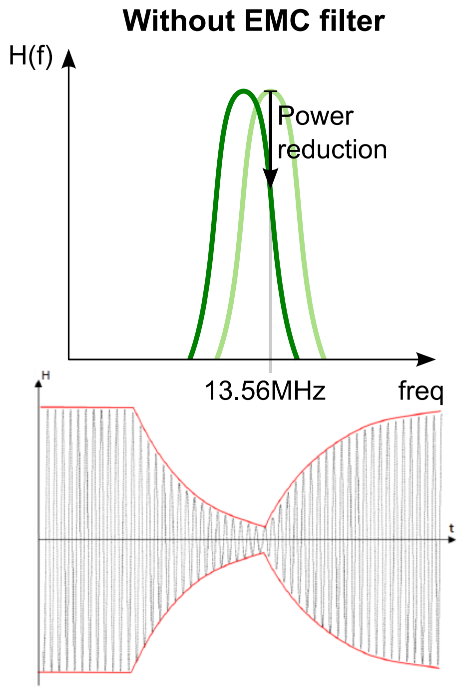
distance 80 … 3 mm

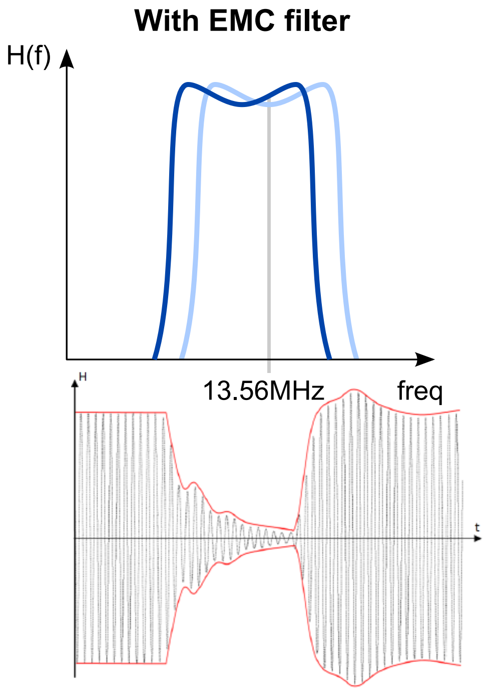


Reasoning to extend the network

- Due to Standard (e.g. ISO/IEC14443) requirements for modulation timing, the allowable Q-factor for the operated system is rather low (~limited bandwidth).
- For efficient *H*-field emission, a higher reactive current is desirable ➔ a 2nd resonance can increase the bandwidth.
- Furthermore, this 2nd resonance (LC low-pass) can suppress unwanted harmonics emission (➔ name EMC filter).



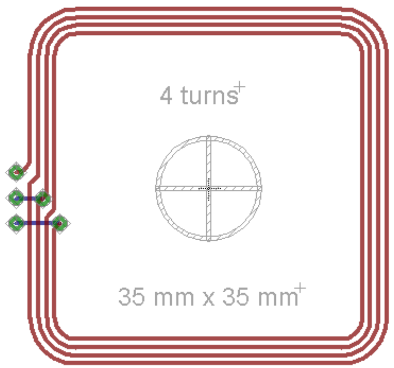


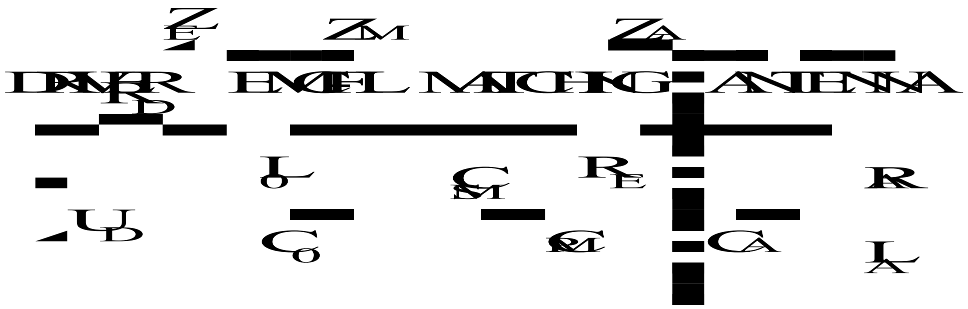


- This comes on the expense of more signal distortions…

Impedance adjustment with EMC filter

- Step by step, let us consider the impedance of loop antenna and matching structure:





$$f_{RES} \equiv f_{CAR}$$

$$\mathrm{Im}\{\underline{Z}\} \equiv 0$$

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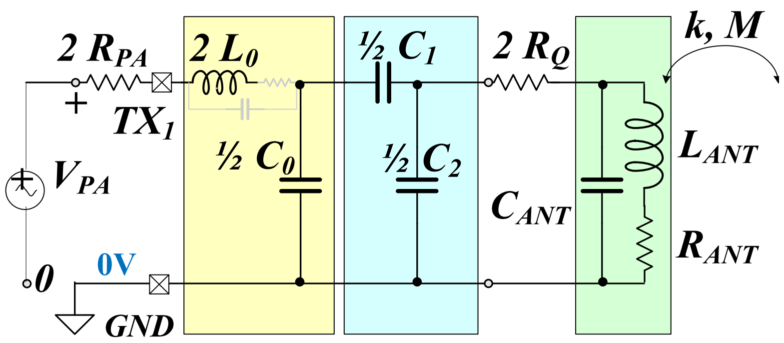
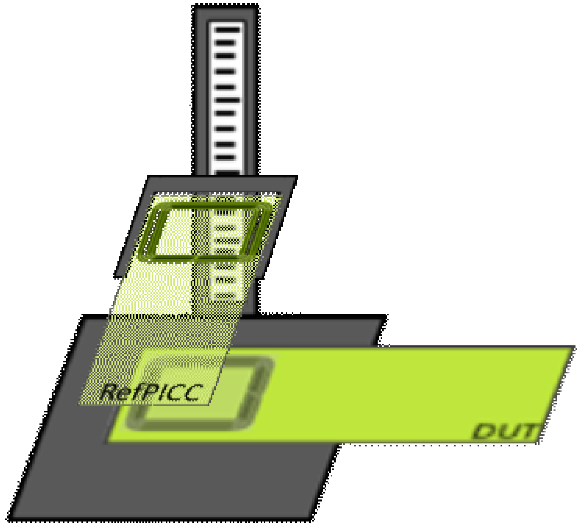
$$Z_A = \frac{R_A + j\omega L_A}{1 + j\omega R_A C_A - \omega^2 L_A C_A}$$

$$Z_M = Z_E = \frac{1 + j\omega Z_A (C_{SM} + C_{PM})}{\omega (jC_{SM} - \omega C_{SM} C_{PM} Z_A)}$$

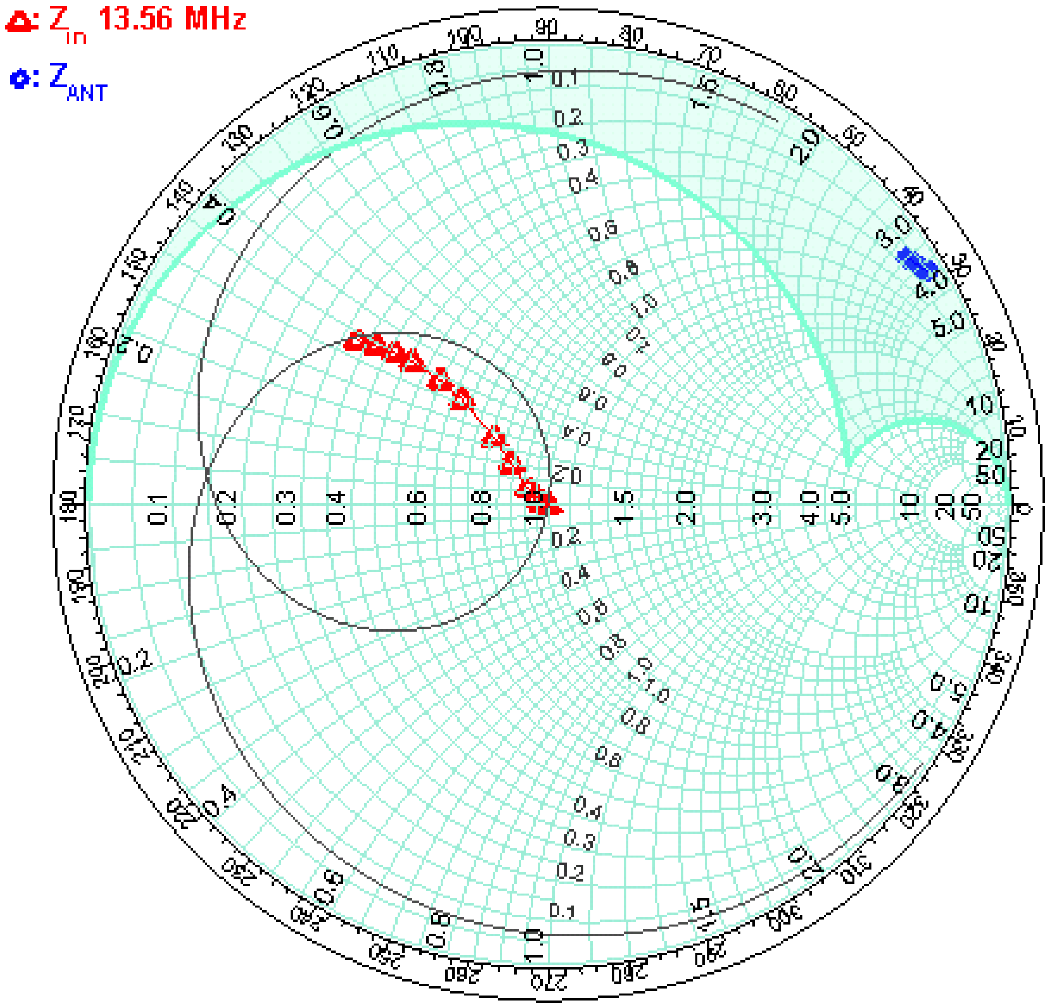
$$Z_E = \frac{Z_M + j\omega L_0}{j\omega C_0 Z_M - \omega^2 L_0 C_0}$$

Antenna coupling – with EMC filter



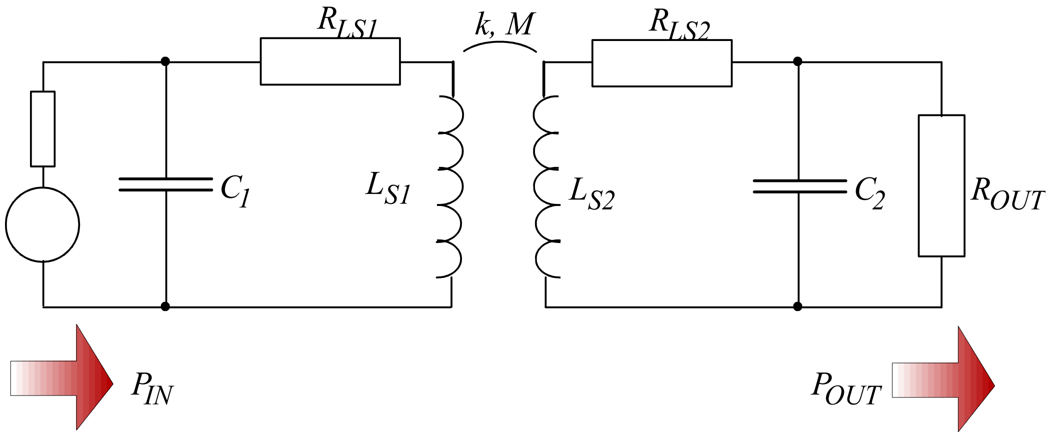


distance 80 … 3 mm



Power consumption and signal distortion in coupling scenario

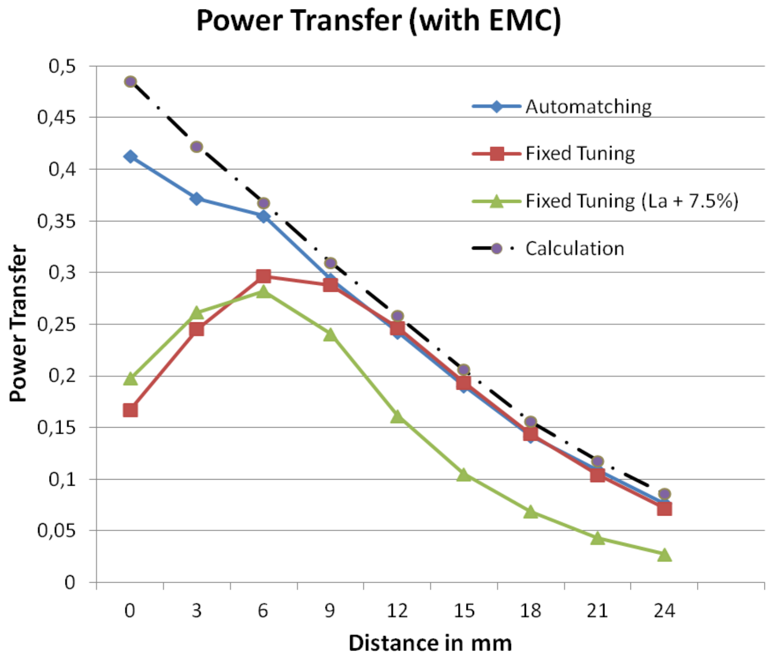
- As we see, coupling to a Transponder Card changes the impedance, connected to the driver output.
- While for the MN without EMC filter, impedance is increased (➔ lower power), in the MN with EMC filter, impedance gets lower (➔ more power required).
- This can be a problem, causing driver overheating, and modulation signal distortion (if driver protection limits the signal).
- This was one reasoning to study automated MN adjustment, and led to the introduction of the xxxx feature.



$$\eta_{MAX} = \frac{k^2\, Q_{LS1}\, Q_{LS2}^{\;2}}{\left(1 + \frac{Q_{LS2}}{Q_C} + k^2\, Q_{LS1}\, Q_{LS2}\right)\cdot\left(Q_C + Q_{LS2}\right)}$$

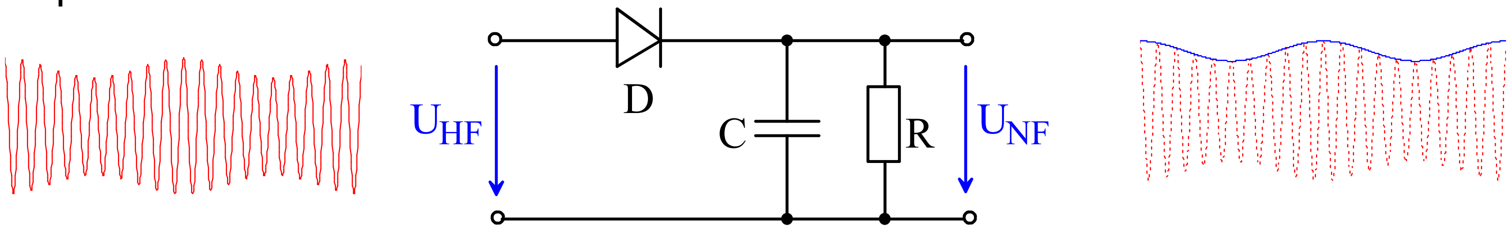
for $Q_L = \frac{\omega_{RES} L}{R_S}$

and $Q_C = \omega_{RES,P}\, C_2\, R_{L2}$

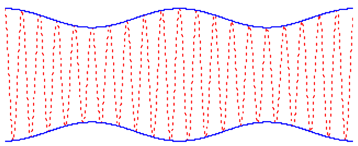


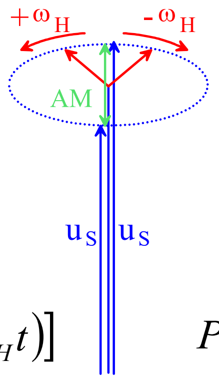
Receiver concepts

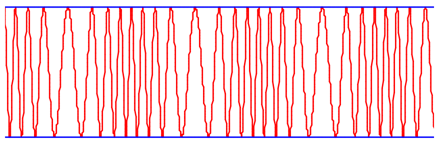
- A very simple option is, to use a diode-based envelope demodulator. Such AM demodulator concepts could be found e.g. in early Asian reader products.

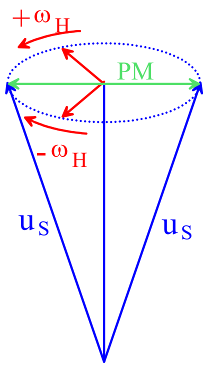


- Load modulation, due to the coupling system, can generate a combination of amplitude modulation and phase modulation. In specific scenarios, only PM may appear, and this is a problem, as it cannot be demodulated in this way.
- This would cause a "communication hole" or fail, for a certain distance between reader and transponder.







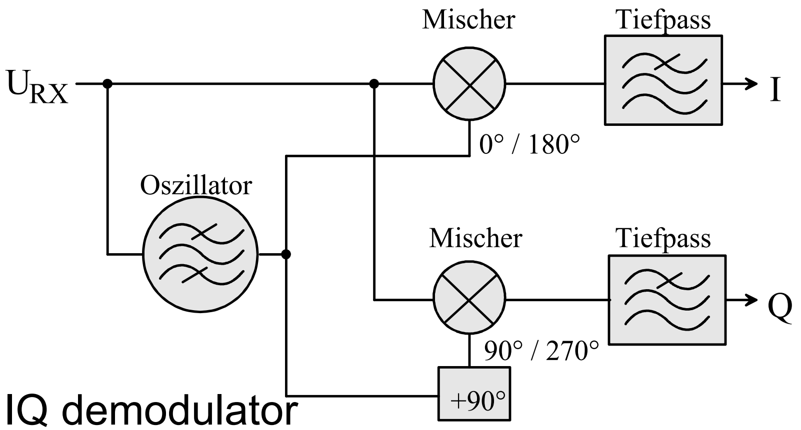


$$AM(t) = A \cdot \cos(\omega_C) \cdot [1 + m \cdot \cos(\omega_H t)]$$

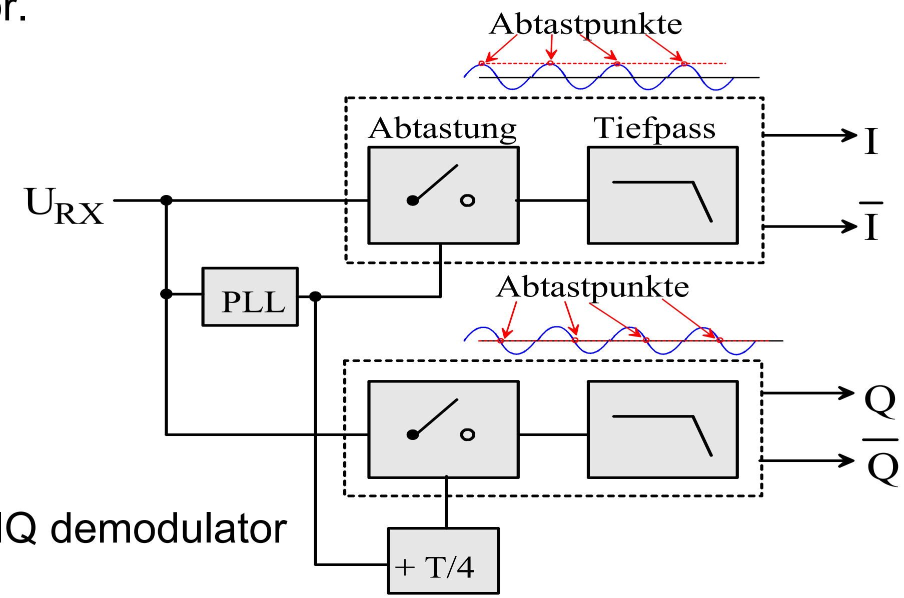
$$PM(t) = A \cdot \cos[(\omega_C) + \eta \cdot \sin(2\omega_H t + \varphi_0)]$$

Receiver concepts

- There are more advanced receiver concepts, which can overcome this problem.
- An IQ demodulator is well suited for these signals. It combines an in-phase channel (I), operating as AM demodulator, with a quadrature-phase channel (Q), operating as PM demodulator.



- Analog IQ demodulator

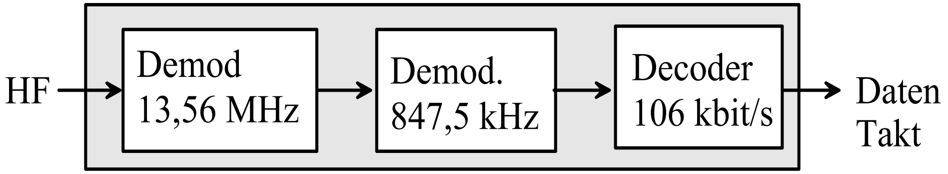


- Digital IQ demodulator

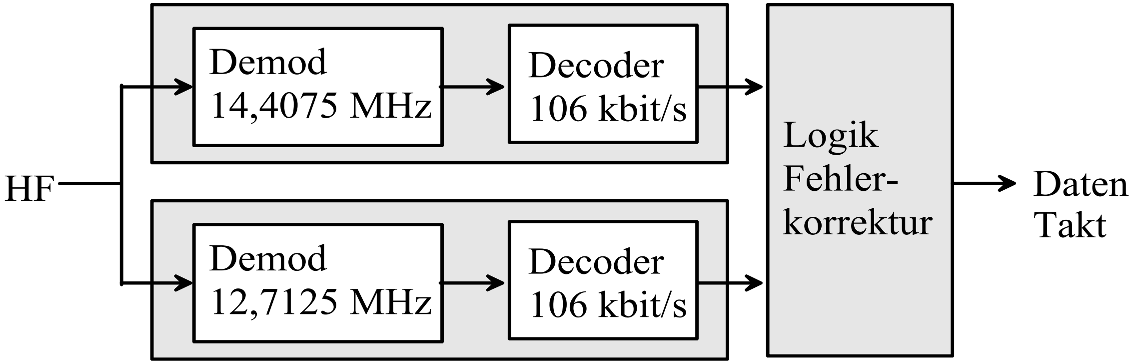
- Combining both demodulator channels, ideally $\sqrt{I^2+Q^2}$ gives the signal phasor, the theoretically best signal.
- As it is difficult to implement this, practically signals can also be simply added. (A variety of names exist at different chip manufacturers).

Data regeneration

- As the transponder information appears in 4 ranges in the frequency spectrum (due to the sub-carrier concept with load modulation), there are different options for signal regeneration. Implemented error recognition techniques can also help, to select the optimum method for the given environment.
- The concept of adding all signals in a 2-stage demodulation gives the most RX energy, but any distortion is in the signal.



- Another option is to have 2 chains with 1-stage demodulation, and to logically combine data, using also error recognition.



- Another option is to have 4 side-band channel demodulators & decoders, or to evaluate symbols in frequency domain.

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Questions for self-evaluation

- How does the Frequency Spectrum look like, for a 13.56 MHz Proximity Reader? Where is the Information content sent back by the Card?
- Which aspects need to be considered, to determine a good value for the Reader loop antenna Q-factor? Which values are typical for Proximity and Vicinity Readers?
- What is the aim of a Matching Network, how is it implemented, how does it work?
- Which power losses are there, from the driver to the loop antenna? Regarding impedance, which options are there to connect the loop antenna to the driver? What is the difference between the antenna Q-factor and the system Q-factor? Calculate the RF feed power as function of the Q-factor of a resonant loop antenna!
- Using an external resistor, how can the Q-factor of a resonant loop antenna be reduced to a desired value? How does antenna inductance change, if 2 instead of 1 turn are used?

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