AUTOMATIC ANTENNA TUNERS

THEORY AND DESIGN

Ettore Lorenzo Firrao

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I would like to dedicate this thesis to my parents and to my family

Preface

After finishing the university, I started to work on my Ph.D. in the IC Design group at the University of Twente, Enschede, The Netherlands. The topic of my Ph.D. is "Automatic Antenna Tuners" which are used for mobile phones, radars and measurement equipments. The project was supported by Philips Semiconductors Nijmegen (Nijmegen, The Netherlands) (nowadays NXP Semiconductors). I chose to carry out a Ph.D. because I like to specialize on microwave and antennas.

To finance my PhD I started to work for Philips Semiconductors Nijmegen. My job was design and verification/characterization of passive circuitry used in RF Power Amplifiers and RF Front-End Modules for GSM applications. In particular the tasks were the design of output matching networks for RF power amplifier for GSM devices.

After the sale of the company and a major reorganization, I continued my career in Thales Hengelo (Hengelo, The Netherlands) where I could further develop my skills by designing a heterodyne receiver with double conversion and the characterization of an RF amplifier in X band and later at Selex Italy, where I designed and programmed FPGAs for space applications.

I restarted my researcher career at the University of Twente in 2012. This thesis describes the work done during my Ph.D. period and during my part-time research at the university in the past years.

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Chapter 1

Introduction

1.1 Wireless communications

Wireless communication deals with transmitting and receiving data (voice, video, etc.) over a (long) distance without the use of cables, wires or any other conductor. Generally speaking there can be more than one transmitter and more than one receiver. For instance for radios and TVs, there is one transmitter (the radio station or the TV station) and many receivers (all the radios and the TVs that we all have at home). For Wi-Fi connections each device (the router and all the computers in the Wi-Fi networks) can transmit and receive (in literature they are known as transceivers). For GSM connection each device (mobile phone) can transmit and receive but not at the same time and there is more than one device in the GSM connection.

Wireless communication began a long time ago with the invention of the wireless telegraphy by Guglielmo Marconi, an Italian inventor who read all relevant work that was published at that time on physics and made portable transmitters and receivers which could work over long distances. Guglielmo Marconi based his work on the work of people like James Clerk Maxwell, who developed the theory of electromagnetic waves, and Heinrich Rudolf Hertz, who studied Maxwell's theory and verified it in a lab. In 1909 Guglielmo Marconi (together with Karl Ferdinand Braun) was awarded the Nobel Prize in physics for the development of the wireless telegraphy.

From the application called wireless telegraphy equipments, wireless communication was improved with the radio (the ability to broadcast voice and music over long distances) and, later on, with TV (the ability to broadcast voice, music and video over long distances). A simple example of wireless radio (voice and music) communication consists of one transmitter and one receiver. In practice there can be more than one receiver but the principle is the same. The first person to broadcast audio was the Canadian inventor Reginald Aubrey Fessenden in December 1900. He also invented a type of modulation called "Amplitude Modulation" or simply "AM". In this type of modulation, the amplitude of the broadcasted electromagnetic signal is modulated or changed according to the signal that has to be transmitted. On the 2nd of

November 1936 the BBC began to broadcast the first regular television service from Victorian Alexandra Palace in London. This date is considered the birth of the television broadcasting as we know today, although experiments were carried out in different places before this date.

Wireless communication further developed with the introduction of handheld mobile phones. The first phone call with a handheld mobile phone was on the 3rd of April 1973 by Dr. Martin Cooper from Motorola. In his first phone call Dr. Cooper called his rival Dr. Joel S. Engel from Bell Labs. The prototype had a weight of 1.1 kg and the battery could last only 30 minutes (after being charged for 10 hours). The first commercial handheld mobile phone was then introduced from Motorola at the beginning of the 1980s.

From the introduction of the handheld mobile phones until present several generations have been developed: 1G or Analogue cellular (introduced on the 13th of October 1983 in U.S. and shut down in 2008 in U.S.), 2G or Digital cellular (introduced at the beginning of the 1990s), 3G or Mobile broadband, 4G or Native IP networks and finally 5G.

Today in our life wireless communication is everywhere: in our radios, in our televisions, in our mobile phones, in our computers, in our tables, etc. (GSM connection, Wi-Fi connection and Bluetooth connection). In all the applications mentioned, antennas are used to transmit and to receive signals. In some cases the same antenna can be used to transmit and to receive like for mobile phones and base stations (GSM connection) and for computers and routers (the device between the ADSL and the Wi-Fi connection). Generally speaking in one device (like a mobile phone, a computer, a tablet, etc.) one antenna is used for each application. For instance one antenna for the GSM connection, one antenna for the Wi-Fi connection and so on. The reason is that it is difficult to optimize a single antenna for all the connections and to allow the use of more than one application at the same time. For each application the antenna is designed to have a predefined behaviour (i.e. a 50 ohm input impedance) but in real life it can be different, and often in an unpredictable way.

1.2 The RF front end

As it has been described in the previous section, a key component for wireless connections is the antenna. There is a wide range of antennas which can be used in a device: monopoles, helical antennas, PIFA antennas, ceramic antennas, etc. Generally they are designed for a 50 Ω input impedance in an ideal environment. The ideal environment often does not include the near presence of the user or other objects. Because of a non-ideal electromagnetic environment, the antenna impedance can differ from its ideal value.

Every antenna is normally connected to an RF front end. A simplified block diagram of a typical RF front end for transmission and for reception is shown in Fig. 1.1.

Fig. 1.1: Simplified block diagram of a typical RF front end. The right hand side Smith chart represents a typical impedance range of the antennas; including an arbitrary phase shift the PA may experience impedances as represented in the left hand side Smith chart.

Generally speaking an RF front end consists of an RF power amplifier (RF-PA) that can drive the antenna in transmit mode, a low noise amplifier (LNA) to amplify the small input signal in receive mode and a switch or a diplexer (depending on the application) to switch between receive and transmit mode of operation. During the design of the RF front end, the antenna impedance is often assumed to be $50\ \Omega$. During the operation, the antenna impedance can differ from $50\ \Omega$ depending on the unpredictable electromagnetic environment. This EM environment can change over time and from user to user making it difficult to optimize the RF front end for a well-defined antenna impedance (even if the antenna impedance is simply different from $50\ \Omega$) ([1] - [21]). Also, the presence of the user affects both input (TX) and output (RX) impedance because it is in the near field of the antenna. In fact, in case of a well-defined antenna impedance different from $50\ \Omega$, a standard matching network could be used to tune back the antenna impedance to an impedance optimum for the RF front end. However a well-defined antenna impedance is not possible with varying near-field conditions, resulting in antenna impedance mismatch. This mismatch will affect transmission and reception in a negative way. The work in this thesis mainly deals with transmission but it can also be extended to reception. In the remaining part of this section antennas and RF power amplifiers are discussed in more details.

RF power amplifiers (RF PAs) are used to drive antennas. An RF power amplifier can be designed using different techniques depending on the specifications such linearity, efficiency, ACPR (Adjacent Channel Power Ratio), etc. The specifications are application related. For example GSM applications, Bluetooth applications, Wi-Fi applications, and many more have different specifications. A technique to design an RF power amplifier is defined as class of operation [22]. There are several classes of

operation. Some of them are: class A, class AB, class B, class C, class E, class F, inverse class F, and more. For class A, class AB, class B and class C the RF transistor is biased and driven with a sine wave. The class of operation is defined by the conduction angle: the part of the sine wave when the transistor in the RF power amplifier is “conducting”. Class A has a conduction angle of 360° , class AB between 180° and 360° , class B of 180° and class C less than 180° . For class E, class F and inverse class F, the transistor is in switch mode: the transistor is acting like a switch (open or closed). For any RF power amplifier, the load impedance of the entire RF power amplifier is typically assumed to be $50\ \Omega$.

A simplified block diagram of an RF power amplifier is sketched in Fig. 1.2 for class A, class AB, class B and class C RF power amplifiers. Generally speaking an RF power amplifier consists of an input matching network, an RF transistor and an output matching network.

Fig. 1.2: Simplified block diagram of an RF power amplifier for class A, class AB, class B and class C RF power amplifiers: the conduction angle determines the exact class of operation.

The input matching network is designed to match the input impedance of the RF power transistor to an optimum input impedance Z_{in} (typically $50\ \Omega$), to bias the transistor base or gate and for harmonic termination.

The RF power transistor can be a bipolar or a field effect transistor. It can be implemented in various technologies, such as Si, SiGe, GaAs, GaN or InP. The choice of the technology depends on the required performance (frequency of operation, gain, noise figure, power capabilities, voltage and current capabilities, etc.) and on the costs.

The output matching network is designed to transform the load impedance Z_{load} (typically $50\ \Omega$) to an optimum impedance Z_{opt} for the RF transistor, to terminate (to short) the harmonics and to bias the RF transistor collector or drain illustrated in Figure 1.2. For class A, class AB, class B and class C RF power amplifiers, this optimum

impedance Z_{opt} (to get maximum output power) is given by

$$Z_{opt} = \frac{(V_{supply} - V_{knee})^2}{2P_{max}} \quad (1.1)$$

where V_{supply} is the supply voltage, V_{knee} (or V_{sat}) is the knee voltage for bipolar transistors (or the saturation voltage for MOS transistors) and P_{max} is the desired maximum output power. This optimum load impedance for the PA usually is not equal to 50Ω . This equation can give a starting value for the optimum impedance for the transistor but the actual value should be found using load pulling. The use of Eq. 1.1 shows that low-voltage high-power implementations are difficult because the optimum impedance becomes too low. The equation also shows that high-voltage low-power implementations are difficult if the RF power amplifier is designed in a conventional way. The main limitations are the quality factor of the passive components (thus power efficiency of the output matching network) and the bandwidth.

For an RF power amplifier the collector efficiency is defined as the ratio between the RF power to the load and the DC power drained from the power supply. For a class A RF power amplifier the maximum collector efficiency is $1/2$ while for a class B RF power amplifier $\pi/4$. Ideal equations of the collector efficiency for a class A and a class B RF power amplifiers as a function of the desired output power normalized to the desired maximum output power respectively are shown below

$$class A \Rightarrow \eta_A = \frac{1}{2} \frac{P}{P_{max}} \quad (1.2)$$

$$class B \Rightarrow \eta_B = \frac{\pi}{4} \sqrt{\frac{P}{P_{max}}} \quad (1.3)$$

where P is the desired output power (achieved by varying the input power of the RF power amplifier and leaving everything else the same) and P_{max} is the maximum output power. A plot of the collector efficiency as a function of the normalized power (P/P_{max}) is given in Fig. 1.3.

Fig. 1.3: Graph of the theoretical collector efficiency vs normalized output power for class A and class B RF power amplifiers.

All these results are based for an optimum load impedance (typically 50Ω). What happens in case the load impedance is not optimal? The RF power amplifier has to work under load mismatch conditions resulting in a lower maximum output power, lower efficiency, stability issues, compromised ruggedness, etc. ([23] – [43])

A lot of work has been carried out on understanding the behavior of an RF power amplifier under mismatch. In [48] a simple model of the output power of a class A, class AB, class B and class C RF power amplifier under load mismatch conditions has been proposed. The derivation is based on constant power circles because the RF power amplifier is either voltage limited or current limited (clipping into impedances with a constant conduction or resistance). An example of a constant output power contour at the output of a class A, class AB, class B and class C RF power amplifier is shown in Fig. 1.4.

In case of a Z_0 transmission line at the output of the RF power amplifier, the constant power contours at the output of the RF power amplifier are rotated around the center of the Smith chart. An example is in Fig. 1.5.

Fig. 1.4: Constant output power contour of an RF power amplifier under mismatch on a Smith chart.

Fig. 1.5: Rotation of the maximum output power contour due to a transmission line at the output of the RF power amplifier.

1.3 Motivation and outline

1.3.1 Motivation

Section 1.1 and section 1.2 described the world of wireless communication and

the fundamental block, RF front end, that is present in wireless RF devices. The RF front end mainly consists of matching networks that are essential key component. Without matching networks our wireless RF devices do not work. Therefore the design and the optimization of these matching networks are very important. Generally speaking these matching networks do not have tunable components and they are referred to as conventional matching networks. This type of matching network has been analyzed for a long time and is well described in the literature.

Because of the continuous and unpredictable changing of the antenna's EM environment and then of the antenna impedance, it is impossible to optimize the conventional matching networks inside the RF front end. Therefore automatic antenna tuners are needed inside a wireless RF device. Automatic antenna (load) tuners can also be used in measurement equipments. An automatic antenna tuner consists of a tuneable matching network, sensors and a control network. Using these automatic antenna tuners, the conventional matching networks, mentioned in the previous paragraph, become tunable matching networks and, in most of the cases, switchable matching networks when switches are used to insert in or out of the circuit the reactive components (capacitors and inductors). Since the switchable matching networks are tunable they can cope with the unpredictable changes of the antenna impedance.

A lot of work has been carried out on automatic antenna tuners for mobile phones but the combination of wideband matching to a variable impedance in combination with an efficient algorithm to find the right match is still a challenging problem. Generally speaking new circuit topologies are proposed in the literature but little attention is paid to their optimization. The problem of optimization is addressed in this PhD thesis.

The work described in this PhD thesis deals with the hardware aspects of automatic antenna tuners but the application is not limited to mobile phones or measurement equipments. In fact throughout the entire PhD thesis it is assumed to have an unpredictable time-varying load which can be very general: antenna, microwave oven, etc. Also in this PhD thesis, the simplification of automatic antenna tuner concepts is addressed in combination with the research questions that address the discrete control of the matching states (such as accuracy, overhead and mapping density).

1.3.2 Outline

In chapter 2, the theory and the state of the art of automatic antenna tuners are described. All the work described in this chapter is based on the existing literature except for the new literature proposed in the remaining part of this thesis.

In chapter 3, as part of research carried out on automatic antenna tuners, an automatic antenna tuner is proposed based on the measurement of three RF signal amplitudes resulting in a low-cost and low-power control loop. This work underlines the importance of power saving in handheld devices.

Chapter 4 is about a theoretical analysis on the derivation of the near minimum number of states for switchable matching networks based on the required input VSWR and on the maximum load VSWR. On a first approach, the switchable matching network is assumed to be lossless. Then the formalism is extended to low loss switchable matching networks. All the derivation is based on analysis and not on brute-force simulations or optimizations.

In chapter 5, an implementation of a switchable matching network is presented trying to achieve the near minimum number of states as described in chapter 4. Several switchable matching networks are analyzed: PI networks, loaded transmission lines, one circulator networks and cascaded circulator networks. In order to benchmark all the switchable matching networks, a Figure of Merit is introduced: the hardware overhead. It is defined as the ratio between the actual number of states of the switchable matching network under investigation and the minimum number of states given in chapter 4. According to our Figure of Merit the cascaded circulator network has the lowest hardware overhead.

Chapter 6 summarizes the conclusions and discusses the future directions that could be followed for a better understanding of the automatic antenna tuners.

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Chapter 2

Automatic Antenna Tuners

2.1 Introduction

Matching is very important in RF / microwave circuits for the transfer of the maximum power to the load, which is especially important due to the limited power and gain available from RF and microwave devices..

Fig 2.1: Definition of some symbols (where $\Gamma_s = (Z_s - Z_0)/(Z_s + Z_0)$ and $\Gamma_L = (Z_L - Z_0)/(Z_L + Z_0)$).

For instance in case of a 50Ω source ($Z_s = 50 \Omega$ or $\Gamma_s = 0$) driving a 50Ω transmission line, the power reflected by the load to the source P_{ref} is

$$P_{ref} = |\Gamma|^2 * P_{source} \quad (2.1)$$

and the mismatch efficiency M (indicating the factor with which the power transfer to the load is reduced) is

$$M = 1 - |\Gamma|^2 \quad (2.2)$$

where

$$\Gamma = \frac{Z_L - 50}{Z_L + 50} \quad (2.3)$$

where Z_L is the load impedance. See Fig. 2.1. As it is possible to see the power going to the load has a parabolic behaviour with respect to the magnitude of the reflection coefficient.

In case the 50Ω source is replaced by an RF power amplifier, the power drop is dependent on the reflection coefficient in terms of magnitude and phase (not only magnitude as for the previous case). The relationship between power drop and reflection coefficient is not as easy as the previous case. In practice load pull data is used. Generally speaking with a return loss of 10 dB (VSWR = 2:1), the power drop can be up to 50 %. This means that in these cases the power amplifier has to drain more current from the battery or to have call drops because the RF power amplifier is unable to reach the required power level.

To cope with the unpredictable variations of the antenna (load) impedance, automatic antenna tuners are used today ([1] - [76]). The automatic antenna tuner is placed between the RF power amplifier and the antenna as shown in Fig. 2.2. An automatic antenna tuner consists of:

- *sensors* ([19] – [27]): the tasks of the sensors is to measure the mismatch / matching of the antenna or load. They can measure the impedance, the reflection coefficient, the return loss, the VSWR or any relevant parameter. The most important parameter for the sensors is the insertion loss which should be as low as possible. Depending on the particular implementation other parameters could be important like directivity, sensitivity, dynamic range. For instance for a directional coupler the directivity defines the minimum value of return loss that can be measured.
- a *tunable matching network* ([28] - [60]): the tunable matching network is the matching network that achieves the tuning of the antenna or load. It is made of reactive components (capacitor, inductors, transformers, transmission lines, etc.) that can be inserted in and out of the circuit. The most important parameter to benchmark a tuneable matching network is the power efficiency. Obviously in the RF front end it is required to have the highest possible efficiency. Other parameters could be the control method (continuous tuning or switchable) and the bandwidth.
- a *control network* ([61] - [75]): achieving the tuning measuring the mismatch /

matching and tuning the tunable matching network. It is typically made of a microprocessor (see Fig. 2.3). The main goal of the software inside the microprocessor is to achieve the tuning of the tunable matching network in the fastest way and with the highest efficiency. A very simple way of tuning could be to try all possible combination until the optimum combination is found. This algorithm would be very inefficient from a time point of view and also from speed and efficiency point of view. Smarter algorithms are needed.

Some commercial products are available in the market [76].

Fig. 2.2: Location of an automatic antenna tuner; the automatic antenna tuner tunes any antenna (load) impedance in a certain large range (see the rightmost Smith chart) to a much smaller region centered and the optimum Z_{load} for the RF PA (see the leftmost Smith chart).

2.2 Antenna tuner example

This section describes a representative implementation of an automatic antenna tuner used for mobile transceivers. The automatic antenna tuner is published by Van Bezooijen ([77] - [79]) and the figures are reproduced here with permission. The first figure, Fig. 2.3, is a “photo” of the actual hardware. The implementation is quite small, but, as it can be seen from the schematic in Fig. 2.4, the implemented matching network contains quite a number of components. Clearly identifiable in this figure are the sensors, built around a sensing coil, and the matching network components (one inductor and two capacitor banks). In total 2^{10} states can be made with this matching network. Not visible is the control electronics and software; these are essential but not treated in details.

Fig 2.3: Photo of the automatic antenna tuner [77].

Fig. 2.4: schematic of the automatic antenna tuner [79].

2.3 Sensors for antenna tuners

2.3.1 Introduction

The sensors are needed to detect the mismatch / matching and they can measure the impedance, the reflection coefficient, the VSWR or any relevant parameter related to the matching.

Fig. 2.5: Placement of the sensors: at the input and at the output.

The sensors can be implemented in many ways ([19] - [27]) and they can be placed at the input, at the output and/or at the input of the tunable matching network (see Fig. 2.5). In relationship to the measured parameter, the system can be feedforward or feedback. For instance placing the sensors at the output of the tunable matching network results in a feedforward system because the load reflection coefficient or load impedance is assumed to not be a function of the state of the tunable matching network. The disadvantage could be high voltages and/or currents in case of highly mismatched load.

While measuring the reflection coefficient at the input of the tunable matching network results in a feedback system because the reflection coefficient and the impedance are a function of the antenna impedance and of the tunable matching network.

2.3.2 Reflectometer

As it is well-known from the literature, the easiest way to measure a mismatch is to use a reflectometer. It is based on the standard theory of the directional couplers which can be used to measure the incident and the reflected wave. Using two log-peak detectors (one for the incident wave and one for the reflected wave) it is possible to measure the return loss or the VSWR as depicted in Fig. 2.6. The disadvantage of this method is that it is relatively bulky (due to the directional coupler) and relatively complex (due to the two log-peak detectors).

Fig. 2.6: Example of reflectometer.

2.3.3 Five-Port Reflectometer

In [27] a five-port reflectometer is suggested for computationally simple measurements. As it is shown in Fig. 2.7; it consists of a directional coupler, a Wilkinson power divider, an attenuator and three power detectors. Using the values of the three power detectors, it is possible to derive the complex load impedance. Compared to the reflectometer from Fig. 2.6, this five-port reflectometer is even more bulky, as it includes not only a directional coupler, but also a Wilkinson power divider and a 90° hybrid coupler.

Fig. 2.7: Five-port reflectometer.

2.3.4 Mismatch Detector

In [21] a phase detector is proposed and it consists of a directional coupler and a detector. The detector is made of a 90-degree phase shifter, two limiters and a multiplier. This detector can be used to measure the sign of the phase between the reflected wave and the incident wave. The magnitude of the output can not be related to the value of the phase.

Fig. 2.8: Implementation of the mismatch detector in [21].

2.4 Matching networks for antenna tuners

2.4.1 Introduction

The tunable matching network can be made with continuously tunable components or with switchable components ([28] - [60]). The component value of the first ones can only be tuned continuously from a minimum value up to a maximum value. The latter can only have a discrete number of values from a minimum up to a maximum (normally binary weighted).

The tunable matching networks can be implemented in several ways. Generally speaking there are two types of switchable matching networks: with lumped elements and with distributed elements. Among the lumped element switchable matching networks there are PI networks, T networks and L networks (which can be seen as a degeneration of a PI or T network choosing a component value equal to zero in a proper way, see Fig. 2.9). Among the distributed element switchable matching networks there are loaded transmission lines, branch line couplers and circulators.

Without entering into too many details, the PI networks, the T networks, loaded transmission lines, branch line couplers and circulators can match any load impedance

on a Smith chart to the optimum impedance (typically the center of the Smith chart). L networks cannot match any load impedance.

The main issue in the design of switchable matching networks is their power efficiency. Generally speaking the higher the better. It is therefore important to do not overdesign the switchable matching network ([59] - [60]) in order to maximize the power efficiency because the higher the number of the components the higher the losses. The exact power efficiency depends heavily on the actual implementation.

2.4.2 Matching networks with two reactive components (L-matching networks)

The simplest two-element matching network is the L-matching network as it is depicted in Fig 2.9.

Fig. 2.9: Typical implementations of L-matching networks.

If we limit both elements to L's or C's, the following eight options can be distinguished.

Fig. 2.10: Examples of L matching networks.

It is clear that Fig. 2.10 (a) and 2.10 (b) can not match capacitive components. To be more precise, the area in the Smith chart that can be matched is shown in Fig. 2.11.

Fig. 2.11: On the left the area on the Smith chart that can be matched by the network topologies in Fig. 2.10 (a) and 2.10 (b); on the right the area on the Smith chart that can be match by the network topologies in Fig. 2.10 (c) and 2.10 (d).

Similarly, for the structures in Fig. 2.10 (c) and 2.10 (d), no inductive impedance can be matched.

The more interesting cases are the figure with one inductance and one capacitance: the matchable areas for Fig. 2.10 (e) and 2.10 (f) are shown in Fig. 2.12.

Fig. 2.12: Area on the Smith chart that can be matched by the network topologies in Fig. 2.10 (e) and 2.10 (f).

This indicates that the circuits of Fig. 2.10 (e) and 2.10 (f) together can match the full Smith chart. The same holds for Fig. 2.10 (g) and 2.10 (h).

If we now add a switch to Fig. 2.10 (e) and 2.10 (f), we can transform this topology into

Fig. 2.13: Circuit topology from merging the circuit topologies in fig. 2.10 (e) and 2.10 (f).

The matching circuit in Fig. 2.13 can then match any impedance. This also

holds for its dual in Fig. 2.14.

Fig. 2.14: Circuit topology from merging the circuit topologies in fig. 2.10 (g) and 2.10 (h).

These structures are hence very attractive for matching networks.

2.4.3 Matching Networks with three reactive components (*PI* or *T* matching networks)

The L-matching networks with a switchable component can be generalized into three-component matching networks, where we simply implement the shunt component or the series component twice.

This results in the following possibilities.

Fig. 2.15: PI and T matching networks.

If we then stick to identical type series or shunt components, we end up with four topologies, see Fig. 2.16.

Fig. 2.16: Different implementations of the PI and T matching networks.

From the discussion in 2.4.3, it is now clear that all these four implementations can match any impedance, if L's and C's can be chosen freely.

2.4.4 matching networks with distributed components (Loaded Transmission Line based matching network)

The simplest matching network, that can be made of distributed components, is the well-known single stub matching network [28]. The main limitation of this implementation is the well defined distance of the stub from the load, enabling us to match only a circle on the Smith chart.

Fig. 2.17: Single stub matching network and its matchable area on the Smith chart.

The follow-up of the single stub matching network is the double stub matching network [28]. But, as it is well-known from the literature [28], it can not match all the Smith chart since the distances from the load and between the stubs are given.

Fig. 2.18: Double stub matching network and its matchable area on the Smith chart.

In order to overcome the limitation of the double stub matching, the triple stub matching network is used as it is described in the literature [28]. The triple stub matching network is able to match any load impedance.

Fig. 2.19: Triple stub matching network and its matchable area on the Smith chart.

In [36] a loaded transmission line based matching network is proposed. The switchable matching network consists of a transmission line loaded with switchable capacitors and/or inductors. The switchable matching network works as follows: in case of load mismatch only one switch is closed leaving the others open; in this way the right

mismatch circle can be mapped into the input circle. The theory is only derived for a symmetrical and lossless switchable matching network with phase shifters at the input and at the output.

Fig. 2.20 Loaded transmission line with switchable capacitors as proposed in [36].

2.5 Adaptive Matching literature

As described above, an automatic antenna tuner consists of sensors, a tunable matching network and a control network. Each block can be implemented in several ways. The goal during the design of these automatic antenna tuners is the simplicity and the microwave power efficiency. All the implementations, described in the literature, have different complexity and different microwave power efficiency. Therefore some implementations are better than others. For the sensors, some sensors only measure RF magnitudes like the return loss meter. Generally speaking the bandwidth of these RF magnitudes is “narrow” making the electronics behind easier to implement. For the tunable matching networks, some circuit topologies can match the entire Smith chart while others can only match a limited area of the Smith chart. The main issue for tunable matching networks is the number of switches (which is equal to the number of switchable components) needed to match a predefined area of the Smith chart. For the same boundary conditions (i.e. same area on a Smith chart), a large number of switches implies, most likely, a lower microwave power efficiency.

Quite some work [29]-[60] has been carried out in recent years on variable matching networks, exploring the field in many directions [29] - [60].

References [29] – [35] and [38] employ matching networks that employ variable-impedance components. This work is a generalization of lumped-element matching networks, but behaves similar to lumped-element networks in terms of complexity.

References [36] and [37] represent the only recent work found that addresses the complexity of switchable matching networks. The goal of the work is to optimize a

tunable matching network that could match an area on the Smith chart defined by a constant magnitude $|\Gamma_{\text{load}}|$ into an area defined by a constant $|\Gamma_{\text{in}}|$. The authors derived a technique based on a transmission line loaded by switchable capacitors. According to their paper, only one capacitor is insert in the circuit at time. With all the switches open, the tuneable matching network behaves like a through. Closing one switch, a capacitor is inserted in the circuit and a predefined circle on the Smith chart can be matched. Several switchable capacitors are used on the transmission line in order to change the rotation of this circle with respect to the center of the Smith. The main limitation of this implementation is the extra number of states derived from closing more than one switch at time. The result is a lot of overhead.

References [39] – [58] and [60] all deal with specific design procedures for matching networks and with implementations in diverse technologies: standard CMOS, PiN diodes and MEMS technologies. The main limitation of the standard CMOS technology is the low quality factor of passive components. Special CMOS technologies should be used but, probably, with no compatibility with the standard technology. PiN diodes are very widely used and they work quite well. The main drawback is the DC power consumption but this is also dependent on the application. MEMS switches and MEMS switchable capacitors can be used to implement switchable matching networks. The main advantage is the zero DC power consumption but the limitation could be speed.

In [59] a method to know the minimum number of states for switchable matching network is proposed but the method is based on brute-force optimization. The foundation of the idea is the constant mismatch circles well known in the literature [28]. The circles are then placed closed together with some overlapping in order to fill the predefined area on the Smith chart. As mentioned at the beginning of this paragraph, the filling of this area is based on brute-force optimization.

Reference [80] describes an impedance tuner topology for load-pulling measurements. The matching network consists of transmission lines and varactors.

2.6 Methodology

Section 2.3 and 2.4 were about two important building blocks of an automatic antenna tuner: the sensors and the tunable matching network. The main goal of the sensors is to detect the matching / mismatching in a reliable way. While the main goal of the tunable matching network is the matching of the antenna (load) impedance. These two blocks can be implemented in many ways.

Generally speaking, in the sensors, RF magnitudes and phases have to be measured. RF magnitudes are easy to measure and the bandwidth of the resulting signal

is “narrow”. Phases are difficult to measure but in the literature often phase measurements are proposed. Therefore systems with only RF magnitudes are preferred.

The remainder of this thesis will first address a practical aspect: In chapter 3 an automatic antenna tuner is proposed based on only RF magnitude measurement. This work was proposed to avoid the need to measure the phase which is not always easy while the magnitude of RF voltages have low bandwidth requiring low speed electronics.

After this simplified implementation is proposed, we will address the theoretical underpinning of the minimum number of states for switchable matching networks in chapter 4 (a tunable matching network based on passive components and switches used to insert a reactive component in and out of the circuit). The proposed mathematical formalism is only based on analytic computation and not on brute-force computation as it has been already proposed in the literature.

Chapter 5 demonstrates the theoretical work of Chapter 4. The theoretical results are verified against several practical implementations, and conclusions will be drawn on the applicability of several matching network topologies in view of the required number of states.

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Chapter 3

An Automatic Antenna Tuning System Using Only RF Signal Amplitudes

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Abstract — The operating environment of mobile phones fluctuates continuously, due to changing handling conditions and nearby objects. The resulting fluctuations in antenna impedance cause both a decrease in link quality and a higher standing wave ratio, that requires more robust and hence less efficient power amplifier implementations. In this chapter, an automatic antenna tuner system for handheld applications is presented that uses two series reactances combined with three simple RF peak detectors to sense both reactive and real impedance mismatches. The control loop only requires low-frequency electronics which makes it low cost, low power and relatively easy to integrate. Measurements on a demonstrator system show correct behaviour for voltage standing-wave ratio up to 10.

Index Terms — Antenna tuning, high-efficiency power amplifiers (PAs), impedance matching, wireless communication systems.

3.1 Introduction

Mobile devices are widely used today, with as main technology drivers — for the RF front-end — better performance at lower power consumption and lower cost. The antennas used are typically narrow band, small, high-Q devices that are detuned by any

nearby impedances formed by e.g., hands, the head, tables and more [1]–[3]. RF front-ends are typically designed under the assumption of a $50\ \Omega$ antenna impedance, for which maximum efficiency, talk-time, quality of the link, and maximum life-time are obtained. Due to detuning of the narrow band antenna, significant degraded efficiency and life-time can be observed [2]–[4]. This chapter presents an automatic antenna tuner system that can tune fluctuations in antenna impedances, resulting in higher (overall) efficiency and lower demands on robustness aspects for front-end RF blocks. Lower robustness requirements translate into lower demands on design margins that may be exchanged for higher front-end efficiency. A number of requirements on antenna tuners can readily be formulated.

- Environmental fluctuation for antennas occur on a time scale of milliseconds, which is slow compared to the RF signals in the gigahertz range [5]. Consequently a relatively slow control loop is sufficient.
- The insertion loss of an antenna tuner must be small for efficiency reasons.
- The demands on the electronics in the antenna tuner system must be low to simplify implementations and to minimize the power consumption of the control loop itself.
- The antenna tuner system must be able to correct for a wide range of (heavily detuned) antenna impedances. Note that by using an automatic tuner, the bulky and expensive isolators that are widely applied to absorb reflected power become redundant [4].

The chapter is arranged as follows. Section 3.2 presents an analysis of the changing antenna impedance and of RF front-end performance. The automatic antenna tuning system is described in Section 3.3. Measurements on demonstrator implementations are shown in Section 3.4. The conclusions are summarized in Section 3.5.

3.2 Antenna impedance and front-end performance

It is well known from literature that the antenna impedance is affected by any nearby impedance, such as the user’s hand and/or head [1]–[3]. For this chapter, many measurements of the antenna impedance were done to clearly identify the antenna impedance range and its sensitivity to the environment. Two types of widely used

antennas were used: a planar inverted-F antenna (PIFA) and a ceramic antenna. A large number of typical handling positions for the two antennas were considered for measurements of impedances and sensitivities.

For both types of antenna the measurements show significant detuning of the antenna mainly due to a dramatic change of the imaginary part of the impedance. The real part of the antenna impedance appears to be much less sensitive to handling conditions.

The impedance of the antenna roughly lies in the range

$$30 \, \Omega \leq \operatorname{Re}[Z_{load}] \leq 100 \, \Omega$$

and

$$0 \, \Omega \leq \operatorname{Im}[Z_{load}] \leq 100 \, \Omega$$

The RF front-end performance depends heavily on the antenna impedance. Fig. 3.1 shows a simplified block diagram for the RF front-end of a single band mobile phone. A single band front-end is considered here only for simplicity reasons; extending the analysis towards multi-band systems is straight forward. The area covered by the reflection coefficient of the antenna is shown in the Smith-chart at the right-hand side of Fig. 3.1 for the GSM 900 and GSM 1800 bands; the maximum voltage standing-wave ratio (VSWR) for this range is about 10.

In a typical design [7], [8] the RF power amplifier (PA) is assumed to be loaded with a real impedance while the antenna impedance is assumed to be $50 \, \Omega$. Transformation of the assumed $50 \, \Omega$ to a lower real impedance level required by the PA is not addressed in this chapter as it is merely an extra (fixed) matching stage that is irrelevant in the context of automatic antenna tuners. In the remainder of this chapter we assume therefore — without loss of generality — that the PA includes a matching network that ideally sees $50 \, \Omega$ loads. Because the phase shift of each front-end block is strongly design dependent, any (design-specific, fixed) extra phase shift θ must be taken into account. Consequently the impedance of the antenna as seen by the PA equals the antenna impedance with some fixed (design dependent) excess phase shift 2θ . The area covered by the possible antenna reflection coefficients on a Smith chart at the antenna connector then translates into a full circle at the RF PA output, see the leftmost Smith-chart in Fig. 3.1.

Fig. 3.1. RF front-end: the antenna impedance seen by the PA equals the transformed antenna impedance with a design-dependent excess phase shift.

An RF PA is designed to deliver power in a certain resistive load, represented here by the center of the Smith chart. With other load impedances, load-pull analysis [7] shows that the maximum output power is lower than the maximum by an amount

$10 \log \left(\frac{1 + \Gamma_{\text{antenna}} ^2}{1 - \Gamma_{\text{antenna}} ^2} \right) \leq P_{\text{drop}} [\text{dB}] \leq 10 \log \left(\frac{1 + \Gamma_{\text{antenna}} }{1 - \Gamma_{\text{antenna}} } \right)$	(3.1)
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where Γ_{antenna} is the (complex) antenna reflection coefficient. For an antenna VSWR = 10, see Fig. 1, the corresponding reflection coefficient $|\Gamma_{\text{antenna}}| = 0.818$ and P_{drop} is then 7 to 10 dB. Note that the two boundaries in (1) correspond to an imaginary and real Γ_{antenna} , respectively. For an antenna VSWR of two the power drop can be already 3 dB which is significant for mobile applications. In Fig. 3 (shown later), a -1.5-dB contour is included. To be able to transmit at full power levels as defined by (e.g., GSM) standards, without compensating impedance variations with automatic antenna tuners, the PA must be designed for much higher than maximum nominal power levels which significantly decreases efficiency and increases costs [7].

3.3 Automatic antenna tuners

An automatic antenna tuner is a system able to transform the unpredictable or badly defined load impedance to a well-defined impedance (usually 50 Ω) automatically, and it is mainly made up of a tunable matching network, an impedance sensor and a control network. The tunable matching network is typically a ladder network, while the

sensor provides some indication of the (tuned) antenna impedance. In this work, the main focus is on efficient impedance measurements and on an efficient control loop; the work is applicable to any tunable matching network.

A number of automatic antenna tuners can be found in literature. These can be categorized by either the way adaptation is done or by the way the impedances (or the impedance-mismatch) is measured. Roughly a number of classes can be identified.

- Adaptation methods based on analytic computation that assume that the impedance can be measured exactly [9]. Usually this requires impedance measurement at RF frequencies and a computationally expensive control loop.
- Adaptation methods that use the sign of the phase difference to tune a reactive component to zero [4], [10], [11].
- These methods usually employ RF phase detectors [4], [12] or Foster–Seeley discriminators [10], [11] and usually are zero-crossing methods: the optimum then is a zero-phase difference signal.
- Adaptation methods that minimize a pre-specified figure of merit e.g., reflected power or VSWR. These are mainly gradient descent like methods [13]–[15] or trial-and-error methods [16], [17].

The disadvantage of exact computation of the state of the adaptive matching network is that the antenna impedance must be measured exactly, which usually involves voltage and current measurements at RF frequencies. This in turn puts high demands on detection circuits and hence results in relatively high power consumption for the detection circuitry.

Gradient decent methods are not pursued because of the need for the derivation of some sensitivity measure of matching network components. Adaptation based on direct information usually is much more efficient in terms of control-loop complexity. The trial-and-error methods are not implemented because any knowledge based method can outperform them in antenna tuner systems.

To avoid computational expensive control mechanisms, in this work we use a zero-crossing adaptation algorithm, while for power efficiency reasons and cost reasons we use quasi-dc measurable properties of RF signals (signal amplitudes) to measure the RF antenna impedance. The impedance measuring electronics and the (strongly related) control algorithm are introduced below.

3.3.1 Concept

The concept of the automatic antenna tuning system is shown in Fig. 3.2. Tuning is done in two steps: first the imaginary part of the load impedance is tuned to (almost) zero using a series (or shunt) reactance, then the resulting real part is transformed to the target real value (i.e., $50\ \Omega$) with a tunable “transformer.” This two-step sequential tuning fits well to the low-power impedance measurement method that is introduced later in this chapter.

The implementation of the impedance sensor block and of the control network blocks is discussed in details in the remaining part of this section.

Fig. 3.2. Basic concept of the automatic antenna tuning system: tune out the imaginary part and tune the real part separately.

3.3.2 Antenna Impedance Sensor: General Setup

Measuring the antenna impedance in some way is essential for any sensible antenna tuner system. In our system we aim at measuring the antenna impedance based on a quasi-dc property of the RF signals—the signal amplitude—because of power efficiency reasons and because of cost. To measure impedance, fundamentally at least three properties must be measured because voltage magnitude, current magnitude and phase must be known in some way. In our system, three signal magnitudes are measured in such a way that the antenna impedance can be derived. The schematic diagram of the total impedance sensor is shown on the left-hand side in Fig. 3.3.

voltage technologies that typically are required to drive the MEMS in switchable matching networks [4].

The sensor system has two outputs: one signal $V_{\Delta X}$ indicating that the imaginary impedance (at some point in the sensor system, see below) is zero, and provides conditional information about the real part of the tuned antenna impedance.

3.3.3 Impedance Sensor: Reactive Part Measurement

Measuring the imaginary part of the antenna impedance is basically done by the two series reactances with values jX_{ext} and the outer (logarithmic) peak detectors in Fig. 3.3. Assuming negligible loading of the middle node, the currents in both reactances are identical. The magnitude of the voltages at nodes V_1 and V_2 are then proportional to the impedance magnitudes at node V_1 and V_2 . Fig. 3.5 shows the impedance magnitudes at nodes V_1 and V_2 (indicated by the radius of the dotted and the solid circles, respectively) for three situations. In Fig. 3.5(b), the impedance magnitudes at V_1 and V_2 are identical; the impedance at node V_3 then is real. Fig. 3.5(a) shows the situation that the impedance at node V_3 is capacitive for which the voltage magnitude of V_2 is larger than that at node V_1 . Fig. 3.5(c) shows the situation where the impedance at V_3 is inductive, resulting in a larger voltage magnitude at node V_1 than that at node V_2 .

Fig. 3.5. Phasor diagrams of impedances at node V_1 , V_2 and V_3 . (a) Capacitive reactance at V_3 . (b) Zero reactance at V_3 . (c) Inductive reactance at V_3 .

It follows that with ideal log-magnitude detectors

$$V_{\Delta X} = \frac{1}{2} \ln \left(1 + 4 \frac{\text{Im}(ZT_{\text{load}})X_{\text{ext}} + X_{\text{ext}}^2}{ZT_{\text{load}}^2} \right) \quad (3.2)$$

This signal can be used to measure the reactive part of the (tuned) load impedance ZT_{load} . The signal is zero if the reactive part of ZT_{load} equals $-X_{\text{ext}}$, corresponding to a zero reactance at node V_3 . Note that a different value corresponds to a zero load reactance at some other (well defined) point: a zero reactance at the RF_{in} node corresponds to

$$V_{\Delta X} = \frac{1}{2} \ln \left(1 + 4 \frac{X_{\text{ext}}^2}{ZT_{\text{load}}^2} \right)$$

3.3.4 Impedance Sensor: Real Part Measurement

Tuning of the real part of the (tuned) load impedance is done using signal $V_{\Delta R}$. Defining the impedance at node V_3 as $Z_{V3} = ZT_{\text{load}} + X_{\text{ext}}$

$$V_{\Delta R} = \frac{1}{2} \ln \left(\frac{\text{Re}(Z_{V3})^2 + (\text{Im}(Z_{V3}) + X_{\text{ext}})^2}{|Z_{V3}|^2} \right) \quad (3.3)$$

In the special case that the impedance at node V_3 is real, this relation simplifies to a relation that directly provides information about the (tuned) antenna

$$V_{\Delta R}(R_{V3}) = \frac{1}{2} \ln \left(1 + \left(\frac{X_{\text{ext}}}{R_{V3}} \right)^2 \right)$$

when $V_{\Delta X} = 0$.

In antenna tuning systems must be tuned to a target value to get a load equal to $V_{\text{target}2} = V_{\Delta R}(50\Omega)$ to get a load equal to $R_{\text{target}} = 50 \Omega$. In Fig. 3.3 this is realized by adding offset to $V_{\Delta R}$.

3.3.5 Tunable Matching Network

The implementation of the tunable matching network system is given in Fig. 3.3; that of the “log peak detectors” is shown in Fig. 3.4. The tunable “transformer” is implemented using a T-network made up by two inductors (with value) and one capacitor (of value). This type of network can be seen as a lumped model of a transmission line; for exact tuning

$$X_{\text{series}} = -\text{Im}[Z_{\text{load}}] \quad (3.4)$$

$$X_0 = \sqrt{R_{\text{in}} \text{Re}[Z_{\text{load}}]}$$

where Z_{load} is the load impedance and R_{in} is the real impedance required by the PA.

3.3.6 Control Network

The signals $V_{\Delta X}$ and $V_{\Delta R}$ give a measure for the imaginary part and of (conditionally) the real part of the (tuned) antenna, respectively. We used these signals in a very simple integrating control loop, see Fig. 3.3. Due to the relatively slowly changing environment of handheld devices (related to the signal frequencies) the performance of this simple control loop is sufficient to keep the antenna properly tuned during typical real-life handling conditions.

3.4 Experimental results

Two demonstrators were developed for matching any load impedance with $\text{VSWR} \leq 10$, at 900 MHz. One has sequential and discretized tuning. The discretization implies that only a limited set of (switchable) matching network states were implemented. With the sequential tuning, first the reactive part of the load is tuned and

then the real part is tuned to 50Ω . In this demonstrator, tuning of the real part is enabled if the reactive part is within a few times the minimum reactive tuning step size. With sequential tuning both the reactive and real tuning is monotonous towards the target impedance.

The second demonstrator implements simultaneous and continuous adaptation of the reactive and real part. Simultaneous adaptation simplifies the implementation as no threshold is required for tuning of the real part. Simultaneous adaptation can be shown to have guaranteed convergence for bounded matching network tuning ranges; the adaptation of $\text{Re}(Z_{V3})$ however can show non-monotonous behavior. The continuous adaptation corresponds to a matching network that either can switch between many states or that implements true continuous matching “states.”

For both examples, the adaptable matching network and the antenna impedance were emulated by a Maury tuner controlled by a pc. For both designs the real and imaginary part of Z_{antenna} are in the ranges

$$R_{\text{load}} \in [5 \Omega, 500 \Omega]$$

and

$$X_{\text{load}} \in [-250 \Omega, 250 \Omega]$$

resulting in (ideally required) tuning ranges

$$X_{\text{series}} \in [-250 \Omega, 250 \Omega]$$

and

$$X_0 \in [15.8 \Omega, 158 \Omega]$$

The discrete (SMD) measurement reactance $X_{\text{ext}} = 19 \Omega$.

3.4.1 Discretized and Sequential Tuning

For the discretized and sequential tuning demonstrator, the target impedance was defined as an impedance region for which the matching was acceptable: a -1.5-dB loss contour [7] was used, see Fig. 3.3. The size of this target area, together with the initial antenna impedance region, determines the minimum number of switchable matching network states.

For the discrete implementation in order to fulfill the requirement of -1.5-dB power contour, it can be derived that 120 matching network states are sufficient; we implemented 500 possible states for the switchable matching network. In the measurement setup a Maury tuner was used to emulate the total impedance of the tunable matching network and antenna. The Maury is driven by a pc that implements the control algorithm, matching network state quantization and actual control of the Maury. The detector circuitry in Fig. 3.4 was realized on PCB using SMD components.

Many measurements were performed for both version of the control loop at various power levels (from 4 to 24 dBm). All measurements started with an untuned system, $X_{\text{series}} = 0 \Omega$ and $X_0 = 50 \Omega$. Fig. 3.6 shows the impedance trajectories as seen by the PA (including sensor circuitry, tunable transformer, series reactance, cables, fixtures, ...) for four extreme antenna impedances. The end of each adaptation is indicated by the arrows; the eventual “state” hops across the three states surrounding this end point due to the discretization.

Fig. 3.6. Measurement results of the PA-load impedance with discretized sequential tuning a) $Z_{\text{antenna}} = 10 - j 60 \Omega$. b) $Z_{\text{antenna}} = 15 - j 18 \Omega$. c) $Z_{\text{antenna}} = 100 + j 100 \Omega$. d) $Z_{\text{antenna}} = 5 + j 18 \Omega$; arrows indicate end region center.

Note that due to the sequential adaptation of the matching network, the impedance curves tend to sharply bend once the (horizontal) real axis is approached. This corresponds to setting in of the real part's tuning as soon as a sufficiently small imaginary residue is reached.

3.4.2 Continuous and Simultaneous Tuning

Whereas for discretized tuning a target region must be defined (to estimate the required number of matching network states) for continuous tuning the target is set to 50Ω : the center of the Smith chart. For simultaneous adaptation, the imaginary part monotonously goes to zero, while the convergence of the real part to 50Ω may show some over-or undershoot.

Fig. 3.7 shows the impedance trajectories as seen by the PA for the same initial settings and antenna impedances as used for Fig. 3.6. The end-point impedance (indicated by the arrow) is close to 50Ω .

Fig. 3.7. Measurement results of the PA-load impedance for continuous simultaneous adaptation a) $Z_{\text{antenna}} = 10 - j 60 \Omega$. b) $Z_{\text{antenna}} = 15 - j 18 \Omega$. c) $Z_{\text{antenna}} = 100 + j 100 \Omega$. d) $Z_{\text{antenna}} = 5 + j 18 \Omega$; arrows indicate end point.

3.4.3 Discussion

The sensor circuitry and the adaptive matching network so far were regarded as separate blocks. Noting that both the adaptive matching network and the impedance sensor block consists of reactances, it is straightforward to (partially) merge these.

The accuracy of the discretized tuning system is typically limited by the limited number of matching network states. In our continuous tuning system the accuracy is limited mainly by the following.

- The residual reactive part seen by the PA. In the proposed system we tune the impedance at node V_3 to zero; by adding an offset to $V_{\Delta X}$ the zero-reactive impedance can however be shifted to the PA output.
- Loading effects of the log peak detector circuitry; in our discrete implementation, the load of the peak detectors is about 0.5 pF.

Due to the low-frequency detection principle, matching requirements for the detector electronics typically are relatively easy to meet.

3.5 Conclusion

A novel automatic antenna tuning system for handheld applications is presented. This tuning system uses only magnitude information of the RF signal (at three positions) to measure the antenna impedance, while the electronics and control loop can operate at relatively low frequencies. These low-frequency properties result in low power consumption and low cost implementation possibilities.

Measurements on a demonstrator implementation at 900 MHz and at various power levels verify correct operation, and show proper convergence of the impedance as seen by the PA to values near the optimum value, starting on any impedance that result in VSWR up to 10. The end-point inaccuracy is ultimately due to the residual series reactance (compensatable) and the load impedance of the impedance sensor circuitry. Driving switchable matching networks, the inaccuracy is typically in the quantization error due to the limited number of matching states.

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Chapter 4

On the Minimum Number of States for Switchable Matching Networks

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Abstract—The impedance of an antenna changes heavily with changing EM environments, while RF power amplifiers (PAs) are optimized for driving a well-defined load impedance. As a solution, switchable matching networks are used in automatic antenna tuners to match the antenna impedance to (about) the desired PA load impedance. This chapter presents a theoretical treaty of the minimum number of required states for switchable matching networks to achieve sufficient matching from a certain load VSWR to a sufficiently low input VSWR. First for an arbitrary passive lossless switchable matching network, the mathematical minimum required number of states as a function of the required input VSWR and of the required load VSWR is derived. Several variants are analyzed and benchmarked: single-stage one-ring configuration, single-stage two-ring configuration, two-stage one-ring configuration and three-stage one-ring configuration showing that single-ring configurations are optimum. An extension towards the required number of states for lossy matching networks is also provided.

Index Terms—Automatic antenna tuners, impedance matching, switchable matching networks, tunable matching networks.

4.1 Introduction

Antenna impedances depend heavily on their EM environment [1], [2] and hence may change significantly during real life operation. The antenna impedance Z is usually expressed in terms of reflection coefficient $\Gamma_Z = (Z - Z_0)/(Z + Z_0)$, where Z_0 is the characteristic impedance, or in terms of voltage standing wave ratio $VSWR = (1 + |\Gamma_Z|)/(1 - |\Gamma_Z|)$. For a typical antenna the VSWR can be up to 10:1 [1]–[5].

This changing antenna impedance causes serious design and performance challenges for the RF power amplifier (PA) driving the antenna due to the associated VSWR. Assuming a certain required lower bound on maximum transmit power, a varying load impedance for the PA (e.g., antenna impedance through a fixed matching network) requires robustness to both maximum voltage and maximum current levels well above those required when driving a nominal load impedance at the same power level [3]–[5]. Consequently, the PA then must be designed to operate properly for a wide load impedance range and hence the PA is necessarily overly robust at nominal conditions and consequently has significantly lower efficiency at nominal conditions. On top of this, matching losses due to non-nominal load impedances decrease the efficiency and radiated power significantly.

Automatic antenna tuners may be used as solution for this. In general, an automatic antenna tuner consists of impedance sensors, a control network and a tunable matching network [6]–[24]. The sensors are used to detect the impedance mismatch or impedance value; the control network controls the state or setting of the tunable matching network that performs the actual impedance tuning from its load impedance Z_{load} into its input impedance Z_{in} .

The tunable matching network can be implemented with continuous tunable components or with switchable components. In case of continuous tunable components it is theoretically possible to achieve a perfect tuning at the input for a certain (limited) load impedance region. In case of ideal switchable components perfect impedance matching can be achieved for only a finite number of load impedances. This chapter presents a mathematical estimation of the minimum required number of states for narrow band switchable matching networks to match a certain load impedance range to within a smaller input impedance range. This problem was implicitly addressed in [22]–[24] for specific implementations, without giving a general approach or mathematical derivation.

Section 4.2 presents a mathematical estimation of the minimum required number of states for a single stage switchable matching network assuming a resistive source driving a lossless switchable matching network terminated by a load impedance (e.g., the antenna). This derivation uses the mismatch efficiency [25] in order to keep the mathematical formalism as general as possible without referring to specific implementations: the only assumption is that the switchable matching network is passive

and lossless; it can be reciprocal or non-reciprocal. Consequently, the results are valid for any (yet lossless) switchable single stage matching network. The resulting minimum required number of states can be used to derive a lower bound on hardware costs to implement switchable matching networks, and can be used to estimate the efficiency of specific hardware implementations in terms of overhead in states.

Sections 4.3 to 4.5 extend the results of Section 4.2 in a number of ways. Section 4.3 discusses the merits and limitations of multi stage matching networks: it shows that partitioning a matching network in multiple simpler matching networks can be very efficient in terms of minimum total required number of states to perform a certain impedance matching. Section 4.4 briefly discusses the impact of using an arbitrary complex source impedance instead of a purely real source resistance, while an extension towards (general) lossy matching networks is discussed in Section 4.5. Section 4.6 summarizes the findings.

4.2 Near minimum number of states for lossless switchable matching networks

This section presents a mathematical derivation of the minimum required number of states for lossless switchable matching networks at a single frequency. The matching network matches any load impedance Z_{load} in a certain impedance range to impedances Z_{in} that are sufficiently close to the complex conjugate of the source impedance. In this section, a real source impedance R_{source} is assumed.

In this chapter, complex impedances are mainly specified in terms of reflection coefficient $\Gamma_z = (Z - Z_0)/(Z + Z_0)$. Some basic properties of this bilinear transformation between Z and Γ form the mathematical basis for the work in this chapter. A first property is that the entire complex Z -plane is mapped to the entire complex Γ -plane with the point $\Gamma = 1$ removed. Secondly, the bilinear transformation is holomorphic: it maps circles in the Z -plane into circles in the Γ -plane; non-overlapping circles in one plane are non-overlapping in the other plane [25,ch. 10].

The “number of states” denotes the actual number of different settings that the switchable matching network can have. Assuming for example a matching network that incorporates in total N switched reactances, the number of states for that network equals 2^N of which a number may be coinciding, thereby effectively reducing the actual number of (unique) states. Assuming e.g., a slide-screw tuner that can be set in discrete settings, the number of states typically equals. In this chapter we, however, do not assume any specific implementation for the matching network.

In the current chapter, the switchable matching network aims to match a load

impedance Z_{load} into a Z_{in} that is sufficiently close to target impedance R_{source} , using a minimum number of required states. This “sufficiently close” means that the associated magnitude of the reflection coefficient $\Gamma_{Z_{\text{in}}}$ is lower than a predefined value $|\Gamma_{Z_{\text{in}}}|_{\text{max}}$.

4.2.1 Single State Impedance Matching

The mismatch efficiency M equals the ratio of actual power in the load P_{load} and the maximum achievable power into the load P_{max} [25]. For a passive lossless matching network, see Fig. 4.1, at a single frequency this is a function of the complex source and load impedances seen at a cross section. At cross section A in Fig. 1 this translates into

$$M = \frac{P_{\text{load}}}{P_{\text{max}}} = \frac{(1 - |\Gamma_{Z_{\text{source}}}|^2)(1 - |\Gamma_{Z_{\text{in}}}|^2)}{|1 - \Gamma_{Z_{\text{source}}} \Gamma_{Z_{\text{in}}}|^2} \quad (4.1)$$

showing that maximum power transfer is achieved for $\Gamma_{Z_{\text{source}}} = (\Gamma_{Z_{\text{in}}})^*$.

The derivations in this chapter are valid both for reciprocal and for non-reciprocal networks. Note that for lossless matching networks the M is the same at cross section A and B in Fig. 4.1. For simplicity reasons we first assume a real source impedance.

Fig. 4.1. Circuit topology and naming conventions in the derivation.

At a single frequency the impedance matching network can be designed to match a load impedance $Z_{\text{load,match}}$, with a reflection coefficient $\Gamma_{Z_{\text{load,match}}}$, to the center of the Smith chart. For this perfect match, $M = 1$ and $\Gamma_{Z_{\text{out}}} = (\Gamma_{Z_{\text{load,match}}})^*$, as depicted in Fig. 4.2.

For this matching network, a certain mismatch efficiency $M < 1$ corresponds to all $\Gamma_{Z_{\text{in}}}$ lying on a circle centered on the center of the Smith chart having a radius $|\Gamma_{Z_{\text{in}}}|_{\text{max}} = \sqrt{(1 - M)}$. Due to the bilinear properties of the matching networks, this circle for $\Gamma_{Z_{\text{in}}}$ corresponds to a circle of $\Gamma_{Z_{\text{load}}}$ values on the Smith chart. This latter circle encloses $\Gamma_{Z_{\text{load,match}}}$, but in general is not centered on due to the compressive character of the Smith

A description of this circle can be obtained by solving (4.1), for all $|\Gamma_{\text{Zload}}|$ values at cross section B (see Fig. 4.1) for a given M . Under assumption that the resulting -circle does not encircle the center of the Smith chart, it can be proven that the center of the circle lies on the line through the center of the Smith chart and $\Gamma_{\text{Zload,match}}$. Using the previously described matching network that matches $\Gamma_{\text{Zload,match}}$ onto to the center of the Smith chart, there are now two impedances Γ_{Zload1} and Γ_{Zload2} that lie on a line in the Smith chart crossing both the center point and crossing $\Gamma_{\text{Zload,match}}$ that are matched onto impedances corresponding to some points on a circle around the origin with radius $|\Gamma_{\text{Zin}}|$ that corresponds to the specific M , see Fig. 4.2 for a graphical representation. Solving (4.1) yields:

$$|\Gamma_{load1,2}| = \frac{\beta|\Gamma_{Zout}| \pm \sqrt{(\beta|\Gamma_{Zout}|)^2 - (\beta-1)(1+\beta|\Gamma_{Zout}|^2)}}{1+\beta|\Gamma_{Zout}|^2} \quad (4.2)$$

with

$$\beta = \frac{M}{1 - |\Gamma_{Zout}|^2}$$

and where $\Gamma_{Zout} = (\Gamma_{Zload,match})^*$.

Fig. 4.2. Mapping of one load impedance circle onto an input impedance circle for a certain matching efficiency M .

The center c_0 and the radius r_0 of the circle that is mapped by the matching network onto a circle around the origin with radius $|\Gamma_{Zin}|_{\max}$ follow from (4.2):

$$c_0 = \frac{|\Gamma_{Zload1}| + |\Gamma_{Zload2}|}{2} = \frac{M \Gamma_{Zout}^*}{|\Gamma_{Zout}|^2 (M-1) + 1} \quad (4.3)$$

$$c_0 = \frac{|\Gamma_{Zload1}| + |\Gamma_{Zload2}|}{2} = \frac{(1 - |\Gamma_{Zout}|^2) \sqrt{1-M}}{|\Gamma_{Zout}|^2 (M-1) + 1} \quad (4.4)$$

4.2.2 Single Ring Impedance Matching

Due to the bilinear nature of impedance transformations of switchable matching networks, circles described by (4.3) and (4.4) can be mapped onto circles centered at the origin of the Smith chart with radius $|\Gamma_{Zin}|_{\max}$. Using many suitably positioned circles a whole band can be constructed, directly outside the circles with radius that can be mapped onto the center circle [22], [23], see Fig. 4.3.

Fig. 4.3. Constructing a matchable area with radius $|\Gamma_{Zload}|_{\max}$ matchable onto a circle around the origin with radius $|\Gamma_{Zin}|$, using multiple (here 10 + 1) number of states for switchable matching networks.

Fig. 4.3. illustrates such a matching network configuration with in total 11 states, having one center circle with sufficiently low $|\Gamma_{Zin}|$, and 10 circles around it that can individually be mapped onto the center circle, and in total forming the largest region matchable region described by $|\Gamma_{Zload}| \leq |\Gamma_{Zload}|_{\max}$. Impedance mapping of two distinct impedances in one of the outer circles onto the center circle is also depicted in Fig. 4.3.

It appears that analytically deriving the required number of matching network states as a function of $|\Gamma_{Zin}|_{\max}$ and $|\Gamma_{Zload}|_{\max}$ is (at least) hard. However, it is possible to derive the region that can be matched to an impedance inside a circle with radius $|\Gamma_{Zin}|_{\max}$ as a function of the number of matching network states.

Using only one ring consisting of circles (each circle corresponds to a single matching network state) around the center (see Fig. 4.3), the total circle that can be matched follows from using (4.3) and (4.4) in plain geometrical constructions to build a larger circle. This yields a 6th order equation with four trivial or not-relevant solutions and one solution leading to a matched overall circle with radius (see Appendix A):

$$|\Gamma_{Zload}|_{\max} = |\Gamma_{Zin}|_{\max} + 2r_1 \cos \left(\frac{\pi}{N_1} + \arcsin \left(\frac{|\Gamma_{Zin}|_{\max} \sin \left(\frac{\pi}{N_1} \right)}{r_1} \right) \right) \quad (4.5)$$

where

$$r_1 = \frac{(1 - h_1^2) |\Gamma_{Zin}|_{\max}}{1 - h_1^2 |\Gamma_{Zin}|_{\max}^2}, \quad h_1 = \frac{2 |\Gamma_{Zin}|_{\max} \cos \left(\frac{\pi}{N} \right)}{1 + |\Gamma_{Zin}|_{\max}^2} \quad (4.6)$$

Boundary condition for this relation include having a real value for $|\Gamma_{Zload}|_{\max}$, $|\Gamma_{Zload}|_{\max} > |\Gamma_{Zin}|_{\max}$ and laying inside the Smith chart of the all circles (i.e., the regions that can be matched onto the center circle). If one or more of these boundary conditions are not satisfied, the parameter N_1 is just too low to find a valid solution. Fig. 4.4 shows the radius of the matchable area on the Smith chart $|\Gamma_{Zload}|_{\max}$ as a function of the number of matching network states $N_{\text{tot}} = (N_1 + 1)$, for a number of $|\Gamma_{Zin}|_{\max}$ -values. The “+ 1” is there because the circle centered at the center of the Smith chart must be included as state.

For low N_{tot} , the matched area does not form a circle larger than the center circle, hence the lower clipping of $|\Gamma_{Zload}|_{\max}$ on in Fig. 4.4. With increasing N_{tot} the radius of the matchable area increases and reaches (for $N_{\text{tot}} \rightarrow \infty$) the maximum radius asymptotically. Note that for clarity reasons, the graph in Fig. 4 shows continuous curves while N_{tot} can only assume integer values. It follows from (4.5) and (4.6) that the maximum radius is:

$$|\Gamma_{N \rightarrow \infty}| = |\Gamma_{Zin}|_{\max} \left(1 + \frac{2(1 - |\Gamma_{Zin}|_{\max}^2)^2}{(1 + |\Gamma_{Zin}|_{\max}^2)^2 - 4|\Gamma_{Zin}|_{\max}^4} \right) \quad (4.7)$$

Fig. 4.4. Radius of the matchable region $|\Gamma_{Zload}|_{\max}$ as a function of the number of matching network states N_{tot} , for various $|\Gamma_{Zin}|_{\max}$.

4.2.3 Two Ring Impedance Matching

The analysis in the previous section can be extended in a recursive way to get, e.g., the circular region on the Smith chart that is composed of a center circle and two rings composed of circles that all can be matched to the center ring. For the two-ring case it can be derived that:

$$|\Gamma_{Zload}|_{\max 2} = \Gamma_{Zload \max 1}(|\Gamma_{Zin}|_{\max}, N_1) + 2r_2(|\Gamma_{Zload, matched 2}|, N_2) \cos\left(\frac{\pi}{N_2} + \varphi(|\Gamma_{Zload, matched 2}|)\right) \quad (4.8)$$

where the first term on the right hand side in (4.8) equals (4.5) and with:

$$\varphi(\gamma) = \arcsin\left(\frac{\gamma}{r_2} \sin\left(\frac{\pi}{N_2}\right)\right), \quad (4.9)$$

$$r_2(x, N) = \frac{(1 - h_2(x, N)^2)x}{1 - h_2(x, N)^2 x^2}, \quad h_2(x, N) = \frac{2x \cos\left(\frac{\pi}{N}\right)}{1 + x^2},$$

$$|\Gamma_{Zload, matched 2}| = \frac{-b + \sqrt{b^2 - 4d}}{2},$$

$$b = \frac{2(1 - |\Gamma_{Zin}|_{max}^2)|\Gamma_{Zload}|_{max1} \cos\left(\frac{\pi}{N_2}\right)}{1 - |\Gamma_{Zload}|_{max1}^2 |\Gamma_{Zin}|_{max}^2},$$

$$d = \frac{|\Gamma_{Zload}|_{max1}^2 - |\Gamma_{Zin}|_{max}^2}{1 - |\Gamma_{Zload}|_{max}^2 |\Gamma_{Zin}|_{max}^2}$$

where again is the required magnitude of the reflection coefficient at the input of the switchable matching network, N_1 and N_2 are the number of circles of the first and second ring respectively. Also in case of two rings some trivial boundary conditions must be satisfied to get a valid solution: e.g., $|\Gamma_{Zload}|_{max2}$ must be real valued and $|\Gamma_{Zload}|_{max2} > |\Gamma_{Zload}|_{max1}$. Fig. 4.5 shows an example of a two-ring configuration for a matching network. In this configuration, any of the rings (here, $N_1 = 10$, $N_2 = 24$) can exactly be matched to the center ring, thereby being able to match any impedance to $|\Gamma_{Zload}| < |\Gamma_{Zload}|_{max2}$ within $|\Gamma_{Zin}| < |\Gamma_{Zin}|_{max}$.

Fig. 4.5. Constructing a area with radius $|\Gamma_{Zload}|_{max}$ matchable onto a circle around the origin with radius $|\Gamma_{Zin}|_{max}$, using multiple (here 35) number of states in a two-ring configuration.

It appears that it is at least hard to derive a closed mathematical expression for the maximum circular region on the Smith chart that can be mapped onto the center circle for the two-ring case. This is basically because an optimum in (4.8), (4.9) for a constant number of total states $N_{\text{tot}} = (N_2 + N_1 + 1)$ cannot readily be derived. These optima can, however, quite easily be derived numerically as only integer N_1 and N_2 occur.

Fig. 4.6 shows a contour plot the radius of the regions that can be matched as a function of N_1 and N_2 , here for $|\Gamma_{\text{Zin}}| = 1/3$. For low N_{tot} this behavior is identical to that shown in Fig. 4.4 for the single-ring configuration of the matching network: the maximum $|\Gamma_{\text{Zload}}|_{\text{max}}$ occurs for $N_2 = 0$. For higher N_{tot} a second ring gets to become beneficial and then the maximum $|\Gamma_{\text{Zload}}|_{\text{max}}$ is achieved for $N_2 > 0$.

Fig. 4.6. Contour plot of $|\Gamma_{\text{Zload}}|_{\text{max}}$ as a function of N_1 and N_2 , here for $|\Gamma_{\text{Zin}}| = 1/3$. On the plot, constant- N_{tot} lines are plotted for $N_{\text{tot}} = 10, 15, \dots 35$. On these lines, the maximum matchable $|\Gamma_{\text{Zload}}|_{\text{max}}$ is denoted with a dot.

A straight forward maximum search for constant $N_{\text{tot}} = (N_2 + N_1 + 1)$ directly yields the maximum magnitude of the load reflection coefficient $|\Gamma_{\text{Zload}}|_{\text{max}}$ as a function of both the total number of states N_{tot} and of $|\Gamma_{\text{Zin}}|_{\text{max}}$. Fig.4.7 shows both $|\Gamma_{\text{Zload}}|_{\text{max}}$ as a function of N_{tot} for (here) $|\Gamma_{\text{Zin}}|_{\text{max}} = 1/3$ and shows the distribution of N_{tot} across the first and second ring. In this example, for $N_{\text{tot}} < 6$ no continuous band can be formed, while for $6 < N_{\text{tot}} < 21$ a single configuration ring is optimum. For larger N_{tot} the two-ring solution is most efficient and can achieve matching of a larger $|\Gamma_{\text{Zload}}|_{\text{max}}$ region compared to the single-ring case.

Fig. 4.7. Matchable region impedances on the Smith chart as a function of the total number of states N_{tot} , for a two-ring network, for $|\Gamma_{Z_{\text{in}}}| = 1/3$ (VSWR = 2:1). The plot also shows the optimum N_1 and N_2 as a function of N_{tot} .

4.3 Multistage lossless switchable matching networks

The previous section showed an analysis of single stage matching networks, that used matching states in one or two rings to match any impedance $|\Gamma_{Z_{\text{load}}}| < |\Gamma_{Z_{\text{load}}}|_{\text{max}}$ to within $|\Gamma_{Z_{\text{in}}}|_{\text{max}}$. It was also shown that a two-ring configuration can match a larger $|\Gamma_{Z_{\text{load}}}|_{\text{max}}$ than a single-ring matching network configuration. This section shows that multi-stage single-ring configurations, see Fig. 4.8 for a two-stage matching network, outperform single-stage multi-ring matching networks in terms of (e.g.) minimum required total number of states.

Fig. 4.8. Block diagram of a two-stage switchable matching network.

In multi-stage networks, the switchable matching network connected to the load achieves coarse tuning from $\Gamma_{Z_{\text{load}}}$ to Γ_{mid} while the left most network achieves fine tuning from Γ_{mid} to $\Gamma_{Z_{\text{in}}}$; this principle is shown in Fig. 4.9 for matching for any impedance within $|\Gamma_{Z_{\text{load}}}| \leq |\Gamma_{Z_{\text{load}}}|_{\text{max}}$ via Γ_{mid} to within $|\Gamma_{Z_{\text{in}}}|_{\text{max}}$.

Fig. 4.9. Matching in a two-stage switchable matching network: the second stage (right hand side Smith chart) matches an impedance to within $|\Gamma_{mid}|_{max}$; the first stage (left hand side Smith chart) matches this to within $|\Gamma_{Zin}|_{max}$.

It follows directly from bilinear properties, and from the analyses in Section II that this approach to match a relatively large $|\Gamma_{Zload}| \leq |\Gamma_{Zload}|_{max}$ to within a circle $|\Gamma_{Zin}|_{max}$ is much more efficient than using multi-ring approaches because mainly the coarse matching stage uses relatively large circles that require a relatively low minimum number of (large) circles to cover a large part of the Smith chart. For illustration purposes, the $|\Gamma_{Zload}|_{max}$ and $|\Gamma_{Zin}|_{max}$ for Fig. 4.9 are identical to the ones used for Fig. 4.5, requiring 35 states in total for the single stage network and 22 in total for the multi-stage solution. This difference in minimum number of required matching network states increases rapidly for larger ratios between $|\Gamma_{Zload}|_{max}$ and $|\Gamma_{Zin}|_{max}$.

4.3.1 Estimation

In case of required good input matching (i.e., low $|\Gamma_{Zin}|_{max}$), a good approximation of the optimum maximum magnitude of the reflection coefficient at the intermediate cross section, $|\Gamma_{mid}|_{max}$, can be derived. Assuming a small $|\Gamma_{Zin}|_{max}$, (4.5) can be simplified to $|\Gamma_{Zload}|_{max} \approx |\Gamma_{Zin}|_{max} (1 + 2 \cos(2\pi/N))$, yielding for a two-stage matching network having one ring per stage

$$|\Gamma_{Zload}|_{max} \cong |\Gamma_{Zin}|_{max} \left(1 + 2 \cos\left(\frac{2\pi}{N_I}\right) \right) \cdot \left(1 + 2 \cos\left(\frac{2\pi}{N_{tot} - N_I - 2}\right) \right)$$

Due to symmetry, equal distribution of the matching network states over the two

stages is optimum, yielding

$$\begin{aligned} |\Gamma_{mid}|_{max} &\equiv |\Gamma_{Zin}|_{max} \left(1 + 2 \cos \left(\frac{2\pi}{N/2 - 1} \right) \right) \\ &= \sqrt{|\Gamma_{Zin}|_{max} |\Gamma_{Zload}|_{max}} \end{aligned}$$

For $|\Gamma_{Zin}|_{max}$ -values that are not very small an easy-to-read expression for $|\Gamma_{mid}|_{max}$ cannot be derived. However, combining the result of two one-stage one-ring results to get multi-stage results is quite easy, with just one numerical (integer) search required to get the optimum $|\Gamma_{mid}|_{max}$.

4.3.2 Benchmarking Lossless Multi-Stage Matching Networks

Fig. 4.10 shows 4 curves, all showing $|\Gamma_{Zload}|_{max}$ as a function of the total number of states in the matching network N_{tot} , for $|\Gamma_{Zin}|_{max} = 1/3$. The curves are for single-stage matching networks with either one of two rings of circles optimally positioned around the center ring and are for cascaded single-stage single-ring configurations.

Fig. 4.10. Maximum magnitude of load reflection coefficient ($|\Gamma_{Zload}|_{max}$) as a function of the total number of states N_{tot} for one-stage one-ring, one-stage two-ring, two-stage one-ring, and three-stage-one-ring networks. For the plot $|\Gamma_{Zin}| = 1/3$ (VSWR = 2:1).

It follows that for maximum load reflection coefficients $|\Gamma_{Zload}|_{max}$ close to the required magnitude of the input reflection coefficient $|\Gamma_{Zin}|_{max}$, it is best to use the one-

stage one-ring configuration. As $|\Gamma_{Z_{\text{load}}}|_{\text{max}}$ (or the total number of states for the matching network) increases, after some point it is more efficient to use any of the other configurations. However, the curves show that a single-stage multi-ring configuration is not efficient: either it degenerates to a single-stage single-ring configuration for low or it is better to use a two-stage single-ring configuration.

Moving to more stages (i.e., three-stage switchable matching networks) appears to be useful only if very $|\Gamma_{Z_{\text{load}}}|_{\text{max}}$ large must be matched onto small $|\Gamma_{Z_{\text{in}}}|_{\text{max}}$. In the case of adaptive antenna matching, these cases typically represent broken antennas or electromagnetically very bad environments. In this case switching off the PA might be a better option.

4.4 Complex source impedances

For simplicity reasons, the derivations in Section 4.2 and 4.3 were done assuming a resistive source, with a straight forward extension towards generators with an arbitrary impedance Z_{source} that require a conjugate Z_{in} of the matching network (see Fig. 4.1) for maximum power transfer.

For RF power amplifiers, complex conjugate matching is typically not used but a specific load impedance is set to be able to get maximum power or maximum efficiency, determined via, e.g., loadpull. Whereas mismatch efficiency contours are circles on the Smith chart, load pull contours usually are not [26]; they generally address a non-linear problem. In this case the analysis shown in this chapter would only give a coarse upper bound estimation on the number of required states for an adaptive matching network.

However, usually in PA design the electrical length between the PA and the actual load is unknown or unspecified which results in having to accept an arbitrary phase shift in the load impedance. This means that actual PAs usually must be designed for the smallest area enclosed by arbitrarily rotated load pull contours. This smallest enclosed area equals the largest circle enclosed by the not-rotated load pull contour. For this circular area, the analyses shown in this chapter hold.

4.5 Lossy matching networks

In Sections 4.2 and 4.3, requirements on the minimum required number of states were derived for general lossless switchable impedance matching networks. In reality, switchable matching networks are lossy, although with low loss to have any significance as impedance matching network. This section presents an extension of the

work in Section 4.2 and 4.3 towards lossy matching networks. Note that also in this section no specific hardware implementation is assumed to keep the results as general as possible.

Any two-port linear network can be described as a 2×2 S or ABCD matrix; in this section first the ABCD-representation is used. As shown in Appendix B, the ABCD-matrix of a lossy matching network can be decomposed in 3 ABCD matrices and an extra phase shift:

$$[ABCD]_{\text{lossy}} = [ABCD]_{\text{attenuator1}} \cdot [ABCD]_{\text{lossless}} \cdot [ABCD]_{\text{attenuator2}} \cdot \quad (4.10)$$

where $[ABCD]_{\text{attenuatorN}}$ are purely real lossy circuits, within total (at least) 3 degrees of freedom, and $[ABCD]_{\text{lossless}}$ is a lossless (reciprocal or non-reciprocal) matching network having 4 degrees of freedom as assumed in Sections II–V in this chapter, see Fig. 4.11.

Fig. 4.11. Equivalent block diagram of a lossy switchable matching network: the cascade of two attenuators, a lossless switchable matching network, and a phase shifter.

In other to extend the results in Sections 4.2–4.5 to lossy matching networks, it is useful to switch back to the S matrix. Using the decomposition of the lossy network into a lossless network with attenuators (and a phase shifter) wrapped around it, the effect of losses can be illustrated and calculated. For the illustrations, relatively high losses are used for visibility reasons.

In Fig. 4.11 the lossless network is between planes (B) and (C). For this part of the lossy matching network, the analyses in Sections 4.2 and 4.3 hold. An example of the corresponding matchable region and circle to match to are shown in Fig. 4.12(c) and Fig. 4.12(b) respectively.

The effect of an attenuator is twofold. Firstly the attenuation scales up the radius of the matchable circle on the Smith chart when moving towards the load side of the matching network. Assuming only losses in the attenuator, so having $S_{11,n} = S_{22,n} = 0$ and having a lower-than-unity magnitude of $S_{21,n}$ for the n th attenuator, it follows that the center point does not shift and $|\Gamma_{Zin}|_{\text{max}} = |S_{21,1}|^2 |\Gamma_{Zin'}|_{\text{max}}$. Secondly, for a more general attenuator $S_{11,n} \neq S_{22,n}$, which results in a change of both the radius and the center point

of the circles according to $\Gamma_{Zin} = S_{11,1} + (S_{21,1} S_{12,1} \Gamma_{Zin'} / (1 - S_{22,1} \Gamma_{Zin'}))$.

The power efficiency of the lossy matching network is

$$P_{load} = |S_{21,1}|^2 \frac{1 - |\Gamma_{Zin'}|^2}{1 - |\Gamma_{Zin}|^2} |S_{21,2}|^2 \frac{1 - |\Gamma_{Zload}|^2}{1 - |\Gamma_{Zload'}|^2} P_{in} \quad (4.11)$$

where P_{load} is the power to the load, P_{in} is the input power and the other symbols are defined in Fig. 4.11.

Fig. 4.12. Transformation of the Smith chart moving from the input to the output: (A) input impedance region of the lossy network: in order to make the plot clear, the effect of the losses has been emphasized; (B) input impedance for the de-embedded lossless matching network part in Fig. 11; (C) matchable region at the output of the de-embedded lossless matching network part in Fig.11; (D) matchable load impedances at the output of the lossy matching network: in order to make the plot clear, the effect of the losses has been emphasized.

4.6 Conclusion

This chapter shows a mathematical derivation for the minimum number of states for passive lossless switchable matching networks to match any load impedance for which $|\Gamma_{Zload}| < |\Gamma_{Zload}|_{max}$ to an impedance for which $|\Gamma_{Zin}| < |\Gamma_{Zin}|_{max} < |\Gamma_{Zload}|_{max}$. In this

chapter, the switchable matching network can be reciprocal or non-reciprocal.

For relatively simple single-stage matching networks, a closed form expression is derived for the largest matchable region $|\Gamma_{\text{load}}| < |\Gamma_{\text{load}}|_{\text{max}}$ as a function of maximum input reflection $|\Gamma_{\text{zin}}|_{\text{max}}$ and of the number of states for the matching network.

Several other configurations were studied: more complex one stage configuration as well as cascaded simple matching networks. It appears that at low ratios between $|\Gamma_{\text{zin}}|_{\text{max}}$ and $|\Gamma_{\text{load}}|_{\text{max}}$ the one-stage one-ring configuration is the best choice. As the mismatch between required input impedance and required load impedance increases, a cascaded configuration consisting of simple matching networks becomes the best option. The analysis has been extended to lossy passive switchable matching networks; the main difference with the lossless situation is a shift and scaling of the matchable region.

Appendix A

This appendix presents the mathematical derivation of matchable region in case of a single-ring switchable matching networks, described by $|\Gamma_{\text{load}}|_{\text{max}}$ in (4.5) and (4.6). The region that can be matched using a single-ring containing evenly spaced circles that individually can be matched onto the center ring can be derived by geometrical construction under constraints imposed by bilinear transformations and e.g., (4.2).

From a geometrical point-of-view, see Fig. 4.13 for an example configuration that shows just the center circle and 2 circles in the first ring, the distance c between the origin of the Smith chart and the center of one of the N_1 circles in the ring is given by

Fig. 4.13. Geometric construction to derive the matchable region for one-stage one-ring switchable matching networks.

$$c = |\Gamma_{Zin}|_{max} \cos(\alpha) + r \cos(\beta) \quad (4.12)$$

where the angle directly follows from the equidistant spacing, $\psi = \Delta/N_1$. The radius of the smaller circles r and the angle ψ are related as $r \sin(\psi) = |\Gamma_{Zin}|_{max} \sin(\Delta/N_1)$. Using this, the region that can be matched to within the inner circle is a circle centered on the origin of the Smith chart, having a radius

$$|\Gamma_{Zload}|_{max} = |\Gamma_{Zin}|_{max} + 2r \cos\left(\frac{\pi}{N_1} + \beta\right)$$

From a bilinear point-of-view, (4.2) can be used to get another set of equations for r and c that also must be satisfied:

$$r = \frac{(1 - |\Gamma_{Zload}|_{max}^2) |\Gamma_{Zin}|_{max}}{1 - |\Gamma_{Zload}|_{max}^2 |\Gamma_{Zin}|_{max}^2} \quad (4.13)$$

and

$$c = \frac{|\Gamma_{Zload}|_{max} (1 - |\Gamma_{Zin}|_{max}^2)}{1 - |\Gamma_{Zload}|_{max}^2 |\Gamma_{Zin}|_{max}^2} \quad (4.14)$$

Substituting (4.13), (4.14) into (4.12) yields

$$\begin{aligned} \frac{|\Gamma_{Zload}|_{max} (1 - |\Gamma_{Zin}|_{max}^2)}{1 - |\Gamma_{Zload}|_{max}^2 |\Gamma_{Zin}|_{max}^2} &= |\Gamma_{in}| \cos\left(\frac{\pi}{N_1}\right) + \\ &\frac{(1 - |\Gamma_{Zload}|_{max}^2) |\Gamma_{Zin}|_{max}}{1 - |\Gamma_{Zload}|_{max}^2 |\Gamma_{Zin}|_{max}^2} \sqrt{1 - \frac{(1 - |\Gamma_{Zload}|_{max}^2 |\Gamma_{Zin}|_{max}^2)^2}{(1 - |\Gamma_{Zload}|_{max}^2)^2}} \sin^2\left(\frac{\pi}{N_1}\right) \end{aligned}$$

Eliminating the square root and using $\sin^2(x) + \cos^2(x) = 1$, the previous equation can be rewritten into an equation that is 6th order in $|\Gamma_{Zin}|_{max}$ and 4th order in $|\Gamma_{Zload}|_{max}$.

$Z_{load}|_{max}$. This equation can be rewritten into an equation that is 4th order in $|Z_{in}|_{max}$ and 3rd order in $|Z_{load}|_{max}$ under the (always satisfied) constraints that $|Z_{in}|_{max} < 1$ and $|Z_{load}|_{max} > 0$:

$$\left(1 - |\Gamma_{Zin}|_{max}^2 |\Gamma_{Zload}|_{max}^2\right) * \left[\left(1 + |\Gamma_{Zin}|_{max}^2\right) |\Gamma_{Zload}|_{max} - 2|\Gamma_{in}|_{max} \cos\left(\frac{\pi}{N_l}\right) \right] = 0$$

Excluding a trivial (and for matching networks invalid) solution $|Z_{load}|_{max} = |Z_{in}|_{max}^{-1}$, this directly yields

$$\left(1 + |\Gamma_{in}|_{max}^2\right) |\Gamma_{Zload}|_{max} - 2|\Gamma_{in}|_{max} \cos\left(\frac{\pi}{N_l}\right) = 0$$

which leads to

$$|\Gamma_{Zload}|_{max} = \frac{2|\Gamma_{in}|_{max} \cos\left(\frac{\pi}{N_l}\right)}{1 + |\Gamma_{in}|_{max}^2} \quad (4.15)$$

with (4.1) and (4.4), (4.15) yields (4.5) and (4.6).

Appendix B

This appendix shows some background on the decomposition of the transfer function of an arbitrary lossy matching network into a cascade of transfer functions, one of which corresponding to the transfer of a lossless matching network [27]–[29]. Any two-port linear network can be described as a 2×2 S- or ABCD matrix. For a lossless network, the on-diagonal elements in the ABCD-matrix are real and the off-diagonal elements are imaginary. For lossy networks the power gain is lower than unity which results in additional constraints which are irrelevant for the decomposition in this Appendix. In general the ABCD-matrix of any lossless matching network therefore has 4 degrees of freedom. For reciprocal circuits, and the number of degrees of freedom would be 3; however, in this chapter no reciprocity is assumed.

The transfer of an arbitrary (lossy) network can be written in a complex 2×2 ABCD-matrix with no constraints on any element in that matrix: the ABCD-matrix then has 8 degrees of freedom. This general ABCD-matrix can be decomposed into multiple ABCD-matrices (one of which corresponds to a lossless network) in a number of ways. Common to all these decompositions is that the total number of degrees of freedom is (at least) 8 and that the decomposition can be made. This last requirement can be reformulated into being able to cover the 8 (4 real and 4 imaginary) dimensional space spanned by the original ABCD-matrix by the cascade of ABCD-matrices in the decomposition.

The hardest decomposition is that of the arbitrary ABCD-matrix into a cascade of ABCD-matrices with in total 8 degrees of freedom. In case that the cascade of ABCD-matrices has more degrees of freedom clearly the mapping can be made more easily. In the derivation below purely resistive attenuator circuits are assumed for simplicity reasons; equivalent circuits including transformers also may be used. Then, starting from basic mathematical rules on determinants:

$$\begin{aligned} |ABCD_{general}| &= \text{complex} \\ |ABCD_{lossless}| &= \text{real} \\ |ABCD_{attenuator}| &= 1 \\ |ABCD_{phaseshifter}| &= \text{complex} \\ |A| \cdot |B| \cdot \dots \cdot |Z| &= |A \cdot B \cdot \dots \cdot Z| \end{aligned}$$

it directly follows that decomposing a general (lossy) ABCD-matrix $[ABCD]_{general}$ into a cascade including one lossless matching network $[ABCD]_{lossless}$ requires at least a phase shifter $[ABCD]_{phaseshifter}$ and 2 lossy attenuators $[ABCD]_{attenuator}$. For a decomposition into a cascade that has 8 degrees of freedom we then get, e.g.,

$$[ABCD]_{general} = [ABCD]_{attenuator1} \cdot [ABCD]_{lossless} \cdot [ABCD]_{attenuator2} \cdot e^{j\phi}$$

Using the inverse function theorem it can be shown that this decomposition can be made with attenuators having in total 3 degrees of freedom (total insertion loss and impedance ratios). Using more degrees of freedom in the attenuators enables infinitely many ways to decompose the general transfer function in the way shown in the above relation.

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Chapter 5

Hardware implementation overhead of switchable matching networks

This chapter is the paper: E. L. Firrao, A. J. Annema, F. E. van Vliet, and B. Nauta, “Hardware Implementation Overhead of Switchable Matching Networks”, on IEEE Transactions on Circuits and Systems — I: Regular Papers, Vol. 64, No. 5, May 2017.

Abstract — Nowadays, more and more RF systems include switchable matching networks to decrease the impact of the environment-dependent antenna impedance on the RF front end performance. This paper reviews the theoretical lower limit on the required number of matching states to match VSWR ranges and then presents an analysis of hardware implementations to actually implement a suitable switchable matching network. A number of matching network topologies are analyzed: PI networks, loaded transmission lines, branch line coupler based circuits, single circulators and cascaded circulators. In our investigation only narrow-band applications are targeted. For the various circuit implementations the required number of matching states for each hardware implementation is compared to the theoretical minimum number of states required for the same matching in order to benchmark their hardware implementation overhead. It appears that a matching network using cascaded circulators is the closest to the theoretical optimum for networks with a relatively low number of states: this type of matching network was implemented and analyzed in more detail.

Index Terms — automatic antenna tuner, impedance matching, tunable matching network, switchable matching network.

5.1 Introduction

Antenna impedances are heavily dependent on their EM environment [1][2]. As a result, antenna impedances may change significantly during operation in e.g. handheld devices. The antenna impedance Z is usually expressed in terms of reflection coefficient $\Gamma = (Z - Z_0)/(Z + Z_0)$, where Z_0 is the characteristic impedance, or in terms of voltage standing wave ratio $VSWR = (1 + |\Gamma|)/(1 - |\Gamma|)$. For a typical antenna the VSWR can be up to 10:1 [1]-[5] which corresponds to $|\Gamma|$ values up to about 0.81.

Typically RF power amplifiers are optimized to drive a nominal load, usually 50 Ω , through a fixed impedance matching network. The VSWR associated with changing antenna impedances cause serious design and performance challenges for the RF power amplifier (PA) driving the antenna. Assuming a certain lower bound on maximum transmit power, varying load impedances for the PA require robustness to both maximum voltage and maximum current levels well above those required when driving a nominal load impedance at the same power level [3]-[5]. Consequently, the PA must be designed to operate properly for a wide load impedance range and hence the PA is necessarily overly robust at nominal conditions which results in significantly reduced efficiency at nominal conditions. On top of this, matching losses due to non-nominal load impedances decrease the efficiency and radiated power significantly.

Automatic antenna tuners are used to match an antenna impedance to an impedance close to the nominal impedance, which is typically 50 Ω . An antenna tuner [6]-[25] is generally implemented through a system consisting of impedance sensing circuitry, a tunable matching network and a control loop that implements the tuning procedure [26]-[30] of the matching network. The tunable matching network can be a continuous tunable matching network or a switchable matching network; typically a switchable matching network is used [6]-[25].

In [31], the theoretical minimum number of states for switchable matching networks was derived, required to match any load impedance for which $|\Gamma_{load}| \leq |\Gamma_{load}|_{max}$ to within a smaller area for which $|\Gamma_{in}| \leq |\Gamma_{in}|_{max}$. The work in [31] targets the theoretical minimum number of required matching network states, explicitly not considering any hardware implementation.

The current paper is about the hardware overhead of implementing a switchable matching network that matches from $|\Gamma_{load}| \leq |\Gamma_{load}|_{max}$ to $|\Gamma_{in}| \leq |\Gamma_{in}|_{max}$ for a number of topologies. This hardware overhead is defined as the ratio of the number of states required for a specific (optimized) hardware implementation and the theoretical minimum number of required states to achieve the same matching performance. This is important because minimization of the hardware implementation most likely leads to easier tuning algorithms and typically results in lower losses. Lower losses directly translate in more efficient transmit systems, while in receive mode the noise figure, noise

matching and losses are all relevant. In this paper, the main focus is on transmit and size and cost aspects are excluded as primary selection criterion.

For the analyses in this chapter, only narrow band applications are targeted and the values of switchable reactances are assumed to be binary scalable (where every switch is controlled by one control line that corresponds to a bit). Furthermore, the passive components in the matching network are assumed to be ideal: linear and without spread. Losses and thermal noise are accounted for by resistive components associated with limited quality factors for (reactive) passive components. Note that this “ideal” is different from [31] where the expression “ideal” denoted the absence of any specific hardware implementation in order to make the theory as much general as possible.

This chapter is arranged as follows. Section 5.2 presents a short review of the theoretical minimum number of required states for switchable matching networks. Section 5.3 reports the hardware implementation overhead – in terms of required number of states – for PI networks, loaded transmission line matching networks, branch line coupler based matching networks, and for matching networks using circulators. For a low number of states, the analyses show that the cascaded circulator topology appears to have the best hardware implementation overhead. In this section the components and switches are assumed to be ideal, linear and lossless.

Section 5.4 shows in depth analyses and measurements on the matching network that is the closest to the theoretical optimum (in required number of states): the cascaded circulator based matching network. Section 5.5 presents a short analysis of the effects of lossless components on the power efficiency and mismatch efficiency of a switchable matching network. In this section the design vehicle is a 4-bit tunable PI network for simplicity and clarity reasons. Finally, the conclusions are summarized in section 5.6.

5.2 Minimum number of states

In [31] it was shown that the minimum required number of states for a switchable matching network can be derived mathematically. It was shown that it is optimum to match radially equispaced circular shaped reflection areas on the Smith chart onto a circular region centered in the origin of the Smith chart. Similar results were obtained recently [25], based on a pure optimization based approach. For higher $|\Gamma_{\text{load}}|_{\text{max}}/|\Gamma_{\text{in}}|_{\text{max}}$ ratios, it was shown that cascaded simple matching networks are more efficient than more complex matching networks, in terms of the theoretical minimum required number of matching network states to match an impedance from anywhere inside $|\Gamma_{\text{load}}| \leq |\Gamma_{\text{load}}|_{\text{max}}$ to an impedance within $|\Gamma_{\text{in}}| \leq |\Gamma_{\text{in}}|_{\text{max}}$.

5.2.1 Theoretical optimum matching properties

The theoretical optimum matching properties of a switchable impedance matching network are illustrated in Fig. 5.1, assuming $|\Gamma_{in}|_{max}=1/3$. Fig. 5.1a shows the $|\Gamma_{load}|_{max}$ as a function of the numbers of states for an optimum matching network. Because the rest of this paper assumes binary controlled matching network states, the number of states is shown on a $\log_2$ -scale.

For a very low number for N_{states} , the optimum configuration appears to be as shown in Fig 5.1b: the $(N_{states} - 1)$ equidistant radially spaced circular regions all encircle the origin of the Smith chart, see also appendix A. For larger number for N_{states} , it is optimum to construct configurations such as shown in Fig. 5.1c-e. In these configurations, around the center $|\Gamma_{in}|=1/3$ circle, circular bands are constructed from circles that each can be matched to the center circle.

5.2.2 Hardware implementation overhead

The analyses leading to the minimum number of states to reach a certain impedance matching, as in Fig. 1, are purely mathematical: explicitly no assumption on actual hardware implementations was made. To compare actual hardware implementations of switchable matching networks, we defined the hardware implementation overhead in equation as:

$$hardwareoverhead = \frac{N_{states}}{N_{min}} \left| \{ |\Gamma_{load}| \leq |\Gamma_{load}|_{max}, |\Gamma_{in}| \leq |\Gamma_{in}|_{max} \} \right| \quad (5.1)$$

where N_{states} is the actual number of states of the switchable matching network circuit implementation and N_{min} is the theoretical minimum number of states.

In this paper we analyze and benchmark various hardware implementations of switchable matching networks. All have different ways of achieving impedance matching and consequently they all have different hardware overhead numbers. We analyze networks that are reciprocal, non-reciprocal, lumped or distributed to cover most conventional classes of switchable matching networks. We introduce a matching network using cascaded circulators because it most closely implements the theoretical optimum [31] expressed in number of states, not in size nor cost. The number of switches (equal to the number of switchable reactances) to implement N_{states} is $N_{switch} = \lceil \log_2(N_{states}) \rceil$ where the $\lceil x \rceil$ operator rounds to the nearest larger integer. Note that there is not necessarily a unique relation between N_{switch} and the total number of reactances, transmission lines and other passive matching network components.

Fig. 5.1. Minimum number of states matching construction to match from any $|\Gamma_{\text{load}}| \leq |\Gamma_{\text{load}}|_{\text{max}}$ to within anywhere inside $|\Gamma_{\text{in}}| \leq |\Gamma_{\text{in}}|_{\text{max}}$. For $|\Gamma_{\text{in}}|_{\text{max}} = 1/3$ this yields the $|\Gamma_{\text{load}}|_{\text{max}}(N_{\text{states}})$ graph in (a) and example configurations in (b) enclosing the center of the Smith chart and in (c)-(e) having respectively 1, 2 and 3 rings of regions that can be matched to the center circle.

5.3 Hardware Implementations

This section investigates the hardware implementation overhead for several hardware implementations. Four cases are treated, which are chosen as examples representative of particular classes. The PI networks are an example of a reciprocal lumped-element implementation. Similarly, the loaded transmission-line networks are an example of distributed matching networks, branch-line coupler networks are an example of reflection-based coupler networks and circulator networks are an example of non-reciprocal networks.

Single-stage matching networks are assumed; combinations or cascaded versions of the matching networks (such as LC-ladder networks) are not treated. Note that the cascaded circulator based matching network in section 5.3.5 uses cascaded circulators, but is not a cascaded version of the matching network in section 5.3.4.

5.3.1 PI networks

The PI network is a reciprocal and lumped-element implementation of a matching network. PI networks are widely used to implement switchable matching networks; a 4-bit switchable implementation example of a low-pass PI network is shown in Fig. 5.2. Its advantages over its T-network equivalent is that parasitic source and load capacitances can easily be absorbed in PI networks while a low pass PI network has advantages over its high-pass counterpart because it allows to filter harmonic content.

Fig. 5.2. Typical implementation of a (here 4 bit switchable) low-pass PI matching network.

The switchable (banks of) reactances in the PI network implement 2^n states of this matching network. Appendix B shows the derivations of the output reflection coefficient for each state which can directly be used to determine the region on the Smith chart that can be matched to the target impedance range defined by $|\Gamma_{in}| \leq |\Gamma_{in}|_{max}$. Due to the compressive nature of the Smith chart and the very non-linear equation for impedance as function of the switchable reactances, there is a considerable overhead in states, compared to the theoretical optimum [31].

An example of the matchable area on the Smith chart is shown in Fig. 5.3. In this figure, every dot can be matched onto the center of the Smith chart, while every circle around each dot can be mapped onto the $|\Gamma_{in}| = |\Gamma_{in}|_{max}$ circle. For this figure, the optimization to match the largest area was done using brute force computing (see appendix C for more details), here targeting $|\Gamma_{in}|_{max}=1/3$ or equivalently VSWR=2:1 and assuming 4 bit settings. For this optimum tunability, the two capacitances and the inductor have 1 respectively 2 bit tunability, see Fig. 5.3.

Fig. 5.3. An example of matchable area of a 4-bit switchable low pass PI matching network based on brute force computing on a Smith chart for $|\Gamma_{in}|_{max} = 1/3$.

For this example 4-bit PI network, for $|\Gamma_{in}|_{max}=1/3$ the matchable region is bounded by $|\Gamma_{load}|_{max} \cong 0.54$ which corresponds to VSWR $\cong 3.3$. With (5.1), the hardware implementation overhead turns out to be about 2. A summary of the $|\Gamma_{load}|_{max}$ and hardware overhead, as a function of the number of bits, for all topologies considered in this paper is shown in Fig. 5.12 and 5.13.

5.3.2 Loaded transmission line based matching networks

Another way to implement a switchable matching network is to load a transmission line with switchable capacitors [31]-[34], see Fig. 5.4 for a 4 bit switchable implementation. This type of impedance matching network is a reciprocal and a distributed element implementation. The advantage of this implementation is simplicity. A disadvantage is the difficulty to make it off-chip at microwave frequencies as the physical length of each transmission line becomes too short, while on-chip the length is relative large.

Fig. 5.4. Implementation of a 4-bit switchable loaded transmission line based matching network.

Appendix D shows the derivation of the output reflection coefficient for each state in a generic loaded transmission line based impedance matching network. To get N uniformly spaced phases of Γ_{out} , we would require N specific transmission line lengths, that directly yields significant redundancy and overhead. An example of matchable area on the Smith chart for a 4 bit switchable implementation, targeting $|\Gamma_{\text{in}}|_{\text{max}}=1/3$ is shown in Fig. 5.5.

Fig. 5.5. Matchable area of a 4-bit switchable loaded transmission line based matching network on a Smith chart for $|\Gamma_{\text{in}}|_{\text{max}} = 1/3$.

For this example the matchable region is bounded by $|\Gamma_{\text{load}}|_{\text{max}} \cong 0.54$ which corresponds to $\text{VSWR} \cong 3.3$. With (1), the hardware implementation overhead turns out to be about 2.

5.3.3 Branch line coupler based matching networks

Branch line couplers can be used to implement impedance transformation [35], [36]-[39], yielding a reciprocal and distributed implementation. An example of a 5-bit switchable branch line coupler based impedance matching network is shown in Fig. 5.6; appendix E shows the derivation of its impedance matching performance. Similar to the PI-network, the very non-linear relations between Γ_{out} and the reactances yields a

significant hardware implementation overhead. The advantage of this topology is that it is well suited for implementation in MMIC technologies; the disadvantage is that the losses are usually higher than that of other topologies.

Fig. 5.6. Implementation of a 5-bit switchable branch line coupler based matching network.

Fig. 5.7 shows the impedance matching capabilities of a 5-bit switchable impedance matching network, for $|\Gamma_{in}|_{max}=1/3$. Again, every dot can be matched toward the center of the Smith chart, while every circle can be mapped onto the $|\Gamma_{in}|=1/3$ circle. For this matching network, $|\Gamma_{load}|_{max} \cong 0.45$ while the hardware implementation overhead is about 3.8 obtained using brute force optimization.

Fig. 5.7. Matchable area of a 5-bit branch line coupler based matching network for $|\Gamma_{in}|_{max} = 1/3$.

5.3.4 Single circulator based matching networks

Circulators can be used to implement (non-reciprocal) impedance matching

network [35], [40]-[43]. In a three-port circulator, ideally all the power incident to a single port is coupled to only one other port, leaving the other port isolated. This property can be used to rotate the phase of the load reflection coefficient. After proper rotation, impedance matching can be achieved using a shunt capacitor. An example of a (4-bit switchable) circulator-based impedance matching network is shown in Fig. 5.8.

Fig. 5.8. Circulator-based 4-bit switchable matching network

A mathematical derivation of the impedance matching properties of this type of matching network is shown in Appendix F. To get uniformly spaced mappable regions as required for the mathematical optimum [31], the appendix shows that the summed value of the switched reactances should be distributed as $X_n = R_0 \arctan(n\pi/N)$. A hardware implementation using a minimum number of switched reactances consequently shows a significant overhead of switchable states. Other disadvantages include the size and cost of circulators.

For $|\Gamma_{in}|_{max}=1/3$, the matching performance of a 4-bit switchable implementation is shown in Fig. 5.9. For this case, the maximum load VSWR is about 4.8 and the hardware overhead is about 6.

Fig. 5.9. Matchable area of a 4-bit switchable single circulator based matching network, for $|\Gamma_{in}| = 1/3$.

5.3.5 Cascaded circulator based matching networks

To get the closest to the theoretical optimum – in number of required matching network states, see [31] – uniformly distributed phase shifts should be implemented in the matching network in section 5.4. The resulting implementation, showing the lowest found hardware implementation overhead in terms of required states to achieve some matching, is shown in Fig. 5.10. Compared to the single circulator based matching networks, more circulators are required which increases cost and size. As advantage (ideal) nicely spaced phase steps can be obtained such as required for matching state configurations such as shown in Fig. 5.1b and c. With this topology, each circulator and associated switchable component section can be designed to achieve a specific phase shift step while the leftmost shunt capacitor can match towards the center of the Smith chart.

Fig. 5.10. Cascaded circulator based 4-bit switchable matching network implementation.

As derived in Appendix F, the output reflection coefficient of this (4-bit switchable) network is

$$\Gamma_{out} = \Gamma_1 \Gamma_{X1} \Gamma_{X2} \Gamma_{X3} \quad (5.2)$$

where Γ_{load} is the reflection coefficient at the output of the three circulators, Γ_{X1} , Γ_{X2} and Γ_{X3} are the reflection coefficients of the switchable jX_1 , jX_2 , jX_3 and where

$$\Gamma_1 = \frac{\frac{1}{Z_0} - Y_1}{\frac{1}{Z_0} + Y_1} \quad \text{and} \quad Y_1 = \frac{1}{R_{source}} + j\omega C$$

and where Z_0 is the reference impedance.

An example of the matchable area on a Smith chart for a 4-bit switchable impedance matching network using the topology in Fig. 11, for $|\Gamma_{in}|_{max}=1/3$, is shown in

Fig. 5.11. In this, each dot can be mapped onto the center of the Smith chart, while the dot at the center of the Smith chart is actually 8 coinciding dots, due to the nature of this network topology.

For this implementation, it can be shown that the hardware implementation overhead as defined in (1) is to 16/9 and it is the lowest achievable value for single ring optimum configurations as described in [31] and as shown in Figs. 5.1b and c.

Fig. 5.11. Matchable area of a 4-bit switchable cascaded circulator based matching network, for $|\Gamma_{in}| = 1/3$.

5.3.6 Discussion

Sections 5.3.2 to 5.3.5 showed a number of hardware implementations of switchable matching networks: PI networks, loaded transmission lines, branch line couplers, single circulators and cascaded circulators. For all of these, the optimum matching performance in terms of $|\Gamma_{load}|_{max}$ as a function of the number of bits tunability for a predefined $|\Gamma_{in}|_{max}$ is plotted in Fig. 5.12 for $|\Gamma_{in}| = 1/3$, $|\Gamma_{in}| = 1/2$ and for $|\Gamma_{in}| = 3/5$, along with the theoretical optimum [31].

It can be seen from these plots that the branch line coupler based matching network, see section 5.3.3, performs worst in terms of (1). In the low-number-of-bit range the implementation based on cascaded circulators performs best in terms of hardware overhead (1). The PI-network and loaded transmission line based implementations that were described in sections III.B and III.C have comparable matching performance in terms of required number of states to achieve some kind of impedance range matching. These two network topologies are optimum when having many states, only to be outperformed in terms hardware overhead at low and medium number of states by the cascaded circulator based matching network topology.

The hardware implementation overhead for the various topologies are plotted in Fig. 5.13, as a function of the number of bits tunability for or $|\Gamma_{in}| = 1/3$. This plot shows that the hardware implementation overhead — in terms of required number of states — increases rapidly with the number of (binary controlled) states in switchable matching networks. Especially the branch line coupler based and the single circulator based matching networks require a relatively large hardware implementation overhead in required states.

The differences in hardware overhead follow from the way states are mapped on the Smith chart, which is very different for all types we analyzed.

The lowest hardware implementation overhead, for a low number of states, is for the cascaded circulator based matching network. Its state distribution is the closest to the mathematically optimum distribution derived in [31]. This matching network is worked out in detail and experimentally in section 5.4. Disadvantages of this type of matching network are the bulkiness and cost of circulators.

The second best type of matching network appears to be the PI network, which does not have the mentioned disadvantages of the cascaded circulator network. An analysis of the impact of lossy reactances (or switches) on the performance of a (4 bit tunable, for simplicity reasons) PI network is shown in section 5.5.

a)

b)

c)

Fig. 5.12. Magnitude of the load reflection coefficient vs Number of bits for: a) $|\Gamma_{in}| = 1/3$; b) $|\Gamma_{in}| = 1/2$; c) $|\Gamma_{in}| = 3/5$.

Fig. 5.13. Hardware implementation overhead versus number of bits for different switchable matching network topologies for $|\Gamma_{in}| = 1/3$. For 4, 5, 6 and 7 bit tunability, the cascaded circulator topology has a lower hardware overhead compared to the other switchable matching network topologies.

5.4 Cascaded circulator topology

5.4.1 Lossy cascaded circulators

Section III showed that the (largely unknown) cascaded circulator based switchable impedance matching network performs the closest to the theoretical optimum [31], in number of required states. In this section, this type of switchable matching network is analyzed in more detail.

The theoretical matching performance of cascaded circulator based matching networks was described in section III.F. In that section, ideal lossless components were assumed. In case of lossy circulators, the circulators rotate phase but also decrease the signal magnitude. Assuming the same insertion loss between ports, the reduction in magnitude is a function of the overall insertion loss of the circulators and of the interconnections and of the quality factor of the switchable reactances. Then (5.2) turns into

$$\Gamma_{cir} = (IL^{3N}) \Gamma_{load} \Gamma_{X_1} \Gamma_{X_2} \Gamma_{X_3} \dots \Gamma_{X_N} \quad (5.3)$$

where Γ_{cir} the reflection coefficient at the input of the circulators, N is the number of cascaded circulators, IL is the linear insertion loss of the circulator (assuming the same for each port and the same for every circulator), Γ_{load} the load reflection coefficient and Γ_{X_1} , Γ_{X_2} , Γ_{X_3} and Γ_{X_N} are the reflection coefficient of the reactances, see Fig. 5.10.

In case of a 50Ω load, and excluding interconnect losses, the overall linear insertion loss is

$$IL_{dB} = N * 20 \log |IL|$$

The switchable matching network under investigation is linear (neglecting the non-linearity of the PIN diodes and circulators) and hence is fully characterized by its two-port S matrix (per state). For this we denote the input port as port 1 and denote the output port as port 2; there are as many S-matrices as there are number of states. For the 4-bit switchable example in Fig. 5.10, 16 S-matrices are needed. Once the overall S-matrices are known, the load contour can straight forwardly be derived for each state based on the required input reflection coefficient contour. Note that as the matching network is linear, load-pull measurements do not provide more information than using the S-matrix does. The input reflection coefficient Γ_{in} and load reflection coefficient Γ_{load} are related as

$$\Gamma_{load} = \frac{\Gamma_{in} - S_{11}}{S_{22}\Gamma_{in} - \Delta}$$

where

$$\Delta = S_{11}S_{22} - S_{21}S_{12}$$

The power efficiency of the switchable matching network is

$$\eta = \frac{P_{load}}{P_{in}} = \frac{(1 - |\Gamma_{load}|^2) |S_{21}|^2}{(|1 - S_{22}\Gamma_{load}|^2)(1 - |\Gamma_{in}|^2)} \quad (5.4)$$

where

$$\Gamma_{in} = S_{11} + \frac{S_{21}S_{12}\Gamma_{load}}{1 - S_{22}\Gamma_{load}}.$$

5.4.2 A practical example

A hardware implementation of a 4-bit switchable cascaded circulator based matching network was built, see Fig. 5.10 for the architecture and Fig. 5.14 for a photo of the prototype.

Fig. 5.14. Photo of the prototype implemented and measured in the lab.

The following circulator was used: CCMTH0801-915-ETL; the measurement frequency was set to 947 MHz and the PCBs were designed for this frequency. Two types of PCBs were designed: one to implement the switchable capacitors and the other to implement the switchable shunt capacitor. Although pHEMT's and MESFET's may be preferred for their lower control power, for simplicity reasons in our experimental setup we used PIN diodes to implement the switches. The switch-on current was 10mA. The PIN diode was biased using two inductors in series (for better isolation), a resistor and a shunt capacitor at the DC input, see Fig. 5.15.

Fig. 5.15. Simplified schematic of the biasing circuit of the PIN diodes.

The measurements were carried out using a Rohde & Schwarz ZVB-20 VNA at a power level of 0 dBm. The measured S-matrices for each state were then used to derive the load contour based on the required input reflection coefficient contour.

Three values for $|\Gamma_{in}|_{max}$ were chosen: 1/3, 1/5 and 1/10 for which the (measurement based) results are shown in Figs. 5.16a, b and c respectively. For these $|\Gamma_{in}|_{max}$ values, the SMD shunt capacitors at the input are 1pF, 1.7pF and 2.7pF. The values of the capacitors that implement X_1 , X_2 and X_3 are 13pF, 3pF and 1.7pF. The measurement results show that the theoretical and measured behavior of the switchable matching network complies nicely. Note that due to the losses, the matchable area is larger than the one derived from the theory using ideal (lossless) components.

5.5 The impact of losses – an example

The analyses in section III assumed ideal switches and ideal reactances, whereas section IV presented analyses and measurement results for the network that is closest to the mathematical optimum in [31]. The results in section IV already show that lossy components have a significant impact on e.g. the matching performance. A major drawback of the system in section 5.4 is in the cost and size of circulators, which makes the cascaded circulator based network unsuitable for e.g. handheld devices.

This section presents the impact of lossy components and switches using a 4-bit tunable PI network as vehicle, for $|\Gamma_{in}|_{max} = 1/3$, corresponding to the situation in Fig. 5.3. This PI network is the second best from hardware overhead point of view and does not have the cost and size disadvantages associated with the matching network in section IV.

For e.g. handheld devices, the matching network could be monolithically implemented on silicon, on GaAs or e.g. using discrete SMD components, that all have losses. To show the impact of lossy components, we assume lossy reactances that have specific Q_L and Q_C that are the same for all inductors respectively capacitors in the switchable matching network. We assume for simplicity that this Q includes the effect of lossy switches. In the comparison, we assume ideal (lossless) implementations with fixed $Q_C = Q_L \rightarrow \infty$, implementations in silicon for which we used fixed $Q_C=20$ and $Q_L=10$, and implementations in GaAs for which we used $Q_C=60$ and $Q_L=35$. As the focus of this paper is on transmit, this section reports on power efficiency (eqn. 5.4) and the impact of impedance matching. For the latter case a variant of (5.4) is used, omitting the second term in the denominator. The impact of lossy matching networks on the noise figure is briefly addressed in appendix G.

Fig. 5.16. Matchable area on the Smith chart for various magnitude of the input reflection coefficient: a) $|\Gamma_{in}| = 1/3$; b) $|\Gamma_{in}| = 1/5$; $|\Gamma_{in}| = 1/10$. The black continuous circle is for ideal circulators with perfect matching at all the ports when the switch of the shunt capacitor at the input is open. Because of the non-ideal behavior of the circulators, in case the switch of the shunt capacitor at the input is open, the circles of the states do not overlap exactly.

5.5.1 Constant power efficiency and mismatch contours

Below we report (for the optimized 4b PI network for $|\Gamma_{in}|_{\max} = 1/3$) constant power efficiency contours and constant matching efficiency contours on the Smith chart. This is done for all three cases listed above. The power efficiency contours are relevant for cases where the matching network is driven by a non-linear source, e.g. an RF power amplifier. The constant mismatch contours assume a linear resistive signal source driving the matching network and its load. For each of the three cases, an optimization was done to reach the largest (circular) matchable region, with radius $|\Gamma_{load}|_{\max}$. The resulting optimum values for the swithable reactances and the radius of the matchable region, $|\Gamma_{in}|_{\max} = 1/3$, are reported below.

Fig. 5.17a shows the constant power efficiency contours for a lossless PI network, where the contour demarks the matchable area. The matching efficiency contours for the same network are shown in Fig. 5.17b.

	value	value	value	tunability
	$Q_C \rightarrow \infty$ $Q_L \rightarrow \infty$	$Q_C = 20$ $Q_L = 10$	$Q_C = 60$ $Q_L = 35$	
X_{CIN}	-41 Ω	-38 Ω	- 40 Ω	1b
X_L	23 Ω	22 Ω	23 Ω	2b
X_{COUT}	-59 Ω	-43 Ω	-51 Ω	1b
$ \Gamma_{load} _{\max}$	0.54	0.66	0.58

Fig. 5.17: For a lossless 4b-tunable PI network for $|\Gamma_{in}|_{\max} = 1/3$ (a) constant power efficiency contours (b) constant matching efficiency contours.

Fig. 5.18a shows constant power contours for a 4-b tunable PI network for $|\Gamma_{in}|_{\max} = 1/3$, implemented on low ohmic silicon ($Q_C=20$, $Q_L=10$). Due to the losses, the matchable area on a Smith chart is bigger and the power efficiency can as low as 10%. Fig. 5.18b shows the matching efficiency contours, demonstrating that matching efficiency is a little lower than the power efficiency.

Fig. 5.18: For a 4b-tunable PI network for $|\Gamma_{in}|_{\max} = 1/3$ in silicon ($Q_C=20$, $Q_L=10$): (a) constant power efficiency contours (b) constant matching efficiency contours for the same case.

Fig. 5.19a shows constant power contours and constant mismatch efficiency contours assuming implementation in GaAs. Since GaAs is a semi-insulator (not a semiconductor like silicon) the quality factor of the passive components is higher: for this example it is assumed that the quality factor of the inductors is 35 and the quality factor of the capacitors is 60.

Fig. 5.19: For a 4b tunable PI network for $|\Gamma_{in}|_{\max} = 1/3$, implemented on GaAs ($Q_C=60$, $Q_L=35$) (a) constant power efficiency contours; (b) constant matching efficiency contours for the same case.

It can be concluded that implementing (here) PI networks using low-Q reactances as in silicon easily yields significantly decreased power and mismatch efficiencies. As a direct result, having no additional impedance matching could outperform impedance matching using low-Q reactances. To illustrate this, Fig. 5.20a and b show the constant matching contours of Fig. 5.18b respectively Fig. 5.19b, showing only the area where the matching efficiency using the matching network is higher than using no matching network at all. For the Si-case a significant part of the mappable region appears to be better off without using a (low Q) matching network.

Fig. 5.20: Constant mismatch contours for a PI network (a) in silicon and (b) in GaAs, showing only the regions where the efficiency with a matching network is better than that without matching network.

5.6 Conclusions

Hardware implementation overhead of switchable matching networks, expressed in required number of matching states to get a predefined impedance matching performance, normalized to the corresponding theoretical lower limit, was worked out in detail. For this, narrow-band applications and binary scaled component values were assumed. Several circuit topologies were investigated: switchable impedance matching networks using PI networks, based on loaded transmission lines, based on branch line couplers, using single circulators and utilizing cascaded circulators.

It follows from the types analysed that (for not too many control bits) the cascaded circulator based topology is the closest to the theoretical optimum, in required number of matching states. This topology requires just up to 2 times the theoretical minimum number of matching states; this type of switchable matching network has been

worked out as a practical example. Second best are PI network and loaded transmission line based switchable impedance matching networks that typically require up to about a factor 10 more matching states than the theoretical minimum.

Measurements on a realized cascaded circulator based matching network confirm its performance. The impact of losses on power efficiency and matching efficiency are shown using a 4-bit tunable PI matching network as vehicle.

Appendix A

This appendix presents an extension of the derivation of the minimum number of states for lossless switchable matching networks [31], for a low number of states. The derivations in [31] excluded the possibility that (except for the target region) matched impedance regions enclose the center of the Smith chart. In this appendix the derivation is extended to encircle the center of the Smith chart.

An example configuration with 5 circular region that can be matched to the centered circular region is shown in Fig. 5.21a. The mismatch efficiency M of a matching network is [35]

$$M = \frac{(1 - |\Gamma_{out}|^2)(1 - |\Gamma_{load}|^2)}{|1 - \Gamma_{out}\Gamma_{load}|^2}$$

in terms of the output reflection coefficient Γ_{out} and of the load reflection coefficient Γ_{load} . Using this equation, it can be derived that the magnitude of intermediate load reflection coefficient $|\Gamma_{load}|_1$ and $|\Gamma_{load}|_2$ as defined in Fig. 5.21a is

$$|\Gamma_{load}|_{1,2} = \frac{\pm \beta |\Gamma_{out}| + \sqrt{\beta^2 |\Gamma_{out}|^2 - (\beta |\Gamma_{out}|^2 + 1)(\beta - 1)}}{1 + \beta |\Gamma_{out}|^2}$$

where

$$\beta = \frac{M}{1 - |\Gamma_{out}|^2}$$

Using goniometric constructions, an expression for the matchable area on a Smith chart can be derived, yielding

$$|\Gamma_{load}|_{\max} = c \cos\left(\frac{\pi}{N}\right) + \sqrt{r^2 - c^2 \sin^2\left(\frac{\pi}{N}\right)}$$

where $c = \frac{\beta|\Gamma_{out}|}{I + \beta|\Gamma_{out}|^2}$ and $r = \frac{\sqrt{\beta^2|\Gamma_{out}|^2 - (\beta|\Gamma_{out}|^2 + I)(\beta - I)}}{I + \beta|\Gamma_{out}|^2}$

The expression for $|\Gamma_{load}|_{max}$ is hence a function of $|\Gamma_{in}|$, N and $|\Gamma_{out}|$. In order to maximize the matchable area, the expression should be maximized for $|\Gamma_{out}|$ which for this paper was done numerically. However, a fair estimation for this maximum can be found choosing $|\Gamma_{out}| = |\Gamma_{in}|$. For this case the states are depicted in Fig. 5.21b while for $|\Gamma_{out}| = |\Gamma_{in}|$

$$|\Gamma_{load}|_{max} = 2 \frac{|\Gamma_{in}|}{I + |\Gamma_{in}|^2} \cos\left(\frac{\pi}{N}\right).$$

Fig. 5.21. a) states enclosing the center of the Smith chart; b) states enclosing the center of the Smith chart for $|\Gamma_{out}| = |\Gamma_{in}|$.

Appendix B

This appendix shows a mathematical derivation of the impedance matching capabilities of a low-pass PI network having binary scaled capacitors and inductor. In the derivation b_i denotes the value of control bit of the i^{th} switchable component. The output reflection coefficient of the PI network is

$$\Gamma_{out} = \frac{\frac{I}{Z_0} - Y_{out}}{\frac{I}{Z_0} + Y_{out}}$$

where

$$Y_{out} = Y_{C2} + \frac{I}{Z_2} = j\omega \sum_{i=1}^{n3} b_i C_{2,i} + \frac{I}{Z_2}$$

$$Z_2 = Z_{L1} + \frac{I}{Y_1} = j\omega \sum_{i=1}^{n2} b_i L_i + \frac{I}{Y_1}$$

$$Y_1 = Y_{C1} + \frac{I}{Z_{source}} = j\omega \sum_{i=1}^{n1} b_i C_{1,i} + \frac{I}{Z_{source}}$$

The Z_{load} for which $\Gamma_{load} = \Gamma_{out}^*$ can be matched onto Z_0 . Similarly, all load impedances that can be matched unto a circle $|\Gamma_{in}| = |\Gamma_{in}|_{max}$ can easily be derived.

Appendix C

This appendix describes the brute force optimization procedure used to compute the maximum mappable area on a Smith chart. In determining the mapping performance of N-bit switchable matching networks, the aim is to get the largest mappable circular Γ_{load} region, composed of $2N$ circular regions that can all be mapped onto the center circle defined by $|\Gamma| \leq |\Gamma_{in}|_{max}$:

1. an initial guess for matching network components is made. In this work, the N bit tunability can be distributed across the M reactances in the switchable matching network in any way that satisfies $N = \sum_M n_M$ where n_M is the tunability (in bits) of the m th tunable reactance in the matching network. In this paper we report the results that yields the largest mappable area; this follows after optimization for each possible distribution (see below).
2. the mapping of the matching network for each state of the 2^N states is determined. In each state, a circular Γ_{load} region on the Smith chart is mapped onto the center circle defined by $|\Gamma| \leq |\Gamma_{in}|_{max}$. As each circular region is uniquely defined by 3 different points on its perimeter, it is sufficient to get a set of three Γ_{load} 's that map onto (e.g.) $\{|\Gamma_{in}|_{max}, -|\Gamma_{in}|_{max}, j|\Gamma_{in}|_{max}\}$. Then getting the circumscribed circle for the three load reflection coefficients using standard algebra yields the full Γ_{load} region that can be mapped onto $|\Gamma| \leq |\Gamma_{in}|_{max}$, per state of the switchable matching network.
3. the circular mappable areas for each the 2^N states of the switchable matching network are plotted together on the Smith chart. This plotting is a numerical/graphics step required for the next step.

4. the largest inscribed circle that only encloses mappable areas is determined graphically. As all mappable areas are plotted together on the Smith chart this is a quite straightforward numerical procedure on the graphics representation derived in the third step.

In the optimization procedure, an M-dimensional sweep of the M matching network component values is executed, starting at a coarse grid, selecting the best performing settings and (about that best point) redoing the M-dimensional sweep at increasingly higher resolution.

For lossless matching networks, getting the Γ_{load} for a specific Γ_{in} is described in Appendix B to E. For lossy matching networks, all our analyses use S-parameter descriptions to derive the matching properties. From these we can straightforwardly derive the matching efficiency and power efficiency. The derivation of the behavior of a few types of matching networks is already described in terms of S-parameters. For e.g. a lossy PI-network, getting S-parameters was done by firstly deriving Y-parameters after which Y-to-S parameter conversion was done [44].

Appendix D

This appendix shows a mathematical derivation of the impedance matching behavior loaded transmission line based matching networks, such as shown in Fig. 5.4. Using the notations in Fig. 5 and working from source towards the load,

$$\Gamma_l = \frac{\frac{I}{Z_0} - Y_l}{\frac{I}{Z_0} + Y_l} e^{-j2\theta} \quad \text{with} \quad Y_l = Y_{\text{source}} + b_l \cdot j\omega C_l$$

where θ is the length of the leftmost transmission line, and b_n denotes the value of control bit of switchable capacitor C_n . Working towards the load impedance, the exact same relations (except for the values of θ_n and C_n) follow, recursively:

$$\Gamma_{\text{out}} = \frac{\frac{I}{Z_0} - Y_n}{\frac{I}{Z_0} + Y_n} e^{-j2\theta_n} \quad \text{with} \quad Y_n = \frac{I}{Z_0} \frac{1 + \Gamma_{n-1}}{1 - \Gamma_{n-1}} + b_n \cdot j\omega C_n$$

where C_n is the n^{th} capacitor.

Appendix E

This appendix shows a mathematical derivation of the matching performance of a branch line coupler based matching network as shown in Fig. 5.6. In the derivation a_1 , a_2 , a_3 and a_4 are the incident waves of the four ports of the branch line coupler while b_1 , b_2 , b_3 and b_4 are the reflected waves. Γ_{x2} and Γ_{x3} are the reflection coefficients of the two (switchable) reactive loads at port 2 respectively 3.

$$\begin{cases} b_1 = -\frac{j}{\sqrt{2}}a_2 - \frac{l}{\sqrt{2}}a_3 = -\frac{j}{\sqrt{2}}\Gamma_{x2}b_2 - \frac{l}{\sqrt{2}}\Gamma_{x3}b_3 \\ b_2 = -\frac{j}{\sqrt{2}}a_1 - \frac{l}{\sqrt{2}}a_4 \\ b_3 = -\frac{l}{\sqrt{2}}a_1 - \frac{j}{\sqrt{2}}a_4 \\ b_4 = -\frac{l}{\sqrt{2}}a_2 - \frac{j}{\sqrt{2}}a_3 = -\frac{l}{\sqrt{2}}\Gamma_{x2}b_2 - \frac{j}{\sqrt{2}}\Gamma_{x3}b_3 \end{cases}$$

The input and output of the matching networks are at port 1 and 4, yielding

$$\begin{cases} b_1 = \frac{l}{2}(\Gamma_{x3} - \Gamma_{x2})a_1 + j\frac{l}{2}(\Gamma_{x2} + \Gamma_{x3})a_4 \\ b_4 = j\frac{l}{2}(\Gamma_{x2} + \Gamma_{x3})a_1 + \frac{l}{2}(\Gamma_{x2} - \Gamma_{x3})a_4 \end{cases}$$

from which readily a two-port S matrix can be obtained:

$$S = \begin{pmatrix} \frac{l}{2}(\Gamma_{x3} - \Gamma_{x2}) & j\frac{l}{2}(\Gamma_{x2} + \Gamma_{x3}) \\ j\frac{l}{2}(\Gamma_{x2} + \Gamma_{x3}) & \frac{l}{2}(\Gamma_{x2} - \Gamma_{x3}) \end{pmatrix}$$

Now the output reflection coefficient of the matching network can be derived, as a function of the (switchable) Γ_{x2} , Γ_{x3} and the switchable shunt capacitor C:

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_1}{1 - S_{11}\Gamma_1}$$

$$\text{with } \Gamma_I = \frac{\frac{1}{Z_0} - Y_I}{\frac{1}{Z_0} + Y_I} \text{ and } Y_I = \frac{1}{R_{source}} + j\omega C$$

Appendix F

This appendix shows the mathematical derivation of matching capabilities of a circulator-based impedance matching network as depicted in Fig. 5.8. In the expression C is the value of the shunt capacitance. A base property of a circulator is

$$\Gamma_{out} = \Gamma_I \Gamma_X$$

where in the circuit in Fig. 5.10

$$\Gamma_I = \frac{\frac{1}{Z_0} - Y_I}{\frac{1}{Z_0} + Y_I} \text{ and } Y_I = \frac{1}{R_{source}} + j\omega C$$

and where Γ_X is the reflection coefficient of the switchable reactance X (the combined effect of jX_1 , jX_2 and jX_3 with their switches). Assuming $Z_0 = R_0$, to get N uniformly distributed phase shifts $\angle \Gamma_X$, the corresponding reactances are given by $X_n = R_0 \arctan(n\pi/N)$ which are far from uniformly spaced. This results in having to have more switched reactances than required for the mathematical minimum as defined in [31] that assumes uniform spacing.

Appendix G

This appendix is about the noise figure of general matching networks, applied to a 4-bit tunable PI network topology. According to [45], the noise figure NF of a passive lossy matching network is related to the available power gain G_{ava} by

$$NF = \frac{1}{G_{ava}}$$

where

$$G_{ava} = \frac{|S_{21}|^2}{1 - |S_{22}|^2}$$

where S_{11} , S_{21} , S_{12} and S_{22} are the S parameters of the matching network. Below are contour plots showing the available power gain G_{ava} for the PI network topology (in dB); dropping the minus sign yields the noise figure.

Fig. 5.22: Available power gain in dB for (a) a lossless 4 bit tunable PI network designed for $|\Gamma_{in}| \leq 1/3$, (b) a silicon implementation, (c) an implementation in GaAs

Three implementations are taken into account: a lossless implementation ($Q_c=Q_L \rightarrow \infty$), implementation in low ohmic silicon ($Q_c=20$, $Q_L=10$) and a GaAs implementation ($Q_c=60$ and $Q_L=35$). The quality factors are assumed to be independent from the component value and they include the switch resistor. Fig. 5.22 a-c show the contours for G_{ava} for the three forenamed situations; the outer contours demark the matchable region.

Clearly the lossless network is noiseless; whereas the noise figure can be as high as 4.4 dB in the silicon implementation and up to 1.5 dB in GaAs, for the 4-bit tunable PI network used as vehicle.

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Chapter 6

Conclusions

6.1 Summary and Conclusions

In **chapter 1**, an introduction to wireless communication was given. Antennas and RF front ends, the basic blocks in wireless communication, are analysed. It is explained how the antenna impedance (assumed to be 50 ohm) can be different from 50 ohm depending on the unpredictable electromagnetic environment. Since the electromagnetic environment can be varied in an unpredictable way, it is not possible to optimise the antenna or to design a static matching network. The effect of the unpredictable antenna impedance is then analysed on the RF front end. An RF front end is typically made of a switch or a diplexer, an RF power amplifier and a low noise amplifier. The RF power amplifier, or simply RF-PA, is used to drive the antenna and the low noise amplifier, or simply LNA, amplifies the weak signal coming from the antenna. In case of unpredictable antenna impedance, the RF power amplifier has to work under load mismatch conditions resulting in a lower maximum output power, lower efficiency, stability issues, compromised ruggedness, etc. Therefore there is the need for automatic antenna tuners in order to "tune" the antenna impedance to its optimum value (typically 50 ohm).

In **chapter 2**, automatic antenna tuners are introduced. They are comprised of sensors, a tunable matching network and a control network. Each of these is described (with most detail for the matching network), after which the state-of-art is discussed. Finally, the methodology of the work is described: to investigate theoretical limits and shows how they can be reached.

In **chapter 3**, a practical implementation of an automatic antenna tuner was proposed. The feature of this implementation is the measurement of only RF magnitudes requiring low frequency control networks. The control loop only requires low-frequency electronics which makes it low cost, low power and relatively easy to integrate. Measurements on a demonstrator system show correct behavior for VSWR up to 10.

In **chapter 4**, a theoretical derivation on the minimum number of states for switchable matching networks was analysed. The word "state" denotes the status (open

or closed) of each switch of the switchable components used inside the switchable matching network. Generally speaking the number of states is equal to 2^n where n is the number of the switches. The theoretical result is only dependent on the load VSWR and the required input VSWR without referring to a specific implementation. In other words no specific implementation was taken into account. The results were based on a mathematical derivation and not on brute-force simulations. Although some other authors have proposed a minimum-number-of-states concept as well, all these are based on numerical simulation as opposed to the analytical approach in this chapter. Configurations were analyzed and benchmarked: single-stage one-ring configuration, single-stage two-ring configuration, two-stage one-ring configuration and three-stage one-ring configuration showing that single-ring configurations are optimum. An extension towards the required number of states for lossy matching networks was also provided. The results can be used to optimise the design of switchable matching networks.

In **chapter 5**, the proof of concept of the theoretical result of chapter 4 was discussed. In order to compare each switchable matching network, a parameter was introduced: the hardwareoverhead. This parameter is defined as the ratio between the actual number of states and the theoretical minimum number of states given in chapter 4. Several switchable matching networks were analysed: PI networks, T networks, loaded transmission lines, branch line coupler based switchable matching networks, single circulator and cascaded circulator based switchable matching networks.

The loaded transmission line was proposed to achieve a lower number of states compared to classical matching networks like a PI matching network or a T matching network. In fact in order to achieve each optimum state, switchable capacitors were used that are equally spaced on a transmission line. The result is quite large number of states because not all the combinations are used and only a small number are really needed: in case of eight states on the ring (where 9 is the total number of states), eight switchable capacitors are needed but the overall number of states is $2^8 = 256$ resulting in an overdesigned switchable matching network. The hardwareoverhead is $256/9 = 28.44$.

According to the analysis carried out, the cascaded circulator based switchable matching networks achieve the closest value to the lowest possible number of states as defined in chapter 4. Referring to the example of nine states depicted above, the overall number of states for the cascaded circulator based switchable matching networks is 16 (16 times less than the loaded transmission line). The hardwareoverhead is $16/9 = 1.77$. This value is also the best value that can be achieved with binary switchable matching networks since 9 is not an integer power of 2. The main limitation of this matching network topology is the integration. Other topologies such as PI and T matching networks can be integrated.

6.2 Original Contributions

This thesis contains the following original contributions to the field:

- A topology of an automatic antenna tuner based only on measurements of RF magnitudes (chapter 3). This work was published as “An Automatic Antenna Tuning System Using Only RF Signal Amplitudes”, E. L. Firrao, Anne-Johan Annema, Member, IEEE, and B. Nauta, Fellow, IEEE on IEEE Transactions on Circuits and Systems — II: Express Briefs, Vol. 55, No. 9, September 2008.
- The derivation of the minimum number of states for switchable matching networks, depending on the maximum VSWR of the antenna and the matched PA, based on mathematical derivation (chapter 4). This work was published as “On the Minimum Number of States for Switchable Matching Networks”, Ettore Lorenzo Firrao, Anne-Johan Annema, Frank E. vanVliet, and Bram Nauta on IEEE Transactions on Circuits and Systems — I: Regular Papers, Vol. 62, No. 2, February 2015.
- The introduction of a Figure-of-Merit for switchable matching networks, based on the minimum number of states necessary (chapter 5). This work was published on “Hardware Implementation Overhead of Switchable Matching Networks”, Ettore Lorenzo Firrao, Anne-Johan Annema, Frank E. vanVliet, and Bram Nauta on IEEE Transactions on Circuits and Systems — I: Regular Papers, Vol. 64, No. 5, May 2017.
- A proof of concept of a switchable matching network with the minimally-needed number of states, based on a cascaded circulator implementation (chapter 5). This work was published on “Hardware Implementation Overhead of Switchable Matching Networks”, Ettore Lorenzo Firrao, Anne-Johan Annema, Frank E. vanVliet, and Bram Nauta on IEEE Transactions on Circuits and Systems — I: Regular Papers, Vol. 64, No. 5, May 2017.

6.3 Future Work

The topic of automatic antenna tuners is a wide topic. They are used to cope with unpredictable changes of the antenna impedance. In fact the performance of the RF front ends are affected by the antenna impedance. In case of antenna mismatch, the result is a degradation of the RF front end efficiency and/or an overdesign of the RF front end in order to take into account the antenna mismatch. There are two types of tunable matching networks: continuous tunable matching networks and switchable matching networks. The former uses continuous tunable components, the latter uses switches to insert in or out of the circuit the reactive elements.

In this thesis the work was focused on the optimization of the switchable matching network through a mathematical formalism that derives the minimum number of states based on the load VSWR and on the required input VSWR at a single frequency.

The findings in this thesis can be extended to:

- *On-chip adaptive matching networks*: the lumped element implementation could be extended to an IC version but the technology must be chosen carefully. In fact CMOS technology is well known for low Q capacitors and inductors resulting in a low efficiency of the switchable matching network. Other technologies could be chosen like GaAs, GaN and InP but they are more expensive. This is possible because silicon is a semi-conductor while GaAs, GaN and InP are semi-isolators.
- *Reconfigurable electronics*: generally speaking the work could be extended to any reconfigurable electronics such as RF power amplifiers. In fact in case an RF power amplifier is used over different frequency bands, there is the need to tune the input matching network and the output matching network to each frequency band. The reconfiguration can be extended to any device that is not possible to design with sufficient bandwidth. In this case, the minimum number of states and hardware overhead can provide guidance to the implementation.
- *Cell phones*: the switchable matching network proposed in chapter 5 can not be used in mobile phones because it is bulky and expensive. Only SMD components can be used or other technologies like PASSI from NXP Semiconductors (formerly Philips Semiconductors). The constraint of using only SMD components leads to a less optimized switchable matching network. The work carried out in chapter 5 on PI networks could be more investigated for cell phones. The PI network could be cascaded into a ladder network.
- *Load-pull measurement equipment*: the switchable matching network proposed in chapter 5 could be used in measurement equipments where size is not a big constrain and measurement time is an issue. The formalism could be used to choose a matching topology that allows to speed up the measurement time.

In chapter 4 and in chapter 5 a mathematical formalism was proposed to derive the minimum number of states based on the load VSWR and on the required input VSWR at a single frequency. The formalism was derived without referring to a specific implementation in order to keep the theory as general as possible. The work could be also extended to include signals with a finite bandwidth. In the end it could be possible to have a formalism that relates the minimum number of states to the required input

VSWR, to the maximum load VSWR and to the bandwidth. During the derivation, the Bode-Fano limit should be always kept in mind since it relates the maximum achievable bandwidth under a predefined load mismatch.

Appendix A

Antenna Behaviour in the Presence of Human Body (*)

(*) “Antenna Behaviour in the Presence of Human Body”, E. L. Firrao, A. J. Annema, B. Nauta, ProRisc 2004, Veldhoven, The Netherlands.

Abstract – Mobile phones are widely used nowadays. The demand in the new generations of mobile phones is better performance. Typically the mobile phone performance is derived assuming an ideal antenna impedance of $50\ \Omega$. Some work has been done in understanding the behaviour of the antenna impedance in a real environment [1], [2], [3]. This work continued the exploration of the difference between the ideal antenna impedance and the antenna impedance in a real environment. In particular the analysis was carried out measuring the antenna impedance in presence of the human body and relating it to the position of the hand and of the hand + head. Two types of antennas were considered: PIFA antennas and ceramic antennas. Several positions of the hand and of the hand + head were considered. These positions were selected after an accurate study about how people on the street normally handle their mobile phones. The study consisted of two steps: 1) observation of the typical positions of the hand and of the hand + head with respect to the mobile phone; 2) their imitation in the lab for the measurements. The outcome of the investigation was the changing of the impedance is primary due to a changing of its imaginary part. The real part is affected but not in dramatic way.

Keywords – antenna behaviour, antenna retuning, human body interaction, mobile phone performance.

A.1 Introduction

In the present days the use of mobile phones has become more and more popular. All the selling specifications, such as talk time, are derived assuming an ideal

behaviour of the antenna. During the design the typical approach is to assume the antenna impedance 50Ω . In reality, due to the presence of the user, the antenna impedance differs from 50Ω resulting in a degradation of performance (shorter talk time and/or call lost). The shorter talk time is the result of a lower efficiency of the RF power amplifier draining more power from the battery in order to keep the connection. The call lost is due to the impossibility of the RF power amplifier to transmit the required power level. As it is well-known both of them are a strong function of the load (antenna) impedance. In order to predict the real mobile phone performance and to improve it, it is relevant to understand the antenna behaviour in presence of human body. The analysis is addressed in this chapter. It is organised as follows. Section 2 starts with a brief analysis of the impact of the antenna impedance on a simplified RF front end. Then the antenna behaviour is described. The section ends giving a simplified model valid also in case of the presence of human body. Finally some conclusions are given in Section 3.

A.2 Analysis

A.2.1 Motivation

A simplified block diagram of an RF front end for dual band GSM mobile phones is sketched in Fig. 2.1. As mentioned in the introduction part, during the design the typical approach is to consider the antenna impedance equal to 50Ω . The importance of the antenna impedance can be evaluated deriving the behaviour of the RF power amplifier under different load conditions. This can be achieved in a general way using the scattering parameters [1]. In the particular case, when the optimum load impedance of the RF power amplifier is real, the scattering parameters can be reduced to normalised scattering parameters [1]. In this case

$$\Gamma_{RFA} = \Gamma_{antenna} * e^{-j2\theta} \quad (\text{A.1})$$

where θ is the phase shift between the output of the RF power amplifier and the antenna connector/input.

Fig. A.1: simplified block diagram of an RF front end for dual band GSM mobile phones.

In relationship with the particular RF power amplifier chosen the output power and the efficiency can be also derived [2], [3], [4]. In some situations with a VSWR = 2:1 a drop of 3 dB was recorded for the maximum output power. The efficiency was also half of the nominal value. The situation was even worse for higher VSWR.

A.2.2 Antenna samples

In the analysis two antennas widely used in mobile phones were considered: PIFA antennas and ceramic antennas. The two samples used in the measurements are shown in Fig. 2.2 and Fig. 2.3, respectively.

A.2.3 Antenna measurements

Several positions of the hand and of the hand + head were considered. These positions were found after an accurate investigation about how people normally hand their mobile phones. The most common six positions are summarised in table 1. For simplicity only the pictures for the PIFA antenna are sketched. The same positions were considered for the ceramic antenna and also with the presence of the hand + head. The

antenna performance was compared with a reference environment denoted as “free space” shown in Fig. 2.4. The measurements were performed using the HP 8510 Vector Network Analyser.

During the measurements the following procedure was adopted: measurements of the reflection coefficient at the samples connector and post-processing of the data in order to plot the reflection coefficient at the antenna connector (removing thus the piece of transmission line on the PCB). The RF properties of the transmission lines were found using the physical dimensions and the electrical properties of the PCB. The outcome of the investigation is summarised in Fig. 2.5, Fig. 2.6 and Fig. 2.7. Fig. 2.5 and Fig. 2.6 are about the VSWR vs frequency. These two figures clearly show the detuning effect because of the presence of the human body. Fig. 2.7 describes the area covered on a Smith chart by the antenna impedance of the PIFA antenna and of the ceramic antenna in GSM 900 and in GSM 1800 bands. The overall ranges for the real and imaginary part of the antenna impedance were:

$$R_{ant} \in [30, 100] \Omega \quad (2.2)$$

$$X_{ant} \in [0, 100] \Omega \quad (2.3)$$

Fig. A.2: PIFA antenna demo used during the measurements.

Fig. A.3: ceramic antenna demo used during the measurements.

Fig. A.4: “free space” environment for the antenna measurements

Table 1: positions of the hand

hand in position 1	hand in position 2

hand in position 3	hand in position 4

hand in position 5	hand in position 6

Fig. A.5: VSWR vs frequency for the PIFA antenna in GSM 900 band.

Fig. A.6: VSWR vs frequency for the ceramic antenna in GSM 1800 band.

Fig. A.7: overall area covered on a Smith chart by PIFA and ceramic antennas in presence of the human body.

A.2.4 Antenna model

A typical simplified circuit model for describing the antenna impedance behaviour around the resonant frequency is an RLC series or parallel tank. For the two investigated antennas a proper model is a series RLC tank [8]. The investigation showed that the same model can be extended to the antenna in presence of the human body. The only difference between the free space model and this model is a new set of parameters (new L, C and R values).

A.3 Conclusions

The antenna behaviour and the RF front end performance were analysed. The conventional approach of using an ideal load is not correct to predict the overall mobile phone performance in presence of the user/human body. In fact this investigation shows that the antenna impedance can significantly differ from the idea value (50Ω). The outcome of the investigation was the changing of the impedance is primary due to a changing of its imaginary part. The real part is affected but not in dramatic way.

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