04 Matching Networks
4th unit in course 3, *RF Basics and Components*

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Content

- Introduction: How to match the chip output to the antenna impedance?
  - Loop antenna
  - Quality factor
  - Low impedance or load matching: Power and Q
  - Modulation envelope and Qrequirement (single resonance)
  - Network transformation: differential to single-ended

- The general matching solution: $\pi$ and $T$-topology networks

- Impedance adjustment with $L$-topology
  - Antenna impedance
  - $Q$-factor adjustment for operating frequency
  - Determination of serial and parallel capacitance for impedance adjustment

- The resonant coupling system
  - How near-field coupling affects the air interface
How to match the chip to the antenna?

- NFC device in Reader Mode (TX), 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**
The loop antenna is a **distributed component** with inductance \( L \) as main element and capacitance \( C \) and resistance \( R \) as parasitic network elements.

For simulation it must be represented by an **equivalent circuit** network of **lumped elements**. Over a wide frequency range this can be a loose coupled reactive ladder network of resonance circuits - it has several resonances in frequency domain.

At 13.56 MHz carrier frequency we use the **fundamental (lowest) resonance**. So we can simplify the equivalent circuit e.g. to a parallel resonance circuit (since losses are mainly determined by chip current consumption in Proximity Systems).

**Note:** This is a **narrow-band approximation** only!
The Quality factor

Originally, the quality factor reports the quality at which the component (coil, capacitor) represents the pure network element (inductance, capacitance).

For a 2nd order resonant LCR circuit, Q can be determined in **frequency domain**. Q is related to the bandwidth and can be measured from the Resistance trace.

\[
Q(\omega) = \frac{\omega L_S}{R_S} \quad \text{and} \quad Q(\omega) = \frac{\omega C_p}{R_p}
\]

**Note:** This is only valid, if the broad band equivalent circuit representation really is a parallel resonant circuit!

For a 2nd order resonant LCR circuit, Q can also be determined in **time domain**. Q is related to the envelope according to

\[
e(t) = Ae^{-t/\tau} = Ae^{-t \cdot \frac{\omega_{RES}}{2Q}} = Ae^{-(\zeta \omega_{RES})t}
\]

\[
Q(\omega_{RES}) = \frac{\omega_{RES} \tau}{2}
\]
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. $\sim 12.5$

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

**Load Matching**

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

$P_{SYS} \sim P_{SOURCE} + P_{ANT} \sim 2P_{ANT} \sim \frac{2}{2Q_{ANT}}$, e.g. $\sim 2 \times 200$ mW

We need a certain operational system $Q$ to achieve time constants for modulation (e.g. $\sim 12.5$). 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.
Power and $Q$ requirement

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

\[ P = U \cdot I = I^2 R = \frac{U^2}{R} \]

\[ Z = R + jX \quad \text{where} \quad X = \omega L \]

\[ Q = \frac{X}{R} \quad \Rightarrow \quad R = \frac{\omega L}{Q} \]

\[ \Rightarrow P = \frac{I^2 \omega L}{Q} \]

\[ \Rightarrow H \sim I_{\text{ANT}} \approx \sqrt{\frac{P \cdot Q_{\text{ANT}}}{\omega \cdot L_{\text{ANT}}} \cdot N} \]
Modulation envelope and $Q$ requirement

- To emit high $H$-field from low driver power, $Q$ should be high,
- To meet modulation timing specifications, $Q$ should be low
- $Q$ is only clearly defined for a single-resonance circuit.
- For this we get for ISO/EC14443 Type B specifications a maximum system $Q$…

- To note: Modulation index and overshoots may be even more stringent!
Modulation envelope and Q requirement

- For Type A specifications...

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<td>0.58</td>
<td>t1/2 +</td>
<td>t1</td>
<td>40.73</td>
</tr>
</tbody>
</table>

(1) quality factor based on fall time t1 - t2
(2) quality factor based on fall time t1 - t5
(3) t_x = (t1 - t5) - 3/fc
(4) t_y = (t1 - t5) + 8/fc
(5) t_z = (t1 - t5) + 4.5/fc
Network transformation: single-ended, differential

- In practice, the TX output is usually differential (to be able to have double output voltage swing from a single supply voltage).

- For simplicity reasons, RF System behaviour can be considered for a single-ended network. Component values for the differential network can then be calculated by the following considerations:
  - Source and load (antenna) impedance is split up for single-ended consideration

- Serial components:

- Parallel components:

\[
\begin{align*}
R_{\text{DIFF}} &= \frac{1}{2} R_{\text{SINGLE}} \\
L_{\text{DIFF}} &= \frac{1}{2} L_{\text{SINGLE}} \\
C_{\text{DIFF}} &= 2C_{\text{SINGLE}} \\
R_{\text{DIFF}} &= \frac{1}{2} R_{\text{SINGLE}} \\
L_{\text{DIFF}} &= \frac{1}{2} L_{\text{SINGLE}} \\
C_{\text{DIFF}} &= 2C_{\text{SINGLE}}
\end{align*}
\]
Network transformation: single-ended, differential

- So for the typical matching network, values for a single-ended equivalent circuit are following quantities of the differential network:

![Diagram showing network transformation and equivalent circuits.](image-url)
Impedance Matching: $\pi$ or $T$ topology

- Any complex load can be matched to a source impedance using an $LC$ element network in $\pi$- or $T$ topology (general solution). As the antenna impedance varies over frequency, this matching is for the carrier frequency only.

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

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

\[ Z_3 = -j \sqrt{R_D R_A} \frac{\sin(\varphi)}{\sin(\varphi)} \quad \varphi \ldots \text{Phase deviation output to input} \]

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

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

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

Literature: F. E. Terman, "Network Theory, Filters, and Equalizers"
Towards a more specific impedance adjustment

- Any complex load can be matched to any source impedance using $\pi$- or $T$- topology.

- However, there is at least one inductor $L$, which introduces losses…

- We may not need to adjust *any* load:
  - antennas are an inductive load ($1^{\text{st}}$ self-resonance $> f_{\text{CAR}}$)
  - phase relation output to input for the carrier frequency is irrelevant

- So there is a more specific solution, consisting only of capacitors
  - HF capacitors of C0G or NP0 have negligible losses and less tolerance
  - Less components also means reduction of costs and PCB area
Impedance adjustment with $L$-topology

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 $\omega$).

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

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

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

$\omega >> f_{\text{RES}}$ $>$ $f_{\text{CAR}}$ $Q_A >> (>) Q_{\text{SYS}}$
Antenna coupling – NFC antenna only

- Coupling affects antenna impedance significantly
  - mutual inductance $L_A$ changes
  - „Loading“ $Q$ changes

- Impedance trace is shown in Smith Chart

distance variation to “loading” antenna $\Rightarrow$ coupling factor $k$ varies
Q-factor adjustment for operating frequency

- The quality factor of the antenna conductor shall be high... above the intended Q-factor for operation at the carrier frequency 13.56 MHz.
- We can neglect losses in the capacitor (if good components are chosen)
  - For one frequency, the serial equivalent circuit can be calculated to an equivalent parallel circuit for the inductor, using the Q equation:

\[
Q_A = \frac{\omega L_A}{R_{SERIAL}} = \frac{R_{PARALLEL}}{\omega L_A}
\]

- This way, an external resistor in series to the antenna allows to adjust the intended \(Q_A\):

\[
R_E = \frac{\omega L_A}{Q_{INTENDED}} - R_A
\]

- To note: This can again be recalculated to a “new” antenna impedance \(Z_A\) (which might also be a parallel equivalent circuit).
Impedance adjustment with $L$-topology

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

\[ Z_M = Z_E = Z \equiv \frac{1 + j\omega Z_A (C_{SM} + C_{PM})}{\omega (jC_{SM} - \omega C_{PM} Z_A)} \]

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

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

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

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

\[ f_{RES} \equiv f_{CAR} \]

\[ \text{Im}\{Z\} \equiv 0 \]

\[ \text{Re}\{Z\} \equiv R_{DESIRED} \]
Annex: Exact derivation of reactance matching with 2 capacitors (I):

- The admittance of the parallel equivalent antenna circuit and $C_P$ is given by

$$Y_A = \frac{1}{R_A} + \frac{1}{sL_A} + sC_A + sC_P =$$

$$= \frac{1}{R_A} + \frac{1}{sL_A} + s(C_A + C_P)$$

- Impedance for the parallel equivalent circuit of antenna and reactance matching network is...

$$Z_{LAST} = \frac{1}{Y_A} + \frac{1}{sC_S} = \frac{sR_AL_A}{s^2R_AR_A(L_A(C_A + C_P) + sL_A + R_A} + \frac{1}{sC_S}$$

- This impedance is set equal to the desired real (⇒ reactance matching) source impedance (e.g. 50 Ω).

$$R_D \equiv \frac{s^2[R_A L_A C_S + R_A L_A(C_A + C_P)] + sL_A + R_A}{s^3R_AR_A L_A C_S(C_A + C_P) + s^2L_A C_S + sR_A C_S} \bigg|\text{Nenner}$$

$$s^3R_D R_A L_A C_S(C_A + C_P) + s^2R_D L_A C_S + sR_D R_A C_S = s^2[R_A L_A C_S + R_A L_A(C_A + C_P)] + sL_A + R_A$$
Annex: Exact derivation of reactance matching with 2 capacitors (II):

- Transition $s \rightarrow j\omega$ and coefficient comparison of real and imaginary parts:

$$- j\omega^3 R_D R_A L_A C_s (C_A + C_P) - \omega^2 R_D L_A C_s + j\omega R_D R_A C_s = -\omega^2 [R_A L_A C_s + R_A L_A (C_A + C_P)] + j\omega L_A + R_A$$

- Imaginary parts:

$$- j\omega^3 R_D R_A L_A C_s (C_A + C_P) + j\omega R_D R_A C_s = j\omega L_A$$

$$C_{S,a} = \frac{L_A}{R_D R_A - \omega^2 R_D R_A L_A (C_A + C_P)}$$

- Real parts:

$$- \omega^2 R_D L_A C_s = -\omega^2 [R_A L_A (C_A + C_P)] + R_A$$

$$C_{S,b} = \frac{-\omega^2 R_A L_A (C_A + C_P) + R_A}{\omega^2 L_A (R_A - R_G)}$$

- Both solutions for $C_S$ are set equal, to solve for $C_P$
Annex: Exact derivation of reactance matching with 2 capacitors (II):

- For $C_{S,a} = C_{S,b}$ follows...

$$\omega^2 L_A^2 R_A - \omega^2 L_A^2 R_D = R_D^2 R_A^2 - \omega^2 R_D R_A^2 L_A (C_A + C_P) - \omega^2 R_D R_A^2 L_A (C_A + C_P) + \omega^4 R_D R_A^2 L_A^2 (C_A + C_P)^2$$

- Sort according the power of $C_P$:

$C_P^2 \left[ \omega^4 R_D R_A^2 L_A^2 \right] + $

$+ C_P \left[ 2\omega^4 R_D R_A^2 L_A^2 C_A - 2\omega^2 R_D R_A^2 L_A \right] +$

$+ \left[ R_D R_A^2 - \omega^2 L_A^2 (R_A - R_D) - 2\omega^2 R_D R_A^2 L_A C_A + \omega^4 R_D R_A^2 L_A^2 C_A^2 \right]$ 

- Resolve the characteristic equation, cancel:

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

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

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

- Knowing $C_P$ allows to calculate $C_S$ using one or the other equation. Note: only positive, real values have a physical meaning!
Antenna coupling – no EMC filter

\[ \text{distance } 80 \ldots 3 \text{ mm} \]
Reasoning for extending 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 $\Rightarrow$ a 2$^{nd}$ resonance frequency can increase the bandwidth.
- Furthermore, this 2$^{nd}$ resonance (LC low-pass) can suppress unwanted harmonics emission ($\Rightarrow$ name EMC filter)

- This comes on the expense of more signal distortions…
Impedance adjustment with $L$-topology

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

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

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

$f_{RES} = f_{CAR}$

$\text{Im}\{Z\} = 0$

$\text{Re}\{Z\} = R_{DESIRED}$
Impedance adjustment with $L$-topology

- For the calculation of $C_P$ and $C_S$ we follow a more systematic approach and re-calculate a real and imaginary impedance for every step:

$$\begin{align*}
\bar{Z}_0 &= R_0 \\
\bar{Z}_M &= R_M + jX_M \\
\bar{Z}_A &= R_A + jX_A
\end{align*}$$

1$^{st}$ step: starting from right side, we calculate an antenna impedance $Z_A$

$$\begin{align*}
\text{Re}(Z_A) &= \frac{R_A X_{CA}^2}{R_A^2 + (X_{CA} + X_{LA})^2} = R_a \\
\text{Im}(Z_A) &= \frac{X_{CA} R_A^2 + X_{CA}^2 L_A + X_{CA} X_{LA}^2}{R_A^2 + (X_{CA} + X_{LA})^2} = X_a
\end{align*}$$
Impedance adjustment with $L$-topology

2\textsuperscript{nd} step: starting from the left side, we calculate the EMC filter impedance and the adjustment network impedance $Z_M$

\[ Z_0 = R_0 \]

\[ \overline{Z_M} = R_M + jX_M \]

\[ X_{C0} = -\frac{1}{\omega C_0} \]

\[ X_{L0} = \omega L_0 \]

\[
\begin{align*}
\text{Re}(Z_M) &= \frac{R_0 X_{C0}^2}{R_0^2 + (X_{C0} + X_{L0})^2} = Rm \\
\text{Im}(Z_M) &= -jX_{C0} \frac{X_{L0}^2 + X_{C0}X_{L0} + R_0^2}{R_0^2 + (X_{C0} + X_{L0})^2} = Xm
\end{align*}
\]
Impedance adjustment with L-topology

- This way, we bring the matching condition in the middle of the network.

**3rd step**: Solving condition for the parallel and serial capacitance (target impedance):

\[
X_{p,2} = \frac{R_m}{R_a - R_m} \left( X_a \pm R_a \cdot \sqrt{\frac{R_a}{R_m} + \frac{X_a^2}{R_m R_a} - 1} \right)
\]

\[
C_p = -\frac{1}{\omega X_p}
\]

\[
X_S = X_m - X_p \cdot \frac{R_a^2 + X_a^2 + X_a X_p}{R_a^2 + (X_a + X_p)^2}
\]

\[
C_S = -\frac{1}{\omega X_S}
\]

- Finally, we need to sort out solutions which have no physical meaning…
  - only positive capacitance values have a representation
  - …not all antennas can be matched (if we violate a pre-condition, it is not possible…

- Note: In practice, the impedance must be checked by measurement, as parasitics may cause deviations from the ideal conditions
Antenna coupling – with EMC filter

\[ \Delta: Z_{in} \quad 13.56 \text{ MHz} \]

\[ \phi: Z_{ANT} \]

\[ k, M \quad L_{ANT} \quad R_{ANT} \]

distance 80 ... 3 mm
Which loads can be matched?

- For the 2 capacitor network, the area in the Smith Chart is following...

Reference: *Electronic Applications of the Smith Chart* by Phillip H. Smith, 1969, p. 124
Which loads can be matched?

- For the network with EMC-filter, it is slightly more difficult
- We need to calculate a target impedance \( Z_T \) first, which depends on our desired input impedance \( Z_{IN} \), and the (loaded) antenna impedance \( Z_A \).

\[
Z_T = \frac{Z_{IN} - jX_{L0}}{1 + jX_{C0}(jX_{L0} - Z_{IN})}
\]
Which loads can be matched?

Antenna impedance $Z_A$ (including coupling)

Region, which can be matched to desired network input impedance

Adjustable range for parallel cap $C_P$

Adjustable range for serial cap $C_S$

Target impedance $Z_T$

Network input impedance $Z_{IN}$ (@ carrier frequency)

Frequency trace of $Z_{IN}$
Which loads can be matched?

- The region of network input impedances $Z_{in}$ at carrier frequency, which can be adjusted to an intended resistance ("matched"), is limited by two circles.

- For the parallel capacitor, from short to $Z_T$

- For the serial capacitor, from $Z_T$ to open.

- The $C_P$ limit circle is given by
  - $x_T$, $y_T$… coordinates of $Z_T$
  - $c$…center, $r$… radius

$$c = \frac{-1 + x_T^2 + y_T^2}{2(x_T - 1)} \quad r = 1 - c$$

- The $C_S$ limit circle is given by

$$c = \frac{-1 + x_T^2 + y_T^2}{2(x_T + 1)} \quad r = 1 + c$$

- Impedance adjustment possible, iff

$$\text{Re}(Z_A) \leq \text{Re}(Z_T)$$

$$\text{Im}(Z_A) \geq 0 \ (\text{inductive load})$$
The resonant coupling system

Reader equivalent circuit properties
L, R, C

Transponder equivalent circuit properties
L, R, C

Resonant system properties
k, f_{RES}, Q

Standard defines properties at the Air Interface
How near-field coupling affects the air interface

Coupling: 0%
Distance: -- mm

Matching-Impedance

EMC-FIL MATCHING ANTENNA

L₀ C₀ Cs Cp

C₀ L₀ A

A

Matching at current distance
How near-field coupling affects the air interface

Coupling: 6 %
Distance: 25 mm

Distance in mm

Distance in mm

LSB fc USB matching at current distance

H_{OVS} 1.076

\begin{array}{cccccc}
\text{t1 in } \tau_c & \text{t2 in } \tau_c & \text{t3 in } \tau_c & \text{t4 in } \tau_c & a & H_{OVS} \\
37.80 & 31.12 & 2.45 & 1.52 & 0.007 & 1.076
\end{array}
How near-field coupling affects the air interface

Coupling: 13%
Distance: 15 mm

Distance in mm

Driver power in mW

Distance in mm

Matching-Impedance

EMC-FIL

MATCHING

ANTENNA

\[ t_1 \ \tau_C \ 37.77 \]
\[ t_2 \ \tau_C \ 31.47 \]
\[ t_3 \ \tau_C \ 2.12 \]
\[ t_4 \ \tau_C \ 1.34 \]
\[ a \ 0.007 \]
\[ H_{OVS} \ 1.12 \]
How near-field coupling affects the air interface

- Coupling: 21%
- Distance: 10 mm

**Graphs and Diagrams:**
- Driver power vs. Distance in mm
- Matching Impedance Circuit Diagram
- Smith Chart with measurements:
  - t1 $\tau_C$: 37.80
  - t2 $\tau_C$: 29.41
  - t3 $\tau_C$: 1.99
  - t4 $\tau_C$: 1.34
  - a: 0.010
  - $H_{OVS}$: 1.16

**Symbols and Notations:**
- $L_0$, $C_0$, $C_s$, $C_p$, $C_A$, $L_A$, $R_A$ (depicted in the circuit diagram)
How near-field coupling affects the air interface

Coupling: 38%
Distance: 5 mm

Distance in mm

Driver power in mV

LSB, fc, USB, matching at current distance

H_{OVS} = 1.15

<table>
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<th>t1 in $\tau_C$</th>
<th>t2 in $\tau_C$</th>
<th>t3 in $\tau_C$</th>
<th>t4 in $\tau_C$</th>
<th>a</th>
<th>H_{OVS}</th>
</tr>
</thead>
<tbody>
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<td>37.78</td>
<td>24.97</td>
<td>1.61</td>
<td>1.04</td>
<td>0.020</td>
<td>1.15</td>
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Matching Impedance

EMC-FIL
MATCHING
ANTENNA
Thank you for your Audience!

Please feel free to ask questions...