Application Note

Impedance Matching for Near Field Communication Applications

ANP084 // Christian Merz / Cem Som

1 Introduction

The number of mobile devices on the market equipped with NFC (Near Field Communication) or RFID (Radio Frequency Identification) technology is increasing exponentially. NFC is a short range, high frequency, low bandwidth wireless communication technology, which enables a standardized communication between two mobile devices like smartphones, smart cards, stickers or tags. NFC and RFID use the same frequency, 13.56 MHz. While RFID is capable of reception and transmission up to several meters, NFC is restricted to very close proximity, less than 10 cm, and is extensively used in NFC tags, data communications, and secure financial transactions. Würth Elektronik already offers a broad portfolio of **WPT coils** (Wireless Power Transfer). As a further innovation, winding an NFC antenna around a WPT coil enables a new communication channel between WPT transmitter and receiver, which can be used for system control and data transfer. With NFC, data rates of up to 848 kbit/s can be achieved. The **WE-WPCC WPT/NFC** series combines a WPT coil and an NFC antenna in one efficient component. For the system designer this combination of WPT and NFC coils in a single component has advantages in terms of simplicity, efficient form factor and cost. In this application note we show how the impedance of the NFC antenna can be matched to the integrated NFC circuit IC to maximize the radiated field and transmission distance [3]. As an example we have used the WPT combination coil **760308101312** from Würth Elektronik but the technique applies to any NFC antenna.

WE currently offers five different WPT/NFC combination coils, as shown in table 1. L_1 and Q_1 are measured at 125 kHz and L_2 and Q_2 are measured at 13.56 MHz.

Part number	WPT		NFC	
	L_1 (µH)	Q_1	L_2 (µH)	Q_2
760308101150	6.3	100	1.2	80
760308103305	8.8	30	1.4	47
760308102306	8	19	1.4	47
760308103307	7.8	19	1.6	47
760308101312	24	125	0.7	30

Table 1: WE WPT/NFC combination coils

2 Complex conjugate impedance matching

2.1 Basics

Complex conjugate impedance matching is a very important procedure in RF circuit design to provide the maximum possible power transfer between a source and its load and to minimize the signal reflections from the load. One example for the need of power transfer occurs in the front end of any sensitive receiver. Obviously, any unnecessary loss in a circuit that is already carrying extremely small signal levels simply cannot be tolerated. Therefore, in most instances, extreme care is taken during the initial design of such a front end to make sure that each device in the chain is matched to its load. [1] In RF technology, loads are often complex, i.e. they have an inductive or capacitive component in addition to their resistive part. For matching, the inductive or capacitive component must be compensated with its counterpart, the so-called complex conjugate component. This means for example that an inductive component must be capacitively compensated. Impedance matching is based on the maximum power transfer theorem. It states that, to obtain maximum external power from a source with a finite internal resistance, the resistance of the load must be equal to the resistance of the source as seen from its output terminals. Additionally, it states that any reactive component of the source and load should be of equal magnitude but opposite sign. This means that the load and source impedances have to be complex conjugate to each other. [2]

In general, the complex conjugate of the impedance $Z = R + jX$ is $Z^* = R - jX$, where R is the real part and X is the imaginary part of the complex impedance Z. Figure 1 shows the complex source impedance Z_S and the complex load impedance Z_L.

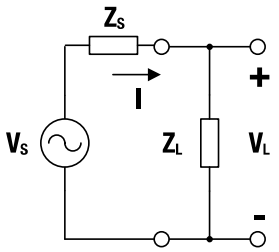


Figure 1: Source and load impedance [2]

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These impedances have to fulfill the following condition, to lead to an optimal match:

$$Z_L = Z_S^* \quad (1)$$

There are many possible network topologies that could be used to perform the impedance reactances. Among them, the simplest one is the L-topology, which comprises of two elements and gets its name from the component orientation, which resembles the shape of an L.[1] Figure 2 shows the two possible implementations of the L-topologies for impedance matching.

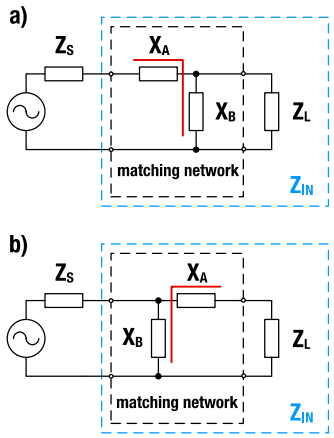


Figure 2: L-topology matching networks

In figure 2, X_A is the ideal reactance of the series leg and X_B is the ideal reactance of the parallel leg. Z_S and Z_L are the source and load impedances. Z_{in} denotes the input impedance and includes the load and matching impedance, which has to be the complex conjugate of Z_S. Before the matching components can be determined, the load and source impedances have to be known. The source impedance is in most cases 50 Ω. In general, the source impedance can also be complex.

For higher frequencies in the UHF band (e.g. 434 MHz, 868 MHz and 2.4 GHz), the matching process is described in the Application Note **ANP057**. Würth Elektronik also offers a customer **antenna matching service** for these frequencies.

2.2 Determination of the complex load impedance

The complex load impedance can be determined by measurement and calculation. The complex impedance measurement for the 13.56 MHz range can be done with a vector network analyzer, which measures the S-parameter of the device under test. S-parameters describe the electrical behavior of linear electrical networks when undergoing various steady state stimuli by electrical signals. For impedance matching, the S_{11} parameter, which is called input port voltage reflection coefficient, is used. The input port voltage reflection coefficient is a complex quantity, whose absolute value is an indicator for the reflection. $|S_{11}| = 0$ means that the circuit is perfectly matched and that none of the incident power wave is reflected. $|S_{11}| = 1$ denotes that 100% of the incident power wave is reflected back to the input. For NFC applications, the load is an antenna. For practical calculations and simulations, the electrical properties of the antenna are represented in an equivalent circuit. The simplified series equivalent circuit of an NFC antenna is shown in figure 3.

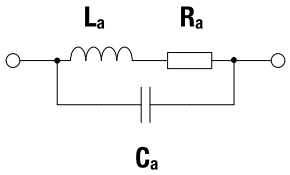


Figure 3: Simplified series equivalent circuit of an NFC antenna

L_a is the inductance and R_a is the equivalent series resistance, which represents all ohmic losses of the antenna. C_a is the parallel equivalent capacitance of the antenna. The values L_a and R_a can be directly measured with a network analyzer or LCR meter. The value C_a is a parasitic value and must be determined by measurement and calculation. The frequency dependencies of L_a, C_a and R_a are not taken into account in the calculations and simulations. Knowing the inductance L_a, the parallel equivalent capacitance C_a can be calculated at the self resonance frequency f_S with equation (2). [8]

$$C_a = \frac{1}{(2 \cdot \pi \cdot f_S)^2 \cdot L_a} \quad (2)$$

The self resonance frequency f_S can be measured as the first point where the complex load impedance becomes real.

The impedance of the antenna Z_L, which is necessary for the impedance matching, can be calculated as follows:

$$Z_L = \frac{R_a}{(1-\omega^2 L_a C_a)^2+(\omega R_a C_a)^2} + j\frac{\omega L_a - \omega^3 L_a^2 C_a - \omega R_a^2 C_a}{(1-\omega^2 L_a C_a)^2+(\omega R_a C_a)^2} \quad (3)$$

$$Z_L = R_L + jX_L$$

The quality factor Q_L of the antenna is defined by the relation of the imaginary part X_L and the real part R_L of the antenna impedance and can therefore be calculated with equation (4).

$$Q_L = \frac{\mathrm{Im}(Z_L)}{\mathrm{Re}(Z_L)} = \frac{X_L}{R_L} \quad (4)$$

2.3 Determination of the matching circuit components

The matching procedure can be performed by calculation and simulation. In general, both impedance networks shown in figure 2 can be used, but the network shown in figure 2a is easier to calculate and has been chosen to demonstrate the matching prodcedure. Because the load has an inductive behavior, the reactances X_A and X_B are capacitive. In general, X_A and X_B can also be inductive if the load has a capacitive behavior. X_A and X_B are considered to be ideal, which means that they are considered to have no resistive or parasitic part.

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2.4 Calculation of the matching circuit components

For the network shown in figure 2a, the perfect match is achieved for $Z_{in} = Z_S^*$, and Z_{in} can be calculated as follows [5]:

$$Z_{in} = R_{in} + jX_{in} = Z_S^* = R_S - jX_S$$

$$Z_{in} = jX_A + \frac{jR_L^2X_B + R_LX_B^2 + jX_L^2X_B + jX_LX_B^2}{R_L^2 + (X_L + X_B)^2} \quad (5)$$

Due to equation (5), the real and imaginary part of the input impedance Z_{in} are:

$$\mathrm{Re}\,(Z_{in}) = R_S = \frac{R_LX_B^2}{R_L^2 + (X_L + X_B)^2} \quad (6)$$

$$\mathrm{Im}\,(Z_{in}) = -X_S = X_A + \frac{R_L^2X_B + X_L^2X_B + X_LX_B^2}{R_L^2 + (X_L + X_B)^2} \quad (7)$$

If the equation (6) is solved for the reactance X_B, we get two different values X_{B1} and X_{B2} because (6) is a quadratic equation:

$$X_{B1,2} = \frac{R_S}{R_L - R_S} \cdot \left(X_L \pm R_L \cdot \sqrt{\frac{R_L}{R_S} + \frac{X_L^2}{R_SR_L} - 1} \right) \quad (8)$$

The reactance X_A also has two values, X_{A1} and X_{A2}, and results due to equation (7) into the following expression, taking equation (8) into account:

$$X_{A1,2} = -X_S - X_{B1,2} \cdot \frac{R_L^2 + X_L^2 + X_LX_{B1,2}}{R_L^2 + (X_L + X_{B1,2})^2} \quad (9)$$

The influence of each capacitance on the impedance Z_{in} can be investigated separately with equations (8) and (9).

The series capacitor C_A can only add a reactive (imaginary) part from the network impedance Z_{in}. The parallel capacitor C_B is needed to adjust the real impedance to the desired value and yield to a circular curve in the complex plane. The influences of the two capacitors in the network impedance is illustrated with the locus curve in figure 4. Because the mathematical way of determining the matching component values can get very complicated, especially for three components matching, the way of simulation is preferred.

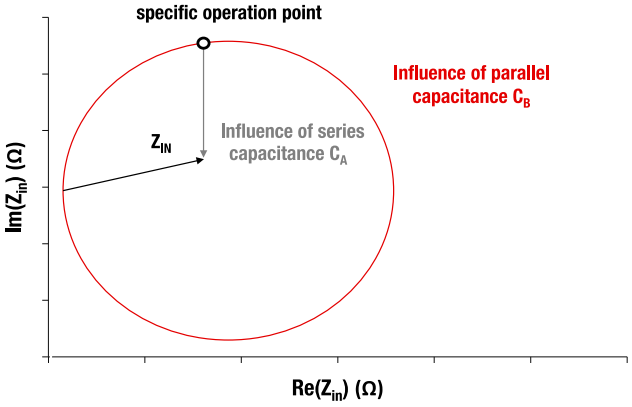


Figure 4: Influence of parallel and the series capacitance C_A and C_B on the network impedance Z_{in} [5]

2.5 Simulation of the matching circuit components

In the field of RF circuit simulation, the software Keysight Advanced Design System (ADS) is the de facto standard. For the determination of the matching components, the ADS optimization tool is recommended to be used. The optimization tool is an automated procedure of achieving the circuit performance in which ADS can modify circuit component values to meet a defined optimization goal. The goal is in this case to minimize the parameter S_{11} by modifying the matching components until the input impedance comprises a complex conjugate of the source impedance. To determine the matching components using the optimization tool and calculation, an example of the complete matching procedure is performed in the next sections. A simplified filter and matching network with L-topology with two capacitors as reactive elements and the equivalent circuit of an antenna as a load is simulated with ADS and the matching components are determined by simulation and calculated using the equations (8) and (9). The calculated and simulated matching components will be compared afterwards.

3 Application of the impedance matching on an output NFC circuit

3.1 Setup of a typical NFC output circuit

An NFC IC normally has a differential output and has an impedance of 50 Ω between the output pins TX_1 and TX_2. These pins are connected with an output circuit comprising of a filter network, a matching circuit, damping resistors and an NFC antenna. The output of the NFC IC is differential to be resistant against electromagnetic noise. A typical differential NFC output circuit is shown in figure 5.

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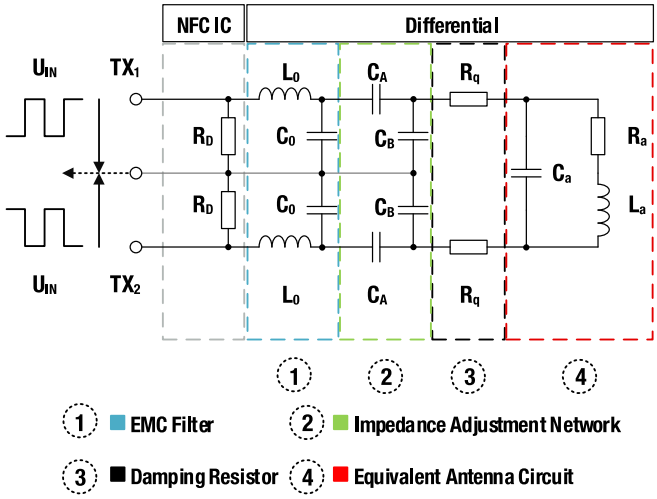


Figure 5: Differential output NFC circuit [5]

To simplify the understanding of the output NFC circuit components the differential circuit has to be transformed to a single ended one. Equations (8) and (9) can be applied to calculate the matching capacitors C_A and C_B. The transformation calculations are performed in [6] and the resulting single ended circuit is shown in figure 6.

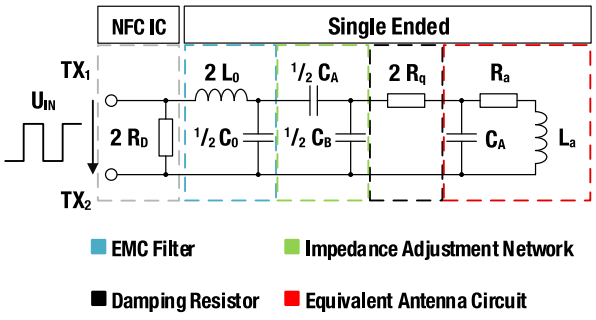


Figure 6: Equivalent single ended output NFC circuit [5]

3.2 EMC filter network

As the output signal of the NFC IC has a rectangular shape, the harmonics have to be filtered out. This is done by the EMC filter network, which is a second order low pass. The low pass filter comprises of the inductance L_0 and the capacitance C_0.

The cutoff frequency f_C of the EMC filter can be calculated with equation (10).

$$f_C = \frac{1}{2 \cdot \pi \sqrt{L_0 \cdot C_0}} \qquad (10)$$

The cutoff frequency f_C must be higher than the upper auxiliary carrier, which is 13.56 MHz + 848 kHz = 14.4 MHz for the highest possible NFC data rate of 848 kbit/s. A NFC load modulation spectra with carrier and auxiliary carriers is shown in figure 7 to visualize the necessity of the filter.

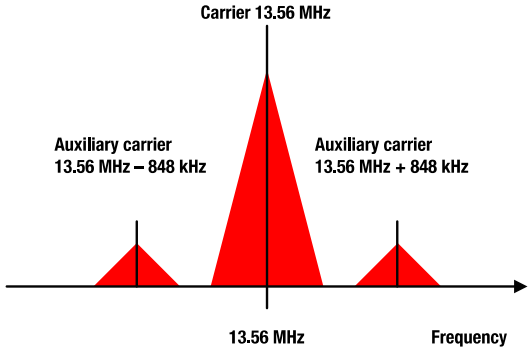


Figure 7: NFC load modulation spectra of carrier and auxiliary carriers [7]

For the impedance matching procedure, the impedance of the filter must be calculated using equation (11).

$$Z_{EMC} = \frac{2\,R_D}{(1 - \omega^2 L_0 C_0)^2 + (\omega R_D C_0)^2}$$

$$-j\frac{2\omega^3 L_0^2 C_0 + 2\omega R_D^2 C_0 - 2\omega L_0}{(1 - \omega^2 L_0 C_0)^2 + (\omega R_D C_0)^2} \qquad (11)$$

$$Z_{EMC} = R_{EMC} - jX_{EMC}$$

3.3 Impedance adjustment network

The matching circuit, also called an impedance adjustment network, shown in figures 5 and 6, has two purposes. The first is to compensate the inductive impedance of the antenna. The second is to provide the impedance transformation from the load impedance to the source impedance. For lossless transformation all components should be reactive elements. In a cost and space sensitive environment like NFC, it is desirable to keep the component count to a minimum, which is the reason to use a L matching topology.

3.4 Damping resistors and equivalent antenna circuit

The damping resistor R_q is an optional resistor, which can be used to reduce the Q factor of the antenna. Selecting the optimum value for R_q is a trade-off. On the one hand, a small value will increase the antenna efficiency of the contactless power transmission, on the other hand, a high resistor value will provide a higher bandwidth for the modulation and decrease the antenna Q factor. The recommended value range for the Q factor is between 20 and 35 as proposed in [8]. The resistor R_q should be integrated, if the Q factor of the antenna, which can be calculated with equation (4), exceeds the value of 35. If the value is higher, equation (4) has to be modified using equation (12). [5]

$$Q_{L,mod} = \frac{Im(Z_{L,mod})}{Re(Z_{L,mod})} = \frac{X_L}{2\,R_q + R_L} \qquad (12)$$

This leads to the following formula to calculate the damping resistor value R_q (for $Q_{L,mod} \geq 35$):

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$$R_q = 0.5\left(\frac{X_L}{Q_{L,mod}} - R_L\right) \quad (13)$$

The last part of the output network is the equivalent antenna network, which is described in section 2.2. This network has been utilized for the simulations and calculations of the load impedance.

3.5 Six Steps to match an NFC IC to a WE NFC antenna

As an example, it is described step by step how the NFC antenna of 760308101312 (see figure 8) can be matched to any arbitrary NFC IC with a differential output impedance of 50 Ω by calculation and simulation. The differential output circuit, shown in figure 5, has been used in the example.

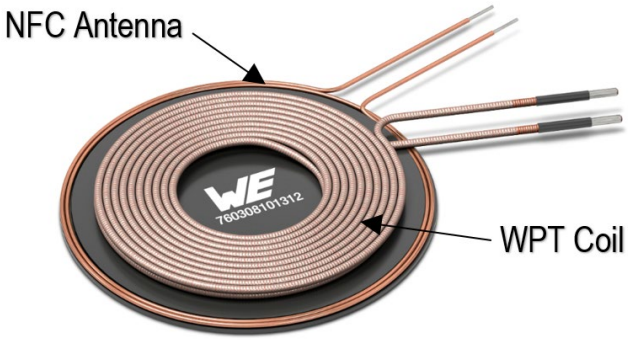


Figure 8: WE WPT/NFC combination transmitter coil 760308101312

The following steps have to be performed, to implement the impedance matching by calculation and simulation:

1. Measure the series equivalent resistance R_a and the inductance L_a at the operating frequency $f_{op} = 13.56$ MHz and the self resonance frequency f_S of the coil 760308101312 with an network analyzer. Calculation of the parallel capacitance C_a with equation (2) and determination of the equivalent antenna circuit (see figure 3).

2. The resulting values of R_a, L_a and C_a have to be inserted into equation (3). The load impedance Z_L with real part R_L and imaginary part X_L has to be calculated. The Q factor has to be calculated with equation (4).

3. The capacitance C_0 and the inductance L_0 of the EMC filter network must be determined with equation (10) and inserted into equation (11), so that the complex impedance Z_{EMC} with real part R_{EMC} and imaginary part X_{EMC} can be calculated.

4. The values for R_L, X_L, R_{EMC} (denoted as R_S in equations (8) and (9)) and X_{EMC} (denoted as X_S in equations (8) and (9)) have to be integrated into the in equations (8) and (9). The capacitor values C_A and C_B have to be calculated out of X_A and X_B.

5. Verification of the matching calculation results by determining $|S_{11}|$.

6. Simulation of the S_{11} parameter of the differential output circuit (see figure 5) and comparison of the results with the calculation.

3.6 Determination of the equivalent antenna circuit

For the NFC antenna of article 760308101312 the following parameters have been determined by measurement and calculation:

Parameter	Value	Unit	Way of determination
f_S	63	MHz	Measurement with VNA
L_a	0.7	µH	Measurement with VNA
R_a	1.7	Ω	Measurement with VNA
C_a	9.12	pF	Calculated with eq. 2

Table 2: Parameters of the equivalent antenna circuit of 760308101312

The values for L_a and R_a have been measured at 13.56 MHz.

The equivalent antenna circuit parameters of the articles 760308101150, 760308103305, 760308102306 and 760308103307 are listed in the appendix A.2.

3.7 Determination of the complex impedance and the Q factor

The complex impedance Z_L of the NFC antenna of coil 760308101312 is calculated with equation (3) by integrating the parameters of the equivalent antenna circuit, which have been determined in the previous section and lead to the following complex impedance value.

$Z_L = (1.87 + j\,62.53)\ \Omega$

This leads due to equation (4) to a Q_L of 33. Because the Q factor of the antenna is lower than 35, the damping resistors R_q can be omitted.

3.8 Calculation of the filter components and the complex impedance of the filter network

For the cut off frequency f_C, which can be calculated with equation (10), a value of 14.8 MHz has been chosen, which is higher than 14.4 MHz. An inductor value of 470 nH is chosen for L_0, which leads to a C_0 of 247 pF. The values L_0 and C_0 have to be inserted into equation (11) to calculate the complex filter impedance. For R_D, the value of 25 Ω is inserted, which is the differential output impedance of a typical NFC IC in relation to the ground signal. These values lead to the following complex filter impedance:

$Z_{EMC} = (165.82 - j\,45.46)\ \Omega$

3.9 Calculation of the matching component values

The matching reactances X_B and X_A can be directly calculated with equations (8) and (9). For R_S, the real part of the EMC network has to be integrated, which is 165.82 Ω. For the imaginary part of the source X_S the imaginary part of the EMC filter has to be integrated, which is -45.46 Ω. For R_L and X_L, the real and imaginary parts of the antenna network have to be integrated, which are 1.87 Ω and 62.53 Ω. The resulting matching reactances X_A and X_B can be calculated to the corresponding differential matching capacitors C_A and C_B with:

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$$C_{A,B} = -\frac{2}{\omega \cdot X_{A,B}} \quad (14)$$

Table 3 gives an overview of the resulting reactance and capacitance values.

Parameter	Value	Unit	From equation
X_{B1}	-69.69	Ω	(8)
X_{A1}	-519.81	Ω	(9)
X_{B2}	-56.8	Ω	(8)
X_{A2}	610.73	Ω	(9)
C_{B1}	337	pF	(14)
C_{A1}	45	pF	(14)
C_{B2}	413	pF	(14)
C_{A2}	-38	pF	(14)

Table 3: Resulting reactance and capacitor values

The capacitor values for C_{A1} and C_{B1} are used for the matching, the values for C_{A2} and C_{B2} are neglected, because C_{A2} yields a negative value. Therefore the resulting matching capacitors for the differential output circuit are:

$C_A = 45$ pF

$C_B = 337$ pF

3.10 Calculation of the resulting reflection coefficient

To verify, if the calculated matching capacitor values C_A and C_B yield to a low reflection, the S_{11} parameter at the output of the NFC IC must be calculated. The smaller the absolute values is, the lower is the reflection and the better is the circuit matched to the 50 Ω output impedance of the NFC IC. The complex input port voltage reflection coefficient S_{11} can be calculated with formula (15).

$$S_{11} = \frac{Z_{in}' - 2R_D}{Z_{in}' + 2R_D} \quad (15)$$

where $R_D = 25$ Ω, which is the single ended output impedance of the NFC IC. The parameter Z_{in}' depends on the input impedance Z_{in} (see equation (5)) and can be calculated with equation (16).

$$Z_{in}' = \frac{2 \cdot Z_{in}}{2 + j\omega C_0 Z_{in}} + 2j\omega L_0 \quad (16)$$

The absolute value of S_{11} can be calculated as:

$$|S_{11}| = \sqrt{Re(S_{11})^2 + Im(S_{11})^2} \quad (17)$$

In RF engineering, $|S_{11}|$ is often given in a logarithmic scale, and can be calculated as:

$$|S_{11}|_{dB} = 20 \cdot \log(|S_{11}|) \quad (18)$$

The resulting values of equations (5), (15)-(18) are shown in table 4.

Parameter	Value	Unit	From equation
Z_{in}	167.15 + j 45.69	Ω	(5)
Z_{in}'	49.7 – j 0.276	Ω	(16)
S_{11}	-0.00297-j 0.00277		(15)
$\|S_{11}\|$	0.004		(17)
$\|S_{11}\|_{dB}$	-48	dB	(18)

Table 4: Calculated values for the reflection coefficient and the input impedance

3.11 Simulation of the differential output circuit with Keysight ADS

An alternative and easier way to determine the matching capacitor values C_A and C_B is to simulate the differential output network shown in figure 5. The optimization tool, integrated in ADS, allows the determination of unknown parameters of the network. A simulation goal and the simulation parameters have to be defined. To determine the matching capacitor values by simulation, the following steps have to be performed:

1. Create the schematic of the differential output network shown in figure 5 with the filter component values calculated in section 3.8 and the equivalent antenna network values determined in section 3.6. An input port with a 50 Ω impedance has been used.
2. Definition of the simulation type and the simulation variables
3. Definition of the optimization goal and the optimization iterations
4. Execution of the optimization and definition of the output plot type

3.12 Schematic of the differential output network

Figure 9 shows the ADS schematic of the differential output network (see figure 5) with input port 1 and source impedance of 50 Ω.

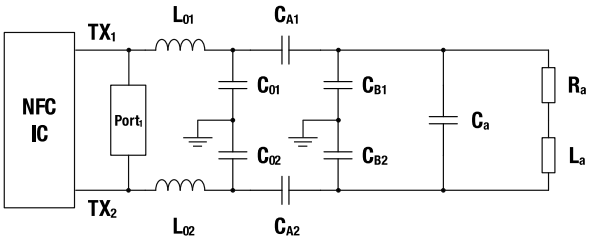


Figure 9: ADS schematic of the differential output network

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3.13 Definition of the simulation type and the simulation variables

The chosen simulation type is the Large-Signal S-Parameter Simulation, which is based on a harmonic balance algorithm. This simulation type computes S-parameters for linear and nonlinear RF circuits. The simulation variables are the matching capacitors C_A and C_B, which have to be determined by the algorithm. The range of 1-1000 pF has been chosen. For the two capacitors, start values have to be defined.

3.14 Definition of the optimization goal and iterations

As an optimization goal, it has been defined that $|S_{11}|$ should be less than 0.0001 in a frequency band between 13.559 MHz and 13.561 MHz. 1000 optimzation iterations have been chosen.

3.15 Execution of the optimization and definition of the output plot type

The simulated matching capacitance values result to C_A = 45 pF and C_B = 337 pF, which are the same values calculated in section 3.9.

The simulated reflection coefficient $|S_{11}|$ has been visualized in a Smith chart (see figure 10) and has been plotted (see figure 11) in the frequency range from 12 MHz to 15 MHz.

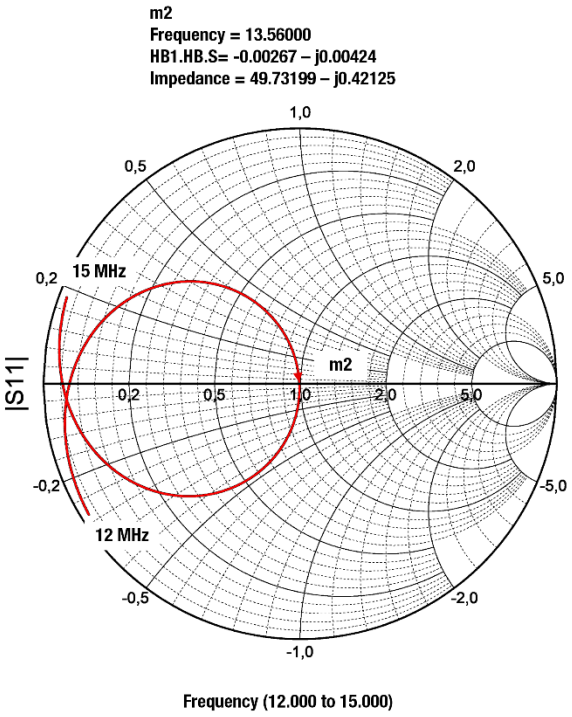


Figure 10: Simulated Smith chart of $|S_{11}|$ over frequency

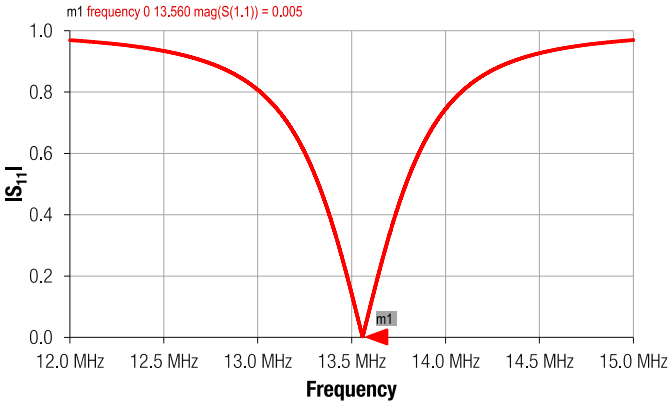


Figure 11: Simulation of $|S_{11}|$ over frequency

For the frequency of 13.56 MHz, the simulated values of the input impedance Z_{in}' and the reflection coefficient S_{11} are shown in table 5.

Parameter	Simulated value
Z_{in}'	(49.73 - j 0.42) Ω
S_{11}	-0.00267 - j 0.00424
$\|S_{11}\|$	0.005
$\|S_{11}\|_{dB}$	-46 dB

Table 5: Simulated values for the reflection coefficient and the input impedance at 13.56 MHz

The simulated values of S_{11}, $|S_{11}|$ and $|S_{11}|_{dB}$ are quite similar to the calculated values, shown in table 4. Because of the precise values of C_A = 45 pF and C_B = 337 pF, the reflexion parameters are very low. In real circuits, these precise capacitor values can not be reached, which results into higher reflexion coefficents.

3.16 Verification of the matching by measurement and matching adaptation measures

The calculations and simulations are done with ideal components, without considering parasitic and tolerance influences. But in reality the capacitors and the inductors have tolerances and parasitic inductances and capacitances. In addition, the components of the equivalent circuit of the antenna are measured and therefore not very precise, thus have tolerances, too. Apart from geometrical and chemical variations induced by the manufacturing process, there are also uncertain parameters about the surrounding or additional ferrite foils for shielding. Any metal in close proximity to the antenna may change the impedance of the antenna and thus also the overall input impedance. How much these variations influence the input impedance mainly depends on the desired value of the input impedance and the quality factor. In general, lower quality factors and higher values of the input impedance are less sensitive to a variation of component values. [5]

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In view of these variations, an additional iteration of the matching is necessary. In the first step, the differential output circuit, which is shown in figure 5, with the calculated matching capacitor values has to be manufactured. The second step is to measure the input impedance with a network analyzer. The matching capacitors C_A and C_B have to be adjusted, so that the input impedance reaches the value of 50 Ω.

3.17 Differential NFC output circuit board and reflection coefficient measurement

To design the PCB of the differential output circuit, the program Altium Designer V.18.1.19 has been used. The Altium schematic of the differential NFC output circuit is shown in figure 12.

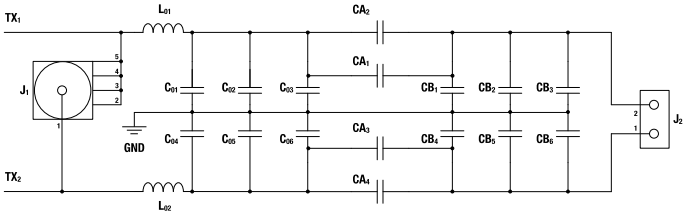


Figure 12: Altium schematic of the differential NFC output circuit

For the filter inductance L_0 = 470 nH (see L_{01} and L_{02} at figure 12) the WE part **744762247GA** has been used. The filter capacitance C_0 = 247 pF comprises of two 100 pF parts of the WE Caps Series with part number **85012006038** (see C_{01}, C_{02}, C_{04} and C_{05} at figure 12) and the 47 pF capacitor with part number **885012006055** (see C_{03} and C_{06} at figure 12). The same capacitor has been used for C_A (see C_{A2} and C_{A4} at figure 12). The capacitance C_B = 337 pF comprises of a 330 pF capacitor with WE part number **885012006041** (see C_{B1} and C_{B4} at figure 12) and a 6.8 pF capacitor with WE part number **885012006050** (see C_{B2} and C_{B5} at figure 12). At the input port J1 of the network, the following impedance vs. frequency dependency has been measured with an Agilent Technologies E5061 network analyzer in the frequency band between 12 MHz and 15 MHz.

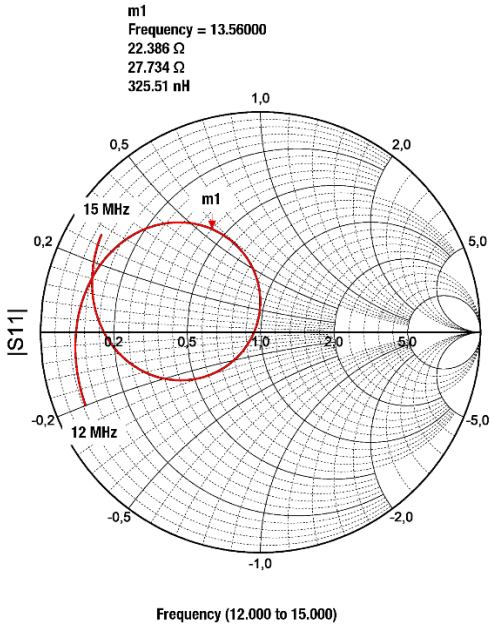


Figure 13: Measurement of the reflection coefficient in dependence of the frequency shown in a Smith chart

A marker at the frequency of 13.56 MHz shows an input impedance of 22.4 Ω + j 27.7 Ω.

Due to equations (15) to (18) the calculated logarithmic absolute value of the input port voltage reflection coefficient $|S_{11}|_{dB}$ yields to -5.94 dB.

The value has been verified by measuring the $|S_{11}|_{dB}$ in dependence of the frequency. The measurement result is shown in figure 14.

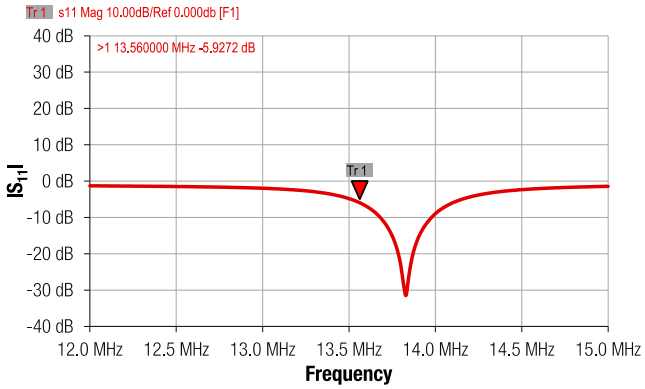


Figure 14: Measurement of the logarithmic absolute value of the input port voltage reflection coefficient in dependence of the frequency

The marker at the frequency of 13.56 MHz yields $|S_{11}|_{dB}$ = -5.92 dB. To improve the matching, at least one of the matching capacitors has to be adapted.

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3.18 Determination of the necessary adaption of the matching capacitance values by simulation

To determine which of the matching capacitors has to be adapted, the influence of both capacitors C_A and C_B on $|S_{11}|$ and Z_{in}' is investigated by an ADS parameter sweep simulation.

Figure 15 shows the Smith chart of $|S_{11}|$ in dependence of C_A and C_B at the frequency of 13.56 MHz.

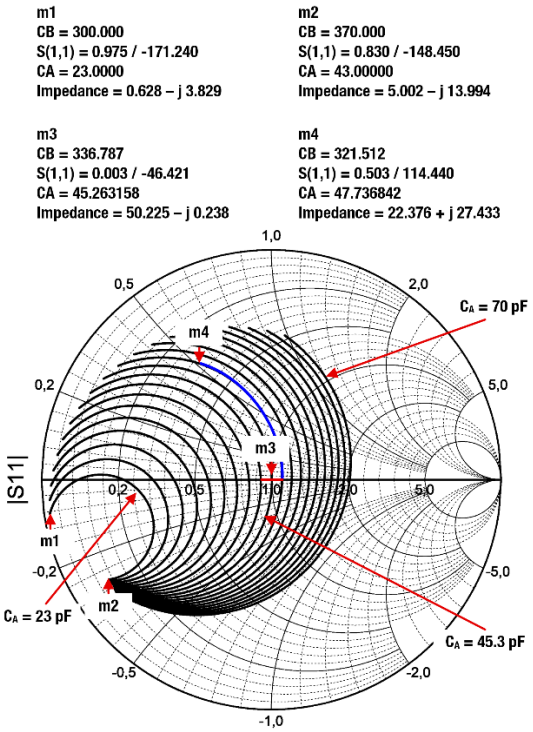


Figure 15: Simulation of $|S_{11}|$ in dependence of C_A and C_B shown in a Smith chart

The parameter sweep of C_B goes from 300 pF to 370 pF in 0.7 pF steps, which is a 10 % variation of the nominal value of 337 pF and the parameter sweep of C_A goes from 23 pF (the left curve) to 70 pF (right curve) in 2.5 pF steps, which is a variation of 50 % of the nominal value of 47 pF. In figure 15, the marker 'm3' shows the desired 50 Ω matching point, which occurs at the curve for C_A = 45.3 pF. The red line indicates the tolerance range of the values of capacitance C_A and the blue line indicates the difference of the capacitance C_B, which leads from the measured impedance (indicated by marker 'm4') to a matching value of about -26.6 dB. At the marker 'm4', the value for C_B is 322 pF, which is about 15 pF smaller than the value of 337 pF, which is reached at the real axis of the Smith chart. This implies, that a capacitor of 15 pF has to be added to C_B, to reach the desired matching point. At the matching test board (see schematic at figure 12) for C_{B3} and C_{B6}, two 15 pF capacitors (one at C_{B3} and one for C_{B6}) have been soldered in (WE number 885012006052). After the integration of the capacitors, the measurement, which is shown in the figures 13 and 14 has been repeated. The results are shown in the figures 16 and 17. The measurement shows, that the parameter $|S_{11}|$dB changed from -5.92 dB to the value of -26.1 dB at a frequency of 13.56 MHz due to the optimization of C_B.

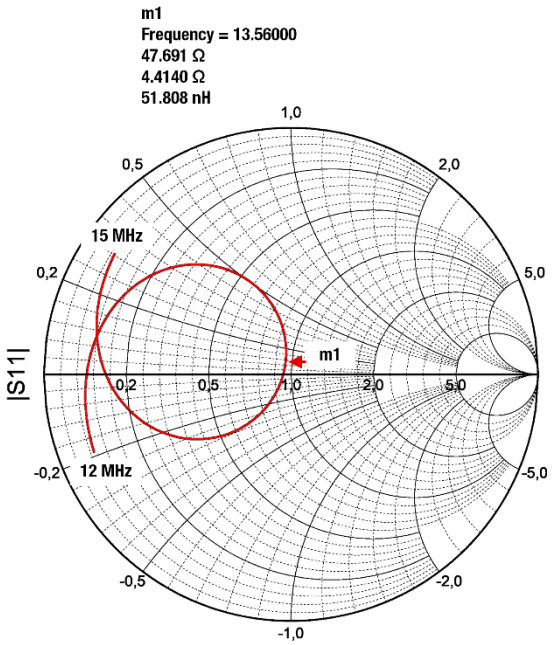


Figure 16: Measurement of the reflection coefficient with frequency shown in a Smith chart after the matching capacitor C_B has been adapted

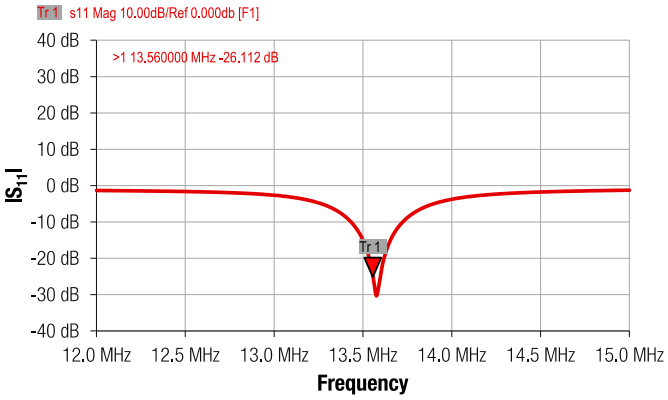


Figure 17: Measurement of the logarithmic absolute value of the input port voltage reflection coefficient with frequency

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3.19 Determination of the necessary adaption of the matching capacitance values by calculation

The change of C_B and C_A to reach the necessary 50 Ω matching impedance can also be done mathematically. To do so, the following calculation steps have to be performed:

1. Integrate the measured value for Z_{in}' (e. g. 22.4 Ω + j 27.7 Ω) into equation (16) and solve the equation for Z_{in}:

$$Z_{in} = \frac{4j\omega L_0 - 2Z_{in}'}{j\omega C_0 Z_{in}' - 2 + 2\omega^2 C_0 L_0} = R_S - jX_S \qquad (19)$$
$$= (87.18 - j\,70.95)\ \Omega$$

2. Integrate the imaginary part (X_S = 70.95 Ω) and real part (R_S = 87.18 Ω) of Z_{in} calculated with equation (19) into the equations (8) and (9) and calculate $X_{A1,2}$ and $X_{B1,2}$.

X_{B1} = -73 Ω

X_{B2} = -54.74 Ω

X_{A1} = -489 Ω

X_{A2} = 347.27 Ω

3. Use equation (14) to calculate the corresponding $C_{A1,2}$ and $C_{B1,2}$ values:

C_{B1} = 322 pF

C_{B2} = 429 pF

C_{A1} = 48 pF

C_{A2} = -67 pF

Because C_{A2} yields a negative value, the capacitances C_{B2} and C_{A2} can be neglected.

4. Determine the difference of the resulting C_A and C_B values to their calculated or simulated values. The calculated values (see section 3.9) for C_A an C_B are:

$C_{A,calc}$ = 45 pF

$C_{B,calc}$ = 337 pF

The values for C_A and C_B determined via measurement in the previous section are:

$C_{A,meas}$ = 48 pF

$C_{B,meas}$ = 322 pF

The differences of the capacitances ΔC_A and ΔC_B are:

$\Delta C_A = C_{A,calc} - C_{A,meas} - = 45\ pF - 48\ pF = -3\ pF$

$\Delta C_B = C_{B,calc} - C_{B,meas} = 337\ pF - 322\ pF = 15\ pF$

With these results, the capacitance C_A has to be decreased by 3 pF and the capacitance C_B has to be increased by 15 pF to reach a perfect impedance match. Because the difference value of ΔC_A = 3 pF is about in the range of the tolerance of C_A, this value has not been changed. In this section it has been shown by simulation and measurement, that the capacitance C_B has to be increased by 15 pF to reach a matching value of about -26 dB. It has to be considered, that the output circuit is differential, which means that the 15 pF has to be soldered twice into the matching circuit, at the places C_{B3} and C_{B6} (see figure 12).

4 Summary

This application note describes how a WE NFC antenna can be matched to a NFC IC. The determination of the matching capacitor values has been described by calculation and simulation. The measurement and necessary adaption of the input impedance to the source impedance has been shown. The basics of the complex impedance matching theory have been stated and typical L-matching topologies are shown. The calculation of the matching reactances and the determination of the Q factor have been performed. The different parts of differential output network have been analyzed and the dimensioning of the single components of the filter and the equivalent antenna components has been performed. Finally, the differential output network has been manufactured and the matching capacitors have been adapted to improve the matching.

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A Appendix

A.1 Bill of Materials

Index	Description	Size	Specifications f_{WPT} = 125 kHz; f_{NFC} = 13.56 MHz	Partnumber
	WPT/NFC Combo Transmitter Coil	Ø65 x 3.5 mm	$L_{1,WPT}$ = 6.3 µH, $L_{2,NFC}$ = 1.2 µH, $Q_{1,WPT}$ = 100, $Q_{2,NFC}$ = 80	760308101150
	WPT/NFC Combo Transmitter Coil	Ø54 x 4 mm	$L_{1,WPT}$ = 24 µH, $L_{2,NFC}$ = 0.7 µH, $Q_{1,WPT}$ = 125, $Q_{2,NFC}$ = 30	760308101312
	WPT/NFC Combo Receiver Coil	45 x 44 x 0.72 mm	$L_{1,WPT}$ = 8.8 µH, $L_{2,NFC}$ = 1.6 µH $Q_{1,WPT}$ = 30, $Q_{2,NFC}$ = 47	760308103305
	WPT/NFC Combo Receiver Coil	44 x 44 x 1 mm	$L_{1,WPT}$ = 8 µH, $L_{2,NFC}$ = 1.4 µH, $Q_{1,WPT}$ = 19, $Q_{2,NFC}$ = 47	760308102306
	WPT/NFC Combo Receiver Coil	40 x 27 x 1 mm	$L_{1,WPT}$ = 7.8 µH, $L_{2,NFC}$ = 1.6 µH, $Q_{1,WPT}$ = 19, $Q_{2,NFC}$ = 47	760308103307
L_0	WE-KI SMT Wire Wound Ceramic Inductor	1008	L = 470 nH, Q = 45	744762247GA
C_0	WCAP-CSGP Multilayer Ceramic Capacitor	0603	C = 100 pF	885012006038
C_0	WCAP-CSGP Multilayer Ceramic Capacitor	0603	C = 47 pF	885012006055
C_A	WCAP-CSGP Multilayer Ceramic Capacitor	0603	C_{A1} = 47 pF	885012006055
C_{B1}	WCAP-CSGP Multilayer Ceramic Capacitor	0603	C = 330 pF	885012006052
C_{B2}	WCAP-CSGP Multilayer Ceramic Capacitor	0603	C = 6.8 pF	885012006041
C_{B3}	WCAP-CSGP Multilayer Ceramic Capacitor	0603	C = 15 pF	885012006050
J_1	WR-SMA SMA PCB THT Jack Straight		SMA connector	6031200211450 3
J_2	WR-TBL Terminal Block		2.54 mm horizontal entry clamp	691210910002

A.2 Equivalent antenna circuit parameters

Artikelnummer	f_S (MHz)	R_a (Ω)	L_a (µH)	C_a (pF)
760308101150	48	1.19	1.16	9.48
760308103305	50	4	1.58	6.4
760308102306	50	2.4	1.6	6.3
760308103307	54	3.7	1.7	5.1

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A.3 Literature

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[4] Würth Elektronik eiSos, ANP057a, WE-MCA Multilayer Chip Antenna Placement & Matching, Application Note, 2018.

[5] T. Baier: Automated Impedance Adjustment of 13.56 MHz NFC Reader Antennas, Master Thesis, 2014.

[6] A. Schober, M. Ciacci, and M. Gebhart, An NFC Air Interface coupling model for Contactless System Performance estimation, International Conference on Telecommunications (ConTEL), June 2013.

[7] Rohde und Schwarz, Near Field Communication (NFC) Technology and Measurements, White Paper, 2011.

[8] NXP Corporation, AN11564, Antenna Design and Matching Guide, Application Note, 2016.

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