added to the antenna equivalent circuit and thus generates a slightly modified formula for the antenna impedance (see formulas 3.3 and 3.4).

$$\boxed{R_A = R_q + \frac{R_a {X_{C_a}}^2}{{R_a}^2 + (X_{C_a} + X_{L_a})^2}} \tag{3.3}$$

$$\boxed{X_A = X_{C_a} \cdot \frac{{R_a}^2 + X_{C_a} X_{L_a} + {X_{L_a}}^2}{{R_a}^2 + (X_{C_a} + X_{L_a})^2}} \tag{3.4}$$

### 3.1.3 Impedance Adjustment Network

The impedance adjustment network, shown as part 2 of figure 3.2, has two purposes. The first is to compensate the inductive impedance of the antenna coil which makes the network resonant at the carrier frequency. However, this would be rather simple and could be done with one single capacitor. The second purpose lays in an impedance transformation to provide a defined load to the voltage source which is supplying the antenna network. In literature, different topologies can be found. The most important are the T, Π and L networks (see figure 3.4). For lossless transformation all components should be reactive elements (capacitances or inductances).

In a cost and space sensitive environment like NFC, it is desired to keep the component

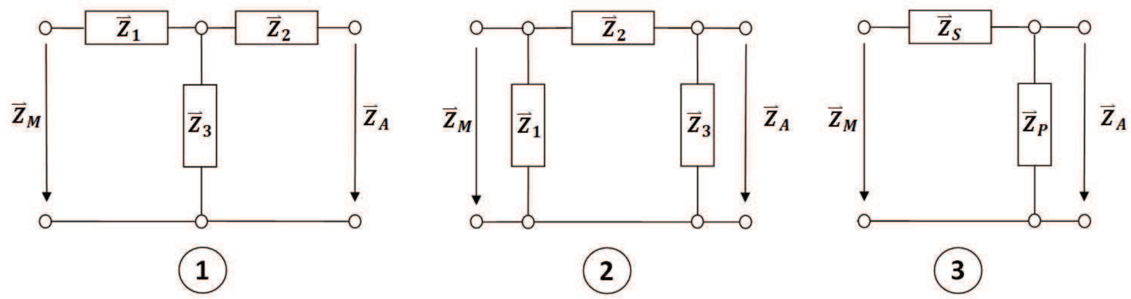


*Figure 3.4:* Different impedance transformation network topologies. 1=$T$; 2=Π; 3=$L$

count to a minimum. For that reason, the L-shaped network is chosen. If a certain matching impedance $Z_M$ and an antenna impedance $Z_A$ are given, then the two impedances $Z_P$, $Z_S$ can be calculated. For simplification we assume these impedance to be pure reactive ($Z_P = X_P$, $Z_S = X_S$) and thus their imaginary impedance value can be calculated as shown in equations 3.7 and 3.8.

$$\begin{aligned}
\vec{Z}_M &= jX_S + \frac{jX_P \cdot (R_A + jX_A)}{R_A + j(X_A + X_P)} \\
&= jX_S + \frac{j{R_A}^2 X_P - \cancel{jR_A X_A X_P} + R_A {X_P}^2 + \cancel{jR_A X_A X_P} + j{X_A}^2 X_P + jX_A {X_P}^2}{{R_A}^2 + (X_A + X_P)^2}
\end{aligned}$$

$$(1)\ real(\vec{Z}_M) = R_M = \frac{R_A {X_P}^2}{{R_A}^2 + (X_A + X_P)^2} \tag{3.5}$$

$$(2)\ imag(\vec{Z}_M) = X_M = X_S + \frac{{R_A}^2 X_P + {X_A}^2 X_P + X_A {X_P}^2}{{R_A}^2 + (X_A + X_P)^2} \tag{3.6}$$

(1) $R_M {X_P}^2 + 2R_M X_P X_A + R_M {X_A}^2 + {R_A}^2 R_M - R_A {X_P}^2 = 0$

$$
\begin{aligned}
&{X_P}^2(R_M - R_A) + X_P(2R_M X_A) + R_M({R_A}^2 + {X_A}^2) = 0 \\
&{X_P}^2 + X_P \frac{2R_M X_A}{R_M - R_A} + \frac{R_M({R_A}^2 + {X_A}^2)}{R_M - R_A} = 0 \\
&X_{P_{1,2}} = -\frac{R_M X_A}{R_M - R_A} \pm \sqrt{\frac{{R_M}^2 {X_A}^2}{(R_M - R_A)^2} - \frac{R_M({R_A}^2 + {X_A}^2)}{R_M - R_A}} \\
&\qquad = -\frac{1}{R_M - R_A} \cdot \Bigg( R_M X_A \pm \\
&\qquad\qquad \sqrt{\cancel{{R_M}^2 {X_A}^2} - {R_M}^2 {R_A}^2 - \cancel{{R_M}^2 {X_A}^2} + R_M {R_A}^3 + R_M R_A {X_A}^2} \Bigg)
\end{aligned}
$$

$$
\boxed{X_{P_{1,2}} = \frac{R_M}{R_A - R_M} \cdot \left( X_A \pm R_A \cdot \sqrt{\frac{R_A}{R_M} + \frac{{X_A}^2}{R_M R_A} - 1} \right)} \tag{3.7}
$$

$$
(2)\ \boxed{X_{S_{1,2}} = X_M - X_{P_{1,2}} \cdot \frac{{R_A}^2 + {X_A}^2 + X_A X_{P_{1,2}}}{{R_A}^2 + (X_A + X_{P_{1,2}})^2}} \tag{3.8}
$$

As we need to compensate the inductive reactance from the antenna, both elements of the impedance adjustment network can be capacitances. As a capacitor represents a capacitive element much better than an inductor represents an inductive element, the lack of inductance in the impedance adjustment network also gives advantages in the accuracy and space requirements of the circuit board. Besides the equations from above, the operating principle can be easily understood by investigating the influence of each capacitance in separate. The series capacitance can only add a reactive (imaginary) part to network impedance. That means, the parallel capacitance needs to adjust the effective (real) impedance to the desired value. Figure 3.5 illustrates this correlation.

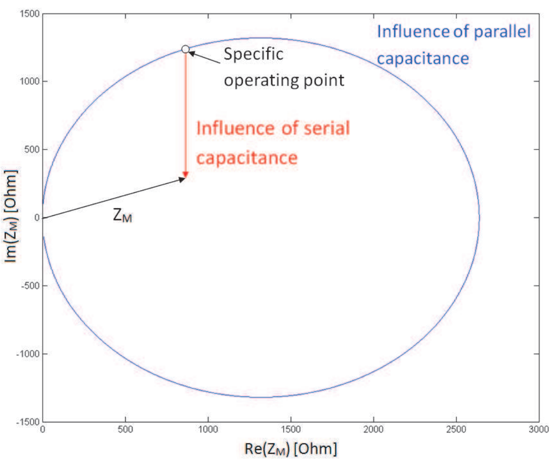


*Figure 3.5:* Influence of parallel and the series capacitance on the network impedance

### 3.1.4 EMC Filter

For NFC reader devices, a very simple and efficient way to generate a 13.56 MHz carrier is to periodically toggle the output voltage of the transmitter between positive supply and GND. This so called Class-D amplifier [1, P. 131] leads to a rectangular signal output which needs to be filtered in order to reduce the harmonic distortion induced by the overtones. A low-pass filter with a cut-off frequency ($f_f$) between 14.5 MHz and 20 MHz does the trick. The lower limit is given to the carrier frequency of 13.56 MHz plus half of the bandwidth needed for modulation and an additional headroom. The specified reader modulation, which is for most protocols ASK, leads to a bandwidth of twice the maximum frequency and in case of 847 kHz to a lower and upper sideband of 12.7 MHz and 14.4 MHz.
The upper limit of $f_f$ should be well below the first harmonic of the rectangular input signal which is three times the carrier frequency or 40.6 MHz. Practically 20 MHz has shown to be the upper boarder for a second order low pass, when electromagnetic compatibility should be achieved. For an ideally loss-less filtering, normally an LC network is chosen (illustrated as part 1 of figure 3.2). If we take NXP's CLRC663 [22] Antenna Design Guide as an example, recommend values for the inductance $L_0$ should be between 390 nH - 1 $\mu$H. The capacitance $C_0$ for the LC low-pass filter can be derived from the cut-off frequency $f_f$ (see formula 3.9).

$$\omega_f = \frac{1}{\sqrt{L_0 C_0}} \rightarrow C_0 = \frac{1}{{\omega_f}^2 L_0} = \frac{1}{(2\pi f_f)^2 L_0} \tag{3.9}$$

Besides filtering of the input signal, this LC low-pass also applies an impedance transformation to the network and also influences timing characteristics during modulation (see section 3.2). To achieve a defined, real-valued input impedance $Z_0$, the impedance adjustment network needs to consider this transformation. Given the component values $L_0$ and $C_0$, the necessary matching impedance may be calculated the following way:

$$\vec{Z}_0 = jX_{L_0} + \frac{jX_{C_0}\vec{Z}_M}{jX_{C_0} + \vec{Z}_M} \overset{!}{=} R_0 \qquad X_{C_0} = -\frac{1}{\omega C_0};\ X_{L_0} = \omega L_0$$

$$(R_0 - jX_{L_0}) \cdot jX_{C_0} + (R_0 - jX_{L_0}) \cdot \vec{Z}_M = jX_{C_0}\vec{Z}_M$$

$$\vec{Z}_M \overset{!}{=} \frac{jX_{C_0} \cdot (jX_{L_0} - R_0)}{R_0 - j(X_{C_0} + X_{L_0})}$$

$$\overset{!}{=} \frac{X_{C_0} \cdot (R_0 X_{C_0} - jX_{C_0}X_{L_0} + \cancel{R_0 X_{L_0}} - j{X_{L_0}}^2 - j{R_0}^2 - \cancel{R_0 X_{L_0}})}{{R_0}^2 + (X_{C_0} + X_{L_0})^2}$$

$$\boxed{R_M \overset{!}{=} \frac{R_0 {X_{C_0}}^2}{{R_0}^2 + (X_{C_0} + X_{L_0})^2}} \tag{3.10}$$

$$\boxed{X_M \overset{!}{=} -X_{C_0} \cdot \frac{{X_{L_0}}^2 + X_{C_0}X_{L_0} + {R_0}^2}{{R_0}^2 + (X_{C_0} + X_{L_0})^2}} \tag{3.11}$$