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Wideband Active and Passive Antenna Solutions for Handheld Terminals

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Abstract

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This thesis presents solutions and studies related to the design of wideband antennas for wireless handheld terminal applications. A method of electrically shortening the terminal chassis length to obtain resonance at high frequencies has been proposed and evaluated, thereby increasing the antennas impedance bandwidth. No significant effect on the lower frequency band in a dual-band antenna prototype has been observed, making the method suitable for multi-band applications. The chassis has further been utilized as a zero-thickness 0.9 - 2.7 GHz high efficiency antenna by inserting a notch in the chassis center, and a feasibility study for typical phones has been performed. Additionally, the effect of talk position on the chassis wave-mode has been investigated, where the standard equivalent circuit model for terminal antennas has been modified to include the presence of the users head. The model has been used to explain measured and simulated effects concerning frequency detuning, efficiency reduction and bandwidth enhancements when the terminal is placed in talk position.

The use of a hands-free earpiece cord is currently mandatory for FM radio reception as the cord is utilized as antenna. However, there is currently a market driven demand for removing the cord requirement since many modern phones are equipped with speakers and Bluetooth headsets. In this thesis, an active ferrite loop antenna is proposed as an internal replacement/complement with a performance of -23 dB (G/T degradation) compared to a full-size lossless dipole in urban environments. Also, a modification to the cord is suggested for DVB H reception.

Complex matching networks have been investigated to increase the bandwidth of dual band PIFA antennas, and a printed dual band dipole has been integrated with a modified Marchand balun for dual resonance at two separate frequency bands, thus covering the commercial cellular bands 824-960 and 1710-2170 MHz with a single antenna.

Keywords: active antennas, impedance matching, dipole antennas, baluns, slot antennas, microstrip antennas, mobile antennas, multifrequency antennas, antenna proximity factors, receiving antennas, ferrite devices

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To Victoria

Appended Papers

This thesis is based on the following papers, which are referred to in the text by their Roman numerals.

I	**P. Lindberg** and E. Öjefors, "A Bandwidth Enhancement Technique for Mobile Handset Antennas using Wavetraps," *IEEE Transactions on Antennas and Propagation*, 54(8):2226-2233.
II	**P. Lindberg** and A. Kaikkonen, "Earpiece cord antenna for DVB-H reception in wireless terminals," *IEE Electronics Letters*, 42(11):609-611.
III	**P. Lindberg**, E. Öjefors and A. Rydberg, "Wideband Slot Antenna for Low-Profile Hand-held Terminal Applications," *Proc. of the 36th European Microwave Conference, Manchester, UK, Oct. 2006*.
IV	**P. Lindberg**, E. Öjefors, Z. Barna, A. Thornell-Pers and A. Rydberg, "A Dual Wideband Printed Dipole Antenna with Integrated Balun," *Accepted for publication in IEE Microwaves, Antennas and Propagation*.
V	**P. Lindberg**, M. Sengul, E. Cimen, B. S. Yarman, A. Rydberg and A. Aksen, "A Single Matching Network Design for a Dual Band PIFA Antenna Via Simplified Real Frequency Technique," *Proc. of the 1st European Conference on Antennas and Propagation, Nice, France, Nov. 2006*.
VI	**P. Lindberg** and A. Kaikkonen, "An Internal Active Antenna for FM Radio Reception in Mobile Handsets," *Submitted to IEEE Transactions on Antennas and Propagation*.
VII	**P. Lindberg**, E. Öjefors and A. Rydberg, "The Effect of Talk Position on The Chassis Mode of 900 MHz Terminal Antennas," *Submitted to IEEE Transactions on Electromagnetic Compatibility*.

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Comments on the author's contribution to the publications

I	Idea, implementation, measurements (except SAR) and writing the manuscript.
II	Idea, implementation, measurements and writing the manuscript.
III	Idea, implementation, measurements and writing the manuscript.
IV	Idea, implementation, measurements and writing the manuscript.
V	Idea, implementation (except optimization of matching network), measurements and writing major parts of the manuscript.
VI	Idea, implementation, measurements and writing the manuscript.
VII	Idea, implementation, measurements and writing the manuscript.

Related Papers

The following papers by the author have not been included as they are not within the scope of the thesis.

VIII **P. Lindberg**, E. Öjefors and A. Rydberg, "A SiGe 24 GHz zero-IF downconverter", *Proc. of GigaHertz 2003*, Nov. 2003, Linköping, Sweden.

IX **P. Lindberg**, E. Öjefors and A. Rydberg, "A SiGe HBT 24 GHz Sub-Harmonic Direct-Conversion IQ-Demodulator," *Proc. of IEEE Topical Meeting on Silicon Monolithic Integrated Circuits in RF systems*, Atlanta, USA, Sept. 2004.

X P. Hallbjörner, M. Bergström, M. Boman, **P. Lindberg**, E. Öjefors and A. Rydberg, "Millimeter-Wave Switched Beam Antenna Using Multiple Traveling Wave Patch Arrays", *IEE Proc. Microwaves, Antennas & Propagation*, vol. 152, no.6, 2005.

XI N. Unno, T. Fujita1, A. Oncu, **P. Lindberg** and B. S. Yarman, "Dual Band Equalizer Realized by Utilizing 0.18μm Si-Processing Technology for a PIFA-900 Antenna", *Submitted to ISACS 2007*.

XII H. Schumacher, P. Abele, J. Berntgen, K. Grenier, J. Lenkkeri, **P. Lindberg**, E. Öjefors, R. Plana, W.-J. Rabe, A. Rydberg, E. Sönmez, and K. Wallin "Compact, low-cost 24 GHz modules using micromachined Si/SiGe HBT technology", *Proc. of the IST Mobile and Wireless Communications Summit*, Lyon, France, June 2004.

XIII M. Boman, M. Bergström, P. Hallbjörner, **P. Lindberg**, E. Öjefors, A. Rydberg "Analysis and Verification of Steerable Microstrip Antenna Concept for Millimeter-Wave Applications", *Proc. of GigaHertz 2005*, Uppsala, Sweden, Nov. 2005.

XIV A. Rydberg, **P. Lindberg** and E. Öjefors, "Towards MEMS-based mm-Wave Radar", *Proc. of Radiovetenskaplig Konferens 05*, Linköping, Sweden, June 2005.

XV **P. Lindberg**, E. Öjefors and A. Rydberg, "A 24 GHz on-chip differential Wilkinson coupler using lumped components", *Proc. of GigaHertz 2005*, Uppsala, Sweden, Nov. 2005.

XVI **P. Lindberg**, E. Öjefors and A. Rydberg, "A SiGe 24 GHz monolithically integrated direct conversion receiver", *Proc. of GigaHertz 2005*, Uppsala, Sweden, Nov. 2005..

XVII **P. Lindberg**, E. Öjefors and A. Rydberg, "LTCC packaging for a 26 GHz SiGe receiver", *Proc. of GigaHertz 2005*, Uppsala, Sweden, Nov. 2005.

XVIII J. Lenkkeri, E. Juntunen, **P. Lindberg**, E. Öjefors and A. Rydberg, "Packaging Issues of 24 GHz Traffic Control Radar Front End Based on SiGe and LTCC Technologies", *Proc. of IMAPS Nordic Conference*, Tönsberg, Norway, Sept. 2005.

XIX E. Öjefors, E. Sönmez, S. Chartier, **P. Lindberg**, A. Rydberg and H. Schumacher, "Monolithic Integration of an Antenna with a 24 GHz Image-Rejcetion Receiver in SiGe HBT Technology", *Proc. of the 35th European Microwave Conference*, Paris, France, Oct. 2005.

XX E. Öjefors, E. Sönmez, S. Chartier, **P. Lindberg**, C. Schick, A. Rydberg and H. Schumacher, "Monolithic Integration of a Folded Dipole Antenna with a 24 GHz Receiver in SiGe HBT Technology", Submitted to *IEEE Transactions on Microwave Theory and Techniques*, 2006.

XXI K. Grenier, L. Mazenq, D. Dubuc, F. Bouchriha, F. Cocetti, E. Öjefors, **P. Lindberg**, A. Rydberg, J. Berntgen, W.J. Rabe, E. Sonmez, P. Abele, H. Schumacher and R. Plana, "IC Compatible MEMS Technologies", *Proc. of the 35th European Microwave Conference*, Paris, France, Oct. 2005.

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XXIII **P. Lindberg**, A. Rydberg and E. Öjefors, "Micromechanically Enhanced Integrated Radar Front-Ends", *Proc. of the International Conference on Electromagnetics in Advanced Applications*, Torino, Italy, Sept. 2005.

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XXV A. Rydberg, **P. Lindberg** and E. Öjefors, "Towards small MEMS based wireless sensors", Submitted to *Smart Systems Integration 2007*, March 2007, Paris, France.

Contents

1. Introduction

1.1 Background

The research field of handheld terminal antennas has witnessed a remarkable evolution during the last decade. Less than 10 years ago mobile phones were used exclusively for voice communication, utilizing a single wireless system (e.g. GSM) and a single frequency band (e.g. 900 MHz). The terminal antenna was external; either a retractable rod, or to make it smaller, folded together into a coil/helix. In the mid 90's, frequency bands at 1800 MHz in Europe and 1900 MHz in the US were allocated to increase the network capacity, thus creating a need for antennas supporting two separate frequency bands, so called dual-band antennas. The first phone[1] with this feature appeared on the market in late 1997. 1998 saw the first internal antenna (single-band however) in a commercial phone[2], a concept that had been proposed by Haruki already in 1982 [2]. These antennas were non-obtrusive and thus retained the aesthetics and increased the mechanical ruggedness of the phone. Dual-band internal solutions was first presented in the literature in 1997 [3] with commercial phones using the technique available in mid-1999[3]. A couple of years later, the consumer requirement of global roaming (e.g. Europeans wanting to use their mobile phones in the US and vice versa) led to the development of triple band phones, featuring triple band antennas (900/1800/1900 MHz). Around the same time, Bluetooth modules, with a separate internal antenna, and FM radio receivers, using the earpiece cord as antenna, started to become standard features in phones. In early 2003[4], a new cellular network called 3G (or UMTS/WCDMA), using a separate frequency band, started world-wide deployment. In about 5 years time, wireless terminals went from single system/single band to multiple systems and multiple bands.

Today, in 2006, phones in the mid- to high price range supports five different cellular bands (GSM850/900/1800/1900 + 3G), Wireless LAN, Bluetooth, Digital TV (DVB-H), FM radio and GPS. In the next few years, several new wireless systems such as RF-ID, UWB, WiMAX etc. will probably also be integrated in the terminal. Meanwhile, some of the "old"

[1] Motorola MicroTAC 8900, using a combined helical-monopole external antenna [1].
[2] Nokia 8810, using a PIFA antenna.
[3] Nokia 8210, using a PIFA antenna.
[4] 2001 in Japan.

systems will most likely gain new frequency bands, such as WCDMA at 2.6 GHz and WLAN at 5 GHz. Additionally, there is an interest in adding diversity capability for some of the systems (requiring extra receive antennas). Several of these antennas are by themselves exceedingly difficult to implement, and the close integration of all the different radiators leads to mutual EM coupling which further complicates the design phase.
As more features have been added to each new generation of wireless terminals, the size of the phones has become progressively smaller. While this trend of size reduction, as demanded by the consumers, was made possible by advancements in battery technology, LCD displays and low-power/high integration circuit technology, the antennas are not as prone to size reduction since their performance (in most respects) is related to their occupied volume by laws of physics. If the volume is reduced, some penalty in performance must be paid. Additionally, during the past few years, an awareness of the possible health effects from mobile phone radiation has led to regulations [4] of the maximum local power densities induced in the users tissue by the mobile phones. Methods to avoid power being radiated in the direction towards the users head is thus needed, with the added benefit of increasing the total antenna efficiency. Also, the multiple possible positions of the user placing the phone in his/her hand should be considered as it will have an impact on the real-life antenna performance. Finally, modern mobile phones comes in a variety of form factors, e.g. bar, swivel, foldable/clamshell, flip, slide etc., each with its particular impact on the antenna design. In short, the professional life of terminal antenna designers has become a lot harder during the past few years. Thus, research in the terminal antenna field to solve the ever increasing amount of problems is more important now than ever.

The following chapters will introduce each of the problem aspects of terminal antenna design together with, where appropriate, typical solutions. These chapters hence forms the context for the appended papers, which primarily presents novel (and more importantly - realistic) solutions to some of the above described problems. Additionally, relevant complementary material that was left out in the appended papers due to space restriction and an errata list is provided here.

It should be noted that although most (if not all) of the examples and papers presented in this thesis addresses mobile phones, the problems and solutions also typically apply to other types of terminals such as PDAs, gaming consoles, laptops etc. Hence, the term 'mobile phone' is mainly used in a broad sense as a synonym to the more general terms 'wireless terminal' or 'handheld terminal'. Furthermore, throughout the thesis, "antenna" usually denotes the complete radiating structure, which typically includes the phone chassis. The part of the phone that is designed by the antenna engineer (e.g. the PIFA, monopole etc.), is referred to as the "radiator".

1.2 Practical limitations of antennas for commercial wireless terminals

The fundamental limitations of electrically small antennas[5] in terms of bandwidth and radiation efficiency were probably first examined by Wheeler in 1947 [5]. A year later, Chu derived an approximate lower limit for the achievable radiation Q [11], which was corrected by McLean in 1996 [12]. Harrington [13] showed the effect of antenna size on gain in 1959. In addition to these fundamental restrictions imposed by physics, a terminal antenna designer also has to respect limits that comes from practical considerations, as the antenna is far from the only subcomponent necessary for making the communication system work. These practical limits, and their consequences, are briefly outlined here (with a thorough treatment of the more significant aspects in the following chapter):

System aspects All mobile phones are built around a multi-layered PCB[6] on which the subcomponents (such as transceivers, DSPs, displays, battery etc.) are mounted. At least one of the layers of this PCB is completely metallized to act as a ground plane for the system. All currently used RF modules have unbalanced I/O ports with the ground plane as a reference terminal, implying that the antennas should also be implemented in an unbalanced configuration [14]. The all-pervading 50 Ω system impedance is exclusively used.

Available Volume Mobile phones are constantly getting smaller. While the length and width of the terminal must be of a certain size to facilitate a wide LCD screen and a conveniently large key pad, the thickness of the phone has no such restriction. The latest trend in terminal design is therefore ultra-thin phones, leading to very small heights above ground plane available to the antenna elements. This has a huge impact on patch type of antennas (such as the popular planar inverted F antenna, PIFA) as the achievable bandwidth and radiation efficiency is proportional to this height [15]. In addition, many modern phones shares the antenna

[5] It should be noted that most mobile phone antennas are not electrically short in a strict sense as all widely accepted definitions requires the antenna to be at least smaller than the radian sphere (introduced by Wheeler in 1947 [5] and further treated in 1959 [6]), i.e. a sphere of radius $a = 1/k$ where k is the wave number, which means that the largest dimension of the antenna must be smaller than $2/k = \lambda/\pi$. Since a typical phone is about 100 mm long, and as the complete phone can potentially function as an antenna, only antennas working at frequencies below 960 MHz, or radiators with no coupling to the chassis, should be considered as electrically small. Other proposed definitions in common text-books have smaller boundaries than Wheeler, e.g. $\lambda/10$ in Balanis [7], $\lambda/8$ in Schelkunoff & Friis [8], $\lambda/4$ in Hirasawa [9] etc. An excellent discussion on the topic is given by Best [10], who suggests a limit half of that of the radian-sphere.

[6] For some form factors, such as swivel, slide and clamshell types, there are in fact two PCBs (one for each part of the phone), but as the two grounds are galvanically connected together the effect is essentially the same.

volume with the the speakers resonance box. While this is an efficient use of available volume it means that the antenna and speaker must be co-designed with the effect of increasing the complexity of both.

Chassis The metallized layer of the PCB, possibly together with other metallized parts of the chassis, functions as a ground plane for the various subcomponents in the terminal. Therefore, the antenna designer typically has no mandate to make major alterations to the chassis to suit his/her needs, rendering e.g. PBG-surfaces [16] and slotted grounds [17][18] unpractical. Recently however, radiators using "ground clearance", i.e. the radiator is positioned above a PCB section without metallization, have been increasingly more common in commercial phones with certain form factors. This is perhaps a first step towards more, for the antenna performance, customized chassis layouts.

Current consumption/Linearity Active antennas, i.e. with electronics (diodes or transistors) integrated together with the radiating element, consumes DC current thus reducing battery life time. This has so far limited widespread deployment in commercial phones. Also, all active components are intrinsically non-linear (even more so at low bias currents) which limits the use in high power Tx systems or systems co-located with high power transmitters. A more practical problem is that the active devices in many cases (such as switches) needs control voltages, implying that the antenna cannot be designed independently from the front-end module.

Price As in any consumer product, price is one of the most important parameters for a mobile phone antenna, thus excluding elaborate production techniques from being viable. Currently, the preferred choice for most radiators is single sided flex-films glued to a plastic carrier, both for internal and external radiators. This means that only 2D metal structures are possible to implement. Since the plastic carrier can be somewhat irregular in shape and the film can be folded around edges, there is however a certain amount of freedom for slightly more complex structures. An interesting technology called LDS (Laser Direct Structuring) has recently become popular, making dual layer metallized plastic carriers (with via holes) possible.

"Exotic" materials such as ceramics and ferrites are preferably avoided for cost reasons, as are active components integrated with the antenna and/or discrete matching components.

Weight One of the key figure of merits for a modern mobile phone is the weight. Hence, the antenna must be light. This is typically accomplished by using hollow plastic carriers and avoiding the use of ceramics and ferrites.

SAR The main limitation of PIFA antennas is the limited bandwidth for low profile phones, and "new" types of antennas, such as monopoles and slots, have been suggested as a remedy. However, most of the non-PIFA/patch type of radiators suffers from excessive specific absorption ratio (SAR) values for high power ($\geq$ 1 W) wireless systems such as GSM850/900. One way of reducing the negative impact on SAR is to place the radiator near the bottom of the phone (as opposed to the much more common placement at the top) which is naturally farther away from the head in talk position. This unfortunately means that the DC connector and external signal interface must be relocated to the side or top of the phone, which is considered unesthetic by many consumers.

Complexity A complete mobile phone is an extraordinarily complex system from an EM point of view. The impact of the battery, display, keypad, various plastics etc. can have a huge impact on the total antenna performance [19]. Perhaps more importantly, it makes modelling of the antenna excessively difficult and therefore nearly impossible to optimize using software. To retain some consistency, the RF community has agreed on using a simple PCB, typically of size 100 x 40 mm^2 using FR-4, for early design evaluation. A very significant part of the scientific literature on terminal antenna design is concerned with synthesization of elaborate radiator shapes to achieve a certain performance. Almost exclusively these radiators are optimized for and evaluated using a naked PCB. However, when implemented in a real phone, these designs will most likely have some change in characteristics. Therefore, for such papers, some means of making simple adjustments to compensate for these effects are necessary to make the proposed design practically useful. This is typically not included in the papers. It should be noted though, that while the immediate practical value is limited, these proposed designs are important for estimating what performance can be theoretically achieved with a certain type of radiator.

2. Challenges in mobile phone antenna development

The design of terminal antennas is in many respects significantly different to the design of antennas for the vast majority of other applications. This is partly related to the smallness of the antennas, which limits the achievable performance and complicates the measurements, with the complex near-field environment of the antenna element, and with the fact that the antenna, in its final implementation, is fully integrated with the front-end. This chapter will discuss these particular aspects of terminal antenna development.

The end objective of any commercial terminal antenna design is the fulfillment of some given specification, typically provided by the terminal manufacturer and/or the network operators (for the case of mobile phones). Additionally, there are government regulations setting constraints on e.g. SAR levels[1]. The specification is typically mainly concerned with electrical properties - bandwidth, efficiency etc., and mechanical properties - size, placement, interface etc. However, from a purely commercial point of view, other qualities are (at least) equally important - low cost, low weight, ease of production and verification (as discussed in Section 1.2) that introduces practical boundaries on the design options. As in the case of most mass-market electronics, a cheap solution with sufficient performance is often preferred over a slightly more expensive solution with much better performance. In particular, this is true for sub-components (such as the antenna) where the performance is not directly noticed by the end-user (i.e. buyer) and which can be compensated for at the system level by e.g. increasing the number of base stations. For the service provider on the other hand, e.g. cellular network operators, the communication capability of the terminal (and hence the antenna) is highly important to relax requirements on base station density. Therefore, it is not surprising that the toughest specifications on the mobile phone antenna are given by the (US) network operators. This chapter is concerned mainly with fundamental challenges related to the electrical properties, though keeping in mind the more practical/commercial aspects.

[1] In many cases though, mobile phone manufacturers have tougher requirements than those set by the regulations.

2.1 Characterization

During the design phase, the antenna is characterized using a coaxial measurement cable (so called passive measurements) connected to the antenna mounted in a preliminary terminal prototype/mock-up. Standard measured characteristics are impedance, antenna efficiency, bandwidth and SAR. Directivity, radiation pattern and polarization, which are three of the most important properties for large antennas, are seldom of interest for terminal applications. Unlike other components such as filters and amplifiers, the antenna is highly sensitive to its nearby environment. Hence, any alteration to some plastic parts or similar during the development of the terminal will require a re-tuning of the antenna. Throughout the development phase of the terminal, minor changes are frequent, leading to many antenna design iterations. For this reason, it is highly important that the implemented antenna is designed for tunability and that the antenna engineer fully understands the design to facilitate rapid retuning. Perhaps this is one of the reasons why commercial antennas are almost exclusively based on fairly conservative design approaches (e.g. PIFAs).

For the final verification measurements, the antenna is implemented in the complete terminal and operated by the transceiver. As the system is fully integrated, it is no longer possible to measure qualities such as bandwidth, efficiency etc. Instead, the antenna is evaluated from over-the-air (OTA) measurements (also called active measurements) [20]:

TIS - *T*otal *I*sotropic *S*ensitivity (Receive mode)
TRP - *T*otal *R*adiated *P*ower (Transmit mode)
SAR - *S*pecific *A*bsorption *R*ate (Transmit mode)
HAC - *H*earing *A*id *C*ompatibility (Transmit mode)

The RF transmitting performance of the complete front-end is determined by the "Total Radiated Power" in an anechoic (and possibly also in a reverberation chamber) by measuring the power density transmitted by the terminal over all angles, with the phone set to operate in its highest power class. TRP is recorded as the largest detected power at any angle (similar to gain being defined as the peak value obtained at any angle). For terminals fitted with external coaxial connectors, it is possible to compare the measured radiated output power to the power available from the transmitter and thus deduce the antenna performance. This is however not necessarily the same performance as measured in passive mode as the transmitter may not operate equally well for all load impedances within the return loss limits (in contrast to a theoretical 50 Ω source). Some specifications only considers the talk-position mode, typically accepting more than 10 dB total antenna loss (for GSM850/900, and around 5-6 dB for GSM 1800/1900), other considers only free-space and yet others specifies both. Also, as TRP is a peak value (in contrast to antenna efficiency in passive measurements, which is an average quantity) it is principally possible to obtain good values by having a narrow beam at some angle. In practice however, terminal antennas always have fairly isotropic radiation patterns.

The receiver performance "Total Isotropic Sensitivity" is measured by lowering the power level from a base station simulator until a specified bit error rate (BER) is reported by the terminal. Again, this might not be identical to the results from passive measurements as the total noise figure of the receiver depends on the source (i.e. antenna) impedance and not only the return loss (which is given in specification). A normal receiver sensitivity in GSM850 is -108 dBm, and a typical specification for TIS is around 10 dB higher in talk position (consistent with the transmit requirement). During measurements in an anechoic chamber, the base station first transmits a signal at a fixed power level and the reported received power by the phone is recorded at all angles and at specified frequencies. For the angle with the highest power level, a specified bit sequence is repeatedly transmitted by the base station with reduced output power for each iteration. The received signal is retransmitted by the phone to the base station and compared to the original sequence, and TIS is recorded at the power level that corresponds to a BER of 2.44% for GSM and 0.1% for WCDMA.

SAR is only relevant to high-power systems such as GSM850/900, and requires a fairly elaborate measurement set-up as described in Section 2.4.1. Since mobile phones sold to the general public in the US needs to comply with FCC (Federal Communication Commission) regulations on SAR (i.e. less than 1.6 W/kg peak value), in some cases, it is the SAR that limits the Tx power rather than the maximum output power that the power amplifier (PA) can deliver [21].

HAC measurements are, in the context of antenna development, conducted to classify a mobile phone into different categories of hearing aid compatibility in a scale from M1 to M4, where a category M4 phone is more likely to work with a given hearing aid than a category M1 phone. A M3 grading meets the standard, and M4 exceeds the standard. Note that the scale is relative, it is not guaranteed that even a M4 phone will work properly with a given hearing aid. The measurement procedure is similar to that of SAR and is typically conducted using an extended SAR measurement set-up. In principle, a probe measures the E-field and H-field 1 cm above the phone (on the speaker side) in a specified area and these values are then compared to standard values specified by FCC, which determines the compatibility category. The exact classification procedure though is somewhat more elaborate, for instance high peak values can be removed from the measurement if they are localized (since the user can compensate by simply shifting position slightly).

As the result of these measurements, as previously mentioned, also depends on other components such as filters, switches, the power amplifier (TRP, SAR and HAC) and the low-noise amplifier (TIS), the system level performance can be quite different compared to what could be expected from the passive measurements, even though each component complies with the specification. This is easily understood by noting that the standard specification on the antenna input impedance accepts the full 17 - 150 Ω range, while the filter, switch and power amplifier (PA) designs all assumes a fix 50 Ω load. In particular, the output power of the PA is strongly dependant on the load impedance (also at harmonic frequencies) and can therefore vary greatly over the frequency band (due to impedance variations). Hence, as a final part of the antenna design phase, re-tuning based on active measurement results are sometimes needed. Needles to say, this re-tuning can be very difficult and time consuming to do.

The US-based Cellular Telecommunications and Internet Association (CTIA) is currently the only organization that produces OTA performance tests. In Europe, ETSI and 3GPP have produced a range of standards, but have yet to publish OTA specifications. Also, the activities in COST 273 is likely to result in OTA specifications but there is currently no known schedule for such publication.

2.2 Crosstalk of Multiple Antennas

With the continuous introduction of new cellular and complementary (WLAN, DVB-H) wireless systems, the number of antennas per handset has been increasing rapidly. The main problem of co-locating several antennas in a confined volume, besides reducing the available space for each antenna and hence bandwidth (see Section 2.3), is the mutual coupling, or crosstalk, between all elements, inevitably leading to an increased design complexity. The design of a single terminal antenna for modern mobile terminals is often extremely challenging, in particular for wideband and/or multiband applications, and taking mutual effects of other elements into account can be very difficult and time consuming. In practice, the process is often iterative: first design one antenna, then design the other antenna taking into account the presence of the first antenna. Then the first antenna is retuned to take the presence of the second antenna into account and so on. It should be noted that it is not clear in the design phase how all non-active (i.e. those currently not being measured) antennas should be terminated to simulate a real world situation. The cellular antenna, for example, will have different terminating impedances depending on if it is in transmitt or receive mode (with either the PA or LNA switched on).

From a system level point of view, crosstalk is a problem in particular for systems that are active simultaneously, such as when a user is talking through a Bluetooth headset using GSM, and is most severe for low-frequency systems that radiates mainly through the chassis mode (i.e. below say 1.5 GHz). Typical problems includes

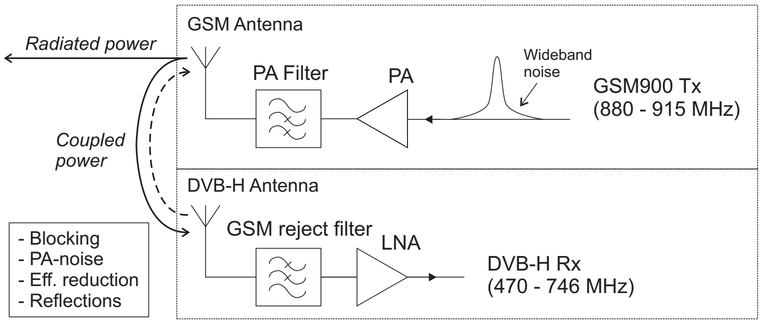


Figure 2.1: Crosstalk between transmitting GSM antenna and receiving DVB-H antenna.

- Reduced efficiency due to power being leaked into other systems instead of being radiated.
- High power signals from transmitters can saturate receivers of other systems, so called "blocking".
- High output noise levels from transmitters can degrade the receiver sensitivity of other systems.

Naturally, the problem is more difficult the closer the systems are in frequency, where currently the most challenging inter-operability issue is between low-band GSM and DVB-H, as seen in Fig. 2.1. Crosstalk can be reduced by using balanced radiators (such as dipoles or loops) with no RF connection to the chassis [22]. The presence of the metallized chassis will however reduce the radiation resistance of the radiators, similar to the case of having a dipole antenna too close to a reflector or ground plane, thus reducing the achievable bandwidth. Also, all currently available transceivers have single-ended I/O interfaces, meaning that an expensive and lossy balun is required to drive the balanced antenna [14]. Mutual coupling can also be reduced by careful orientation of the radiating elements [23], by cutting band-notch $\lambda/4$ slots in the ground plane [24], or by introducing an inductive link between two antennas [25].

2.3 Bandwidth

The single most challenging task in terminal antenna design is the attainment of a sufficient impedance bandwidth without sacrificing efficiency. Standard requirements for mobile phone antennas today is a return loss of <-6 dB within the frequency band. To compensate for manufacturing tolerances, some extra margin is typically required in the design phase (e.g. <-7 dB return loss at the band edges).

Currently, the most difficult systems are the low frequency bands GSM 850/900 and DVB-H, where even the terminal length (which could potentially be used as part of the antenna) is shorter than half of the operating wavelength. The bandwidth of the mobile antenna is for such systems theoretically limited by the resonance frequency and the radiation quality factor of the chassis, and in practice by the amount of achievable coupling from the radiating element to the chassis. The fundamental limit for electrically small antennas was first presented in 1948 by Chu [11] (and later corrected by McLean [12]), who derived an approximate lower limit for the achievable radiation quality factor Q_r. The corrected formula (usually called 'the Chu limit' or 'Chu-Harrington' limit), which relates the bandwidth to the physical size, is given as

$$Q_r = \frac{1}{(ka)^3} + \frac{1}{(ka)} \tag{2.1}$$

where $k = (2\pi)/\lambda_0$ is the wave number in free-space and a is the radius of the smallest sphere in which the antenna could be fitted. Eq. 2.1 is only valid for single resonant structures exciting a single (lowest) mode (TE or TM); when both the lowest TE and TM modes are excited (circular polarization) a modified expression is given in [12].

The radiation quality factor relates the stored electric or magnetic energy W to the radiated power P_{rad} through $Q_r = \frac{\omega W}{P_{rad}}$, where ω is the angular frequency $\omega = 2\pi f$. Q_r only includes power dissipation through radiation (there is probably no fundamental limit for lossy antennas). Conductive losses P_c and dielectric losses P_d can be added to the radiated power to obtain the total power loss P_{tot}, and since ω and W is fixed, an unloaded quality factor Q_0 can be defined as

$$Q_0 = \frac{1}{Q_r} + \frac{1}{Q_c} + \frac{1}{Q_d} = \eta \left(\frac{1}{(ka)^3} + \frac{1}{(ka)} \right) \tag{2.2}$$

where Q_c and Q_d are the conductive and dielectric quality factors, respectively, and η is the radiation efficiency. Material losses are more commonly expressed using dielectric ($\tan\delta_\varepsilon$) and magnetic ($\tan\delta_\mu$) loss tangents, which relates to their respective quality factors as [26]

$$\frac{1}{Q_d} = \tan\delta_\varepsilon \quad \text{and} \quad \frac{1}{Q_m} = \tan\delta_\mu \tag{2.3}$$

It should be noted that the quality factor rather than the bandwidth is the fundamental property of the small antenna (or any resonator). The realized (fractional) impedance bandwidth $B = (f_2 - f_1)/f_r$ of the complete antenna on the other hand is also related to (besides Q_0) the source/system impedance through [27]

$$B = \frac{1}{Q_0}\sqrt{\frac{(TS-1)(S-T)}{S}} \tag{2.4}$$

where

- f_1 is the lower frequency limit
- f_2 is the upper frequency limit
- f_r is the resonant frequency
- S is the bandwidth criterion $VSWR \leq S$
- T is the coupling coefficient, $T = Z_0/R_0$ for a series resonance and $T = R_0/Z_0$ for a parallel resonance
- R_0 is the input resistance at resonance
- Z_0 is the characteristic impedance

Maximum bandwidth, for a certain return loss (or VSWR) criterion, can be achieved by selecting the coupling coefficient T for so called "optimal overcoupling" (instead of a perfect match at the resonance frequency, called "critical coupling"). The value of T_{opt} for optimal overcoupling is obtained by finding the zero of dB/dT from Eq. 2.4 with the solution

$$T_{opt} = \frac{1}{2}\left(S + \frac{1}{S}\right) \tag{2.5}$$

For maximum bandwidth in typical mobile phone applications, $S = 3$ (equivalent to a return loss of -6 dB) results in an optimum input series resistance of R_0=30 Ω in a Z_0=50 Ω system. Entering T_{opt} from Eq. 2.5 into Eq. 2.4 gives the maximum achievable bandwidth for a single resonator without matching network [28]:

$$B_{max} = \frac{S^2 - 1}{2Q_0 S} \tag{2.6}$$

Compared to critical coupling, i.e. $R_0 = Z_0$, the bandwidth improvement for -6 dB return loss can be calculated using Eq. 2.6 and 2.4 (with $T = 1$), giving a 15% increase. Naturally, the increase in bandwidth compared to using critical coupling is larger for higher values of S (or return loss).

The total input impedance of the antenna close to resonance is related to the quality factor as [27]

$$Z_{in} = R_0(1 + jQv) \quad \text{or} \quad Z_{in} = \frac{R_0}{1 + jQv} \quad \text{with} \quad v = \frac{f}{f_r} - \frac{f_r}{f} \tag{2.7}$$

for the series (left) and parallel (right) resonant cases, respectively. Using Eq. 2.7, the quality factor can be extracted from measured or simulated impedance data by curve-fitting. This method however assumes a specified feeding structure and is thus not applicable to analyzing e.g. the Q of coupled resonators such as the chassis (which is typically excited by e.g. a high-Q PIFA element). A method that circumvents the need for port excitation is the Characteristic Mode Theory, where a discrete set of orthogonal current densities (*characteristic currents* or *eigencurrents*) are given as eigenvalue solutions of the

Z-matrix eigenvalue problem for the modes on conducting bodies [29][30]. The Z-matrix, or generalized impedance matrix, is given by direct application of the standard Method of Moments. While the modes are dependent of the body's geometry only, the relative excitation of each mode can be controlled by careful selection of feed or radiator type and placement. Unfortunately, unlike cavity type of resonators for which eigenmode solvers are readily available in commercial software, open boundary eigenvalue problems (such as antennas) is generally not directly supported in standard field solvers. In [31] this problem was solved by utilizing a plane wave illumination of the structure (in IE3D [32]) and then post-processing the resulting surface current density to identify the chassis mode resonances. By chassis mode resonance it is meant an equiphase current distribution over the (lossless) metallic surface, which when excited (and undisturbed) behaves as [31]

$$\mathbf{J}_S(\mathbf{r}',t) = \mathbf{J}_{S,i}(\mathbf{r}')e^{j\omega_i t - \omega_i t/2Q_{rad,i}} \tag{2.8}$$

where i means the i-th resonant mode with angular resonance frequency ω_i and $Q_{rad,i}$ as before. Other presented applications of characteristic mode theory for terminal antenna analysis and design are given in [33][34][35]. In particular, the method was used by Rahola [36] to validate the optimum placement of patch type radiating element at the short-edge (voltage maximum) of the chassis for monoblock (or bar) type of phones. Rahola also demonstrates the important lower-order modes of the chassis, where the first mode is the well known dipole-type distribution (resonant at 1.3 GHz) and the second mode is a full wave mode with a current minimum at the center (at 2.95 GHz). The third and fourth modes (at 4 GHz) have currents that are mainly in the direction of the short-edges (from long-side to long-side). A similar analysis was performed in [37], which also included the radiating element.

Methods of measuring the quality factor of antennas are provided in [38][39][40].

In the previous equations and arguments, the antenna has been assumed to be sufficiently modelled as a simple series or parallel resonance circuit. This implies in particular that the total resistance (radiation plus losses) is constant throughout the frequency band of interest, which is not always the case for real antennas. For example, the radiation resistance typically increases with frequency for small antennas. For the typical case of coupled resonators, e.g. the chassis and the radiator with the chassis resonance above the operating frequency band, a tighter bound of the achievable bandwidth was presented in [41]. For more complex cases, such as dual band operation [42], the Real Frequency Line Segment Technique of Carlin [43] or Simplified Real Frequency Technique (SRFT) [44] provides an insight to matching problems with an excellent estimate of the upper level of flat transducer power gain (or return loss) over the selected frequency bands. SRFT was applied in *Paper V* for a standard dual-band PIFA antenna to estimate the achievable bandwidth using lumped components.

2.3.1 Impedance Matching

The maximum achievable bandwidth, for a given antenna Q value and a certain acceptable reflection coefficient Γ, using an infinite number of tuned circuits (i.e. series LC and/or parallel LC circuits) was presented by Bode in 1945 [45]

$$B_{max\_\infty} = \frac{1}{Q}\frac{\pi}{\ln\left(\frac{1}{\Gamma}\right)} \tag{2.9}$$

In 1948 Fano [46] presented the achievable bandwidth for a given number n of tuned circuits, expressed as a set of coupled trigonometric equations. Lopez developed more accessible equations in 1973 [47] for the case of n tuned circuits, where for the typical case of $\Gamma > 1/3$

$$B_{max\_n} = \frac{1}{Q}\frac{a_n}{\ln\left(\frac{1}{\Gamma}\right)} \tag{2.10}$$

The coefficients a_n are provided from numerical solutions of the Fano equations [47], as tabulated below together with the bandwidth enlargement factor $\Delta B = B_n/B_{n-1}$.

n	1	2	3	4	5	6	7	8	...	∞
a_n	1	2	2.41	2.63	2.76	2.84	2.90	2.94	...	π
ΔB (%)		100	20	9.1	4.6	3.3	1.8	1.7	...	0.0

As can be seen, one extra resonator (the antenna counts as the first (n=1)) doubles the bandwidth while the next one provides 20% extra bandwidth. Further resonators are probably not motivated in practice. For the particular case of coupled resonators in mobile phones, the bandwidth enlargement factor was thoroughly investigated in [48].

2.3.2 Chassis Influence on Impedance Bandwidth

The characteristics of small internal (e.g. PIFA) antennas mounted on handheld terminals are very different compared to when placed on an infinite ground plane, and depends on both the antenna position on the terminal chassis and the dimensions of the chassis (the length in particular). This is due to the existence of radiating surface currents on the terminal ground plane induced by the antenna element. While the typical bandwidth of a patch type antenna on an infinite ground plane is in the 1-3 percent range (depending mainly on the height above ground), more than 10% is regularly achieved in standard size terminals. Although this has been known since (at least) the mid

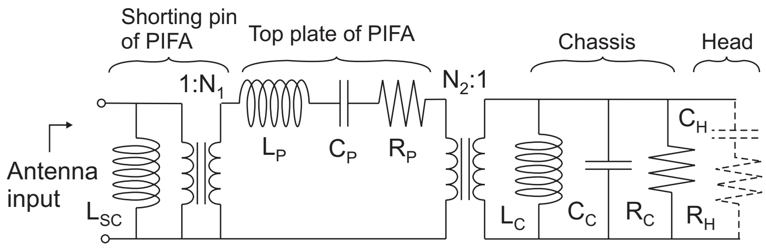


Figure 2.2: Equivalent circuit model of PIFA-chassis combination, including the effect of the head in talk position.

80's [49], the significance of the chassis was not fully appreciated and analyzed until late 90's - early 2000's [50][51], which was probably due to the introduction of internal antennas in mobile phones.

The standard model for studying (single band) terminal antennas, as reported by Vainikainen *et al* [50] in 2000, analyzes internal antennas as two coupled resonators - the antenna element itself, which supports a high-Q quasi-TEM transmission line wave-mode in the case of patch type antennas (e.g. PIFA), and the phone chassis, which supports a low-Q thick-wire dipole type current distribution. A circuit model of the PIFA and chassis combination, based on [52] and [53], is shown in Fig. 2.2. The PIFA element, being essentially a wide monopole bent towards the chassis where the short-circuit functions as a gamma-match [7], is modelled as a high-Q series resonant circuit. The chassis of a typically sized (100 x 40 mm^2) handheld terminal, being slightly less than $\lambda/2$ long at 900 MHz and coupled to the PIFA at the edge (voltage maximum), is modelled as a low-Q parallel resonant circuit. As an extension of the standard model (from [52]), it is proposed in *Paper VII* to represent the presence of the users head in talk position as a lossy capacitance ($C_H + R_H$) parallel to the chassis resonator.

Bandwidth maximas for the coupled resonators in Fig. 2.2 are obtained when both resonance frequencies are equal. For the chassis, this frequency is mainly determined by the terminal length and it therefore exists a relationship between chassis length and antenna bandwidth. As was previously noted, a 100 mm long chassis is resonant at 1.3 GHz (without loading from dielectrics and radiating element; it reduces to about 1.1 GHz in typical mobile phones [40]) and so some method of increasing the length is needed (e.g. by introducing a center slot as in *Paper III*). At higher frequencies, such as GSM1800/1900 or UMTS 2.1 GHz, the chassis is too long for resonance (the second chassis resonance is at around 3 GHz) and needs to be shortened (e.g. by wavetraps as proposed in *Paper I*). In Fig. 2.3 the simulated impedance bandwidth of a 2.1 GHz PIFA antenna as a function of chassis length is presented. As can be seen, a bandwidth minima is obtained for the typical length 100 mm at frequencies around 2 GHz. It is interesting to note that several antennas have recently been proposed that displays spectacular bandwidths

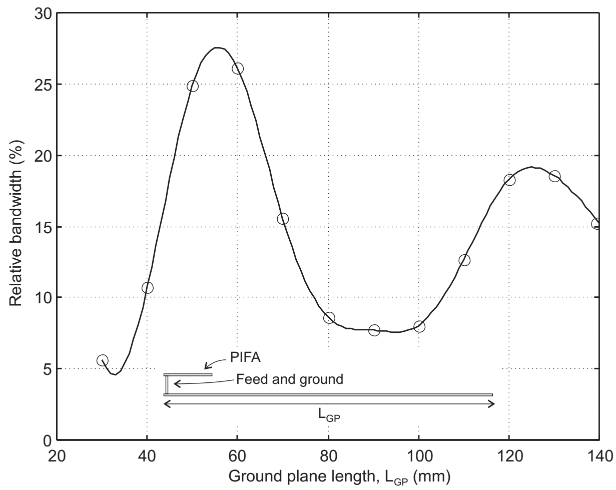


Figure 2.3: Simulated relative impedance bandwidth (RL < -6 dB) of a critically coupled PIFA antenna as a function of chassis length.

at around 2 GHz (like 1700 - 2200 MHz) [54][55][56][57][58]. A common feature of all these antennas is a chassis length of 60-70 mm, which, intentionally or not, provides a high coupling to the chassis resonance (and hence wide bandwidth, see Fig. 2.3). Also interesting is the fact that none of the papers motivates the particular choice of chassis length. The lack of standardized lengths therefore makes the comparison between the different proposed radiators difficult, or even impossible.

To obtain maximum coupling to the chassis resonator, for maximum bandwidth, the placement of the radiating element (e.g. PIFA) and feed/short pins are highly important [59]. For a patch type of antenna, the coupling is mainly through the E-field (or voltage) at the edge of the chassis, meaning that the best placement is at the short edges [52]. Different locations of radiating elements and feed/ground pins was studied in *Paper VII*.

2.3.3 Reconfigurable Frequency Tuning

The use of electrical switches, implemented using transistors, PIN diodes or MEMS, has been suggested to reconfigure the properties of an antenna. The main benefit for terminal applications is frequency band agility, mainly in order to enable smaller (and hence more narrow-band) antennas to be used. Typically, this is realized by changing the electrical structure of the antenna (e.g. by shorting a slot [60][61]), by selection of different matching components/networks at the input port [62][28] or by loading the antenna by reactive tuning components [63][64]. As an added benefit, these antennas are likely more efficient in talk-position due to their lower coupling to the chassis, as indicated in *Paper VII*. However, there are many problems with this technique that has so far limited it from widespread deployment in commercial phones:

- On/off switches are typically implemented using PIN diodes due to their high current capability and better linearity compared to varactor diodes. At microwave frequencies, PIN diodes are basically variable resistors. High resistance is achieved with no DC current while low resistance is achieved by a high DC current. High currents (> 5 mA) are necessary to reduce the insertion loss of the switch (i.e. reduce the series resistance) and to increase the linearity of the device, thus reducing battery life time. FET based switches are preferable as they do not consume DC current, here the main limitation is instead the linearity [65]. MEMS switches are naturally potential candidates for frequency band selection [66] due to the high linearity and low loss, today however such devices are not commercially viable at reasonable prices.
- Switches needs control voltages, implying that the antenna can not be designed as an ad-hoc component independent of the RF module. An exception to this is exemplified in *Paper VI* where the 'channel select' voltage from the receiver VCO is used (by re-mapping) for frequency agility of an active ferrite loop antenna.
- The switches are, relative to the cost of a passive antenna, expensive.
- The complexity of design, production and verification is increased.
- The switch package can become very hot, which can lead to e.g. melting of the carrier plastic.

2.3.4 Adaptive Impedance/Frequency Tuning

As an alternative to simply switching between different frequency bands, there are also more ambitious proposals where a multitude of switches, placed between the antenna port and the transceiver input/outpu port, could be used for adaptive impedance tuning to compensate for e.g. hand/head effects [67][68][69][70][71]. Such modules, typically employed at low frequencies, are called ATUs (Antenna Tuning Units) or ITUs (Impedance Tuning Units). In contrast to frequency band switching, this however requires a continuous monitoring of the antenna impedance, which must be done with low losses (RF and DC) and at a low cost. Alternatively, varactors can be used to control the resonance frequency of e.g. patch elements [72][73][74]. The problems associated with frequency switching, e.g. non-linearities, current consumption, complexity etc., are also valid for adaptive tuning systems. More importantly, none of the current proposals in the literature are viable for commercial applications as the antenna impedance/VSWR sensor is not possible to integrate (due to reasons of cost, size, current consumption etc.).

2.4 Talk Position

Three major effects (excluding HAC), all related, are introduced when the mobile phone is placed in talk position, i.e. pressed against the users ear and cheek:

- The increased effective dielectric constant as experienced by the antenna compared to free space reduces the antennas resonance frequency (so called 'detuning'). This will, for the typical case of a narrow band antenna designed for (and characterized in) free-space, lead to increased mismatch losses. The exact position of the hand and head are both effecting the magnitude of the detuning, and as this is different for different users, it is very hard to compensate for in the design phase. Cognitive antennas able to adaptively adjust the resonance frequency (or more generally, the input impedance) for different near-field environmental effects has been suggested in the literature [73] but has so far not been commercially implemented. As has been shown in *Paper VII* the detuning effect is however not as severe (for certain types of internal antennas) as one might initially guess from the high relative permittivity of the tissue ($\varepsilon_r = 43$ [59]) at the frequencies of interest.

- As the human head is highly lossy at microwave frequencies (due to the high water content), the power radiated towards the user is nearly completely absorbed by the tissue. This creates a deep null in the radiation pattern, and more importantly, reduces the total antenna efficiency. The problem is exacerbated by the low wave impedance presented to the antenna in the direction towards the head compared to the free-space side, amplifying the ratio of power delivered to the head compared to radiated into free-space [75]. In *Paper VII* it is suggested that the severity of this problem is less than one might initially suspect for patch type of radiators due to a re-distribution of power between the two resonators compared to free-space.

- SAR.

2.4.1 Specific Absorption Rate (SAR)

Due to an increasing awareness of the potential health effects from the near-fields and radiation of mobile phones since the 1990s, limits have been introduced by governmental regulating agencies in terms of peak power densities induced in human tissue, so called SAR values [76]. In the US, the Federal Communications Commission (FCC) specifies a maximum of 1.6W/kg taken over a volume equivalent of 1 g tissue [4]. The corresponding limit in the

EU, recommended by the International Commission on Non-Ionizing Radiation Protection (ICNIRP), is 2 W/kg taken over a volume corresponding to 10 g tissue. Recently, IEEE has adopted requirements similar to that of the EU [77].

It has been reported that spatial peak SAR values are located in areas close to the antennas current/H-field maxima [78]. Therefore, for 900 MHz terminal antennas with $\sim$100 mm chassis length, SAR peaks are generally obtained close to the vertical center of the chassis and at the short circuit of the antenna element (typically at the top-most short edge) [59][79]. At 1800 MHz, peak SAR is normally obtained at the short circuit of the antenna element [79]. For small chassis lengths, i.e. close to a $\lambda/2$ resonance at 1800 MHz, a SAR maxima is also seen at the chassis vertical center [59]. As a contrast to the common belief that SAR is related to peak magnitudes of antenna currents, a recent paper explains the SAR peaks by the boundary conditions of the quasi-static E-field of the antenna at the air-tissue interface [80]. The idea is here that the high real part of the tissues permittivity ε_r attenuates the perpendicular E-field component from the antenna and so the peak SAR value can then be found in regions with significant E-field components parallel to the tissue (which is less affected by the high ε_r). For the simple case of a dipole antenna (or terminal chassis), the parallel component of E-field is strongest at the center of the dipole, which of course also corresponds to the H-field (or current) maxima, meaning that both theories (E- vs H-field) predicts the same peak SAR region (at least for this simple case). Finally, it should be noted that SAR peaks and distributions can be significantly affected by metallic objects, such as piercings, worn by the user [81].

In general, SAR is a greater problem for GSM900 compared to GSM1800 due to the higher maximum output power (+33 dBm at 900 MHz in 1/8 time slots, +30 dBm at 1800 MHz in 1/8 time slots). For passive measurements on antenna mock-ups, a 250 mW CW signal is applied by a coaxial feeding cable to the antenna at 900 MHz, and 125 mW at 1800 MHz. For active measurements, the phone is set to output maximum power. SAR is measured or calculated from the r.m.s. electric field strength inside the human tissue, the conductivity σ and the mass density ρ from:

$$SAR = \frac{\sigma E^2}{\rho} \tag{2.11}$$

Typically, a Dosimetric Assessment System (DASY) is used to determine the SAR distribution inside a hollow phantom of the human body, see Fig. 2.4. The phantom is filled with a liquid with similar electrical properties as tissue, and a robot controlled probe measures the field densities in the volume of interest. For active measurements, the mobile phone is mounted in a specified talk position on the outside of the hollow phantom and is set to operate at maximum output power. The SAR values are calculated using the measured E-fields using the known σ and ρ of the liquid.

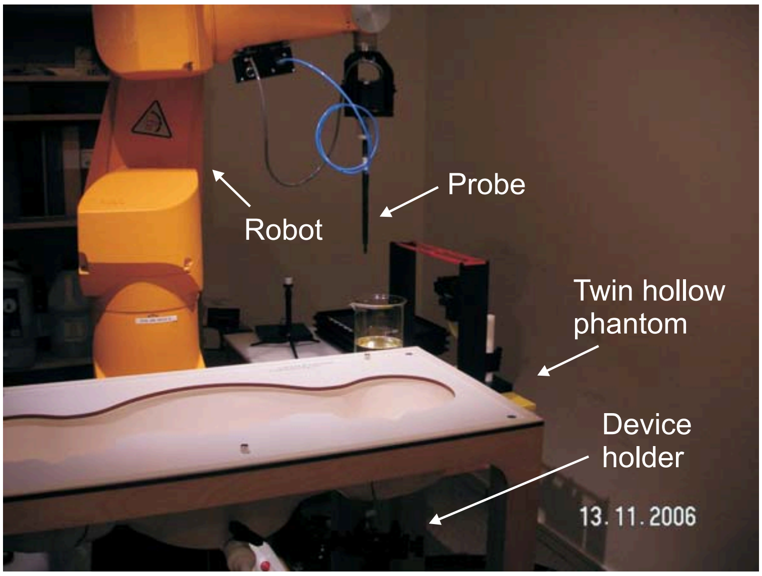


Figure 2.4: The DASY4 measurement set-up by Schmidt & Partner for SAR measurements at Laird Technologies, Åkersberga, Sweden.

2.4.2 Gain/Efficiency

In talk position of mobile phones, the gain (or more relevant - the efficiency (or average gain)) is reduced mainly due to power absorption in the lossy tissue, and to a lesser extent due to a detuning effect from the high permittivity of the tissue in the near-field of the antenna. The reduction of efficiency is naturally related to the SAR value, but is different in two respects. First, SAR is concerned with peak densities while efficiency is an average effect. Hence, an antenna with high SAR values does not necessarily have a low efficiency in talk position (although it is likely). Secondly, SAR depends strongly on the output power available from the power amplifier (through the filter/switch). Here it should be noted that SAR values are absolute while the efficiency reduction is relative. The low SAR value of a certain phone could be caused by a low output power available from the PA, or due to high losses in the antenna, or in an optimum scenario due to the radiation being directed away from the head. From the SAR value alone, it is not known which of these effects is responsible. Additionally, as the output power in many wireless systems is adaptive (e.g. in GSM), the SAR value by itself is not a sufficient measure of the total power induced into the operator during real usage, and it does also not provide any information about the reduction of communication range due to the reduced efficiency. For these reasons, TCO Developments have suggested the complementary (to SAR) figure of merit "Telephone Communication Power" TCP. Together with SAR, TCP fully characterizes the emission characteristics of the mobile phone. Currently, an average TCP of 0.3 W for GSM phones (with the phone operating in maximum output power mode) for

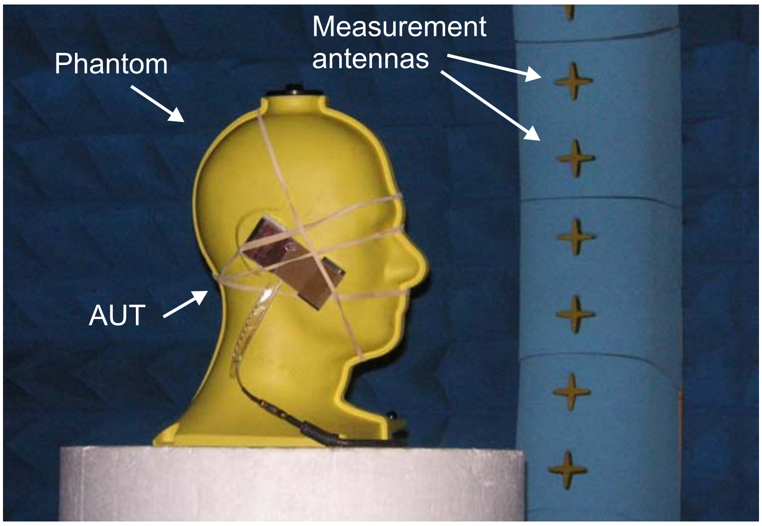


Figure 2.5: Measurement set-up for passive characterization of the efficiency in talk position of an antenna mock-up mounted on the phantom head "Gunnar" inside a near-field measurement chamber.

each band/mode/antenna over 4 different telephone positions is mandated. It should be noted that TCP is actually identical to the previously mentioned TRP (Total Radiated Power) which is regularly specified by mobile phone manufacturers, it is however not clear if the requirements are identical.

As a final note, there is an non-obvious connection between efficiency reduction in talk position and impedance bandwidth [59] in free-space. This is due to the fact that large bandwidths can only be obtained by strong coupling to the low-Q chassis resonator, which has a dipole type current mode distribution and hence similar efficiency reduction. In contrast, the radiating element is often shielded by the very same chassis (e.g. for patch type of radiators) and also has a larger distance to the tissue.

2.5 Form factors

Modern handheld terminals comes in a variety of shapes, or so called form factors. Concerning mobile phones, the "bar" or "monoblock" used to be the dominant type, mainly competing with the "flip" type which had a cover face for keypad protection and holding the microphone. Today, other forms such as the "swivel", "foldable/clamshell" and "slide" are fairly common together with various other, more experimental concepts. See Fig. 2.6 for the most popular versions. The swivel etc. have become popular as they allow a large keyboard (sometimes even QWERTY-style) and a large display while preserving a small in-pocket size. From an antenna designers view-point, these additional forms presents both challenges as well as opportunities. As an example of the latter,

(a) Bar (b) Clamshell (c) Slider (d) Swivel (e) Flip

Figure 2.6: Common form factors of modern mobile phones.

antennas for the clamshell type, which has two separate chassis sections (like a wide dipole), can be advantageously implemented by simply feeding the two chassis sections, thus realizing extremely large bandwidths. This technique has been proposed for DVB-H reception [82][83] and dual-band cellular applications [83]. As an example of the former, most forms except the bar can be operated in two states: open and closed. Preferably, the implemented antenna should work equally well in both states.

2.6 Active Receive-only Antennas

The direct integration of a low noise preamplifier with the antenna element is well known to provide advantages in many receive-only (broadcasting) applications. Examples include VHF/UHF television antennas with LNA:s integrated to compensate for the typically long and low quality interconnecting transmission lines [84], compact AM/FM car antennas [85], high sensitivity GPS modules where the effect of cable losses must be minimized etc. Concerning terminal antennas, active implementations can provide significantly superior performance to passive alternatives in particular for low frequency systems such as FM and DVB-H.

At about 100 MHz, the typically used 50 Ohm system impedance is no longer a suitable choice. At these frequencies, modern low-noise microwave transistors are intrinsically very high ohmic at the (gate/base) input and are therefore difficult to match to 50 Ohms without (noisy) resistive loading or resistive feedback. Additionally, internal antenna elements are difficult to make self-resonant and typical radiation resistances achievable are in the milli-Ohm range, again making 50 Ohms a far from optimum system impedance. By direct integration of the antenna and amplifier, both can instead be co-optimized for maximum performance.

In general, two types of active pre-amplifying antennas are utilized: resonant and non-resonant implementations. An example of a resonant solution is given in *Paper VI*, where a multi-turn loop is used both as an antenna element, and also to compensate for the capacitive input impedance of the integrated transistor amplifier. Non-resonant solutions are realized by e.g. connecting a short monopole type antenna directly to the high impedance gate of a FET transistor. The operation of these antennas are based on voltage division (or equivalently current division) between the source (i.e. antenna) and load (i.e. gate-source of a FET transistor). While resonant solutions obviously provides better performance (higher gain and lower noise figure), the bandwidth of these are very limited for terminal applications due to the high Q of the antenna element. Conversely, non-resonant antennas can provide a very wide bandwidth (in terms of gain flatness), but also lower gain and higher noise figure.

Active antennas, while achieving better performance than passive versions, suffers from several problems:

- The transistor(s) consumes DC power and the circuits occupies valuable PCB space.
- Due to the physical closeness to high power Tx antennas (like GSM850/900), the linearity of the amplifier must be very high unless the antenna/amplifier combination has intrinsically high filtering. This becomes particularly troublesome for systems closely spaced in frequency, e.g. DVB-H and GSM850. The problem is greater for non-resonant active antennas due to their large bandwidth. Filtering is here very difficult in practice as the impedance levels are so high, in particular because of the limited Q values of available inductors.
- The design complexity is much higher, greatly increasing the development time and production/verification time.
- To maximize performance, modern low noise transistors are preferably used (if the target prize permits). These transistors typically have gain up to 10-100 GHz, making stability a difficult issue, in particular as the antenna impedance is varying depending on its near-field environment, and as the antenna dissipates very little power at low frequencies (due to the low radiation resistance). Stability is, if possible, ensured by increasing the reverse isolation of the amplifier, i.e. reducing S_{12}, by selection of active device, topology and layout, instead of using resistive loading. In particular, the first amplifying stage should have as low voltage gain as possible to minimize the feedback (or Miller) capacitance between collector-base or drain-gate. This can be realized by e.g. connecting two transistors in a cascode, or by using a common emitter/source topology. I/O traces on the PCB must also be kept well separated to minimize EM coupling. Feedback through the parasitic source lead inductance can be avoided by thin substrates and hence short vias.

An example of an active receiving antenna for FM radio is given in *Paper VI*. Here, the transistors were assumed to operate at a specified quiescent point (as set by I_{DS} and V_{DS}) by setting the gate-source voltage V_{GS}. In reality, V_{GS} is not identical over temperature, time and device selection, so some means of adaptivity (feed-back) is needed. Such schemes, while not directly related to the RF part of the active antenna design, is important for a successful implementation and is therefore briefly addressed in the following sub-section. Also, the challenges associated with reliable and repeatable measurements of small active antennas are addressed here.

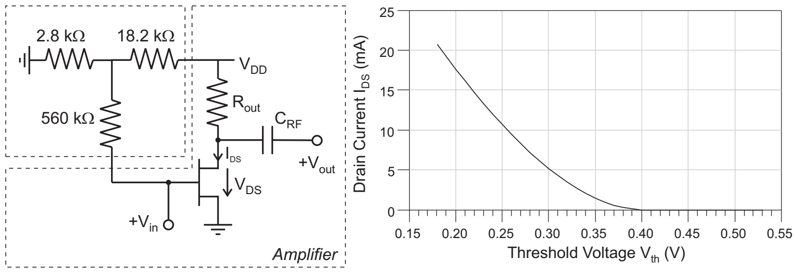
2.6.1 Biasing of Transistors in Active Antennas

DC biasing of the active devices sets the required operation condition (or quiescent point) which should be stable over input power, time, temperature and technology process variations. Furthermore, the bias circuitry should not affect the RF performance. For the case of FET transistors, the quiescent point is set by the drain-source voltage V_{DS} and the drain current I_{DS}, see Fig. 2.7. V_{DS} must be high enough so that the transistor operates in the saturated region (above the linear region). For most transistors, including the one used in *Paper VI*, this means $V_{DS} > 1V$. V_{DS} is simply set by selecting I_{DD} and the drain resistance R_{out}, possibly also a source/common-mode rejection resistance if used. R_{out} is typically chosen to be about 50 Ω to provide a proper interface to the receiver.

I_{DS} is controlled by the gate-source voltage V_{GS}, where the gate potential can be set by e.g. a voltage divider from V_{DD}, see Fig. 2.7. To reduce the thermal noise from the biasing resistors, a large (560 $k\Omega$) resistance is connected between the voltage divider and the gate input. This resistor also increases the RF/DC isolation. Since the source in many cases must be grounded for stability reasons, the gate potential sets I_{DS} through $I_{DS} = I_{DSS} * (1 - (V_{GS}/V_{th}))^2$, where I_{DSS} is the saturation current and V_{th} is the threshold voltage (also called 'pinch off' voltage V_P). For the E-pHEMT transistor used in *Paper VI*, V_{th} is positive and is specified to be in the 0.18-0.53 V range (varying from device to device), with 0.37 V being the typical value. Hence, for high volume applications some means of adapting the gate voltage to compensate for variations in DC parameters is necessary. Two examples - passive and active - of such feed-back circuitry are presented and compared here.

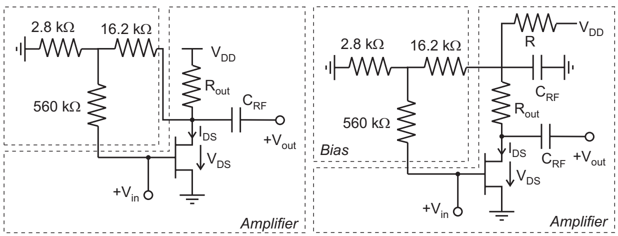
Passive feed-back

By connecting the voltage divider input to the drain of the transistor instead of V_{DD} (see Fig. 2.8a), the drain current (converted into a voltage through R_{out}) is fed back to the gate voltage. Now, if the current increases (due to e.g. a change of temperature or device), the drain potential and hence the gate potential decreases, thus reducing the current increase. Naturally, this is not limited to DC currents and so the gain of the amplifier is reduced. The feed-back can



(a) Schematic (b) Drain current vs V_{th}

Figure 2.7: Schematic of amplifier section of active antenna without bias feed-back (a), and drain current sensitivity to threshold voltage $V_{th}(b)$.



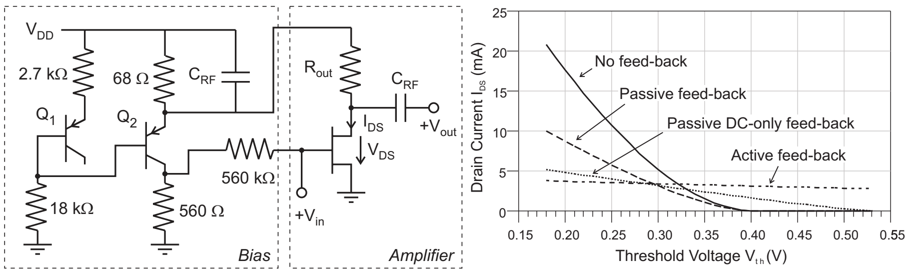
(a) With RF feed-back (b) Without RF feed-back

Figure 2.8: Schematic of amplifier section of active antenna with passive drain-gate bias feed-back.

be limited to DC only by connecting a second drain resistance, called R_{FB}, in series with R_{out} and use this terminal for feed-back instead, see Fig. 2.8b. A large capacitor C_{RF} to ground between R_{FB} and R_{out} decouples the RF signals from DC. R_{FB} can be chosen arbitrarily large (if voltage headroom allows) to increase the amount of feed-back, unlike for the previous case where the resistance is set to 50 Ohms. On the downside, the DC only feed-back implementation requires two extra components (one resistor and one large capacitor) and also reduces the available voltage headroom, which typically is fairly restricted ($\sim$ 3 V) in battery driven applications.

Active feed-back

Active biasing provides the best compensation of variations in DC parameters at the cost of several extra components and increased current consumption. A typical schematic is shown in Fig. 2.9a, where two bipolar PNP transistors Q_1 and Q_2, four resistors and one capacitor have been added. Two resistors (2.7 $k\Omega$ and 18 $k\Omega$) in series with a diode connected PNP transistor Q_1 provides a fixed voltage at the base of Q_2. The diode compensates for variations in the internal base-emitter voltage of Q_2.



(a) Schematic (b) Drain current vs V_{th}

Figure 2.9: Schematic of amplifier section of active antenna with passive drain-gate bias feed-back (a), and drain current sensitivity to threshold voltage $V_{th}(b)$.

If the current I_{DS} increases (due to e.g. a change of temperature or device), the emitter potential of Q_2 is decreased, and since the base voltage is fixed the collector current through Q_2 is reduced by ΔI_C. ΔI_C is translated into a gate voltage change $\Delta V_G = \Delta I_C * R_C$ (where $R_C = 560\ \Omega$ here) which compensates for the change in I_{DS}. The capacitor C_{RF} in parallel with the 68 Ω resistor decouples the RF signals.

2.6.2 Measurements on Active Receive-only Antennas

The main problem of measuring the performance of small antennas is the influence of the coaxial feed cable [86]. Not only is the far-field pattern distorted, the cable can (if not properly decoupled by baluns) become a large contributing part of the total radiating structure [87]. As the size of the antenna goes down, the problems are increased. In particular, measurements on electrically small single-ended active internal antennas for FM reception have presented serious challenges concerning reliable characterization. Here, the problems are related both to the smallness of the antenna, but also to the fact that it is active. Since the antenna is active, any amount of gain is obtainable (for instance by cascading several amplifiers), making gain by itself an irrelevant figure of merit (unlike for passive antennas). Instead, the metric 'gain divided by noise temperature at the output', G/T ("G over T") [88], is typically used. Here, the gain is normalized by the noise temperature rendering the absolute value of gain unimportant. With G/T as figure of merit, the output noise power must also be measured, further complicating the antenna characterization. A detailed discussion of these aspects is provided in *Paper VI*. It should be noted that at FM frequencies, the background noise temperature is much higher than the thermal noise floor of 290 K (or -174 dBm/Hz) [89] due to manmade noise (e.g. industrial equipment, consumer products, power transmission etc.). The increased noise level means that the effect of the noise contributions from the active devices and resistors is reduced, unless the efficiency of the antenna element is so low that the physical temperature of the antenna dominates the noise temperature. Additionally, the high background noise level means that the efficiency requirement of the antenna element can be reduced without as significant a reduction of G/T as for the ideal, low noise case (but then the noise contribution of the preamplifier must be low).

To reduce the effects of the feed cable during measurements of terminal antennas, several suggestions have been made. Generally, the cable should be connected to an E-field minima of the antenna (if such can be located) [90]. High impedance ferrite beads are commonly employed along coaxial cables, which will partly reflect and partly absorb (depending on operating frequency) the induced power on the cable. Also, several types of baluns suitable for terminal measurements have been proposed [91], including a balun for dual band applications in *Paper V*. At wavelengths much longer than the size of the terminal chassis, such as for FM radio, it is no longer possible to decouple the cable by such means. Instead, cable-less procedures have been suggested as a remedy. For transmitting antennas, a (battery operating) voltage controlled oscillator (VCO) with known output power is typically integrated on the terminal and the received power level is recorded with a reference antenna. For receiving antennas, the receiver power level (using e.g. diode detectors) is typically transmitted using (non-metallic) optical fibers to a measurement computer. Also, Tx/Rx fiber-optic systems have been used for accurate impedance measurements on small antennas [92].

To measure the antenna gain of small single-ended active monopole-type antennas for FM applications, a measurement system based on fiber-optics has been developed at Laird Technologies, Åkersberga, Sweden, as outlined in Fig. 2.10. The terminal antenna (Device Under Test, DUT, in Fig. 2.10) receives RF CW signals within the 70-120 MHz frequency range. An Electrical-Optical (E/O) converter with 50 Ohm input interface is mounted on the terminal and transmits a modulated optical signal through the fiber, where the modulation depth is proportional to the power of RF CW signal. A Signal Analyzer detects the RF signal through an Optical-Electrical (O/E) converter, with the power being proportional to the received signal by antenna. Thereafter, the DUT is substituted by a reference antenna with known gain, and the DUT gain is calculated.

The E/O converter, placed on a 100x40 mm^2 PCB, is about 30x30 mm^2 size, and is fed by +3.7 V from a battery. Both the E/O and O/E converters are sufficiently linear for the application, no compensation/calibration is used. The complete system is portable and thus suitable for field tests. The measurement system specification is the following:

- Measurement range: -70 dBm -20 dBm input power level at DUT
- Measurement frequency: 70-120 MHz
- Linearity: 25 dB harmonic level suppression
- Continuous operation time: 2 hours

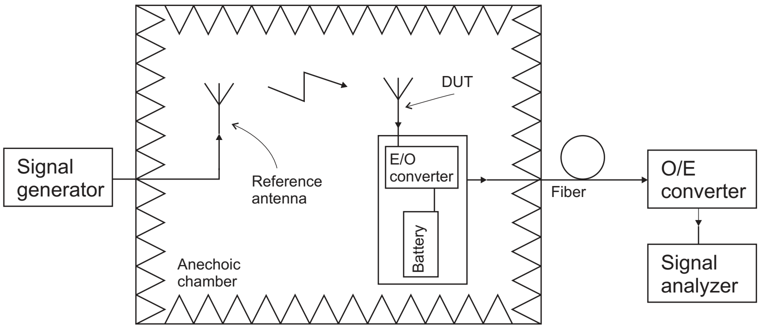


Figure 2.10: Measurement set-up for internal active FM antennas.

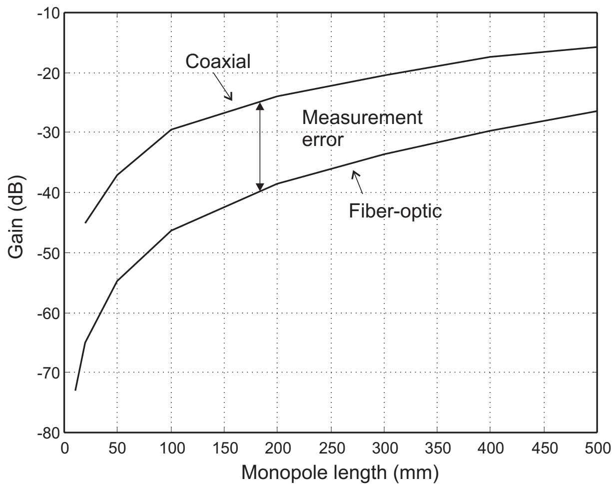


Figure 2.11: Comparison between measured average gain on different length monopole antennas using a coaxial feed cable and optical fiber.

- Accuracy: 0.5 dB

The fiber-optical setup was used to measure the gain of monopole antennas (or rather asymmetrically fed dipoles as the ground is small) of different lengths. The monopole was implemented as a 1 mm diameter copper wire extending along the long side of the terminal and fed against the chassis at the top short edge. Different lengths (10-500 mm) of the monopole were tested. For comparison, the same antennas were also measured using a coaxial cable (without balun) and the two results were compared. As can be seen from the average gain measurement (i.e. antenna efficiency) in Fig. 2.11, the measurement error is worse ($\sim$ 20 dB) for shorter antennas, but larger than 10 dB even for long lengths.

Naturally, the largest gain reduction for the shorter antennas stems from the reflected power due to large impedance mismatch. While even very small antennas could in theory be matched to any impedance (like 50 Ohms) and thereby achieve (almost) the same average gain as a full sized dipole in a nar-

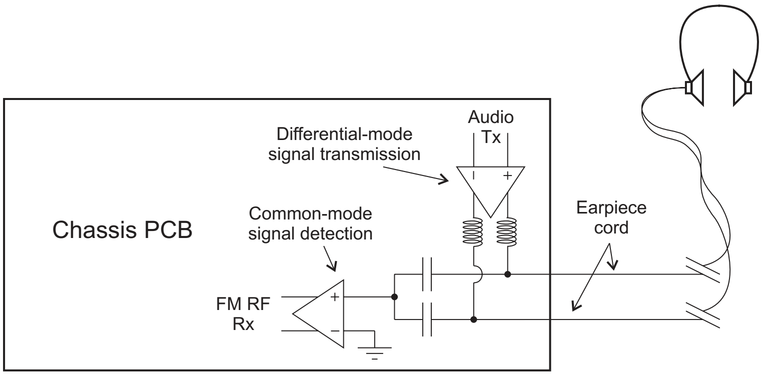


Figure 2.12: Conceptual schematic of simultaneous RF reception (in common-mode) and audio transmission (in differential-mode) using the earpiece cord.

row frequency band (assuming exceptionally low-loss antenna material and matching components), it is anyway interesting to note the absolute values of measured gain of the unmatched monopole antennas in Fig. 2.11. Hands-free cords, which are typically used as FM receive antennas for hand-held terminals, are usually in the $\sim$500 mm range and as they normally do not employ matching components[2], the expected gain of such solutions are in the -25 dB range. The lack of matching of such antennas is motivated by the cord placement and surrounding environment being fairly random in real use, and hence the input impedance is not predictable or stable. However, by placing a RF choke on the cord at a $\lambda/4$ distance from the chassis edge, the cord can sustain a stable impedance at UHF frequencies for DVB-H reception, as suggested in *Paper II* (assuming that the $\sim$13 cm cord section between phone and choke remains fairly straight). At FM frequencies, the noise level is much higher than the typical thermal floor of $kT = -174$ dBm/Hz (where k is Boltzmann's constant and T the absolute temperature) encountered at gigahertz frequencies. Due to this high noise level, in particular in urban environments, the gain requirement (for receivers) can be significantly relaxed without sacrificing system performance, as is discussed in *Paper VI*. Hence, the low efficiency of earpiece cords (and naturally also for internal solutions) can in practice be tolerable. Naturally, while the high ambient noise level relaxes the requirements on the receiving antenna in terms of efficiency, it of course significantly degrades the total system performance and can only be compensated for by increasing the transmitting power and/or the directivity of the transmitter and/or receiver.

The simultaneous use of the cord as receiving antenna and for audio signal transmission to the earpiece is shown conceptually in Fig. 2.12. The RF signal is received in common-mode and connected through a simple high-pass

[2] Actually, some matching components are usually employed for earpiece cord antennas. However, they are typically only employed to transform a supposed 50 Ω antenna impedance into a differential 100-200 Ω interface that is more suitable for the FM transceiver.

filter (here shown as two series capacitors) to the LNA. The audio signal is transmitted in differential mode on the two-wire cord with the power amplifier connected through a low-pass filter (here shown as two series inductors). For stereo systems, one more audio PA is used, and all four terminals are connected (through the filter) to the RF LNA. In a real application, the implementation can be slightly different, perhaps using a volume control unit on the cord. In the case of three-wire systems (left, right and common ground), either one of the wires is used as a monopole type of antenna, or two of them are used in a dipole-type configuration. If a coaxial metal shield is used, this shield is utilized as the (monopole) antenna. The principle though remains the same in all cases.

3. Conclusions and Future Outlook

In this thesis, solutions and studies related to modern requirements on handheld terminal antennas have been presented. A major theme has been the role of the chassis resonator - how to utilize it for maximum bandwidth, and how it is affected by the users head in talk position. In addition, an internal antenna for FM radio reception has been demonstrated, together with a method of utilizing the earpiece cord for DVB-H. Though each of these proposed solutions, as well as many other found in the scientific literature, are competitive in terms of development ease and cost effectiveness compared to traditional techniques, the industry has up until now taken a fairly conservative approach concerning the antenna implementation. Nearly all phones today contains a standard internal PIFA for the cellular bands, which has almost completely replaced the previously popular external helix or retractable rod/whip. For complementary systems, such as Bluetooth or WLAN, a simple monopole or IFA is typically employed. The reason for this, of course, is that there has not been a need to make any radical changes - standard solutions have, with some minor tweaking, been good enough for the job. Today however, in late 2006, we are seeing a paradigm shift that has been imposed by the new trend of ultra-thin terminals and by the increasing demand for coverage of all GSM and UMTS bands. The needed bandwidth is simply not achievable using a standard passive PIFA antenna. In fact, many modern antenna road-maps do not contain any PIFA solutions for phones in the mid- to high price range at all.

So, what new exotic radiator is now replacing the all-pervading planar inverted F antenna? It turns out that the answer is not very exciting - we are currently witnessing the grand come-back of the conventional external monopole [93][94][54][95]. This time however, it is hidden within the phone cover and has been relocated to the bottom part of the phone. Similar to the normal-mode helix concept, the monopole is usually folded together to save space, and typically consists of two unequally long branches for dual band functionality. Sometimes, the monopole is called something else, like "semi-PIFA" or "coupling element"[1] which is merely a matter of semantics. Common to all

[1]The term "coupling element" was introduced to emphasize the role of the internal "radiator" merely as a structure to induce currents onto the chassis, and possibly also to act as a matching circuit, rather than actually radiating power. The term is also used for a particular class of non-resonant structures which are typically implemented as wide patches (with no short-circuit) [96] extending beyond the ground plane - hence they qualifies as monopoles (albeit very wide, and short, such.) Where PIFA:s are typically meandered to obtain sufficient length

these antennas is the lack of ground plane directly beneath the antenna, or so called "ground clearance". As many of these designs require a shunt inductance as matching component, which essentially has the same functionality as the short circuit pin in a PIFA, a more appropriate name would be inverted F antennas (IFA). Again, it is merely a question of semantics.

In parallel with the return of the monopole, a higher acceptance among phone manufacturers concerning reconfigurable/steerable antennas for frequency band selection has been observed. A few pins with programable outputs are made available from the front-end which can (optionally) be used as control signals to switches on the antenna. The purpose of this is not so much to gain large bandwidths, but mainly to allow for high miniaturization of the radiating elements. Adaptive solutions, which somehow detects a large impedance mismatch (from e.g. the influence of the users hand), or steers the antenna beam away from the user, are still not commercially viable. For the case of mismatch compensation, there are two main practical problems. The first is related to how to do the VSWR/impedance measurement using cheap, small size and low loss components. The other is how to implement the tuning circuitry with sufficient linearity (which typically excludes varactors), low loss, low price and high power handling capability. Naturally, the losses due to the adaptive impedance tuning unit must be much lower than the mismatch loss. As for beam control, the problem of talk position is highest for low frequency and high power systems such as GSM900. Here, the radiation pattern is mainly determined by the chassis geometry, which is not accessible to the antenna designer. Possibly, adaptive beam steering might become feasible for the emerging high frequency systems such as HiperLAN and WiMAX. Additional trends include full metal covers (for aesthetic reasons), which demands either some metal clearance around the radiating element, or some type of slot solution. Other current/future challenges includes internal DVB-H antennas, internal FM antennas, RFID, MIMO (Multiple Input Multiple Output) and diversity (or SIMO - Single Input Multiple Output). Diversity and MIMO will probably be utilized mainly for high data rate systems such as WLAN, 3G HSDPA or WiMAX, and introduces new characterization challenges besides increasing the design complexity. Several new systems are also expected to be implemented in terminals in the near future, such as UWB (3.1 - 10.6 GHz), WiMAX, HiperLAN etc, leading to increasing problems with cross-talk and sharing of volume (besides introducing new challenges by themselves).

To summarize, there is still a lot to be done in the area of handheld terminal antenna design.

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for self-resonance, coupling elements instead uses a (lumped) series inductor. The short-circuit is similarly replaced by a (lumped) shunt inductance.

4. Summary of Papers

4.1 Paper I: A Bandwidth Enhancement Technique for Mobile Handset Antennas using Wavetraps

It has recently been shown that the achievable impedance bandwidth of internal mobile handset antennas is mainly determined by the size and shape of the phone chassis (or ground plane) as opposed to the geometry of the antenna element (e.g. planar inverted-F antenna, PIFA) for a fixed allocated antenna volume. In particular, bandwidth maximas are obtained at frequencies where the chassis effective electrical length (adjusted for open-end effects) corresponds to a half wavelength resonance. At frequencies above 1.5 GHz, the typical chassis length of 100 mm is too long for resonance, and in some cases, such as at the frequency band allocated for 3G (1920 - 2170 MHz), it even corresponds to a bandwidth minimum.

This paper presents a general technique for extending the impedance bandwidth of terminal antennas by connecting resonant short circuit transmission lines (called wavetraps in this context) to the long sides of the terminal chassis, thereby electrically shortening the ground plane lengths to obtain resonance and hence maximum bandwidth. The wavetraps occupies minimum area/volume and can be applied to any already designed antenna element and PCB layout in an ad-hoc fashion. Three typical examples of applications have been proposed and analyzed, with various performance enhancements achieved. The wavetraps have been shown through measurement not to reduce the antenna efficiency and is also applicable to dual band applications. Finally, SAR measurements have been presented.

4.2 Paper II: Earpiece cord antenna for DVB-H reception in wireless terminals

In this paper, the utilization of the commonly employed earpiece chord as an antenna for DVB-H reception at UHF frequencies is proposed, similar to what is now common practice for handheld FM radios. As the cable length of a typical earpiece chord is several wavelengths ($\sim$1.2 m) long, the input impedance (when fed against the phone chassis) is multi-resonant and hence difficult to match, and the radiation pattern displays several deep nulls. Therefore, the cord has been shortened by an RF choke at a quarter wavelength distance from the phone chassis to obtain an asymmetric dipole-type radiation pattern and input impedance. By using two chip components, the input impedance has been matched to S11 $<$ -5 dB over the frequency band 470-700 MHz. As the RF choke can be implemented by winding a short section of the cord into a coil, this solution presents no extra cost compared to a standard hands-free.

4.3 Paper III: Wideband Slot Antenna for Low-Profile Hand-held Terminal Applications

A wideband slot antenna for ultra-thin handheld terminal applications has been proposed and evaluated in this paper. By feeding the slot with a microstrip line, a triple-resonance response is obtained from the chassis, slot and microstrip stub interaction, providing sufficient impedance bandwidth and radiation efficiency to cover the popular 0.9-2.7 GHz frequency range. Simulations and measurements of return loss, radiation efficiency and radiation patterns on a prototype antenna implemented on a standard 40 x 100 mm^2 ground plane have been presented, indicating satisfactory performance in the targeted cellular (GSM and UMTS) and complementary (WLAN) frequency bands. By routing signals between the upper and lower parts of the PCB over the ground connection at the rightmost part of the chassis, using a low height above ground plane and optionally a shield box, $>$30 dB isolation between antenna and signal line is obtained without extra filtering. This implies that the antenna design and signal routing can be done independently without significant mutual coupling. As the antenna is realized by simply removing parts of the chassis metallization, the solution involves no extra cost.

4.4 Paper IV: A Dual Wideband Printed Dipole Antenna with Integrated Balun

Wireless terminals today are typically required to support several communication systems and bands, such as GSM 850 & 900 MHz, PCS/DCS 1800 & 1900 MHz, UMTS 2.1 GHz, WLAN 2.4 & 5 GHz etc. Usually, a single multi-band RF transceiver module covers all related systems (e.g. cellular) supported by the terminal to reduce component count, prize, volume and power consumption. To be compatible with the transceiver and from size constraints, a single antenna using a single (typically unbalanced) feed point is needed that covers all bands supported by the transceiver.

In this paper, a dual wideband printed dipole antenna concept with integrated balun for cellular applications is presented. The radiating structure is a two element dipole with resonances at 900 MHz and 1800 MHz, with the balun providing one extra resonance at each band. By co-designing the radiating element with a Marchand balun modified for dual band operation, an antenna covering the cellular bands 824-960 MHz and 1710-2170 MHz has been realized. Measurements on a fabricated antenna on low-cost FR-4 substrate shows antenna efficiencies of $> 80\%$ and $> 70\%$ in the lower and upper band respectively. Despite the use of a dual element radiator, close to omnidirectional H-plane radiation pattern is obtained.

4.5 Paper V: A Single Matching Network Design for a Dual Band PIFA Antenna Via Simplified Real Frequency Technique

In this paper, the matched performance of a dual band PIFA antenna is reported. The return loss of the antenna has been optimized over the popular commercial wireless communication bands of 824-960 MHz and 1710-1990 MHz using a single matching network. The network was synthesized using the Simplified Real Frequency Technique, which yields an "easy to implement" circuit topology and realizable component values. With the implemented matching network, a simultaneous measured bandwidth enhancement of 58% and 127% has been achieved in the low and high frequency band, respectively, without significant reduction of radiation efficiency. Simulations and measurements are reported together with analysis of the power handling capabilities of the used components. Also, details on how to decouple the measurement cable without difficulty at dual frequency bands are provided.

4.6 Paper VI: An Internal Active Antenna for FM Radio reception in Mobile Handsets

This paper presents design parameters and measured results of an internal active FM antenna for mobile phone applications. The antenna is based on an electrically short multi-turn ferrite loaded loop that is co-optimized with a low noise differential amplifier, thereby avoiding the use of a 50 Ω system impedance and hence large losses from matching networks. Due to the high ambient noise temperature at FM frequencies in urban environments, the low antenna efficiencies obtainable with such miniaturized antennas is not as detrimental to the system performance (i.e. G/T) as for the low ambient noise case typical for higher frequencies. However, the low efficiency motivates the use of low-noise transistors in the amplifier as the active devices together with the biasing resistors mainly determines the output noise power level. Measured G/T degradation compared to an ideal dipole, using a receiver with 6 dB noise figure, is < 42 dB and < 23 dB at noise temperatures of 290 K and 23.000 K (urban environment), respectively. This is at least 6 dB better than for a passive loop antenna using the same ferrite material, and in practice the difference will be substantially higher due to the inevitable matching losses for the passive case. Measurements are in good correlation with simulations. The measured isolation from a co-located GSM900 PIFA-type transmit antenna is 40 dB including amplifier gain, making the active antenna sufficiently insensitive to GSM blocking. The 1 dB output compression point is reached with +35 dBm input GSM power (CW).

4.7 Paper VII: The Effect of Talk Position on The Chassis Mode of 900 MHz Terminal Antennas

The effect of talk position on terminal antennas consisting of electrically small internal radiating elements is investigated in this paper using a proposed circuit model of the antenna and head interaction. From this model, predictions are made on the effects on radiation efficiency, impedance bandwidth and frequency detuning in talk position. Validating measurements and simulations of three versions of the popular planar inverted F antenna, together with a reference dipole, is presented. It is shown that the power distribution between the antenna element and chassis wave-modes is altered by the presence of the lossy head, with corresponding effects on all antenna parameters. In particular, the antenna efficiency in talk position is much higher and the detuning effect is significantly lower than expected from previously proposed antenna models.

5. Summary in Swedish

Bredbandiga aktiva och passiva antennlösningar för mobila terminaler

Mängden användningsområden för den vanliga mobiltelefonen har ökat dramatiskt de senaste åren. En typisk modern telefon kan, förutom att användas till samtal (över olika system över hela världen), fungera som TV och radio-mottagare, till datakommunikation via WLAN (internet) och Blåtand, som GPS-mottagare etc. Inom en snar framtid kommer dessutom telefonen antagligen att användas (via RFID) som bussbiljett, för att köpa biobiljetter och betala parkeringsavgifter, som ersättning till passerkort o.s.v. Samtidigt som fler funktioner har adderats till telefonerna så har storleken minskat dramatiskt, i synnerhet i samband med den senaste trenden med ultra-tunna telefoner. Denna miniatyrisering har möjliggjorts av ett antal teknikgenombrott relaterade till strömsnål kretsdesign, högre elektronikintegration, förbättrad halvledarteknik samt förbättrade skärmar och batterier. Till skillnad mot vanlig elektronik vars prestanda i de flesta avseenden förbättras vid förminskning så är funktionen hos antennerna direkt kopplad till dess allokerade volym. Den tillgängliga volymen för antennerna har dock kontinuerligt minskat dels i samband med minskningen av terminal-storleken, dels p.g.a. den ökade mängden antenner för alla nya system.

I denna avhandling har ett antal antennlösningar relaterade till de senaste trådlösa systemen föreslagits, tillsammans med en generell metod för att utöka bandbredden på befintliga antenner för höga (runt 2 GHz) frekvenser. Dessutom presenteras en studie av hur antennen för de cellulära banden runt 900 MHz påverkas av användarens huvud. Slutligen så har en integrerad antenn för FM-radio föreslagits, som ersätter eller kompletterar handsfree-kabeln som traditionellt fungerat som antenn. Varje bidrag presenteras kortfattat och populärvetenskapligt nedan.

För mottagning av digital TV i mobiltelefoner (via DVB-H) krävs en extremt bredbandig antenn (470 - 700 MHz) som dessutom, i full skala, är väldigt stor (runt 20-30 cm). För att få plats med en sådan antenn internt i telefonen, måste den på något sätt krympas ihop, vilket oundvikligen leder till en reduktion av bandbredden, vilken även i full skala är svår att uppnå. Som ett alternativ till en intern antenn har i denna avhandling föreslagits en modifierad handsfree-kabel som antenn-element med prestanda långt över nuvarande specifikation. Förändringen innebär en introduktion av en högimpedant sektion, t.ex. genom att linda en del av kabeln i en spiral, vid en specifik längd efter telefon-kontakten. Således är inte priset för en sådan lösning högre än för en vanlig handsfree.

I samband med att telefonerna blivit allt tunnare de senaste åren så har den tillgängliga bandbredden hos antennerna blivit allt mindre (enligt grundläggande fysikaliska lagar). Samtidigt har kravet på bandbredd hos antenn-elementen gått upp p.g.a. introduktionen av nya trådlösa system och frekvensband. Som alternativ till traditionella lösningar presenteras här en antenn integrerad i chassit på telefonen, och som därmed inte är beroende av en viss telefon-tjocklek och som heller inte utgör någon extra kostnad. En vidare fördel med den föreslagna antennen är att en extraordinär bandbredd erhålles p.g.a. att lösningen på ett optimalt sätt använder hela chassit som antenn. Lösningen täcker däremed in samtliga av dagens cellulära system (inklusive 3G) både i Europa och i USA, samt GPS, WLAN och Blåtand samt kommande utökade frekvensband för 3G.

Det har nyligen visats att de flesta traditionella antenn-element i huvudsak fungerar som kopplare och matchningskomponent till chassit, som utgör den egentliga antennen (åtminstone på lägre frekvenser såsom GSM 850/900 MHz). Maximal bandbredd uppnås om antenn-elementet och chassit har samma resonans-frekvens, vilket i chassits fall bestäms av dess fysiska längd. I de flesta terminaler är längden ca. 10 cm, vilket är för kort för låga frekvenser (t.ex. GSM 900 MHz) och för långt för höga frekvenser (t.ex. 3G runt 2.1 GHz). I synnerhet för 3G-applikationer så har chassit en sådan längd så att ett bandbredds-minima fås. I denna avhandling har en generell metod presenterats där chassits längd, för höga frekvenser, kan förkortas till optimal storlek för att erhålla maximal bandbredd. Metoden är enkel att implementera, kan introduceras ad-hoc till redan befintliga konstruktioner och innebär ingen prisökning då endast en mindre modifiering av chassits metalliseringen behövs.

Strålning och närfältseffekter från mobiltelefoner har påvisats vara ansvarig för bland annat temperaturhöjningar i användarens vävnad, så kallade termiska effekter, på grund av absorption av stora delar av den utstrålade effekten. Detta har till följd att telefonen i många fall adaptivt höjer sin uteffekt vilket leder till reducerad batteritid, och dessutom begränsas det maximala kommunikationsavståndet. Vidare så har användarens huvud en påverkan på telefonens antenn på sådant sätt att antennens resonansfrekvens förändras, vilket leder till ytterligare förluster. I denna avhandling presenteras en modifiering av en populär modell för typiska interna terminalantenner för att inkludera närvaron av användarens huvud. Modellen används sedan för att förutsäga ett antal icke-uppenbara effekter på antennen, bl.a. att storleken av absorptionen och övrig påverkan på antennen är betydligt mindre än vad som kunnat förutses med tidigare modeller. Effekterna har påvisats genom mätningar och simuleringar.

I många fall behövs en antenn för cellulära frekvens-band med egenskapen att strålningen fördelas jämt över alla kompassriktningar då det ej är känt i vilken riktning närmaste basstation ligger. Exempel på sådana applikationer är bilantenner till handsfree-enheter, telematik- och M2M-moduler, repeatrar, fjärravlästa elmätare, trådlösa access-punkter etc. Den vanligaste antennen med sådana egenskaper är den så kallade dipolen. Nackdelen med en traditionell dipol är dess begränsade bandbredd samt att den ej direkt kan kopplas till en koaxial-kabel. I denna avhandling presenteras en dipol med en integrerad enhet som möjliggör användning av koaxialkabel samt även utökar bandbredden i två separata frekvensband för att passa de mest populära cellulära frekvensbanden.

Den absolut vanligaste typen av interna antenner till mobiltelefoner är den så kallade PIFA:n (Planar Inverted F Antenna). Denna antenn har många fördelar såsom enkel produktion, låg kostnad, möjlighet att anpassa till flera frekvensband samt riktad strålning bort från användaren (d.v.s. lågt SAR-värde). Med tiden så har PIFAn utvecklats från att fungera för ett smalbandigt system (t.ex. GSM900) till att samtidigt täcka 2-3 olika band av totalt 5 önskvärda (GSM850/900/1800/1900 + 3G). Av intresse är att undersöka vilken bandbredd som en sådan antenn teroretiskt kan erhålla (helst samtliga 5 band), samt påvisa möjligheten att praktiskt realisera detta. En sådan studie har genomförts här och en prototyp har implementerats som täcker 4 av den önskvärda banden.

Det sista bidraget addresserar integreringen av en antenn för FM radio internt i telefonen, som ersättning för handsfree-kabeln som traditionellt fungerat som antenn. Lösningen bygger på ett aktivt koncept där en lågbrusförstärkare har designats ihop med antenn-elementet för optimal funktion. Förutom att presentera och karakterisera den specifika lösningen har relaterade mätmetoder samt isolation av strålningen från GSM-sändaren studerats. Med hjälp av den interna antennen kan användaren lyssna på radio både i sin Blåtands-handsfree och även genom den interna högtalaren utan att behöva koppla in hörlurs-sladden.

6. Errata and Addendum of appended papers

6.0.1 Errata and Addendum of Paper II

In the published paper, Fig. 3, page 610, the legend description is identical for two different efficiency requirements. The correct figure is given below:

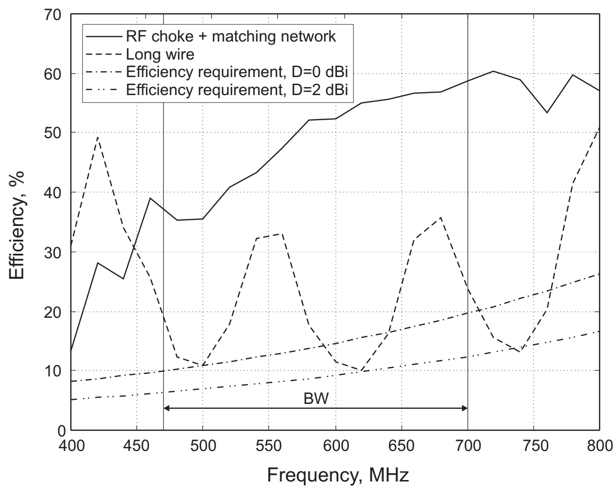


Figure 6.1: Measured antenna efficiency (including mismatch loss) of cord antenna, including 1.2 m long wire as reference. Gain specification also included by conversion to efficiency.

Also, in the published paper, the width of the chassis is not included in Fig. 1, page 1. This width is 40 mm (as stated in the text).

On page 1, Section II, the $0805CS-271X-L$ coil used as choke has an inductance of 270 nH, a Q value of 48 (at 250 MHz) and a self-resonant frequency of 730 MHz.

6.0.2 Addendum of Paper III

Some extra material related to *Paper III* was provided at the oral presentation during the conference and is provided here for completeness. In particular, measured SAR values, an explanation of the exceptional measured impedance bandwidth and an implementation variant suitable for slightly wider terminals are presented. It should also be noted that a similar concept has

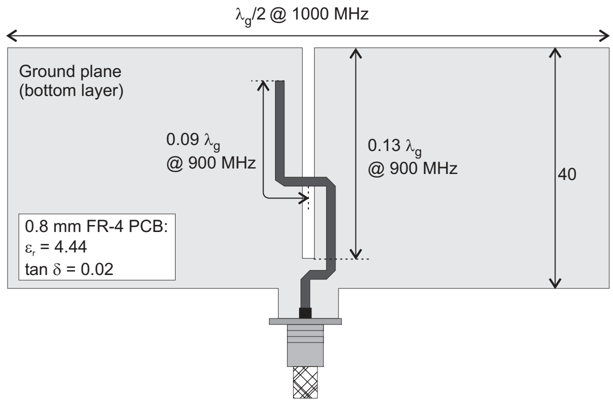


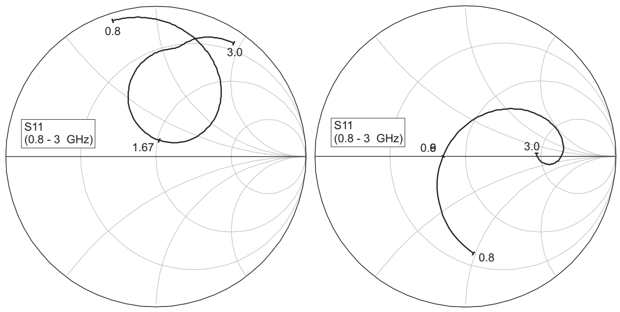
Figure 6.2: Layout of slot antenna and microstrip feed indicating electrical lengths of microstrip stub, notch and chassis.

since been proposed also for low frequency (<1 GHz) UWB applications [97].

The layout of the antenna is reprinted in Fig. 6.2 with the electrical lengths of the slot, microstrip stub and chassis, as provided by IE3D simulations, now indicated.

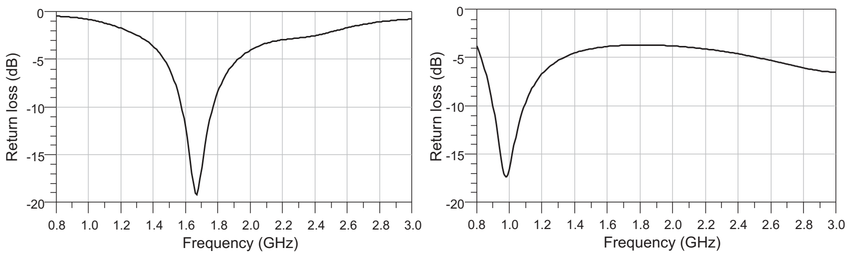
The notch, analyzed by making the ground plane (or chassis) large, is resonant at approximately 1.7 GHz and provides inductive reactance at all other frequencies, see Fig. 6.3a. The chassis, simulated by feeding the two wide dipole halves separated by the notch with a centralized gap source and chamfering each dipole leg to avoid the effects of the slot providing extra shunt capacitance at the feed point (see *Paper I*), is resonant at about 1 GHz, see 6.3b. Corresponding return loss plots are shown in Fig. 6.4. The combination of the two resonators, analyzed using a gap source at the microstrip-slot crossing, is shown in Fig. 6.5 where, as expected, the dual resonant response and the overall inductive behavior is shown. Clearly, the low radiation resistance at 0.8 GHz and 3.0 GHz of the slot is limiting both lower and upper frequency limits of the antenna. Later, a modified version suitable for wider terminals, featuring a longer slot, is presented to reduce the lower cutoff frequency. Also, it can be seen in Fig. 6.5a that series capacitance is needed at the feed point, which is provided by a microstrip stub below resonance (i.e. below 2.22 GHz, see Fig. 6.6a). This stub further provides an extra resonance. However, as all three resonators are strongly coupled and overlaps in frequency, it is difficult to attribute each resonance (e.g. in a return loss plot) to a specific resonator. The input impedance of the complete antenna, with reference plane at the microstrip-slot crossing, is shown in Fig. 6.6b.

One of the main concerns of non-PIFA type antennas is the SAR value and antenna efficiency in talk position, in particular for low frequency cellular bands (e.g. GSM900). As was shown in [59] and *Paper VII*, the SAR value increases with higher coupling to the chassis resonator for internal (PIFA type



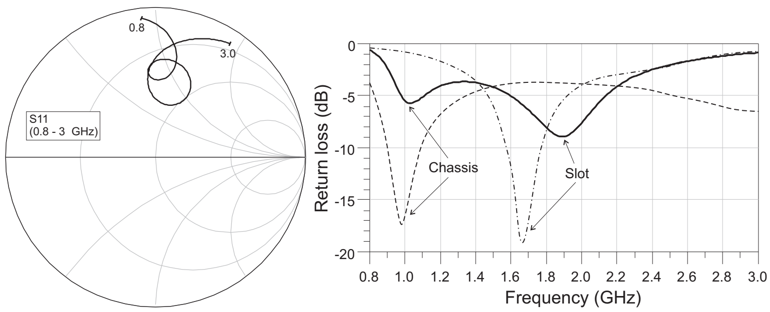
(a) Slot only (b) Chassis only

Figure 6.3: Simulated input impedances of slot and chassis.



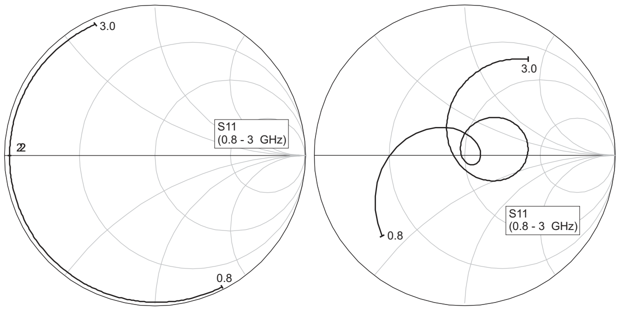
(a) Slot only (b) Chassis only

Figure 6.4: Simulated return loss of slot and chassis.



(a) Slot+chassis (b) Slot+chassis

Figure 6.5: Simulated input impedance and return loss of the combination between slot and chassis.



(a) Microstrip stub (b) Complete antenna

Figure 6.6: Simulated input impedance of the series microstrip stub and the complete slot antenna.

of) radiators. Hence, it is expected that radiators such as the slot presented in *Paper III* that relies solely on the chassis for radiation should have excessive SAR values thus rendering it commercially unacceptable. SAR was measured using a Schmid & Partner DASY4 dosimetric assessment system, similar to the set-up described in [59]. In accordance with specifications, a human hand was not included in the measurement. To provide a representative distance to the phantom head, thin dielectric spacers of height 5 mm (simulating a typical terminal cover) were mounted on the back side of the chassis. The antenna was measured in touch position on the left side of the phantom at 890 MHz where a very low return loss was obtained. A 250 mW CW signal was applied to the antenna, corresponding to +33 dBm in 1/8 time-slots. The SAR distribution is shown graphically in Figure 6.7 with a peak of 4.1 mW/g averaged over 1 g tissue, and a peak of 2.68 mW/g averaged over 10 g tissue. The Federal Communications Commission (FCC) limit for public exposure from cellular telephones is a SAR level of 1.6 mW/g, which is applicable in the US. In EU (and Japan, Brazil and New Zealand) the recommended SAR limit by the International Commission on Non-Ionizing Radiation Protection (ICNIRP) is 2 mW/g averaged over 10 grams of tissue. This limit has also recently been adopted by IEEE [77]. While the prototype slot antennas peak value of 2.68 mW/g over 10 grams is above all recommended limits, the value for an implemented phone antenna could be significantly lower due to additional losses (from the cover and other nearby objects) and also due to a less than +33 dBm output power available from the power amplifier (which is fairly common in reality).

The proposed configuration in *Paper III*, using a naked FR-4 PCB, has a lower cut-off frequency of 915 MHz, making the antenna unable to support GSM850 (824-894 MHz). While the cut-off frequency is expected to be lowered by dielectric loading (from cover and other plastic parts) when

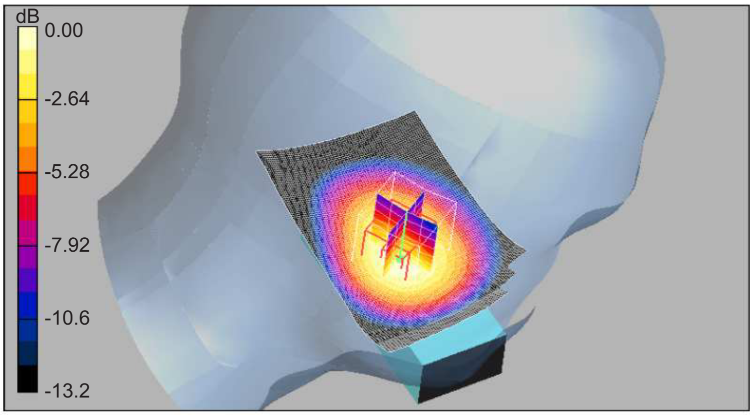


Figure 6.7: Measured SAR distribution of slot antenna.

implemented in a real terminal, it is anyway interesting to examine if/how the structure can be modified to enable also GSM850 coverage. As was concluded previously in this section, the notch length is mainly responsible for the bandwidth of the antenna. By increasing this length and the total width of the chassis by 5 mm, it was possible to significantly decrease the lower frequency limit. The layout of the modified antenna is shown in Fig. 6.8 and the simulated return loss is provided in Fig. 6.9. As can be seen, the bandwidth of this version is (RL < -6 dB) 810 - 2790 MHz.

Concerning the feasibility of this type of radiator in a commercial application, consider the inside of a typical bar-type mobile phone depicted in Fig. 6.10. As can be seen, the phone is already, from a metallization point of view, partitioned into two equal-size sections with the keypad and battery (although not visible here) in one section and the display in the other. Hence, the technique should be commercially feasible at least for some phone models and form factors. Also, many ultra-thin phone models are wider than the standard width 40 mm, so the version suitable for 45 mm wide phones could possibly also be realized in real terminals.

6.0.3 Errata of Paper IV

On page 5, Section 3, line 28, "$\eta_{rad} = P_{rad}/P_{loss}$" should be "$\eta_{rad} = P_{rad}/(P_{rad} + P_{loss})$".

6.0.4 Errata of Paper VI

In Fig. 10, page 11, the input power is denoted S_i. This should be replaced by S_{in}. Also, S_i should replaced by S_{in} in Eq. (4) and (5) on page 10.

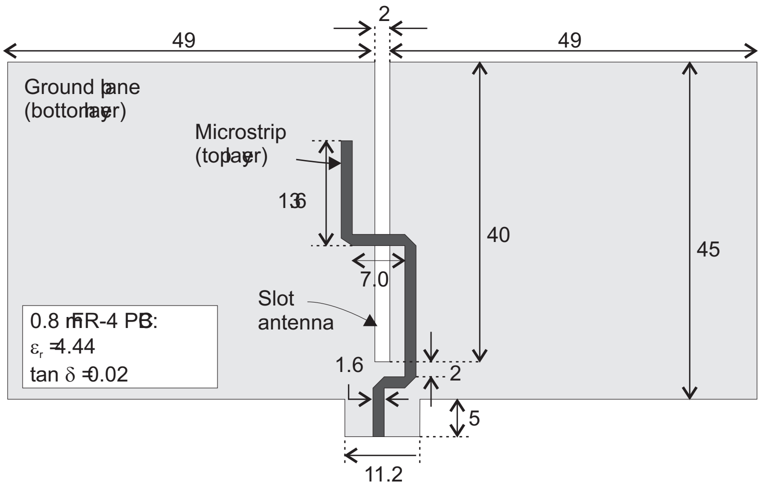


Figure 6.8: Layout of modified slot antenna for wide chassis applications. All dimensions in mm.

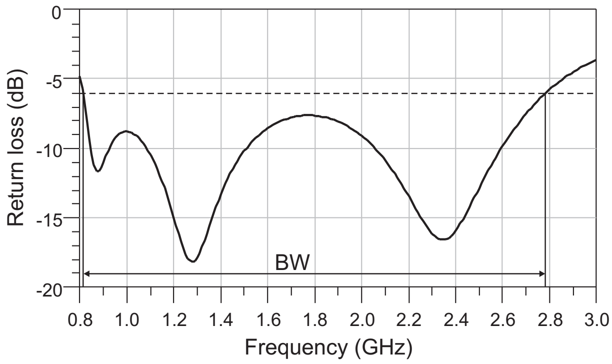
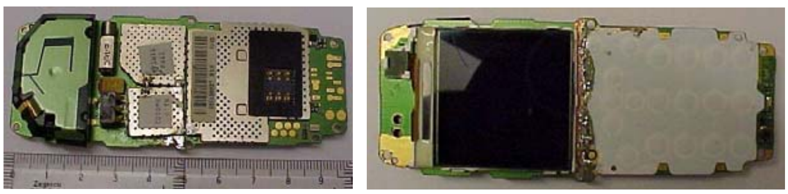


Figure 6.9: Simulated return loss of slot antenna with 40 mm long slot in 45 mm wide chassis.



(a) Back side (b) Front side

Figure 6.10: Pictures of front and backside of typical bar-type mobile phone without plastic cover.

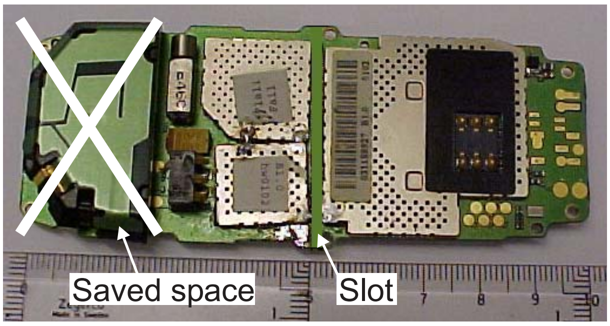


Figure 6.11: Simulated return loss of slot antenna with 40 mm long slot.

Eq. 4, page 10, is misprinted (it spans slightly more than one column) and is reproduced here (with S_i replaced by S_{in}):

$$\begin{aligned}\left(\frac{S_{in}/T_a}{S_{out}/T_{out}}\right) &= \frac{S_{in}(G_{amp}(\eta T_a + (1-\eta)T_0 + T_{amp}) + T_e)}{S_{in}\eta G_{amp}T_a} \\ &= 1 + \frac{(1-\eta)}{\eta}\frac{T_0}{T_a} + \frac{T_{amp}}{\eta T_a} + \frac{T_e}{G_{amp}\eta T_a}\end{aligned} \tag{6.1}$$

6.0.5 Errata of Paper VII

In Table I, page 7, all cells using the % sign indicates true percentages and not 'percentage points'. For instance, the dipole has a 18.4% bandwidth in free-space, which is increased by 45.1% to 26.7% in talk position.

In the introduction on page 1 and in the discussion on page 7, antenna efficiencies of two different antennas are compared using the term "octave", which is typically only used for frequency relations. In the paper, one octave means a doubling of the efficiency.

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