

component

circuit

device

piece of  
equipment

system

# Introduction to RF Equipment and System Design



**Pekka Eskelinen**

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Pekka Eskelinen



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# Preface

Every year, tens of thousands of young engineers and university graduates enter the fascinating professional field of *radio frequency* (RF) design. Most of them have a reasonable understanding of applied mathematics and physics, circuit theory, electromagnetism, and electronics as well as computers and programming. Despite the comprehensive courses and overwhelming educational literature, however, many of these talented young people have to face the crude practical project environment of systems and equipment without much prior knowledge of, or tutorials about, how and why things are done the way they are done. I was once in that situation. Typically, nobody in the office has time enough to explain things—and not that much time to listen, either. Often, young graduates are not acquainted with “neighboring” sciences, because the amount of information is simply too large for inclusion in any reasonable university course structure. The scientific goals of universities might also encourage both students and lecturers to concentrate on relatively narrow topical areas within which the available resources are most likely to yield academic merit. Universities emphasize publications and dissertations rather than organizational skills or system-level thinking.

The target audience of this humble, entry-level book is definitely those young, recently graduated RF engineers. Additionally, university students in the fourth year or so should find the case examples and working schemes interesting. In fact, an elementary course in RF systems design based perhaps on this book and related material might well be justified. My goal is to highlight the problems and selected solutions that make participating in complete radio systems projects so challenging and motivating. Readers may find individual encouragement even in the less successful trials as well. In addition, I hope that readers with a diverse scientific background can make use of the text, which includes examples ranging from mechanical vibrations of antenna towers to computer-controlled test systems. Although the invasion of numerical processing into radio systems shall continue and expand, there will be areas that long remain “RF-proprietary.” In fact, some areas of digital processing are approaching RF in the sense of continuously increasing clock frequencies. Many of the examples in this book have a direct or indirect connection to national defense, which obviously indicates the application area where the most complicated RF problems tend to appear. Nevertheless, the majority of the practices and design principles are similar to those needed for successful civilian communication systems or scientific test instrumentation.



# Acknowledgments

Despite the fact that many of the practical designs included in this book are reflections of my own work and desperate experiments, several young project scientists have had a remarkable effect on the results. I especially want to thank Jukka Ruoskanen, Arttu Rantala, Teemu Tarvainen, Suvi Ahonen, Jussi Saily, and Ville Mottonen for granting me permission to use portions of their research findings as examples.

The majority of my almost 30-year career in RF engineering would have looked quite different if my wife Tuula had not entered the scene. Her support throughout the more and less successful design projects has definitely been indispensable—not forgetting the early years when she had to take care of our sons when their father was “out in the field.” These two youngsters, Ari and Jussi, have not only been a source of inspiration, but also during the past 5 years they have actively participated in selected design tasks, mainly for software, of course. Jussi has additionally edited most of the photographs in this book. Boys, you have made a really good start! In addition, my brother Harri has been an efficient and vigorous coworker in a number of projects and has greatly contributed to many of the mechanical designs for my RF gadgets. I always appreciate his generosity and fruitful ideas.



# Introduction

This chapter aims to clarify some of the fundamental concepts of systems engineering in general and, in particular, their use in RF design projects. First, the chapter defines some elementary terminology and briefly describes essential background information, which the reader should collect prior to delving more deeply into this book. Primarily, this introduction is devoted to showing some of the goals and working methods of systems engineering as applied in various RF and microwave design tasks in civilian and, especially, military application areas. Additionally, the chapter highlights the importance of the reliability and availability aspects of RF—common to most engineering problems—and connects the human user into the technical field of interest. Finally, the author attempts to give readers a short glimpse into the realistic project environment in which a novice RF engineer might be put without prior warning.

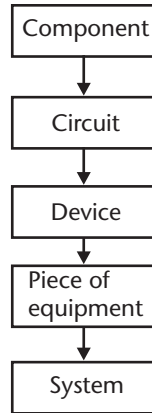
## 1.1 Definitions

Many scientific disciplines try to arrange things into some kind of a logical order and give definitions to functions, features, and processes. Almost every time, such definitions fail—at least to some extent—but the practice continues. Despite apparent deficiencies and inaccuracies, a large number of systematic structures has helped scientific conversation—and caused even more. Most importantly, the education of newcomers into the field of interest has become much easier.

Electrical and electronics engineering has traditionally divided building blocks into four or five categories. However, the meaning of words can still be confusing. In a simple electronic design, the lowest level of hierarchy is often comprised of components, such as resistors, inductors, and transistors. They are connected to form circuits [e.g., a voltage regulator or an *automatic gain control amplifier* (AGC)]. If such an amplifier is put into a physical enclosure and is furnished with coaxial connectors, many colleagues call it a device. Putting together a number of devices, some of which are often actually just circuits next to each other on the very same *printed circuit board* (PCB), makes a piece of equipment. This could be, for example, a receiver (RX), a transmitter (TX), or a spectrum analyzer. Finally, when the designer has a bundle of RXs and TXs, he or she can configure an entire RF system, which might be used to transfer digital terrestrial television signals or to track a potential hostile aircraft. This five-step process is further clarified in Figure 1.1.

An RF system can fail due to a bad resistor or the fact that the operating point of an amplifier has drifted out of the appropriate region. Such things have indeed

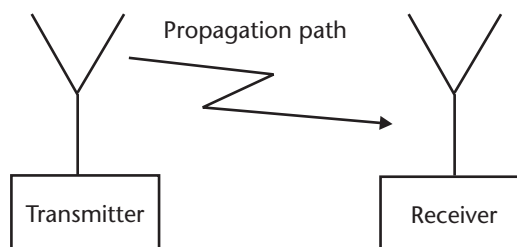




**Figure 1.1** The five steps of design. The boundaries between adjacent blocks are not rigid and sometimes all the different phases are not needed.

happened several times in real life and are sure to happen again. However, this book generally assumes that components and circuit design aspects are covered elsewhere. Here, the focus is on the design and analysis of complete pieces of equipment (e.g., RXs and, to a larger extent, systems such as radar or radio communication networks). The word *system* has had many definitions, some perhaps better than others. *System engineering* is even more difficult to describe in brief, but we might try by saying that it is a clever way of combining the capabilities of different engineering disciplines for a successful result. It also takes into account the varying levels of harmful side effects and attempts to ensure that the set goals are met evenly. Unavoidably, system engineering has to handle tasks in which there is a strong mutual coupling between the various subsectors of interest. An unsuccessful systems engineering endeavor is quite easy to recognize once it happens. It could be described as a set of rocket explosions punctuating a research program [1] or as the wrong interpretation of right-hand circular polarization during the world's first satellite-TV relay session [2]. The technical history of the World War II is full of examples of more and less lucky systems engineering [3] as is the life and operations of the former Russian MIR space station [4].

For this book, Figure 1.2 serves as a good example of an RF system. We normally have at least two antennas, one at each end of the propagating path. One site has a TX, the other an RX. TX building blocks include some master oscillators, a modulation input and an adjacent modulator, and some power amplifiers (PAs). A



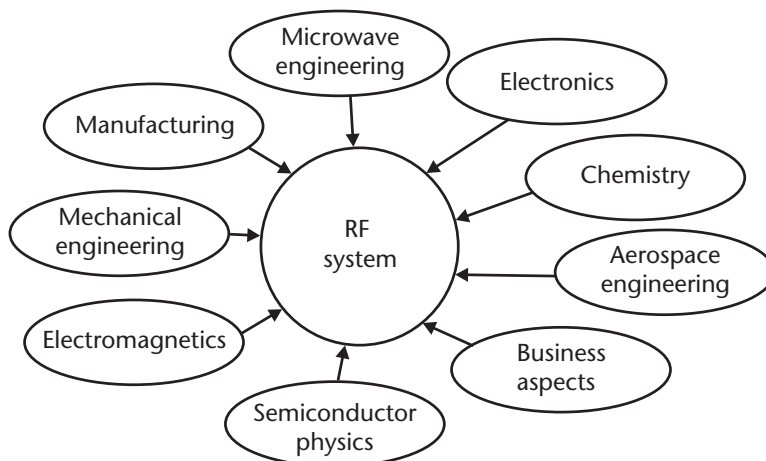
**Figure 1.2** A basic RF system comprised of an RX, a TX, two antennas, and the propagation path in between.

power supply is a must, too. In the RX we often have a low-noise preamplifier; some mixing or downconversion, which again needs master oscillators (normally called local oscillators); a demodulator; and after that a set of baseband amplifiers and processors.

## 1.2 What the Reader Should Already Know

Many RF engineers graduating from universities have obtained a solid background in radio-wave propagation, antennas, transmission line analysis, and electronic circuit theory and also have at least some capability in programming. In fact, because most RF systems are real physical constructions that one can see, touch, and feel, a basic understanding of mechanical engineering fundamentals would be a great help to a reader of this book. Electromagnetic theory forms the basis of most electrical engineering and RF design makes no exception. These topics are all assumed in this presentation to be a foundation for understanding the discussions (see Figure 1.3), although many of the items appear as headings in coming chapters. There, however, the treatment may omit details or focus on a limited topic, thereby giving a misleading impression to those without the proper background. In addition, due to the lack of task-specific people in the radio field, engineers from related scientific fields have entered the radio business and may not have been able to gather all the necessary start-up information. I, therefore, briefly list some of the key elements of each topical area so that interested readers can consult suitable textbooks for further studies.

Radio-wave propagation is generally treated as a set of processes, which start from the relatively simple free-space case—an approximation of which might be a radio link between two deep-space probes traveling in an unobstructed part of outer space. For more usual circumstances, the model is supplemented to take into account the effects of the troposphere and the ionosphere. Reflection from the Earth's surface and from man-made structures must be modeled in most cases as



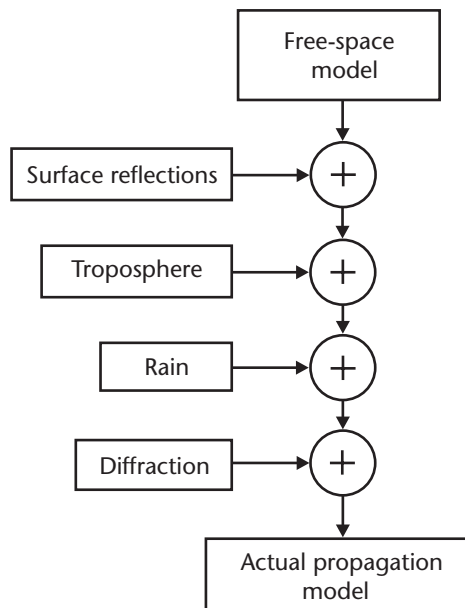
**Figure 1.3** RF equipment and system design relies on a number of different technologies, without which a successful process is hard to maintain.

well. Multipath is the key phrase used in this context within today's short-distance mobile radio communication. Further important propagation issues are diffraction and, more precisely, edge diffraction and smooth-sphere diffraction and ducting, which are occasionally observed as super-long radio connections at *very high frequencies* (VHFs) and *ultrahigh frequencies* (UHF). Rain attenuation and scattering are important special questions, particularly encountered at higher microwave and millimeter-wave bands. Figure 1.4 further highlights the process of dealing with various propagation topics.

RF transmission lines and various components based on them must be analyzed and designed according to their distributed nature [5]. Voltage and current are no longer only functions of time but also depend on the physical location of observation. Therefore, transmission line analysis with the Smith chart and related operations are essential [6]. These include the scattering parameters of two-ports, return loss, attenuation, standing wave ratios (SWRs), and group delay [7]. Impedance matching in coaxial cables, waveguides, and microstrip or stripline structures is another important field [8]. Propagating modes in waveguides, transitions between different guide shapes and forms, and transmission line power-handling capability must be understood.

Many RF systems require antennas to operate properly. In some industrial microwave systems they may be called transducers. Antenna characteristics form the basis of many higher-order performance figures [e.g., *low probability of intercept* (LPI) in military radar or communications]. Moreover, despite the abilities of modern digital signal processing, antennas can sometimes be vital for success. Different antenna types, their dimensioning basics, and pattern parameters like gain, beam-width, and sidelobe level are some of the necessary features [9].

Sometimes the first university courses of electronics can substantially dampen students' interest in circuit design due to their relatively heavy emphasis on



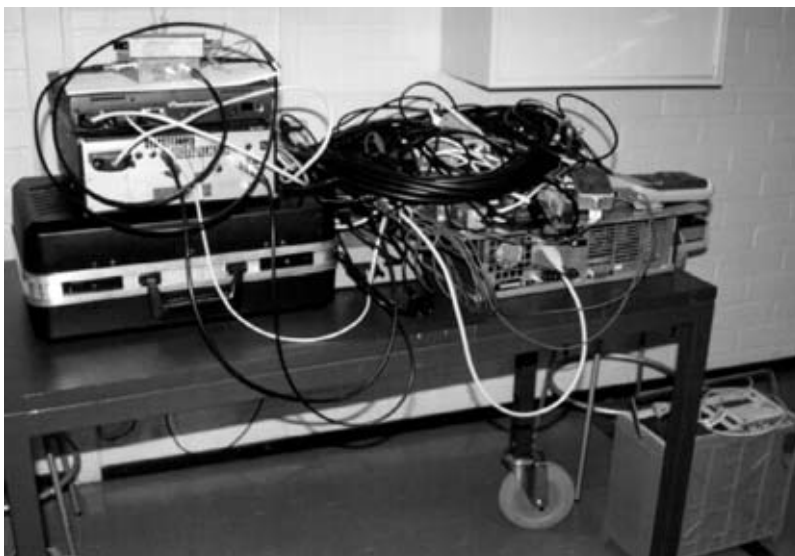
**Figure 1.4** Starting from the elementary free-space model, a system designer estimates the actual propagation characteristics by taking into account supplementary factors.

semiconductor physics [10]. For our purposes, however, the more relevant topics include amplifier [11] and oscillator [12] design principles, power supplies and voltage regulation [13], and different diode detectors. Some understanding of elementary logic design and microwave materials [14] is also helpful. Again, in the systems design phase, it is even more important to have an idea why things happen the way they do—which is very often the most unexpected way. Knowing how to apply, for example, a *field effect transistor* (FET) stage might be more valuable than to be able to precisely calculate the drain current in that circuit.

### 1.3 Style of Approach

Many of my colleagues believe that a proper book in the field of engineering sciences should be based on a strict mathematical formulation of processes and phenomena. In my opinion, such an approach does not necessarily yield a better understanding of the topics but might instead increase confusion. Accordingly, the order of learning and discussion should rather be such that one first acquires a sufficient overall view of the problem and only after that starts to formulate it as a set of complicated mathematical equations. This book tries to follow that principle and takes into account the practical problems of RF projects, often appearing in the form shown in Figure 1.5. A large amount of mathematical manipulations have been omitted or at least considerably shortened. Moreover, a number of factors having a mathematical origin are presented in graphical form only, because this creates a longer-lasting memory for the reader.

Topics in this book are often treated from a problem-oriented point of view. This means that we first define a task to be handled by a specific RF design and subsequently try to figure out what kind of equipment is needed and how it should be organized. My emphasis is on processes and phenomena, and I often describe actual



**Figure 1.5** RF systems work has its practical aspects. This mixed pile of hardware must first be put into operation; only after that we can expect results for numerical analysis.

systems that have been constructed for a specific job. Readers who need detailed information on specific component-level issues should consult one of the many top-quality sources that exist—see, for example, [15]. Unlike some other books in this same field, this text also points out cases where the system design was faulty or even a complete failure. In this way the book attempts to document a technical heritage for the following engineering generations. Just as one historian said, “The purpose of military history is to explain why things went wrong in order to make it possible to avoid the same mistakes happening again.”

I have purposely selected a slightly casual writing style, which I believe will make reading slightly more fluent; it also allows me to tell about some of the less successful experiments in the original style—in other words, in the manner in which they were once discussed internally in the field or in the lab. Maybe this approach is encouraging, too. Readers need to learn that it is not so important if their first radio monitoring RX system does not work initially or gives astonishing output. The early warning radar system of the U.S. military in the past detected the Moon as a hostile target [16], and the British contribution to the world’s first satellite television experiment failed due to an incorrect interpretation of circular polarization. Russian World War II radio-controlled mines, one example of which is shown in Figure 1.6, were ingenious pieces of engineering hardware, but could not blow up the second-largest city in Finland in 1941, because somebody had decided to use the very same broadcasting band frequency for every single detonator, thereby making the whole arsenal relatively easy to jam.



**Figure 1.6** A radio system might easily fail even if its components are perfect. During World War II, these Russian radio-controlled mines all used the same carrier frequency and were easily jammed.

Computer simulation is today one of the cornerstones of RF equipment and system design. Several efficient software products have been released by commercial vendors, and some large enterprises have even had resources to develop and maintain their own. This book will, however, not discuss RF design software issues except in a couple of examples where something special once turned up. I have chosen to emphasize the more physical side of things and devices, and incorporating anything useful from the software world would have doubled the book's number of pages.

## 1.4 Goals in System Design

It is relatively common that an RF systems project has in the broad sense a multitude of targets, only some or one of which is actually known to an individual designer. This is particularly the case if the project is large in terms of time, manpower, cost, or geography, and if there are lots of newcomers in the team. First of all, there is or should be, in some cases, a “pure” technical goal or a set of technical goals. Here the word technical means something that can be described as an RF parameter or as any other technical parameter. This might be, for example, a TX output power of 1 MW at 200 GHz—a task in itself. On some rare occasions, the specification may be as vague as “being the best—no matter what it costs,” which could be the case in military electronic countermeasures or which was the case when humans first went to the Moon.

However, a real project environment normally has additional goals that cannot easily be put into technical form. Currently, one of the most frequently encountered goals is financial. It can be defined as the lowest manufacturing cost per unit, the largest revenue per year, or, in some government projects, just staying within the budget. Another issue is time. Almost every real-world technical project has a definite deadline before which the desired results must be available. Only work near or in the fundamental research area can enjoy partial freedom from schedules. Thus, system design normally has to achieve the primary technical goal but at the same time meet other restricting requirements. Unfortunately, poor management can lead to a case where the technical goal is intentionally or unintentionally neglected in favor of budget or schedule. An experienced project manager should also understand that design engineers are primarily motivated by the technical challenges and, if left working alone, will surely use all resources to meet them.

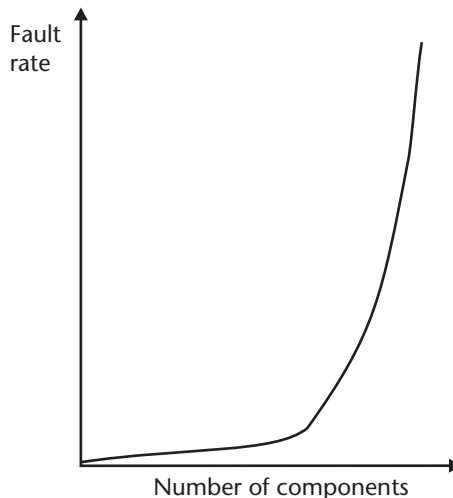
## 1.5 The Spirit of System Design

Before jumping into the individual parameters that influence the performance of RF equipment and systems, we can first briefly outline some very general statements governing the task area. To start, a normal systems project seldom has surplus time. This means that every effort must be taken to speed up the design and evaluation phases. In this sense, nothing new has appeared since the 1940s. “Keep it simple” is a very good general working motto that not only speeds up a process, but also reduces the number of faults. The fewer the elements, the fewer the things to break

down. This effect can be quite dramatic, as illustrated in Figure 1.7. The size of the project typically has a drastic effect on the final output in terms of performance figures. A system can often be described as a compromise or a collection of compromises. Maintaining the direct current power limitation might mean a slightly lower output power, or reducing the rack height to fit the aircraft cabin could imply throwing away a couple of secondary displays. Amused colleagues and coworkers can even suggest that a system is made of mistakes, but that should be an exaggeration. Nevertheless, we should remember that a median system has some mistakes in it—although we hope not very many. The key thing is that the system we are discussing must be able to deal with the “built-in-faults.” This is of paramount importance when tracing the weakest link in a process chain. A low-rated fuse in the wrong bus will jeopardize a whole space mission.

Overengineering is another threat. This not only consumes time, manpower, and money but also often will deteriorate the overall quality. A too sophisticated system easily has a lower *mean time between failures* (MTBF) and a much longer *mean time to repair* (MTTR). A better way to work is to optimize performance so as to meet the target with a suitable margin but well within the expected time. Recently, enthusiastic software designers have been very keen on continuing their activities far beyond the practical. Care should be taken not to ignore the cyclic nature of a design activity. Sometimes, depending on the complexity of the task and the experience of the team, a total relaunching of a mission is mandatory. In long-term projects this can cause further harm due to the rapid renewal of modern semiconductor components. Actually, it is very typical that a large system has components or devices with varying levels of novelty and that the system itself is not necessarily as up-to-date as some of its individual blocks.

A reasonable rule of thumb for commercial systems design is to take performance from where it is cheapest. This indicates that if, for example, we have problems in meeting a radio link distance requirement, we could increase TX power, lower the RX noise figure (NF), increase antenna gain, or change modulation. If no other constraints exist, we might figure out which of these four choices gives the needed



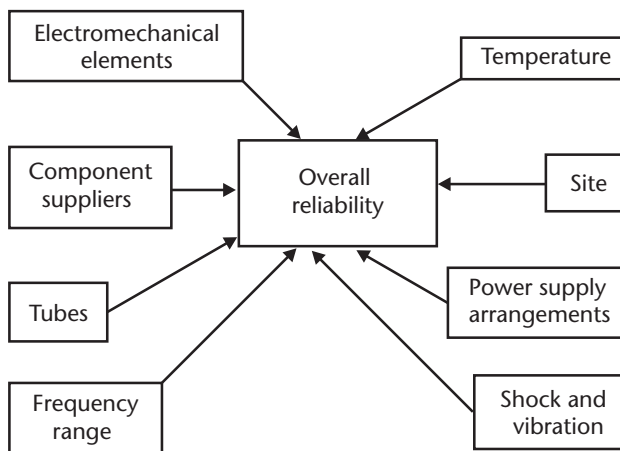
**Figure 1.7** If the number of blocks or components in a system increases, the occurrence of faults grows, too. However, the function is highly nonlinear and depends on the application.

improvement at the lowest cost. Naturally, as will be demonstrated later, the problem is not that simple. Modulation and coding may be bound to standards. Antenna size has an adverse effect on tower stress, and TX power may put too much of a burden on the batteries. Moreover, working at the lowest practical output power level shows very good RF engineering skills.

The author's home country has been one of the few nations that has been able to test and integrate both western and Soviet-based equipment into complete systems, mainly but not solely for defense purposes. These projects have shown a number of astonishing similarities in thinking—despite the cultural and political discrepancies—but have also highlighted a couple of notable differences. In the years of the Cold War, NATO authorities often used a time delay parameter to describe the gap between Russian and western electronics and weapon technology, but as we now see, that might not have been completely justified due to the adopted system concepts. For example, the maintenance principles of aircraft looked very much the same regardless of manufacturer. Large nations can use practically endless amounts of manpower and organizational effort to run a depot, whereas small countries have to adapt to the available number of men and women. On the other hand, the trend of using individual subcontractors has pushed western electronics more toward internal interoperability, which is also of great benefit if upgrades have to be made in the field or if supplementary units from third parties have to be added. Pieces of former Soviet equipment generally offer few possibilities for later fine-tuning unless their owners are willing and prepared to perform major refurbishment actions.

## 1.6 Reliability and Availability

If some very exceptional scientific instruments are excluded, the user community generally expects a certain level of operational reliability and availability from any RF system or device. Both parameters are fundamentally defined during the initial design process although some of the designers may not recognize the fact [17]. Some of the factors affecting system reliability are shown in Figure 1.8. The selection of operating principles can already be important (e.g., rotating reflector antenna or an



**Figure 1.8** Selected factors that influence the overall reliability in an RF system.

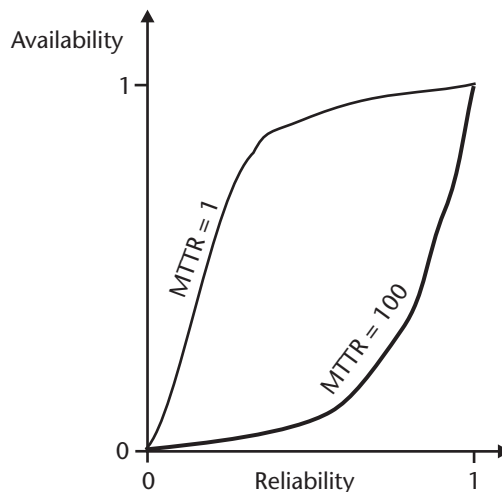


adaptive array in a gun-laying radar), and components have a definite role, too. Designers can often select such parameters as the operating voltages and temperatures. Availability is connected to reliability but depends on the amount of and time needed for essential corrective actions [18]. If the MTBF is low, reliability is bad, of course, but if at the same time the MTTR is very low, the overall availability can be acceptable. This is highlighted in Figure 1.9. In some other contexts (e.g., air navigation) availability is seen, for example, as a function of the geographical area or volume. Such issues are discussed separately.

## 1.7 Effects of User Profile

Typically, project managers and team leaders should take care of the proper design philosophy with respect to the expected user community of a specific piece of equipment or an entire radio system. Almost regardless of functions, features, or frequency ranges, entertainment gear has its scope of user interface character [19] just as military systems have theirs. If properly understood, a project creates equipment that not only fulfills the primary technical specifications but also provides end users with a friendly, suitably dimensioned man-machine interface. If setting up a simple VHF two-way voice radio link, we might as well omit the *graphical user interface* (GUI) and extensive *bit error rate* (BER) test facilities.

The fundamental question is how much a potential user is assumed to understand about the working principles of the system to be designed in our project and, additionally, what is the wanted or needed level of operator intervention in the system usage. The more possibilities that are given, the higher the probability of errors and technical difficulties is [20]. On the other hand, if we as engineers design a system to be used by other engineers, we can anticipate a lot of talent but—generally—minimal sympathy in the event of malfunction. Despite of extensive training, military troops and individual soldiers or officers can perhaps not be treated as technical professionals, but their user environment sets very high requirements (e.g., for



**Figure 1.9** Availability and reliability in a system are connected, but if the time for each corrective action is very short, even unreliable systems can have reasonable availability.

thermal and mechanical sustainability). Broadcasting people are a diverse breed because both highly qualified engineers and artists will use a system, depending on the specific application and production team. Of course, large broadcasting station networks do have proper engineering manpower but what about a small, local one-person station?

In addition to carefully considering the operation of a system, we must take into account the possible challenges in setting it up or configuring it. This means, for example, that an RF transmission line configuration should be designed for easy on-site assembly and that antenna towers of mobile tactical military microwave links should be erectable without hydraulic cranes. Special requirements concerning mounting places (i.e., loading and fixtures) are often frustrating if one has to set up a satellite ground station on top of a 1920s building, the roof of which may be a collection of tar-coated wood chips!

## 1.8 Project Working

Several comprehensive and up-to-date handbooks and manuals are currently available for a novice project engineer. It is, therefore, not necessary to present here a thorough set of instructions. Instead, I want to point out a couple of topics that may sometimes have specific importance in an RF systems project.

The larger a design project is, the more important is an exchange of information and design data between team members. This is particularly true in a radio project, because most blocks of a system have a direct influence on others [21]. For example, raising the RX NF will generally require more TX power or a larger antenna. An antenna having higher gain needs a more sturdy support, which no longer fits on the original trailer, and thereby necessitates a change in the tow vehicle. This, however, means that the system does not fit into the cargo room of the transport aircraft, and finally the system is not transportable anymore.

If personal relations are not managed, cooperation will easily cease. We engineers inside the technical sciences carry the stigma of possessing poor communication skills. Individuals are—and in my opinion should be—selected for projects mainly based on their scientific and technical merits, rather than on the width of their smile. Still, suitable constructive humor and a positive working attitude are often all that is needed to maintain a fruitful working atmosphere. It is sad indeed that many of us more experienced fellows do not give the younger team members a proper chance to show what they can do. Personal encouragement, which can be just a couple of kind words to recognize good work, does not cost anything but may improve daily creativity significantly. In a further word of warning, complete misunderstanding is not as rare as we might think. Did the team members really get you right?

Dedicated special components—particularly those coming from overseas—are a continuous cause for worry. If possible, avoid them. You might get a couple of samples and start designing based on them but when the real need comes, the vendor may turn totally silent or the delivery times may easily exceed the remaining project duration, no matter how long it is. Then, in the end, when you get something, it may not be the same component or block anymore. Homemade specialties

are no better: Unless technical specifications clearly dictate creating such curiosities, stick to the proven concepts. Even the slightest modification to an existing in-house building block can take months or years. As organizations and institutions do not have memory, design data tends to walk out of the lab forever on the very day that a coworker decides to quit or is happily retired.

Therefore, documentation in all forms is essential. Each design detail must find its way to a permanent medium—that is not a computer hard disk. Despite the fact that paper drawings and calculations are easily mixed up on a crowded office table and may even disappear forever, the disaster caused by a PC collapse is much worse. It is mandatory to document administrative data in addition to technical facts. Make accurate notes of who said what, when something should be completed, and whose responsibility a particular task is. A well-known joke says that “in the end of a project, those who did nothing will draw applause, and those who did it will be doomed.” Be sure that you are not one of those who are doomed!

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# Available Parameters

This chapter examines different possibilities for adjusting the difficult system matrix of inbound and outbound signals and the propagation path in between. First, the chapter provides some information about current standardization and regulation schemes both for military and commercial usage of RF signals. Subsequently, the chapter discusses each main parameter in turn. Carrier frequency or frequency range is perhaps the first and most natural thing to choose, even though this is one of the most severely restricted things to play with. TX power has similar limitations. Much more can be done in the RX, where NF issues and RF transmission line arrangements become important. Transmission lines are naturally involved in the transmitting blocks as well and could thus be utilized. The geographical environment in which our system should work sets difficult requirements but can sometimes be seen as a design option as well. This is true both in communication networks and in radar systems. Naturally, the baseband signal, modulation and demodulation concepts, and, briefly, signal processing all have their input as well. Finally, there are a number of factors outside the RF engineering discipline, through which system performance can be easily enhanced or degraded. Many of these fall inside the discipline of mechanical engineering.

## 2.1 Standardization and Regulations

Fortunately most RF equipment design activities are currently controlled at least partly by two regulatory functions. First of all, the usage of RF frequencies, modulations, and output powers is strictly guided by international telecommunication bodies, such as the globally working *International Telecommunication Union* (ITU), the *European Telecommunications Standardization Institute* (ETSI), and the *Federal Communications Commission* (FCC) in the United States. Those not working, for example, on a classified military countermeasures project, should consult first these specifications. However, it is quite obvious that a jammer design team possibly does not obey such limitations.

The second level of harmonization comes through system-specific standards, which typically are one or two orders of magnitude more detailed than those given by telecommunication authorities. Typical examples are the *Groupe Spéciale Mobile* (GSM) and *Universal Mobile Telecommunications System* (UMTS) standards for cellular communications and secondary surveillance radar specifications from the *International Civil Aviation Organization* (ICAO) and applied by national bodies like the *Federal Aviation Authority* (FAA) in the United States. Naturally, if

the system to be designed falls into one of these categories, one should check the respective documentation first.

## 2.2 Frequency

RF spectrum is a scarce natural resource and should thus be treated with utmost care. The available spectrum extends currently from about 10 kHz up to 400 GHz and even above. As indicated in Section 2.1, government authorities and international regulating bodies have tried their best to prevent interference and the waste of frequencies, but the final responsibility rests in the design office. Fortunately, two distinct cases exist. If equipment or systems are to be designed according to an existing standard, for example into a *secondary surveillance radar* (SSR) network [1] or to the UMTS [2], the designer has practically nothing to choose from in terms of carrier frequency. Interoperability must be maintained. Large portions of the entire spectrum are actually allocated to specific services and functions. There are wide gaps that are “not in use,” but they are actually there because current component technology does not allow feasible (cost-efficient) systems to be constructed. Once the components exist, systems start to appear. The good thing here is that the regulatory authorities have taken or are about to take responsibility for *electromagnetic compatibility* (EMC), and as long as the design complies with the regulations, there is only a minor risk of problems.

The situation is much more difficult when we are dealing with something totally new, a system or piece of equipment that does not have existing counterparts to work with and whose mission requires a decision of operating carrier frequency or frequencies. Such a situation is common to different scientific measuring instrumentation and to various military systems. As far as can be estimated from unclassified publications, different nations follow slightly different practices in defining frequencies for these tasks. In some regions there seems to be a predefined set of bands across the entire usable spectrum reserved to scientific and military use. Others have apparently preferred to rely on the use of “adequate” shielding. This means tests and evaluations in totally closed test chambers preventing *electromagnetic interference* (EMI). At the same time they give the option of experimenting with unexpected operating frequencies as seen by a potential enemy during an open military conflict.

The carrier frequency, or perhaps more appropriately the operating frequency of an RF system, has several mutual interconnections to other performance-related parameters. The obtainable transmission path length depends on the frequency in use. If we want to push a wide information bandwidth such as live uncompressed video or high-speed data through our link, we should consider a proper relation between baseband and carrier frequencies if we do not intend to create an *ultrawide band* (UWB) system. That is, high data rates are well-suited to high carrier frequencies [3]. Generally, system design and circuit design in particular gets difficult at higher frequencies due to the evident distributed nature of transmission lines and even components. Physical elements are often more expensive at higher microwave and millimeter-wave frequencies and getting high output power or low NFs tends to be more tedious. On the other hand, antennas and other hardware as well are typically a lot smaller if the wavelength is short. One trick is to balance these

interconnections in the most favorable way. Note, however, that the yardstick for success can be a pure technical characteristics or some more derived parameter such as survivability in combat or cost of series production.

Despite the fact that frequency selection in these cases is not only (and sometimes it is to very small extent) a technical matter, I show next a brief scheme that could be extended to give at least intelligent guesses for frequency selection. The following example is based on the assumption of a monostatic [4] tactical battlefield radar, but similar thinking can be adapted to a communication system. Let us assume that we want to maximize the *signal-to-noise* (S/N) ratio in the RX input by selecting the most suitable carrier frequency. [Note that sometimes communications people use the *carrier-to-noise* (C/N) ratio, but C/N would stand for clutter-to-noise for the radar community. Therefore, we here use S/N, but it is measured prior to detection.] The radar's *intermediate frequency* (IF) bandwidth is 100 MHz, which is rather suitable for target detection and gives quite nice possibilities for signal processing in the digital part of the RX. The radar uses medium pulse repetition frequency (PRF) and pulse compression. Doppler processing is foreseen as well. Let us further assume that we want the radar to be mobile and that its antenna's maximum diameter or dimension should thus be about 1m.

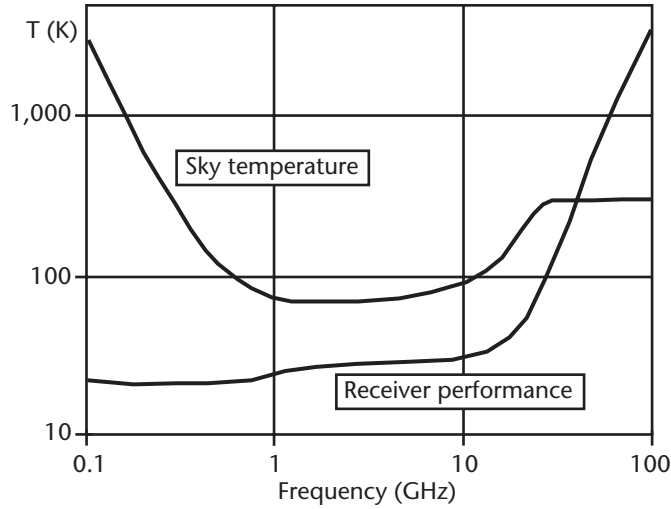
The things we should take into account as a function of carrier frequency include the following.

- Free-space attenuation;
- Antenna gain;
- Target's radar cross-section (RCS);
- Atmospheric attenuation (no rain);
- Rain effects, including backscattering;
- Available TX power;
- Available RX NF.

The way we proceed is to create a tiny database that contains numerical values for those parameters that we cannot express in closed form. After this, we define semi-empirical approximate equations for all parameters and combine them to get one graphical presentation of S/N as a function of frequency. The data that follows has been collected and edited from multiple sources and represent the author's synthesis of the current state of the art. Only unclassified references have been included. The reader must observe though that in most cases there are exceptions to the suggested "typical" values, and in certain cases respective examples will be highlighted.

Selected initial presentations of available NF values, TX output power, and target RCS [5] are shown in Figures 2.1 to 2.3. One needs to examine the RX NF together with the prevailing external noise contribution coming from the antenna input. This power is composed of man-made and natural background noise and often expressed as the equivalent noise temperature. For this reason Figure 2.1 includes a second curve, where the inherent RX noise temperature is compared to the power coming from the feed as a function of frequency. This procedure has been adapted from [6]. Here, sky noise characteristics for an antenna elevation of 1° are

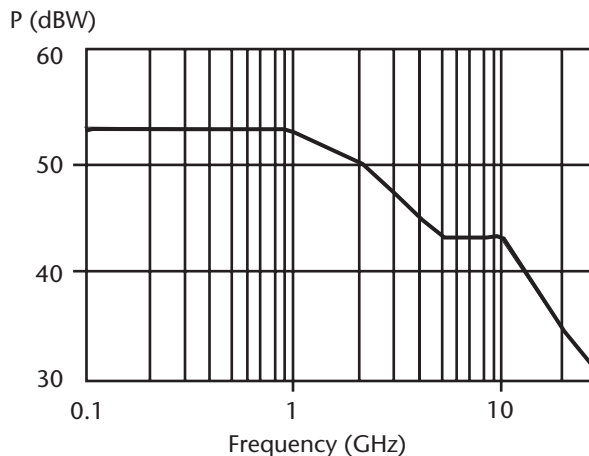




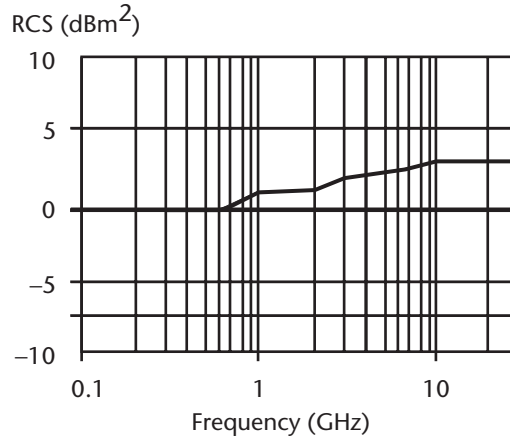
**Figure 2.1** RX noise behavior as a function of operating frequency on a logarithmic scale. Environmental noise exceeds component performance up to the lower microwave bands as demonstrated further in Figure 2.9.

used. It turns out that front ends better than 1 dB can seldom be fully utilized below the Ku-band except in some high-elevation satellite downlinks. Many millimeter-wave RXs achieve much better NFs, if they are cryogenically cooled to, say, 20K or so. Such dedicated devices can be mostly used in space probes traveling to outer space.

Also, TX output power must be judged carefully. Figure 2.2 shows average powers that are generally feasible in mobile or transportable systems. It does not, however, illustrate the ultimate limits of technology at a particular frequency band. Fixed stations have for decades achieved almost 10 times these levels: the missile site radar of the Safeguard system operated in the S-band with 300 kW average power, while the haystack radar operates in the X-band with up to 500 kW average power [7]. There are other specific examples of higher powers in various



**Figure 2.2** Transportable radar TX average output power possibilities as a function of carrier frequency on a logarithmic scale. Different cooling technologies are combined in this graph.

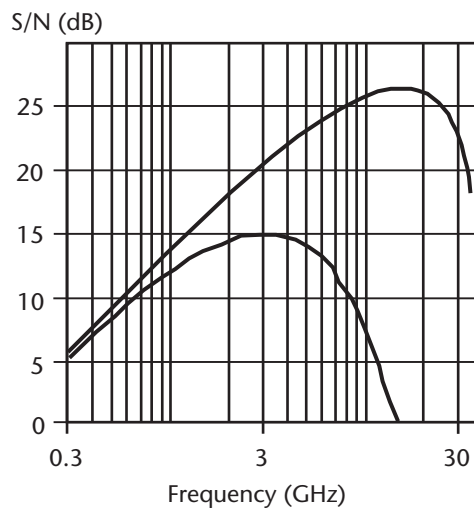


**Figure 2.3** The median RCS of a fighter aircraft at different frequencies (300 MHz, logarithmic scaling) as given by commercial simulation software. A median value of  $1 \text{ m}^2$  is justified.

bands, including millimeter-wave bands. The issue is one of size, weight, and economics, rather than pure technical feasibility.

Certain nations have continued or relaunched radar development at the VHF and UHF bands. This is partly aimed against RCS reduction techniques, because lower frequencies will usually generate structural resonances in the airframe, and effective absorbers tend to be far too large to be a feasible solution. Figure 2.3 cannot show any simulated results of these radars, because the software set in use did not cover frequencies below 300 MHz. However, work documented elsewhere [8] suggests that much larger RCS figures are to be expected, perhaps up to  $20\text{--}100 \text{ m}^2$ .

What we get out of all this is shown in Figure 2.4, excluding backscattering from raindrops. There are actually two different cases corresponding to two target

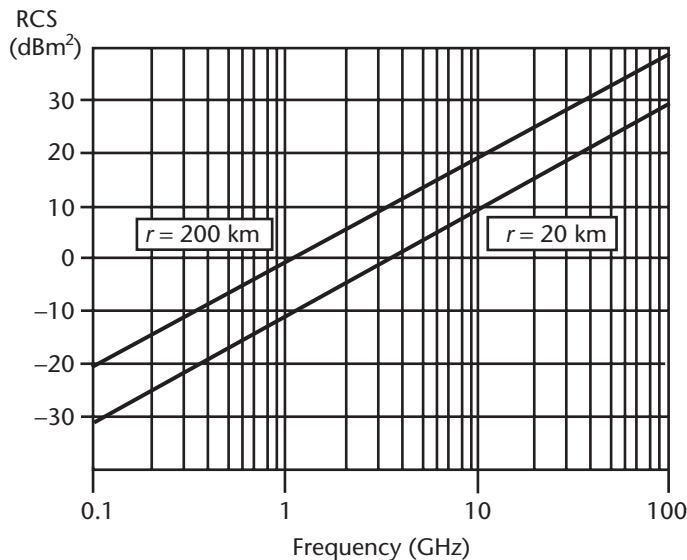


**Figure 2.4** Radar RX S/N before detection as a function of carrier frequency (logarithmic scaling) for target distances of 20 km (upper curve) and 200 km (lower curve). For short ranges, 5 to 25 GHz might do but for longer target distances 2 to 7 GHz give the best possibilities. Constant antenna diameter (1 m) and target RCS ( $1 \text{ m}^2$ ) have been assumed.

scenarios. If we want to build a gun-laying radar, an approximate maximum distance to the potential target could be approximately 20 km. A surveillance radar, on the other hand, needs much more—at least 200 km. We observe that a very short-range radar might well use a frequency anywhere between 5 and 25 GHz and give the same S/N (assuming a fixed antenna size) but for longer ranges something between 2 and 7 GHz gives the best results. Note that I do not say anything about the possibilities for handling the available power in such a small antenna and that I have also excluded the classified data for such factors as coupling losses. Practical radar frequency selection is unfortunately not as straightforward, because we have to consider such parameters as antenna scan rates, physical sizes of antennas, spatial resolution, and clutter rejection. Lower frequencies are somewhat easier in this respect. If a radar must fulfill several functions, the carrier frequency range becomes a compromise. This is the case, for example, in vehicles that have only one radar system available.

Backscattering from rain changes the situation drastically. The effective RCS of rain as seen by our fictitious radar is plotted in Figure 2.5 for two distances and two antenna sizes. We observe that a moderate rain of 10 mm/hr creates an RCS comparable to that of our median target already at 7 GHz (shorter distance) or at around 2 GHz (200-km distance). Of course, this treatment did not take into account any processing functions in the RX.

A second way of approaching frequency selection is based on the fact that in most cases the physical size of an efficient antenna has to be comparable to the operating wavelength or larger. If we work at 100 MHz and the wavelength is thus 3m, we hardly get good performance from a tiny 5-cm whip. This means that higher frequencies yield smaller antennas. As long as we are operating below a few gigahertz, there is nothing that would seriously disrupt this scheme, but when going higher, we

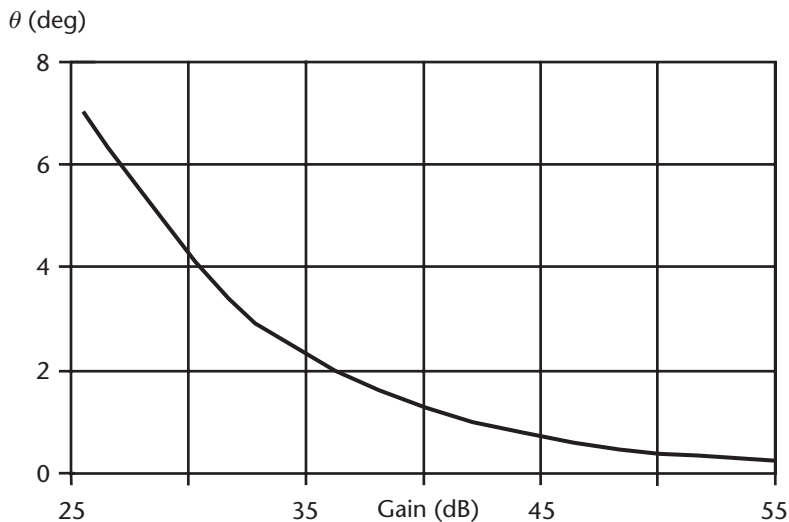


**Figure 2.5** RCS of rain as a function of radar frequency (logarithmic scale). Rain intensity is 10 mm/hr and two target distances, 20 and 200 km, are evaluated. Assumed pulse width is 100 ns, which defines the “length” of the rain clutter volume. The two curves differ by only 10 dB because the antenna diameter assumed for the 200-km radar is 3.3m instead of 1m.

observe the increasing attenuation in the propagating media and that caused by vegetation and man-made obstacles. Also, as seen in Figure 2.2, we start losing output power. Although the plot is for the maximum power, it is rather correct for sensible power, too, if scaled by about 1/100. Alternatively, we get more antenna gain by increasing the frequency if the size of the antenna is kept constant. However, this will cause a potential drawback as well, particularly for mobile communications applications. The beamwidth of an antenna will generally get smaller when the gain increases (see Figure 2.6). Now, if the alignment of the antenna is not perfect or is moving (e.g., to platform vibrations) we may not get the full *equivalent isotropically radiated power* (EIRP) toward the receiving station. Here, the term *effective radiated power* (ERP) is often used, too.

Sometimes the frequency selection is dictated by components only. For example, if we want to construct an all-digital system up to the input of the PA, the *digital-to-analog converters* (D/As) in use must be capable of producing the wanted output waveform with the required resolution, and, of course, in the respective RX we must have sufficient speed and dynamic range to make the thing work at all. At the time of writing, 1 to 2 GHz is feasible in direct digital processing but higher microwaves still wait new component technologies for mass-produced radio systems.

Scientific test instrumentation needs frequencies that are most suitable for the interesting phenomena. This means that there must or should be some a priori information about the characteristics of the process into which the new system should appear. If, for example, we are interested in detecting small items buried in a dielectric substance, it might be a good idea to select a wavelength that corresponds, for example, to the dipole resonance of those small particles. For spherical objects, the wavelength could be equal to the circumference of the particles. Alternatively we have to design a wideband tunable device, which typically increases overall costs at least by an order of magnitude if not more and makes frequency-related components (e.g., amplifiers, oscillators, and modulators) rather complicated.



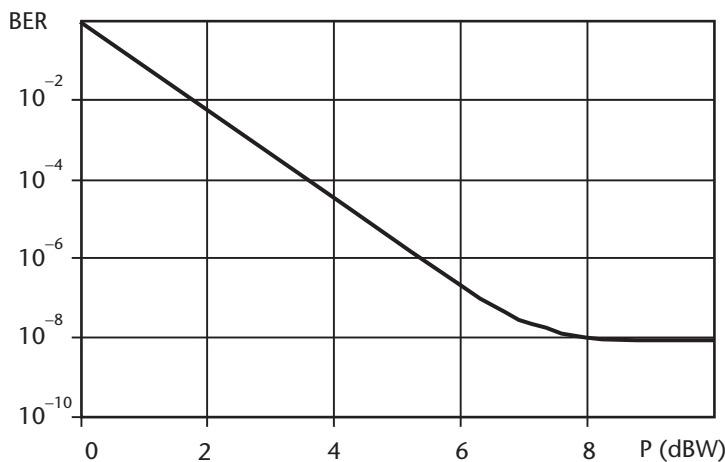
**Figure 2.6** Antenna beamwidth as a function of gain. A parabolic reflector antenna with a horn feed and about 55% efficiency is used here as an example.

## 2.3 Power

We now examine the output power in a system. First, we must distinguish between average power and peak power. Communication equipment and frequency modulation (FM) broadcasting, for example, use TXs whose output power stays fairly constant regardless of the length of an observation interval. Pulsed radar units, on the other hand, must have a peak power specification as well. Some special communication systems—and the ubiquitous *amplitude modulation* (AM) radio and television—use the term *peak envelope power* (PEP) or just envelope power to describe the variation of output amplitude due to modulation. Typically, we have some kind of a PA that is connected to an antenna or a transducer. As long as available components allow, we can in principle play with this parameter to get the desired overall performance. In practice, however, when high-power TXs are used in radar or in satellite uplinks it is because such powers are needed at a reasonable cost in addition to high-gain antennas to meet operational requirements.

There is a wide range of products and designs to choose from, ranging from about 0 dBm at the very high millimeter-wave devices up to several megawatts at the microwave and VHF and UHF bands. Higher power means that we can tolerate more propagation loss (one-way in communications and two-way in normal radar). Alternatively, if the loss is kept constant or can be assumed as such, we have a better C/N ratio at the receiving site, which typically yields to an improved S/N or BER after detection. Depending on applied coding and modulation, the results vary a lot, but Figure 2.7 illustrates one possible relationship between power and error rate. Alternatively, if we expect hostile jamming, more power allows possibilities to overcome the adverse effects.

The efficiency of conventional RF output stages is very low; a rate of around 30% is often considered to be good performance. Of course, this figure strictly follows the operating frequency. Nevertheless, an increase in the RF output means a higher burden for the power supply but also higher thermal stresses all over the equipment. An increased power is often a reason for spectral impurities caused by the overwhelming nonlinearities of PAs whereby RF pollution becomes a problem.

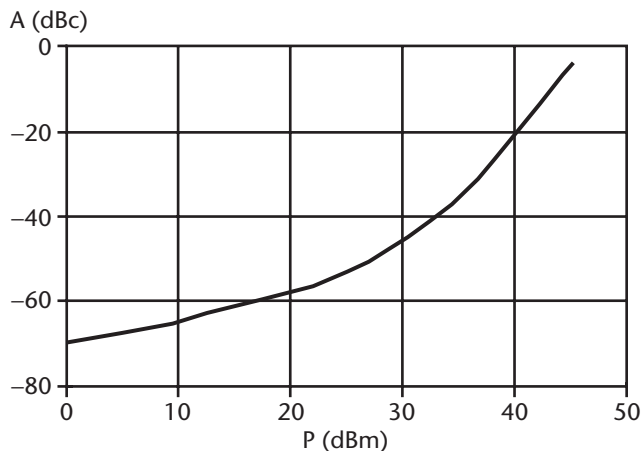


**Figure 2.7** If other parameters are kept constant, an increase in TX power will give an improved BER in the RX. The exact characteristic depends on the selected modulation and coding algorithm.

Figure 2.8 shows one example of unwanted spectral components as a function of wanted output power for a specific RF amplifier device, but should not be understood as a technical limit. There are many radars, for example, whose output powers operate at a rate that is tens of decibels higher and that show considerably lower harmonics. Unfortunately, such factors as the aging of power supply filters may turn a perfect design into an unintentional jammer. Communication and broadcasting gear tends to have somewhat lower spurious levels, too, but those units seldom handle hundreds of kilowatts.

Even if unwanted signal components are not generated, a higher output power means that we prevent or hamper the use of the same frequency in a wider geographical area. Actually we often work in a three-dimensional volume, and other potential users may well be flying above us. In military systems, we easily give the hostile opponent better chances of finding our transmitting site and thus provide a suitable target for surveillance RXs and *antiradiation missiles* (ARMs). It is very well possible that we also make our own system design and later the operation of this system more complicated due to increased internal interference problems.

Recently, the interest in *specific absorption rate* (SAR) and related human health issues relating to RF signals has been enormous, particularly when related to mobile communication *base stations* (BSs) and cellular handsets. Public discussion has been fierce, but at the time of this writing, only thermal effects on the human body have found scientific proof. Extensive and costly studies continue. Radar people have been familiar with the topic since the late days of World War II, and appropriate precautions, based on thermal effects, were taken long ago. The connection between output power density and SAR is self-evident. Making the power density less will also drop the SAR, and the potential risks to human safety will be minimal. In some rare cases the real problem is not the direct radiation from an antenna interface or the antenna itself. An inadequately shielded very high-power klystron or magnetron output stage in the immediate vicinity of the operator room can have sufficient stray X-ray (ionizing) radiation to be a safety concern. Normally, the power densities that come from the effects of the power stage and the antenna, which have been found technically and economically feasible, are not known to



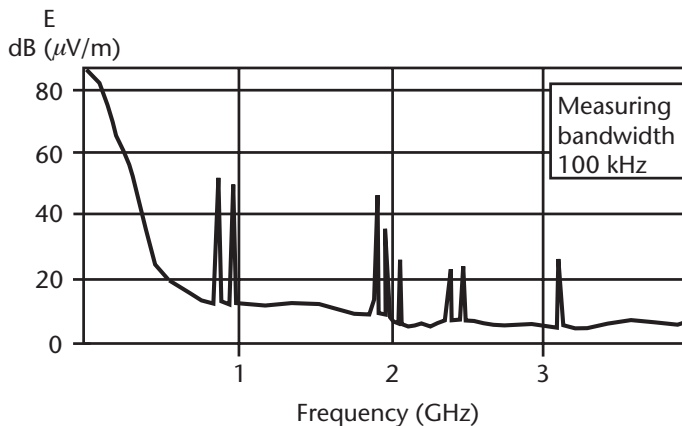
**Figure 2.8** An example of PA harmonic output as the average output power is increased. The level of harmonics is shown as decibels referred to carrier (dBc) scale.

have caused any health problems to the general public. Putting one's head into the feed horn when the TX is on is another story.

## 2.4 NF

This book uses the traditional way of dealing with NFs and assumes that it is a characteristic of a receiving part of a system—be it radar or communications or a scientific test instrument. Very often there is a special *low-noise amplifier* (LNA) as a first item in the block diagram, but sometimes this performance figure is entirely defined by factors such as the mixer conversion loss or the passband attenuation of the pre-selector filter. In numerical values, practical NFs vary from something below 0.3 dB up to 6 dB and even more. In wideband equipment we may find it hard to achieve levels below 12 to 15 dB or so. As briefly indicated in Figure 2.1, we must accept a dependency between available NF and frequency. However, we can also think of NF as an independent parameter, the effects of which can be used to tune the performance of a system. As stated in conjunction to Figure 2.1, radar design often uses the convention of noise temperatures, but radio communication designers normally prefer to use the prevailing interference levels expressed as field strength values. Generally, low NFs are preferred, because they allow us to lower the output power at the transmitting equipment and make smaller antennas feasible. Alternatively, we may want to keep the EIRP as it is and utilize the lower noise level to enhance the signal characteristics after detection or demodulation.

At this juncture, it is necessary to make three major points. First, reducing the system noise level below that of the prevailing environment seldom makes sense. Figure 2.9 shows one example of measured RF field strengths across the lower *high frequency* (HF), VHF, and UHF bands. We observe that man-made interference is much higher than the typical performance of even low-cost front ends and thus very little improvement—if any at all—can be expected. Typical cellular mobile phone bands are clearly distinguishable, and frequencies below about 100 MHz are very crowded due to such factors as FM broadcasting and unlicensed devices. Individual carriers cannot be viewed, because the resolution bandwidth in the measurement has



**Figure 2.9** There is not much sense in lowering the system noise level below that of the operating environment. In this particular case, frequencies below 100 MHz suffer from severe interference.

been quite large, exactly 100 kHz. The observed external noise very much depends on the site and time of measurement. Distant rural locations, far from overhead power lines, radar stations and communication equipment, can show 20- to 30-dB better results at least above 30 MHz. The *short-wave* (SW) and *medium-wave* (MW) bands are almost everywhere very crowded—both due to true broadcasting and due to spurious emissions. Readers should compare the plot in Figure 2.9 with Figure 2.1. Although the scaling of vertical axes is different, the importance of strong intentional emissions and unintentional interference in the lower VHF bands comes out. The conversion between noise temperature and effective field strength is left to the reader as an exercise.

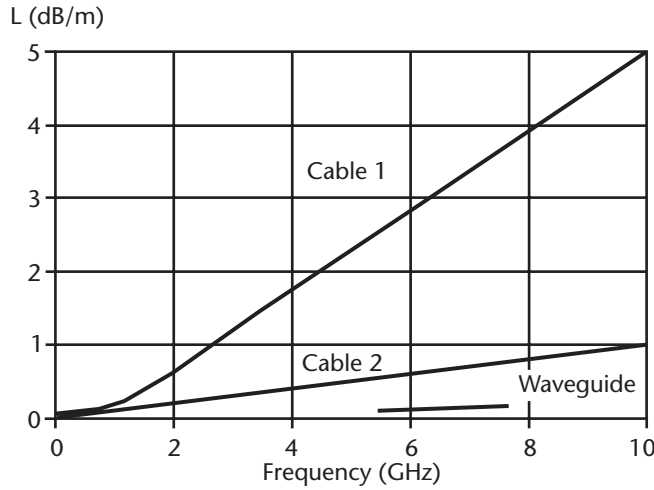
The second thing about low NFs is the way in which we get them. As long as it is just a matter of circuit design and component selection or choosing the proper semiconductor process, there is not very much to worry about. The extreme performance, however, is typically obtained with cooled front ends, which implies the use of cryogenic equipment. If possible, system design should try to avoid specifying such low noise levels in favor of better reliability in operation and in order to keep running costs reasonable. Twenty-Kelvin LNAs with liquid helium cooling are precious instruments indeed and require talented operators almost throughout their entire life.

The third remark regards dynamic range in RXs or related components of a system. Typically, if the NF is kept low, particularly in a wideband arrangement, the front end seldom handles high occasional input levels without severe distortion or blocking. Both radar systems and some cellular-type RXs may face this problem. Before setting the target NF for a LNA, we should first evaluate the entire range of input power levels possible in a real operating scenario. Generally, a compromise, which prevents severe blocking but does not handle the weakest input, is a preferable choice. Due to the rapid evolution of *digital signal processing* (DSP) algorithms and components, we can often relax the hardware NF specification to some extent and rely on processing gain, which can be understood as “digging” our wanted signal out of surrounding noise through mathematical manipulations. Unfortunately, even DSP cannot always provide a rock-solid solution, and we are forced to install the helium tank.

## 2.5 RF Transmission Lines

A first approximation is to consider the often unavoidable pieces of RF transmission lines as an extension of the propagation path whereby the two major effects are attenuation and group delay or phase change as a function of frequency. Figure 2.10 shows a typical example of two coaxial cables and one waveguide type. Thus, it is natural to aim at the lowest practical loss whereby we can increase the allowed attenuation across the air interface. In receiving equipment, excessive transmission line attenuation between the antenna and the LNA will totally spoil the NF. Figure 2.11 illustrates what this could mean on the system level by showing the required TX output of a tactical military UHF link as a function of selected cable attenuation. The assumed cable run length is 30m, which can be considered typical for field-transportable equipment. As discussed in Section 2.3, the increase in TX power is

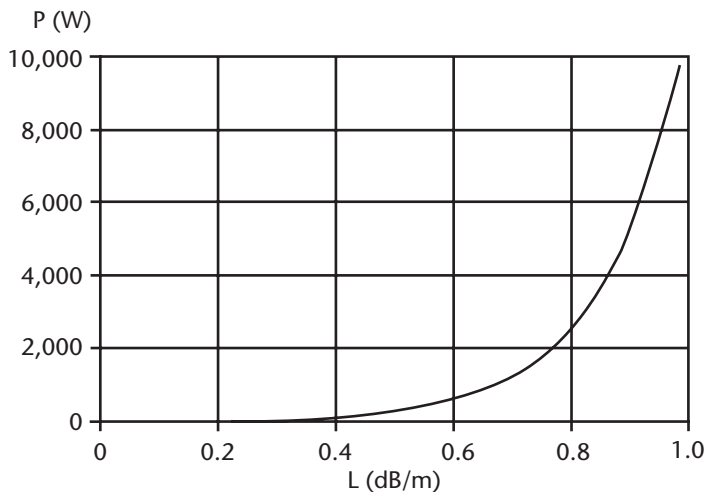




**Figure 2.10** Attenuation characteristics of two coaxial cable types as a function of frequency. For comparison, the performance of a rectangular waveguide of specific dimensions is illustrated as well. Note the limited bandwidth available in the waveguide due to the possibility of unwanted propagation modes. However, a suitably dimensioned waveguide can be used for any part of the practical radio spectrum, as long as 25% to 35% bandwidth is acceptable.

further converted to a need of a larger dc supply. The heavy power supply might jeopardize the whole concept of highly mobile tactical units.

Unfortunately, real transmission lines typically have other secondary effects such as dispersion, which causes different frequencies to have different propagation velocities and *passive intermodulation* (PIM). Practically all line types have a maximum usable power level above which either the field strength exceeds the breakdown value of the dielectric or the thermal heating starts to deform the insulation. Lines other than TEM-mode lines are restricted in bandwidth and even coaxial cables exhibit a higher limit of operating frequency due to the appearance of

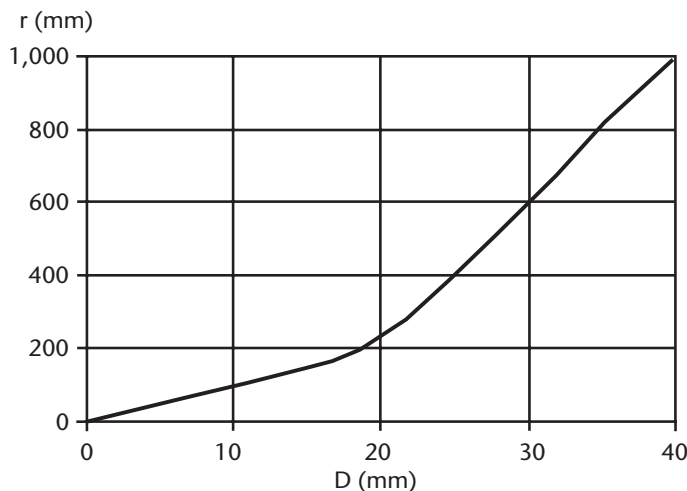


**Figure 2.11** Cable attenuation may have an adverse effect on the required TX output in a tactical UHF link. The cable length is 30m. The vertical axis shows the attenuation as decibels per meter whereas the Y-axis shows the TX power, which is further converted to supply loading.

parasitic modes. One of the most apparent characteristics of RF lines is their varying mechanical nature. Cables generally allow some bending whereas rectangular waveguides normally do not. Figure 2.12 indicates the dependency of bending radius of selected cables as a function of cable diameter. Discontinuities in a transmission line are always problematic as are transitions between different line types (e.g., from waveguide to coaxial or from microstrip to waveguide). The phase and group delay characteristics are further a function of bending, which might be a serious problem in systems involving continuous mechanical movement (e.g., in scanning radar antennas or industrial measuring robots).

We can seldom use a transmission line without some special kind of a mechanical interface to the remaining equipment. For example, we seldom solder an antenna cable straight to the legs of an RF *integrated circuit* (IC). Instead, such arrangements are usually done through connectors or transitions, and in some cases also by waveguide flanges only. The best connector is the one that is not there. The overall RF performance of a system is practically always degraded by any connector or transition type, but we are often forced to use them, just because of physical requirements such as assembly and maintenance, or due to subassemblies coming from different vendors, for example. Additionally, even if the design performance is initially met, connectors and related hardware tend to wear in use, and this normally causes adverse effects in terms of RF behavior. Such processes are faster in severe operational environments but cannot be totally avoided in clean and warm laboratories either.

The general recommendation related to RF transmission lines is to keep them short and select types having the lowest feasible attenuation—as long as bandwidth, physical size, cost, and possible requirements of moving subassemblies permit. The less transitions, bends and angles, the better the system. Environmental stresses may be severe. Humidity can get into cables and waveguides and may even be absorbed by a less suitable dielectric. This happens with certain foams, which provide lower attenuation for the same cable diameter. However, additional means of keeping



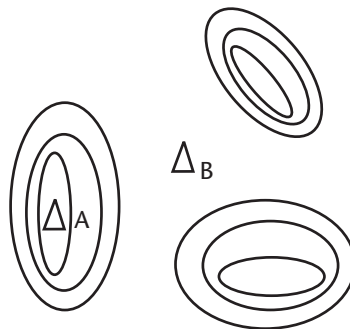
**Figure 2.12** Continuous bending of coaxial cables must obey the type-specific limits of the bending radius. Otherwise permanent deformations of the dielectric or the outer shield or both will occur. This will change the impedance and is observed as mismatch, increased attenuation, and unwanted modes.

water out of the lines are often based on pressurization, which means nitrogen bottles or pumps and valves. Despite solving at least temporarily the acute problem, such ingenious arrangements create a handful of new challenges, particularly during the operational phase.

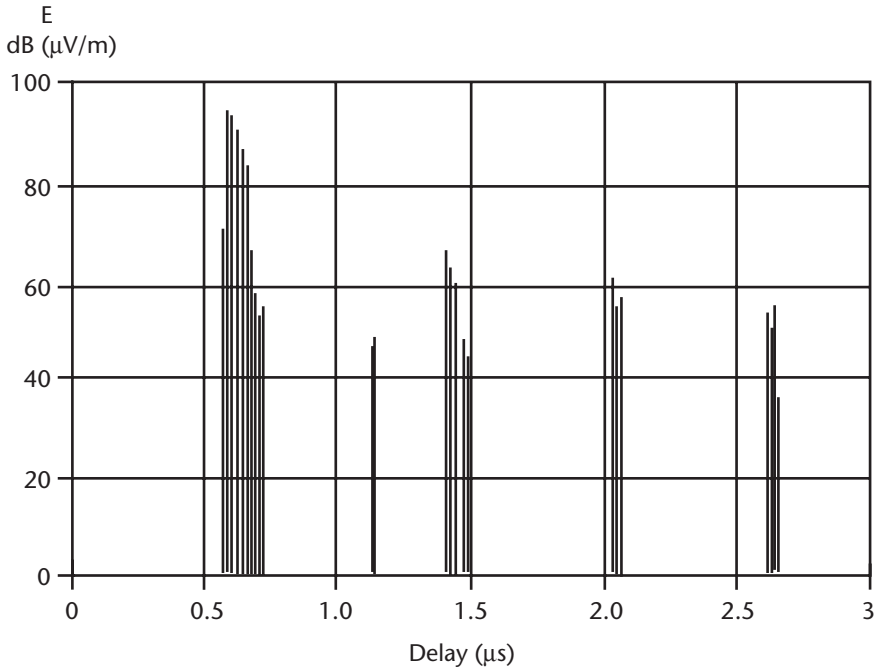
## 2.6 Geographical Topology

Many RF systems can make use of the characteristics of their operating environment to perform better but more often the site and its neighborhood are a nuisance or may even prevent proper functioning. This is true both of communication and radar systems. One example is highlighted in Figure 2.13. Radar, however, is normally doomed to fail against a jammer just because the propagation path is traveled twice by the own signal. The harmful signal from a jammer does the one-way trip and will thus be far larger by default. This means that hiding behind terrain obstacles does not help, if we cannot create a situation where the radar sees the target but the jammer is not able to see the radar. Of course, operating frequency sets further constraints. Bistatic radars [4] can give an additional degree of freedom, because we normally need to care about the receiving site only in terms of jammer cancellation. Unfortunately, ARMs have to be taken into account as well, which may also limit the possible TX site selection.

Both subcommunities of the RF engineering field have adopted their dedicated practices and schemes to live with the real world. In communications, this is called network planning. Naturally, the main effect of the terrain and different obstacles is to change the propagation path. Attenuation is often larger (e.g., due to vegetation) but can show a virtually lower value, too, because of multipath and suitable phase differences. The time-domain characteristics of the real radio environment are highly complicated, and diverse computerized methods are currently in use for the prediction of such factors as UMTS network capacity. Figure 2.14 shows an example of urban propagation results, which were obtained by Suvi Ahonen, who was one of my students while she completed her master's project [9]. Many radar installations have to face backscattering from adjacent ground, hills, and obstacles in the horizon and from a sea surface, the topology of which has one of the widest fluctuation ranges of all. The term *clutter* is used to define this adverse phenomenon.



**Figure 2.13** Radar A uses a mountaintop site to get the largest range but is easily jammed from a greater distance. Radar B is shadowed by the nearby hills and gets obviously less interference. However, its range may be limited.

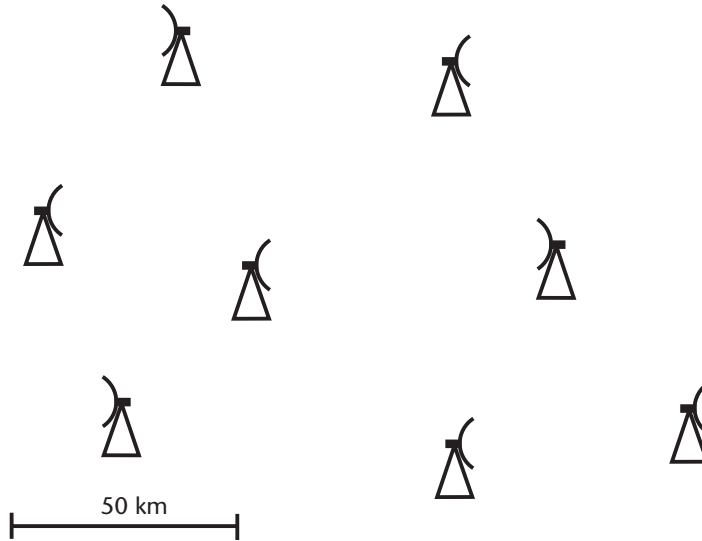


**Figure 2.14** Numerous reflected and diffracted signal components are observed in an urban UMTS simulation. Sometimes there is no direct path (the first peak in the plot) at all.

System design can make use of the expected site topology if that information is readily available, for example in the form of a digital three-dimensional map. We can think of working with shorter antenna towers, less transmission line attenuation, and perhaps even lower TX output power in case we can find a hill or mountain to set up our link station. In some cases, we could increase radar range by a similar means, but we might run into trouble due to severe clutter. Additionally, hilltop radars are a delicious target for distant jammers, surveillance RXs, and ARM attacks. Of course, the horizontal distances are also a parameter to consider. Putting radar stations closer each other as indicated in Figure 2.15 makes possible smaller antennas and lower output power but requires much more hardware—yet, this may still be a cheaper option. Small cellular base stations the size of a shoebox are frequently put at distances shorter than 100m if modern urban microcell or picocell networks are constructed. This improves capacity and signal quality as seen by the end user and the network but may be a drawback in terms of network connections. These are typically based on optical or copper cabling or short-range microwave links from the *BS controller* (BSC)—when applicable—to the individual BSs.

## 2.7 Modulation

Total freedom in choosing the applied modulation or demodulation method is seldom seen, even in completely isolated systems. The reason is twofold: Regulations giving us the spectral portion to work in normally assume a predefined modulation,

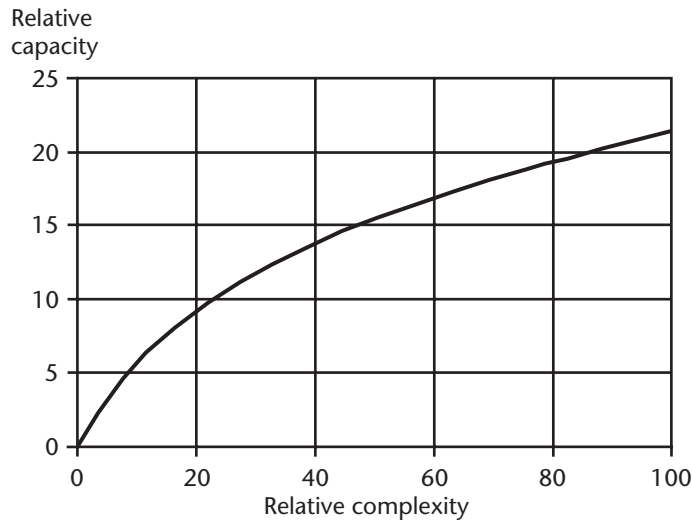


**Figure 2.15** A dense network of short-range radars may give the benefit of lower unit cost, but the number to be manufactured is considerably higher. There may be operational differences as well when compared to a sparse matrix of large radars.

or they are defined so as to restrict the available bandwidth to a minimum. Then the question is more of finding a way to put all the data or other baseband information to be transmitted into the all-too-small corridor. The other reason for limited possibilities lies in the components. Systems designers often want to make use of ready blocks instead of starting the trial-and-error process of constructing their own circuit. Therefore, the only feasible schemes and bandwidths are those for which such commercial hardware is available.

Let us first have a look at modulation aspects of communication systems. In general, the choice of the modulation method is a compromise between performance and complexity. We should, of course, select a scheme, which enables us to put all the wanted baseband information through it with an adequate margin, but we must also pay attention to the robustness of the design. If we have to consider moving installations in which multipath and fading are continuously present, this is of particular importance. Military systems are another natural example of such requirements. On the other hand, redundancy or error correction in transmission can take more than half of the initial capacity. This is a high price for reliability, but we often cannot avoid paying it. The risks in using very complicated modulation algorithms lies partly in making the subsystem too expensive even in volume production and in having it potentially unstable (e.g., regarding timing issues, in severe operating conditions). The difference between simple narrowband FM and 256-stage quadrature AM (256-QAM) is large. Figure 2.16 gives an example of obtainable capacity for a set of different modulations. In real systems, capacity is often expressed as transferred bits per second and hertz of bandwidth.

Radar systems make extensive use of modulated waveforms, too. There, however, the main purpose is not to transmit information from the radar to the external world, but to enable the gathering of more precise data from the environment or to get otherwise more favorable signal characteristics (e.g., against hostile electronic warfare actions). *Continuous-wave* (CW) radars are similar to CW communication



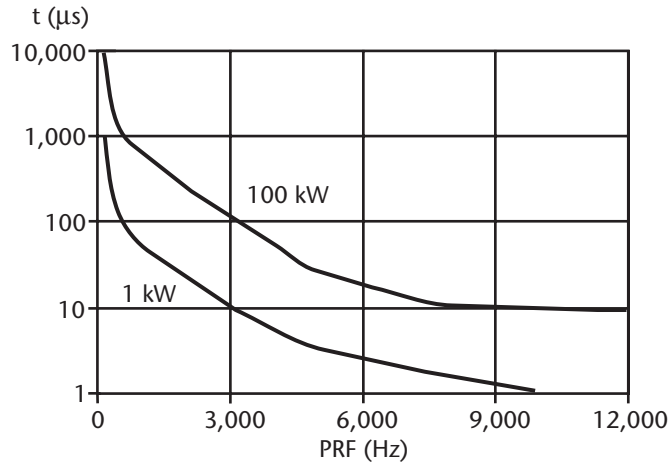
**Figure 2.16** Going to more complicated modulation schemes can multiply the available transmission capacity within the same spectral window but at the expense of circuit and equipment-level difficulties.

TXs in many respects, but without any modulation at all. Pulsed radars just chop the carrier. More advanced schemes included pulse compression, which can be realized through different FM or frequency sweep arrangements or by intelligent coding. Examples of these are Barker-type codes used for binary phase modulation of the carrier waveform. These are means to improve the range resolution without sacrificing pulse energy. Additionally, clever modulation patterns can be used—to some extent—as a watermark whereby hostile fooling of our radar requires more talent and equipment.

## 2.8 Effects of the Baseband Signal

There is a great difference in the expected performance of real-time voice-only links and those transferring digital maps on a delayed basis. Some signal and user types are more tolerant of interruptions and disturbances than others. As long as the end user is a human being, we can partly rely on the adaptive and filter-like post-processing of our vision and ears, but automatic, totally autonomous equipment normally will not allow this. Contrary to this, retransmitting is of no use in a live phone conversation to correct for a failed handover but can be very effective for nonreal-time data.

Elementary communications theory tells us that the bandwidth needed by various baseband signals can be simply calculated based on their own initial spectral width. This of course is true only when we forget all intelligent coding and data-reduction or compression schemes. Analog voice-only will need the traditional 3 kHz, and analog color video will need about 5 MHz. Radar signals can have very different bandwidths ranging from about 100 kHz up to 150 MHz or above. Today's digital world is different—both in radar and in communication systems. Generally, bandwidth is wasted in favor of better or expected characteristics. Let us perform a



**Figure 2.17** Radar TX PRF as a function of pulse width at different average output power levels. Commercially available TWTs from a selected vendor are assumed as the output amplifier.

simple calculation. The analog voice spectrum is 3 kHz. We want to satisfy the sampling theorem and set the *analog-to-digital* (A/D) converter clock to run at 30 kHz whereby we get about 10 samples for each cycle of the maximum frequency. The dynamic range is—as a first approximation—six times the number of bits. In case we are satisfied with average phone quality, we may select eight bits giving 48 dB—something similar to the old C-cassette before the introduction of noise-reduction techniques. This means that we create 30 KB of data every second and thus need at least 60 kHz of free radio spectrum to transmit it instead of the 6 kHz once required by the old AM systems. As we all know, the quality of reception is better as long as the bits go through. After that, there is no reception at all—a clear drawback we must take into account when checking our baseband signal characteristic.

In simple pulse-Doppler radar, the baseband signal can be described by the pulse width and the PRF. These, of course, have their known effects on the detection performance of the entire system but have unavoidable implications on the block-level design, too. Figure 2.17 illustrates how commercially available *traveling-wave tube* (TWT) amplifiers from a selected vendor handle various combinations. The interpretation is rather simple, because the plot indicates that, using the technology of this manufacturer, different duty cycles are generally feasible in different output power classes. Products from other component suppliers may well behave differently, based, for example, on their thermal and electrical design.

## 2.9 Signal Processing

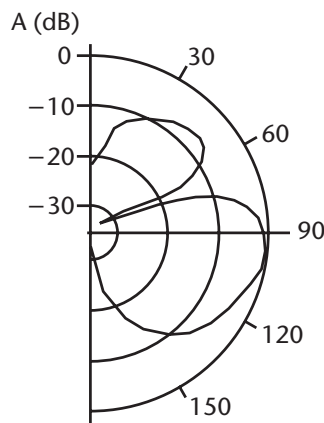
Modern DSP is one of the greatest inventions of the past few years in the eyes of a true RF engineer, because we finally have an effective tool to help us tackle the unfortunate problems of bandwidth limitations, noise, interference, and spurious emissions. However, we must not fall asleep and forget good RF engineering practices while thinking about all the fancy features of digital processing and intelligent algorithms.

Basically, the good idea is to go to digital as close as possible to the antenna interface of any RF system. Possible here means when component performance, power supplies, and cost allow. The critical parameters are speed and dynamic range, and in many applications, the power consumption too. Application areas include adaptive antenna arrays, entire RX blocks [e.g., those containing level-adjustment (former gain control)], and detection stages and modulators [10]. DSP blocks have shown their power [e.g., in *fast Fourier transform* (FFT) calculus, in Doppler processing, in waveform recognition through various embedded algorithms, and in automatic target classification] [10]. Particularly promising results have been obtained in pattern-shaping tasks, which are utilized in a couple of military radar constructions and recently in some cellular networks, too. Figure 2.18 illustrates how four patch antenna elements can be used to provide a set of steerable nulls or maxims, according to operator wishes or tactical requirements [11]. True adaptive beamforming has not yet been fully integrated into the systems level because each antenna element would need its own dedicated RX module. Another suitable task is in various correlation-type processes, which give the RF designer some relief in the  $C/N$  budget and can be used for camouflage as is done in the *global positioning system* (GPS).

## 2.10 Nonelectrical Factors

Some of the remaining typical parameters in RF systems design are not normally included in the university curriculum for electrical engineering but are more related to the mechanical department or perhaps to advanced physics, which is the case in satellite environments [12]. The most noteworthy are antenna size, tower height, equipment weight and size, and thermal stresses.

Almost every antenna seen until today has been too large in some respect. Of course, as mentioned earlier, gain and beamwidth, the key electrical factors, depend on size, but, on the other hand, size may set limitations (e.g., to survivability in wind or to the supporting pedestal if a tracking system has to be constructed). In general,



**Figure 2.18** Sample results from an experimental adaptive patch antenna array. Four elements are in use, with each element connected to a phasing network that is controlled by a DSP system. Wanted minima and maxims can be created as required by the operational status.

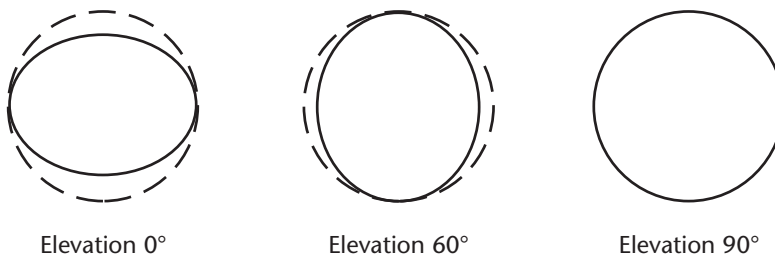


smaller size is better, but we are often pushed against the wall, because there is not enough TX power or not a sensitive enough RX available, or because our targets are far away, perhaps deep in the outer space [13]. Maintaining the initial surface accuracy for better performance at higher frequencies during manufacturing is more difficult if the diameter is several tens of meters. Some typical shape deformations are seen in Figure 2.19. Large antenna structures tend to suffer from thermal deformations whereby their aperture efficiency gets lower and sidelobe levels no longer meet the design specifications. If used in an arctic climate, large parabolic antennas often need huge heating systems or a warm radome shelter, which is not a straightforward thing either. Mesh-type reflectors survive sometimes without such precautions particularly, if their structure does not allow the cumulative growing of snow and ice.

As mentioned in Section 2.6, the geographical topology might enable us to put an antenna higher up. Of course, we can “repair” the local topology by using very tall towers, perhaps extending up to 300m to 600m. This often gives us a *line-of-sight* (LOS) path to our expected propagation scenario but should be carefully considered before implementation. High antenna masts are difficult engineering tasks, and their survivability in wind, particularly when furnished with the most extraordinary antennas, is an entire field of science itself [14]. Short, tactical military antennas, which are often telescopic in design, do not look that long but they have to be lightweight and are assumed to withstand temporary blasts of air and mud due to nearby explosions [15].

Towers are not cheap items, although made of steel and bolts or often of reinforced plastics, carbon fiber, or aluminum, and a high mast needs considerable space around it due to safety regulations. Network planning regarding adjacent station interference gets complicated and jamming may turn out to be exceptionally easy—as is the case for hilltop radar sites. Arctic environments, which are encountered also outside the normal tundra area if going up to the mountains, set special requirements regarding ice formation. A 300-m tower can easily accumulate several tens of tons of ice and icy snow on it and in between the antenna structures. The result often resembles a crow’s nest in the eyes of a mechanical engineer. The tower must be designed to withstand this load; also we must design for times when the wind starts to blow hard, and we have lost all the initial aerodynamic features of a “clean” tower. Many specialists consider the risk height to be somewhere between 60m and 80m for sea-level towers located  $60^\circ$  north of the equator.

Sometimes we have an initial desire regarding the overall size and weight of our RF system or piece of equipment, or we can balance other costs or technical

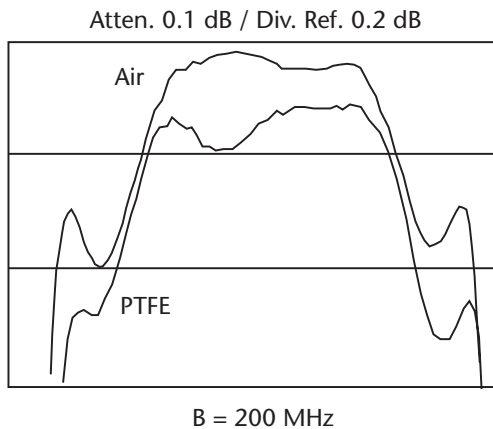


**Figure 2.19** Depending on the design of the supporting structure, a paraboloid can suffer from different shape deformations only because of gravity. When the antenna points to the zenith (straight above it, elevation angle  $90^\circ$ ), the circumference is round, but the parabolic cross-section gets disturbed. The dashed line shows the theoretical circumference as seen along the focal axis.

parameters by allowing a change in the initial size expectation. For example, getting the lowest possible filter attenuation may mean going to the traditional all-metallic design, which is perhaps a coaxial or stripline construction. The overall size is large, because the wavelength is longer compared to a filter on *polytetrafluoroethylene* (PTFE), but we get an improvement in the NF or output power. Figure 2.20 shows a comparison between two such filter constructions. Waveguides [16] tend to be larger than coaxial cables, and they need additional space due to their rigid nature, but they yield far smaller attenuation and a better shielding effect as well. In certain occasions, the selection of a particular material might be critical. Aluminum and related light-alloys are known to pass low-frequency magnetic fields, and therefore, if shielding against these is necessary, we have to look for ferrous materials. A practical example could be the protection of the fundamental carrier oscillator against power supply fields in an air navigation beacon. Of course, the increase in equipment weight can be considerable.

The thermal design for RF equipment tends to be complicated—perhaps due to the inherently low efficiency of our circuits and components. However, both direct and indirect consequences encourage us to trying for improvement. If our equipment runs cool, it will have fewer faults during the operational lifetime [17]. Many active circuits yield better performance if their temperature can be maintained within reasonable limits, and the design margin of operating points (e.g., available gain) can be reduced accordingly [18] if we can handle the waste power properly. On the other hand, large cooling surfaces and fins need space. Forced cooling—whether just blowing air or circulating distilled water—is less reliable. Dust and small particles including insects like bees and even butterflies strangely get in to devices despite the various filters and strainers we use to protect them.

We can think of a trade-off situation in which it is possible to produce the needed output power—for example, with an oversized TWT amplifier that does not need any external cooling but costs a lot and has a complicated power supply arrangement. Alternatively, we might be able to push a different cheaper electronic design that yields the same power but that glows a little bit reddish and thus requires fierce cooling all the time. A further challenge might be the distribution of thermal



**Figure 2.20** If we have to make it small, we may use a high dielectric constant, which means a shorter wavelength. However, lowest attenuation comes from using air as the insulating material. Here, the passband performance of a conventional (small) and a larger stripline design are compared. The small difference in losses may be vital in high-power applications.

loads inside an equipment rack or shelter. There are typically so-called hot spots and then a lot of subsystems, which initially run cool [19]. Shall we start cooling the entire cabin or try to configure dedicated things for the low-efficiency items? How then do we ensure that the heat removed from them is not transferred to the neighboring units?

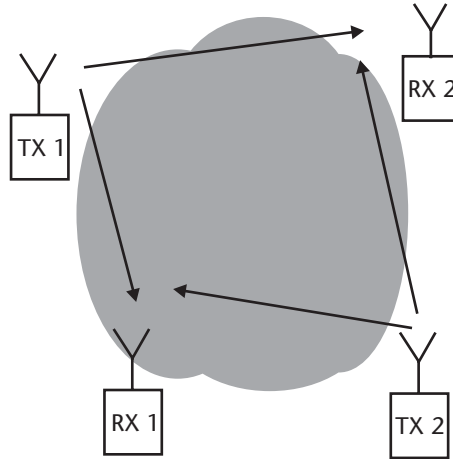
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# Systems Problems Involving Wave-Propagation Mechanisms

One of the most uncertain elements of RF systems is the path across which the signal has to propagate to go from the TX to the RX. Of course, those designs that for some specific reason employ a direct transmission line, such as coaxial cable (e.g., in cable television systems) or a piece of rectangular waveguide for this purpose, might be excluded from consideration. In the more typical scenario, the radio signal propagates freely (e.g., between two sites on the Earth's surface or from a satellite to the ground). The transmitting and receiving installations can be completely stationary, which is often the case in high-capacity long-haul microwave links. Perhaps the most challenging propagation situations are, however, met in those systems where the TX or RX or both are moving, perhaps at supersonic speeds. This chapter presents some quite often encountered problems of such arrangements and suggests means to overcome them or—in practice—to circumvent them. The emphasis is on noncellular concepts, because intense research and related publication activities have been going on within that particular area since the 1980s; see, for example, [1–4]. Although we normally first think about our wanted signal and try to arrange the best possible propagation scheme for it, we must also observe the unfortunate case where some interfering components appear on the scene. They can either be unintentional or intentional, with the latter meaning jamming. It is assumed that the reader already has an initial background in radio-wave propagation concepts; thus the discussion will focus on very select topics.

The overall situation is clarified with a simple example in Figure 3.1. Here, two independent extremely primitive radio networks, assigned numbers 1 and 2, are operating in the same geographical area. The first wanted signal route, indicated schematically by an arrow, is from TX 1 toward RX 1. Further on, there is a wanted path from TX 2 to RX 2. Unwanted signal paths go from TX 1 to RX 2 and from TX 2 to RX 1. The gray background presents the inhomogeneous propagation environment, which includes, for example, topographical features, vegetation, man-made obstacles, and ionospheric and tropospheric particles. As shown schematically in Figure 3.1, the four routes go through unequal propagation path characteristics. For example, the attenuation between TX 1 and RX 2 might be less than the attenuation between TX 2 and RX 2. Most often, neither network 1 or 2 tries to hamper the communication of its neighbor, but interference between them is hard to avoid for example due to an overefficient usage of the spectrum. We demonstrate in Section 3.3 one practical example of such unintentional and highly unexpected interference due to unexpected propagation conditions and show potential solutions as well.



**Figure 3.1** TX 1 and RX 1 form the first wanted communication pair. A similar setup exists between TX2 and RX 2. There may be unintentional or intentional interference from TX 1 to RX 2 or from TX 2 to RX 1. All four path characteristics (e.g., attenuation, delay, and scattering), shown with the gray background, are different.

### 3.1 Propagation Models in Brief with Reference to System Design

The two main parameters defining the propagation scheme in an RF system are its operating frequency and the environment between—and in the neighborhood of—the TX and the RX [5]. Additionally, the selected antennas have some clear effects on the dominant characteristics—for example, through their polarization. A number of cases are clearly nonstationary, and we have to accept that the propagation mechanism may not be the same throughout the entire time span of interest. If the TX and RX trying to communicate with each other are located in completely empty space, we can easily calculate the input power by following the simple free-space propagation model, for which

$$P_{\text{IN}} = \frac{P_{\text{OUT}} G_T G_R \lambda^2}{(4\pi r)^2} \quad (3.1)$$

where the receiving antenna gain is  $G_R$ , the transmitting antenna gain is  $G_T$ , and the TX output power is  $P_{\text{OUT}}$ . The term

$$L_{\text{fs}} = \frac{(4\pi r)^2}{(\lambda)^2} \quad (3.2)$$

is called the free-space loss. However, (3.2) is not actually loss, because the radio wave is not losing its energy to heat. It comes instead from the fact the even the sharpest antenna beams have to accept the spreading of the wave as it propagates away from the point of origin, and thus we cannot get all that power density back with any reasonably sized receiving antenna. Intuitive interpretations from (3.1) may lead to confusion. For example, it might look as if any increase in frequency (meaning a decrease in wavelength) would be seriously harmful to the received

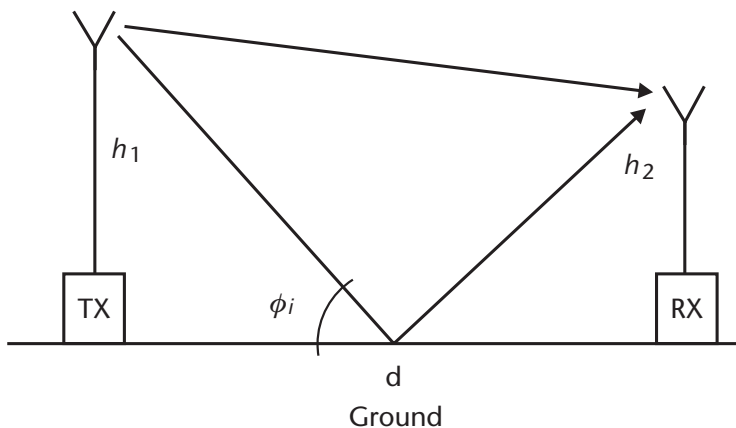
power. However, in case we are allowed to use, for example, reflector antennas of a fixed diameter, we actually benefit from somewhat higher frequencies, because the gain of those antennas is inversely proportional to the square of the wavelength as will be shown in Chapter 5. Of course, we must have constant output power as well, which may not be possible if go up to the higher microwave bands.

The model of (3.1) gives only a rough basis for most evaluations due to the very exceptional simplifying assumptions made during its creation. Therefore, (3.1) seldom works in true radio networks or radar setups. The system designer should be aware of the fact that the signal amplitude as seen by the RX front end may be severely degraded from that of the pure free-space case but also take into account that the phase of the received signal or its time-delay is not necessarily as calculated from the simple group velocity [6]. For example, the main “environmental” things that have an effect on the received signal amplitude in and above the VHF and UHF bands include the following:

- Reflection or reflections from ground;
- Reflections from other obstacles (e.g., buildings);
- Wave propagation through vegetation;
- Heavy snowfall or rain;
- Refraction or “bending” of the wave in the lower atmosphere;
- Ducting;
- Rotation of polarization when a wave propagates through the ionosphere.

Most systems above the lower VHF limit, which is defined as 30 MHz by the ITU, mainly operate in an environment where both the direct LOS signal and a number of reflections jointly form the propagation model [7]. If we assume just one reflection path (in addition to the direct signal) as illustrated in Figure 3.2, we have the resultant electric field strength  $E$  at the receiving antenna site as

$$E = 2E_0 \sin \left[ 2\pi \frac{h_1 h_2}{\lambda d} \right] \quad (3.3)$$

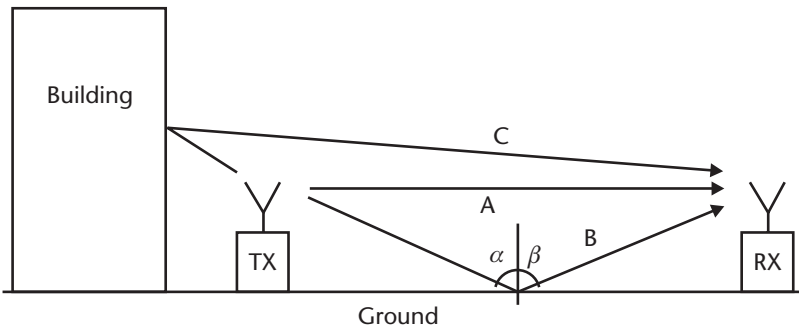


**Figure 3.2** The simple case of just one reflected signal (in addition to the direct path) can be analyzed by using the two antenna heights  $h_1$  and  $h_2$  and the distance  $d$  between TX and RX.

where  $E_0$  is the field strength, which would be prevailing in the free-space situation;  $h_1$  and  $h_2$  are the two antenna heights above the ground level; and  $d$  is the distance between the two antennas. Such a simple case is quite rare in practice, because numerous reflecting objects and surfaces are typically located along the propagation path.

Therefore, we initially want our signal to follow just the straight path, but normally it is instead reflected from several objects and surfaces close to the original LOS route. In these cases, calculations often assume incidence angles  $\phi_i$  below  $10^\circ$  (see Figure 3.2 for its definition) and a perfect reflection coefficient ( $= 1.0$ ) of the Earth's surface for horizontal polarization, but for vertical polarization this factor will be less [8]. Both polarizations experience a phase change of  $180^\circ$ , if the angle of incidence stays below  $10^\circ$ . At the point of reflection on ground, the angle  $\alpha$  of the incident wave as measured from the normal of the surface is equal to the respective angle  $\beta$  for the reflected wave—a well-known rule from geometrical optics. The mechanism is often described as if there exist several “rays” between the receiving and the transmitting antennas; see Figure 3.3 for a simple example and the definition of the angles. Here, path A is the LOS route and has the shortest delay. Path B involves a reflection from the ground and path C comes from a less obvious direction, because this signal is reflected from the wall of a building, which is located behind the TX. Distances involved are typically less than 50 km or so, and often much shorter. Cellular phone networks may use about 100m as a design radius and even less in indoor microcells. The key word is “multipath,” which implies that we have to accept a continuous merging of different rays, whereby the resultant amplitude and apparent time delay of the signal are far from fixed. If very long path distances are involved, we must also take into account the effects of the changes in the refractive index of the troposphere. The electromagnetic wave fronts will “bend” down slightly along their paths, and their LOS range will be larger than that calculated from the pure geometry.

Even if we were able to evaluate and compute the multipath situation precisely, there is no exact analytical approach available to handle the attenuation for example due to vegetation that happens to grow too close to our signal path. Therefore, the analysis and design of many radio networks relies on statistical computer simulations and extensive on-site measurements [9]. One way to approximate the effects of the true propagation environment is to use the free-space formula to calculate the



**Figure 3.3** A typical but simplified VHF and UHF propagation scenario having three paths between the TX and the RX. Path A is the LOS route; path B involves a reflection from the ground; and path C is initiated by a second reflection from a wall, which is located behind the TX.

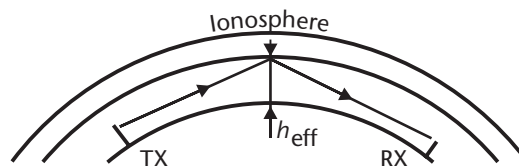
attenuation  $L_0$  at a selected reference distance  $r_0$  but to add into this an environmental term, which has an exponent  $n$  different from 2 [10]. This empirical approach gives the attenuation in decibels at a distance of  $r$  as

$$L_r = L_0 + 10n \log_{10} \left[ \frac{r}{r_0} \right] \quad (3.4)$$

According to [11], the values of  $n$  range from 2 (in the free-space case) to 16 and even more. For example, some commonly encountered semiurban communication scenarios can be dealt with  $n$  values between 3 and 5.

Scattering and diffraction are also processes mainly found in the higher frequencies. An urban built-up area is one example of a difficult environment where the initial wave has to go “around the corner,” or it will meet a number of spatially dispersed objects like cars, streetlights, and window frames before reaching its target. Naturally, these effects appear combined in the multipath scenario [12]. In radar work, scattering has a specific meaning. Although the process is basically the same as in communication, we here are normally in the monostatic case interested in the specific signal that comes back to our radar antenna. Thus, we speak of backscattering, which is of course a parameter of our wanted targets, too, but which is here understood as an unwanted process in the propagating medium, caused, for example, by rain clouds.

Low-HF signals, typically in the megahertz range, can use the so-called ground wave. This mode of propagation is valid in the vicinity of the Earth’s surface and is therefore affected by the conductivity and permittivity (dielectric constant) of the soil (or sea, for example), terrain, and refraction in the lower atmosphere [13]. Alternatively we can assume an ionospheric path of propagation (or a sky wave) as indicated in Figure 3.4. In this case the electrons of the E and F layers of the ionosphere form a reflecting “cloud,” the effective height of which ( $h_{\text{eff}}$  in Figure 3.3) varies with the time of day. There are also notable long-term changes up to a cycle of 11 years. Our Sun’s activity (so-called sunspots) is the main reason, because its radiation has a dramatic effect on the number of free electrons in the ionosphere. If we point our transmitting beam at a suitable angle toward this electron cloud and select a suitable carrier frequency, we can get a very low-loss link. Above the so-called *maximum useful frequency* (MUF), reflection is not possible, but our signal will penetrate through the entire ionosphere toward outer space. In fact, there is practically always an ionospheric component when operating in this frequency range, but its relative amplitude may be low enough so that it goes undetected [14]. Alternatively, even if it is detectable, its level is far below that of the ground wave



**Figure 3.4** HF communications and radar can make use of the ionosphere as a reflecting surface located at an effective height  $h_{\text{eff}}$ . We have to point our antennas at a predefined angle and select an optimum frequency to get the best performance. Due to the instability of the ionosphere, the path characteristics (e.g., time delay and attenuation) are of a statistical nature.



component. For very long-range communications at a low VHF, for example, the ionospheric path is a must. The signal typically reaches distances of several thousand kilometers and can actually go round the Earth. However, as known by every SW listener, the ionosphere is far from stable, which means level fluctuations and also fading due to several reflections coming at slightly different instants of time and at different relative amplitudes. This instability of the ionosphere is again due to changes in its electron density. For those applications, where we desire and design the operation based on a ground wave, the ionospheric alternative is a nuisance, or it may even totally hamper system reliability or availability. This is the case, for example, in certain air navigation systems [15]. However, when properly designed, HF systems are valuable for many long-range tasks [16].

### 3.2 Means to Counter Adverse Conditions (Stationary and Nonstationary)

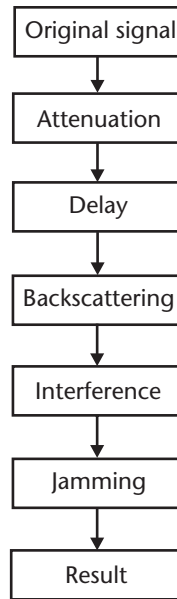
There are very seldom any real possibilities for changing the operating environment of an RF system to provide better propagation characteristics for the signal. We cannot move the Alps, nor can we stop the rain. People want to use their portable phones anywhere they go, and military forces must be able to utilize their systems regardless of time and place. Therefore, the system design has to rely on other available parameters to circumvent the possible or probable adverse effects. We are mainly concerned about excessive attenuation caused by absorption, which is one-way in communications and two-way in radar, and its variation as a function of time. Scattering is a problem particularly in radar but could be difficult in a communication system, too. Additionally, we may be seriously hampered by the often unpredictable time domain transfer function of the path. We may also want to take measures to reduce the effect of interference caused by sudden changes in the propagation conditions. This complex problem is further illustrated in Figure 3.5.

#### 3.2.1 Attenuation

As suggested in Chapter 2, the radio link budget has a number of parameters to be used to adjust the received signal level. If attenuation in the propagation medium looks too high, we might first check if there are chances for selecting another frequency range to overcome the issue. Of course, reducing the RX noise floor could help, but often this is difficult due to the wanted transmission capacity or due to the external noise contribution. A simple expression to clarify this is

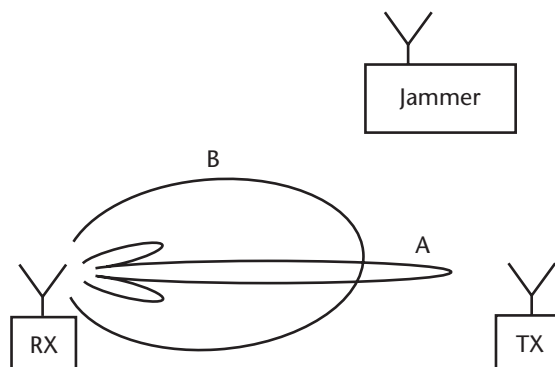
$$P_n = kT_n B \quad (3.5)$$

where  $P_n$  is input the noise power,  $T_n$  is the equivalent noise temperature as defined by our hardware and the noise from the antenna,  $B$  is the processing bandwidth that is directly proportional to the desired data rate, and  $k$  is Boltzman's constant. Now, if we must fulfill a certain transmission rate requirement, we typically have to keep some minimum value of  $B$ . On the other hand, as briefly discussed in Chapter 2,  $T_n$  includes the noise coming to our RX antenna from the external world whereby there might be no use in lowering our front end's contribution to it.



**Figure 3.5** The problems caused by the propagation process can be thought of as a stepwise flowchart. Part of the signal is lost due to absorption; a small fraction may be scattered away; there is an unavoidable delay; and we even have interference added. However, these steps may reoccur and their order is arbitrary.

Sometimes a totally different modulation and coding approach solves the issue because we might actually be able to live with a lower C/N (or S/N prior to detection) in our system. Increasing the antenna gain is often a relatively good approach, particularly at the receiving site, because at the same time we tend to obtain a narrower beamwidth and thus can get rid of some potential sources of interference; see Figure 3.6. Here the antenna in case A has high gain and narrow beamwidth, thereby reducing the effects of the jammer. In case B the gain is low and virtually constant for a large range of azimuth angles; thus no significant reduction of the jammer signal can be anticipated. The main challenge here is the overall size of the

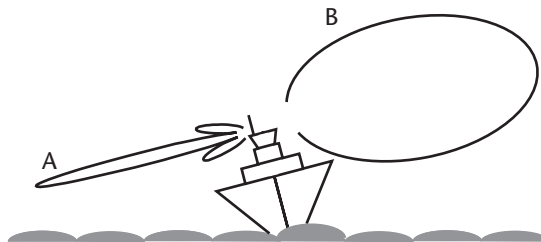


**Figure 3.6** Increasing the receiving antenna gain can often also reduce the possibilities of external interference to our system, because the beamwidth is typically smaller in high-gain antennas. In case A the gain is high toward our friendly TX, and the effects of the jammer get smaller. Pattern B is wide and the difference in gain toward the jammer when compared to the wanted direction is not enough to suppress the harmful signal.

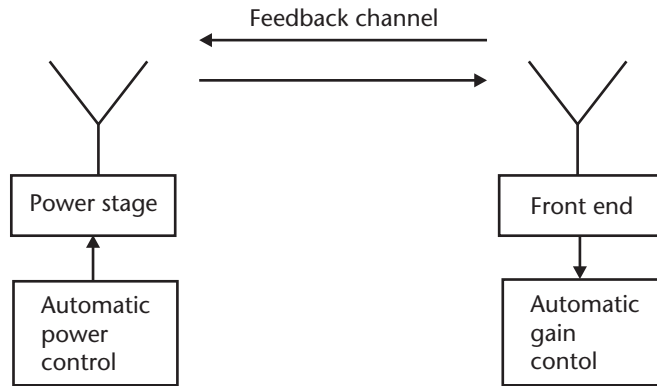
equipment. Larger gain typically means larger antennas and more weight. It might also be impractical to allow a gain increase due to problems in maintaining the antenna alignment. Sidelobe levels tend to get larger when we try to have maximum gain, and jammers may intervene through them. *Ultralow sidelobe antennas* (ULSAs) can be tried as a remedy against sidelobe jamming, but some decrease in gain must be accepted in favor of interference reduction.

Mobile platforms may simply require almost omnidirectional patterns. An example of this situation is indicated in Figure 3.7. The ship's momentary yaw, pitch, and roll angles make the narrow antenna pattern of case A point, for example, far too low (or high), but the broader beam in case B always gives moderate gain toward our friendly link. However, if we can afford a mechanical or phased-array-type stabilizing and tracking system, we can also try narrow beamwidths. Finally, we have the possibility of increasing the TX output power or the power density of the system at the receiving antenna. This alternative requires special care in order to not cause interference to other users of the same or nearby frequencies and to stay within the cost and dc power margins of the system design. The transmitting antenna gain is generally a preferred choice compared to output power when trying to overcome attenuation, but as discussed in Chapter 2, we often need power and gain simultaneously, or we cannot use gain (e.g., due to scanning time constraints). This is partly a question of overall feasibility—not only tied to technical possibilities. Depending on the actual pattern we may be able to reduce spatial interference caused by our system, but the same alignment problems appear as in the case of RX site gain.

Fluctuations in the path loss, either of slowly varying or rapid characteristics, are a major concern if high operational reliability or data rates are expected. Part of the phenomena observed as changes in path attenuation are, in fact, caused by multipath propagation, but some of the remedies are effective anyhow. The first solution is to design the system with enough of a margin in the link budget to sustain any expectable attenuation values. In practice this easily yields huge transmitting powers or antennas, because real operational scenarios show fluctuations of 50 dB and even more. A far better way is to use AGC circuits and blocks. If applied as a feedback loop across the entire path, we can have adjustable gain both at the transmitting and at the receiving site. This is shown in Figure 3.8, where the implementation of the feedback electronics has been omitted for better clarity. The TX function is often called *adaptive power control* (APC), because it is typically employed in modern

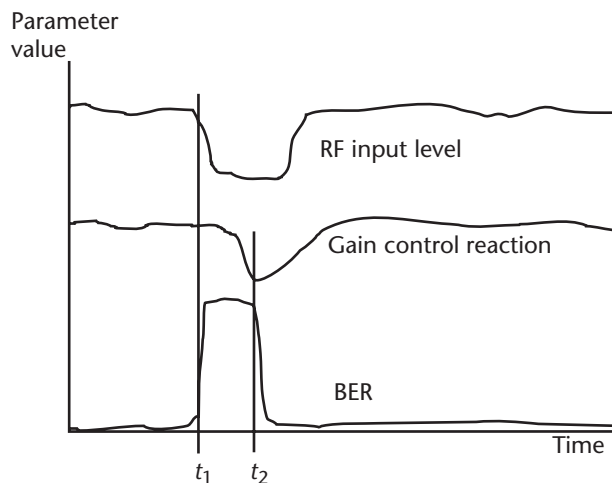


**Figure 3.7** If the platform is moving and we cannot stabilize the antenna position, the narrow beamwidth of a high-gain antenna in case A may turn out to be impractical. Thus, mobile transceivers often have an omnidirectional antenna or an antenna with broad beamwidth (and low gain) as shown in case B.



**Figure 3.8** An intelligent feedback loop can make use of an AGC amplifier in the RX and power control in the mating TX. Control information is transferred through the reverse feedback channel.

cellular networks to set the mobile TX output power within suitable limits. However, geographically distributed gain has potential instability problems, particularly if severe jamming or unintentional interference is encountered. For example, the enemy might be able to fool our system to use the lowest output power simply by jamming our RX, because the distributed AGC thinks that the incoming signal level is adequate. Simple commercial AGC modules provide about 30 to 60 dB of control range, and they are cascadable. Their utilization requires a careful analysis of the necessary attack time constant. If too slow, the  $C/N$  before demodulation (in radar,  $S/N$  before detection) will drop below the acceptable threshold, and the baseband output will be temporarily lost as indicated in Figure 3.9. Here the RF input drops at  $t_1$ , but the corrective process of the AGC circuit is accomplished only at  $t_2$ . Therefore, the bit error rate is very high between the two instants. If the loop works too fast, it can start following the modulation envelope if that exists or to correct for irrelevant level changes.

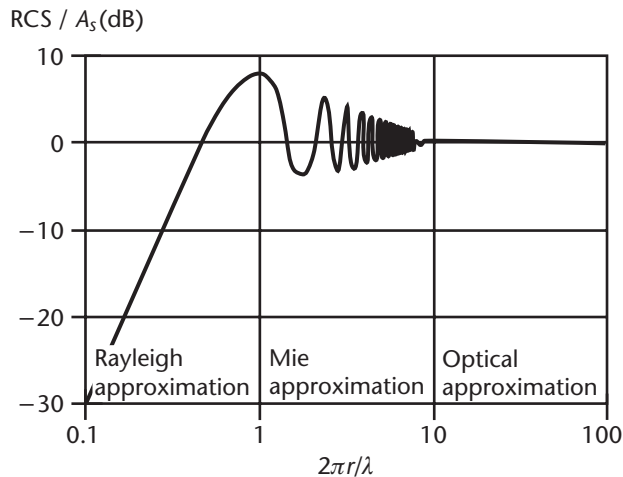


**Figure 3.9** Too slow AGC attack times cause temporary loss of signal between  $t_1$  and  $t_2$  because the system cannot track a sudden increase in path loss. Scaling is arbitrary on both axes and depends on application.

### 3.2.2 Scattering

Scattering is a relevant term for most RF systems transmitting into their external world, but its effects depend to a great extent on the operating frequency. In this context we deal with unwanted particle scattering as observed in radar systems or in millimeter-wave communication networks. The main effect of scattering, for example, in UHF or L to Ku microwave bands is to throw part of our transmitted waveform back as unexpected clutter. At higher millimeter-wave frequencies, though, scattering can also prevent part of our transmitted waveform from reaching the interesting target, or once the signal is returning from the target, scattering further diverts a part of that wave from reaching our radar RX. In communication systems the effect is naturally a reduction in received signal level. Note the difference between scattering and absorption. Scattering causes the wave to “split” into numerous new waves that continue their own propagation toward new directions. Absorption means that part of the energy from the original wave will be “taken away” during the interaction process with the propagation media—and will be converted to heat. Backscattering is the special case where we are interested only in that portion of scattered waves coming toward our RX.

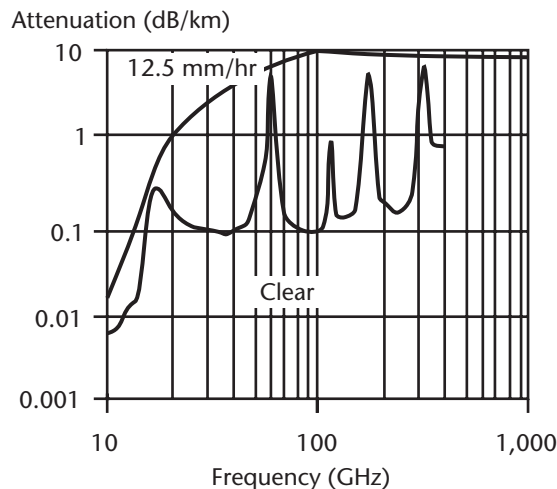
Millimeter waves propagating through rainfall or fog are attenuated or depolarized because of scattering and absorption resulting from the particles [17]. Up to the lower millimeter-wave bands the attenuating effect of absorption is much larger than that of scattering. The main cause of the scattering problem is water, either in the form of rain, mist or—if temperature permits—as ice or snow. Fog or clouds consist of water particles that have diameters of about 0.1 mm or less. This means that the so-called Rayleigh scattering approximation is valid for the lowest millimeter-wave frequencies [18]. If the drops are really large compared to the wavelength, for example if we are working at or above 100 GHz, we can use the geometrical approximation instead. Mie scattering approximation is recommended for most millimeter-wave radar cases by [19] with a note that “wide variation in the data and poor predictability tend to be a rule.” The behavioral difference in drop RCS for various scattering approximations is highlighted in Figure 3.10. The drop



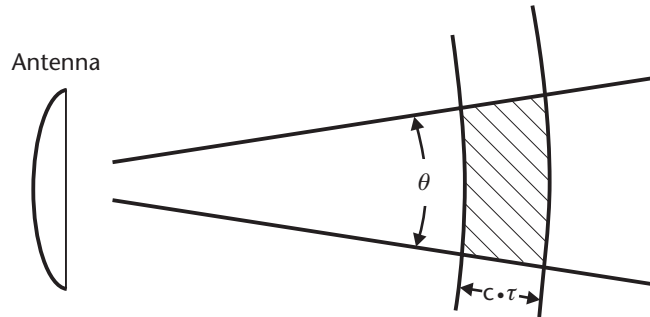
**Figure 3.10** Normalized drop RCS for various scattering approximations as a function of drop circumference  $2\pi r$  in wavelengths  $\lambda$ .  $A_s$  is the physical area inside the equatorial circle.

radius in rain is reported to vary between 0.25 and 3.5 mm [20] but is expected to vary with region and time. Already, rain intensities of the order of 10 mm/hr tend to cause noticeable scattering above 32 GHz, if the average drop size is large. At 100 GHz, scattering forms about one-third of total rain attenuation, if the average drop radius is 3 mm [20] as in thunderstorms. Depending on the vertical temperature profile and related turbulence, we may face wet snow, scattering from which is one of the most difficult forms to handle [21]. Polarization has a considerable effect on the backscattering characteristics, too. Generally, circular polarization shows less rain clutter than vertical or horizontal. For example, at 35 GHz this difference can be almost 10 dB [22]. Besides rain, other macroscopic or microscopic obstacles such as sandstorm particles or dust clouds caused by explosions can be a reason for scattering, but their importance in real applications is much less significant than that of water. Figure 3.11 illustrates the attenuation in clear sky conditions and during 12.5 mm/hr rain as a function of frequency. The rain attenuation plot is a sum of absorption and scattering results.

To reduce the clutter caused by scattering, we could theoretically start by reconsidering the radar carrier frequency, but normally this is of no use. If we have run into rain scattering difficulties, there has obviously been already a good reason to choose that high operational frequency [23]. For example, if we want good angular resolution with a single small airborne antenna or with one installed onboard a ship, we just have to use reasonably high frequencies. Since clutter, which is caused by scattering, is typically a volume-oriented process, the next trick is to try to shorten the apparent pulse. Of course, depending on our system principle, this is feasible only if we have a pulse. Maybe we could make the beamwidth still narrower without losing too much scanning performance. This is illustrated in Figure 3.12, where the beamwidth  $\theta$  and pulse duration  $\tau$  multiplied by the speed of the wavefront jointly form a cell of specific volume. Polarization of the transmitted and received waves is often used successfully to counter rain scattering. This is, however, a little complicated, if we want to maintain a wide RF bandwidth, for example.



**Figure 3.11** Attenuation in the Earth's atmosphere in clear sky conditions and during rain (12.5 mm/hr) as compiled from data of [18, 20]. Rain attenuation includes both absorption and scattering effects.



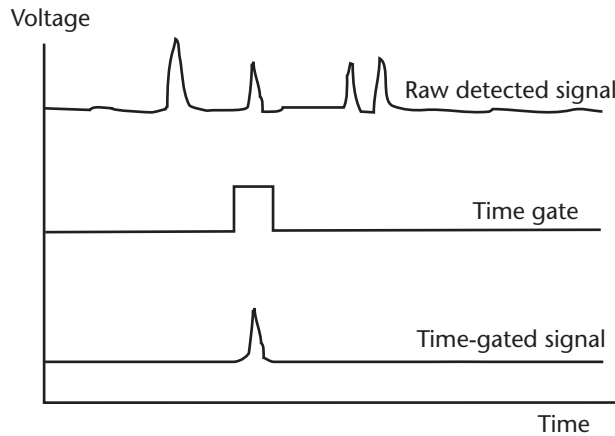
**Figure 3.12** Cross-sectional view of an attempt to reduce the backscattering volume in a pulsed radar system. We use a narrower beamwidth  $\theta$  and shorter pulses of duration  $\tau$ , which affect the radial length.

The final suggestion is to accept rain backscattering and go to the higher system levels for a solution. We might be able to use other target detection technologies in rainy conditions, or maybe we could change our monostatic system into a bistatic or multistatic arrangement. Raindrops are isotropic scatterers, if their shape is not extensively distorted by airflow and surface tension, and therefore, we do not get a real benefit from illuminating them from different directions. However, if we are able to select the direction of the transmitted wave and that of observation so as to circumvent looking directly “through” the heaviest rain front, there is less clutter as well. Of course, this works only if our radar sites are geographically not too close to each other. Further complications may arise due to unknown bistatic RCS of the target.

### 3.2.3 Multipath Problems

The simplest multipath scenario is often presented as the two-way theory on a flat ground. If it were only a reflection and there were no phase changes due to unequal path distances, we would not have a serious problem at all. The different propagation delays and the possible phase change during reflection are combined to cause the amplitude function (3.3) of sinusoidal character. Besides this, the signals really arrive at different moments of elapsed time. This has become a major problem after the introduction of pulsed *time division multiple access* (TDMA) and *code division multiple access* (CDMA) communication techniques. Of course, old all-analog television suffered from this symptom as well as almost overlapped pictures on the RX screen. Additionally, the same problem is encountered in certain radar systems if, for example, target echoes start arriving several times through paths of different length. Intentional jamming is often used to create such a situation [24].

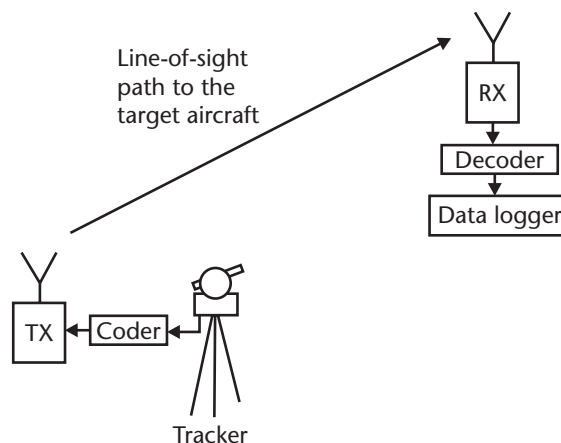
If we know the time window inside which the wanted pulse or waveform should appear, we can use time gating to prevent other delayed signal copies from reaching the detection as suggested in Figure 3.13. Unfortunately, this delay data is seldom available. The other alternative is to limit the amplitude of multipath components. This is often done through antenna pattern shaping or through antenna alignment. We can also try an adaptive antenna concept or antenna diversity, but again, this would preferably use some a priori information. Simple pattern shaping, which



**Figure 3.13** A time gate can be used to reject pulses appearing too late or interference coming too early to be our wanted signal. However, this approach needs prior information about the correct time of arrival. Scaling is arbitrary and depends on specific application.

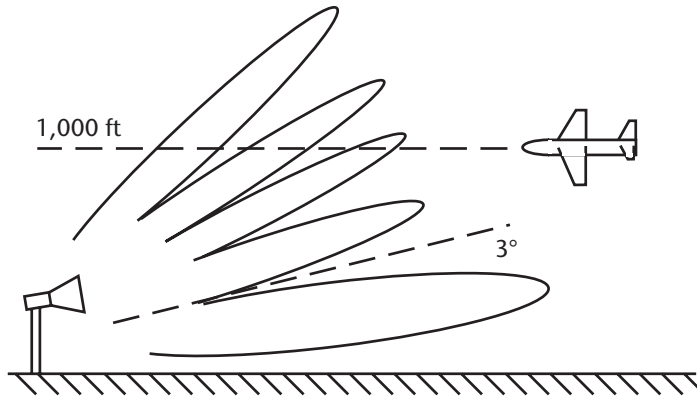
limits the RF illumination of the reflecting plane (e.g., the ground in front of the antenna), is the most effective way in many cases. Severe environmental restrictions may prevent the usage of this action.

Let us briefly look at an example of a digital telemetry link used for certain airborne tests. The simplified system block diagram is presented in Figure 3.14. In this case, the link operates in the 2.4-GHz *industrial, scientific, and medical* (ISM) band and transmits tracking data from the ground up to the approaching aircraft. A simple coder converts the parallel data format of azimuth and elevation transducers into a *pulse width-modulated* (PWM) waveform, which is used to frequency-modulate the carrier. After respective demodulation in the aircraft, the data stream is decoded and fed to a data logger together with the primary test signals. The mission implies that the airplane must follow a precisely aligned path, which forms an angle of  $3^\circ$  above the reference ground (runway). Figure 3.15 illustrates the problem. Due to a lack of systems engineering, the ground antennas (two short



**Figure 3.14** A simplified block diagram of a digital ground-to-air telemetry link, which is used to transfer azimuth and elevation data to a test aircraft at 2.4 GHz.





**Figure 3.15** Improper system design yielded an antenna mount high above the ground, yielding severe multipath problems that appeared only at a 1,000-ft level flight or at 3°, which happened to be the most often used flight profiles.

rectangular horns having only moderate gain of about 10 dBi) were mounted at a height of about 1.2m. This caused severe lobing of the resultant radiation pattern—or a severe multipath scenario, in which a deep null appeared exactly at 3° above the horizon. A brief quantitative calculation of this situation starts with the apparent field strength  $E$  at the location of the aircraft

$$E = 2E_0 \sin[2\pi(h_1 h_2)/\lambda d] \quad (3.6)$$

where  $E_0$  is again the free-space field strength,  $h_1$  and  $h_2$  the height of the ground station and aircraft antennas above ground level,  $\lambda$  the wavelength (in this case 0.125m), and  $d$  the distance between aircraft and ground station. The first pattern null above horizon appears when

$$(h_1 h_2)/(\lambda d) = 1/2 \quad (3.7)$$

and because  $d \gg h_2$  and  $h_2 \gg h_1$  in practical flight test cases, we can rewrite this by taking advantage of the fact that the trigonometric tan and sin functions of small angles approach the value of these angles expressed in radians. Thus, the mutual distance and aircraft altitude disappear and (3.7) becomes

$$(h_1/\lambda)\phi = 1/2 \quad (3.8)$$

where  $\phi$  is the angle above the local horizon. Substituting the given values for wavelength and antenna height gives exactly 3° for the first null angle.

Thus, no signal was received in the aircraft although the antenna elevation angle and TX output power were both furiously adjusted. Fortunately, the test flight procedure included level flights at 1,000 ft as well. A look at the obtained RX AGC plot immediately revealed the reason of difficulties. The cure was to lower the antenna pedestal as close to ground as possible, which is obvious based on (3.7). When the antenna height is small, the first null appears at high angles that are outside the region of flight test activities. A description of the entire flight test system can be found in Section 7.5.4.

### 3.2.4 Interference Issues

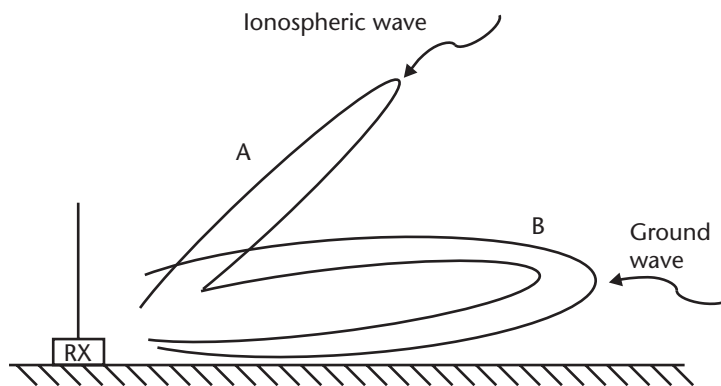
RF interference can enter our system from only a few meters distance. So, the general protective measures and constructions are often not very specific if we are discussing interference arriving from distant locations due to exceptional propagation conditions (e.g., ionospheric anomalies or ducting at VHF and UHF). Adjusting RX bandwidths to reject everything unnecessary and shielding our equipment against unintentional stray coupling is currently mandatory. Naturally, if we are working with the ground wave below 30 MHz, we should consider antenna pattern shaping so as to reduce the gain of higher elevation angles. This is obvious, because the ionospheric wave would arrive from there, as suggested in Figure 3.16. If we can guess certain *directions of arrival* (DOAs) from where the interference might come, we can also try azimuth pattern adjustment, but this works better in the VHF and higher bands because the physical sizes of directive antennas are more practical. However, simple nulling is very easy at HF, too (for example, just with a loop antenna).

## 3.3 Examples

Two results from the real world are shown next to introduce the reader to the multitude of possible layouts where radio systems might be blocked by interference. First we discuss a dramatic case in which the author had to perform measurements after an airplane crash to find out possible sources of continuing complaints against poor navigation signal quality. The second problem was studied in depth by Arttu Rantala under my supervision and led to his master's thesis about unexpected interference in short-range microwave links. The third example briefly illustrates how large propagation path attenuation values are handled in a test satellite system.

### 3.3.1 Unexpected Ionospheric Disturbances at HFs

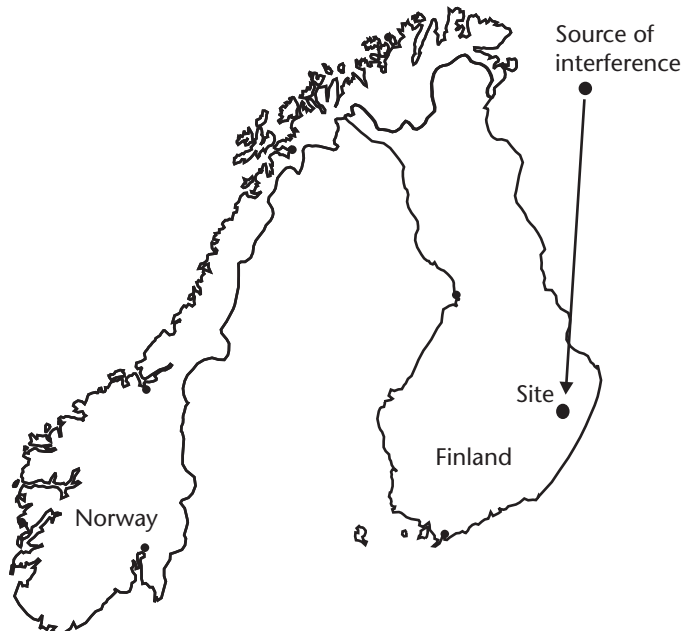
During the winter holiday season in 1986, a courier aircraft crashed in bad weather in eastern Finland. The pilot was killed, and an extensive campaign was launched to reveal possible problems in the airport's navigation facilities. Flight tests indicated



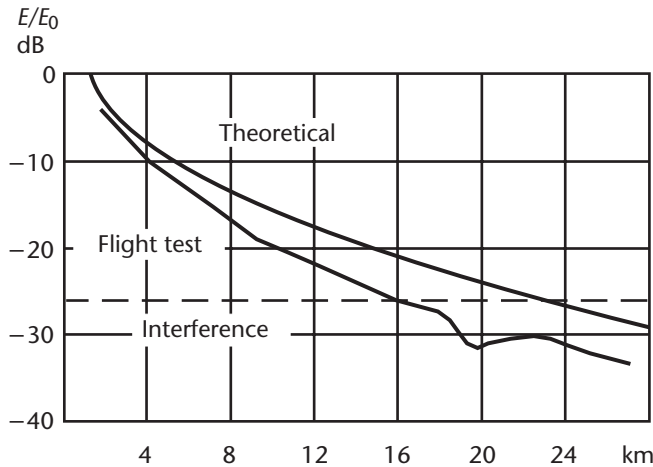
**Figure 3.16** Sometimes we can reduce propagation-related interference by shaping our antenna patterns (from A to B) so as to reject the arrival of the ionospheric wave component.

that the reception quality of the navigation signal from the main approach beacon was exceptionally poor due to an unknown interfering component at the same frequency. HF *direction-finding* (DF) equipment was set up at the site, and dozens of cross-bearings were obtained. Figure 3.17 shows this real-life map scenario with the location of the airport facilities and the approximate position of the interfering TX. As can be seen, the *nondirectional beacon* (NDB) frequency was severely disturbed and actually masked from a distance of about 1,000 km. The exact configuration of the interfering foreign TX remained unknown, but there was some evidence that it, too, was primarily used for navigational purposes. Therefore, its output power was presumably on the same order of magnitude or larger than that found in commercial NDBs. Further investigations revealed that Finnish national authorities had already registered that particular signal source in 1979, but it had showed only low to moderate field strength. Thus, there was no reason to worry about interference at the time of commissioning that navigation facility. However, due to the exceptionally favorable propagation conditions in the ionosphere during the 1986–1987 winter season [25] and the momentarily suitable skip distance, this sky wave signal showed a considerable field strength of about 30 dB( $\mu\text{V}/\text{m}$ ) at the time of the accident in this specific region. Unfortunately, in this kind of a case, interference reduction can not make use of antenna patterns because the airborne *automatic direction finder* (ADF) antenna may be in an arbitrary position as seen from the NDB.

However, further investigations suggested that the airport's *navigation aid* (NAVAID) maintenance had failed to keep the antenna system and the NDB TX itself in top condition [26]. The field strength was an order of magnitude less than stipulated in international recommendations, about 20 to 25 dB ( $\mu\text{V}/\text{m}$ ) [27]. Results of flight tests of that particular installation are shown in Figure 3.18 together with a plot, which is based on a simple calculation. The measuring aircraft was



**Figure 3.17** This map shows one unexpected real-life interference scenario where a TX about 1,000 km away blocked the reception at a particular airport's NDB frequency.



**Figure 3.18** Results from flight tests compared to a theoretical field strength profile near the accident site. The level of the interfering signal stays constant due to the minimal change in relative distance to that foreign TX.

flying along the appropriate approach route. We see that the flight measurement follows the general characteristic that could be expected from the theory, with received power decreasing as a function of distance approximately as  $1/r^2$ . However, additional losses close to the beacon cause a reduction in field strength. There is also a dip at a critical point (20 km) due to a severe local anomaly in soil conductivity. Note that the measured level of the interfering signal stays constant, as the relative change in propagation path length from the Kola Peninsula is minimal.

One, but not the only, reason for poor performance is illustrated in Figure 3.19, which shows the relatively short vertical antenna buried among tall, snow-covered trees. The grounding network, which was composed of 16 copper ropes extending radially from the antenna pedestal, was mostly cut in pieces by the frozen soil. Because ground conductivity in that area was poor (see Section 7.5 for related



**Figure 3.19** This installation had very few possibilities for competing with the interfering foreign transmission. Snow-covered trees absorbed part of the RF energy and caused frequent tuning difficulties. The antenna feeder is positioned incorrectly, too.

discussions), the TX faced an unfavorable input impedance at its antenna connector. Additionally, the feeder of the antenna was improperly mounted, which caused a partial short circuit. All these factors caused the true RX input power to be less than the theoretical estimate. There are two lessons to be learned here. First, wave propagation must be treated as a statistical process, and we must set the performance figures of our own system design to cope with unexpected but possible situations as well over extended periods of time. In HF bands this may mean 10 to 15 years of monitoring. Second, we seldom can control the propagation of interfering emissions, but we should take care to do whatever is possible to assure the highest quality of our wanted signal at the receiving site.

### 3.3.2 Interference Problems in Microwave Links

BS systems of mobile phone networks require cost-effective and easily installed wired or wireless access systems. This becomes even more important because we work with shorter cell ranges below 100m or so, and the total number of installations grows respectively. One alternative is to connect a cellular BS (BTS) through an ISM band—fixed radio link operating within the unlicensed 2.4 GHz band to the mobile access network. There is no need for digging up the road or surveying for the availability of existing lines and half a day is enough if the coffee breaks are not excessively prolonged. Therefore, the majority of current BSs use microwave links and modern short-range sites need only  $n^*64$  kbit/s connection capacity as indicated in [28] and partly in [29]. Such sites require cheap but high-quality and reliable link hardware, otherwise a wireline connection would be superior. The initial goal of this study was to investigate the bit error performance—one part of *quality of service* (QoS)—of such commercial ISM links in a real operating environment, but the focus rapidly turned toward the RF spectrum [30] and unexpected propagation of interference. Partly due to practical measuring arrangements and partly because of expected density of potential users, the tests were carried out in a semi-urban area where tiny parks separate buildings from each other. However, the civil engineering infrastructure was otherwise typically urban.

The 2.4-GHz ISM band is quite heavily used [e.g., by *radio local area networks* (RLANs), Bluetooth units, telemetry systems, and microwave ovens]. They are all potentially harmful regarding an access link, but when using direct-sequence spread-spectrum schemes and different channels, for example, both colocated RLANs and links should work. Of course, the larger output power of an outdoor RLAN version is more threatening. Theoretically, frequency hopping Bluetooth indoor devices that operate within some tens of meters or more should be harmless. A microwave oven leaks a frequency sweeping signal over tens of megahertz (see [31]) and could become a problem when the door gasket gets deteriorated. Remote controllers, medical heating appliances, test instruments, industrial heaters, and scientific research systems might hamper the proper operation of an ISM link, too. Recently declassified reports indicate that a considerable number of military surveillance radars also utilize this frequency range. Their intelligent modulation schemes and mission-specific frequency selections pose a high risk to link performance.

The commercial half-duplex link radio interface used in this example operated at 11 Mbps. Its frame length was 263  $\mu$ s, in which the preamble and header parts

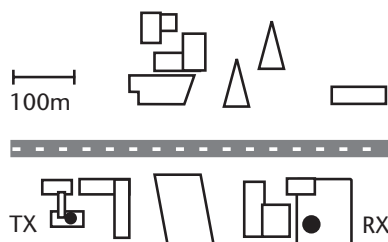
used *differential binary phase shift keying* (PSK) (DBPSK) and a bit rate equal to 1/11 of that of the quadrature  $m$ -ary biorthogonal keying (QMBOK). These features improve multipath behavior—see [32]—and security. The essential performance criteria and related limits are documented in [33]. A particularly important parameter is the *unavailability time* (UAT). Typically, 17 *errored seconds* (ESs) per day and one severely ES (SES) per day can be tolerated.

Five carrier frequencies are in use from 2,416 to 2,468 MHz, but only three can be in operation simultaneously. In Europe, ETSI has specified an EIRP of 20 dBm. Because the typical TX power is around 0 dBm, the antenna gain must stay below 20 dB. The antennas used in these measurements had a beamwidth of  $25^\circ$  to  $30^\circ$ , as suggested, for example, by [34]. A simplified test range map is shown in Figure 3.20. All antennas were attached some 5m above roof level on top of buildings marked Tx and Rx and their mutual distance was about 500m. Thus, perfect conditions for LOS propagation were available.

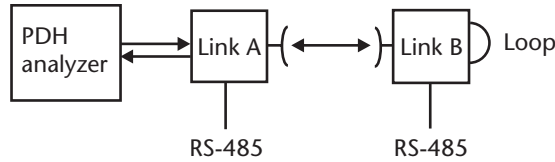
The links were operated in transparent mode without frame synchronization or cyclic redundancy checks of the incoming data; see Figure 3.21 for the configuration. Commercial high-end bit pattern generators and link analyzers were utilized in BER measurements. In this special case the generator/analyzer was connected at link A, and a loop cable was used to return the baseband data from link B. International telecommunication recommendations suggest test periods up to 1 month, but only a couple of days were actually sufficient. A spectrum analyzer was used to check the transmitted signal and to verify possible interference sources.

Synchronization is the most critical point in any radio link system because the first link takes the clock from the incoming, normally wireline, data, but the second must extract it from the radio interface. If the signal quality drops, which is shown as severe *alarm indication signals* (AISs) in the link's diagnostic system, the second—in this example receiving—link loses clock. The communication connection will be completely gone as well. In our case, tests indicated that the highest performance degradations occurred mainly around noon on every weekday; see Figure 3.22. There were long periods of clear data transmission and then sudden bursts of fatal errors. The total BER was  $1.6E-04$ . It turned out that most of the problems were caused by a microwave oven.

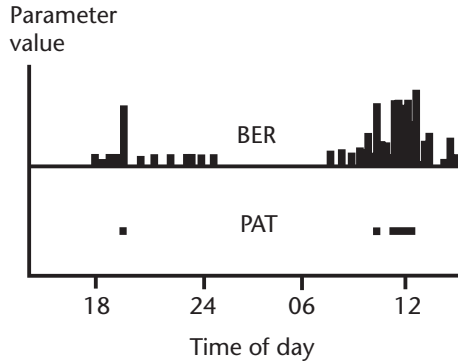
Spectrum measurements were made on top of building Rx. A 1,000-W microwave oven having a mug of water inside was located on the top floor near the window in house Tx. The spectrum illustrated in Figure 3.23 was measured using the peak hold function. Typically the center frequency was between 2,450 and 2,460 MHz, but the narrowband signal was continuously sweeping over tens of



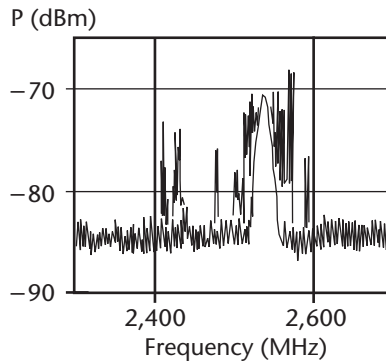
**Figure 3.20** The radio links were installed on the roof of buildings at sites marked as Tx and Rx on this plan view. There was a clear LOS between them above all adjacent structures.



**Figure 3.21** Bit error measurement setup. A pattern generator/analyzer is connected at A, and link B is looped with a cable. The RS-interface is used for housekeeping functions only.



**Figure 3.22** A 24-hour measurement result from the pattern analyzer showing severe errors mainly around noon. The total BER is 1.6E-04. PAT indicates periods where the entire pattern synchronization is lost. At night the transmission quality is perfect.

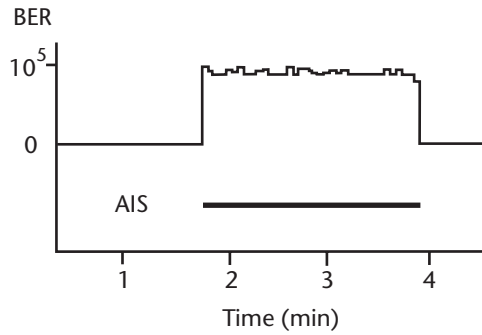


**Figure 3.23** The measured spectrum of a microwave oven (black) versus the link (gray) at the Tx site. The frequency sweep of the oven is clearly visible.

megahertz. In the next, confirming measurement the access link was switched on at Tx. Bit errors were measured at Rx. After a few minutes of perfect transmission the oven was turned on again. The system failed immediately. The link reported continuous AISs due to a loss of timing as indicated in Figure 3.24.

Let us next briefly analyze the obtained power level measurement results. First of all, the wanted signal from the looped TX would cause an input power  $P_{\text{IN}}$ , which is

$$P_{\text{IN}} = \frac{P_{\text{OUT}} G_T G_R \lambda^2}{(4\pi r)^2} \quad (3.9)$$



**Figure 3.24** Transmission errors during the oven's heating cycle. BER counting actually stops if an AIS is generated, because there is no useful bit pattern reference.

where antenna gains  $G_R$  and  $G_T$  are both 16 dBi, and the TX output power  $P_{OUT}$  is 0 dBm. At 2.4 GHz, (3.9) gives about  $-62$  dBm. This is 8 dB above the measured level. Because cable losses were calibrated out prior to setting the analyzer reference level, and both antennas were carefully aligned in azimuth and elevation for maximum signals, the discrepancy must be due to excessive attenuation during propagation. As will be seen soon, the link antennas had such wide beamwidths that multipath propagation with reflections from the ground is the most obvious reason.

The maximum level of the unwanted interference component caused by a microwave oven can be estimated from the allowed power densities. The limit set by the U.S. Food and Drug Administration (FDA) for microwave ovens is  $5 \text{ mW/cm}^2$ . This is defined at a distance of 5 cm from the oven's outer surface. Assuming a typical oven width of 50 cm, we can make a first rough approximation, based on the inverse square law behavior and on (3.2) for a distance of 500m as

$$S_{500} = S_{0.3} \left( \frac{0.3\text{m}}{500\text{m}} \right)^2 \quad (3.10)$$

where 0.3m is the distance from the oven's center to the  $5 \text{ mW/cm}^2$  limit, and  $S_{0.3}$  is  $5 \text{ mW/cm}^2$ . With these values (3.10) gives about  $18 \mu\text{W/m}^2$ . Now, the input power to the RX will be

$$P_{IN} = S_{500} A_{\text{eff}} \quad (3.11)$$

where  $A_{\text{eff}}$  is the effective area of the receiving antenna. This is obtained from its gain as

$$A_{\text{eff}} = \frac{\lambda^2}{4\pi} G_R \quad (3.12)$$

Again, as the gain is 1 dBi and the wavelength 0.125m, the area will be  $0.05 \text{ m}^2$ . Combining the results of (3.10) and (3.12) into (3.11) yields an input power of  $-31$  dBm, which is almost 40 dB above the measured interference level. This suggests that the oven used in our experiment quite apparently had a large margin to the power density specification. On the other hand, the level of oven interference might in the worst case be almost 40 dB above that of the wanted signal.



We can finally estimate the angular sector within which the interfering microwave oven might be as seen from the antenna and still cause interference. In our case the antennas were small paraboloids, and for them a rough approximation combines gain and diameter as

$$G = \eta \left( \frac{\pi D}{\lambda} \right)^2 \quad (3.13)$$

Here,  $\eta$  is the radiation efficiency of the antenna. It varies a lot, but many paraboloids fall within 0.5 to 0.6. On the other hand, the 3-dB beamwidth of such antennas is often expressed as

$$\theta_{3\text{dB}} = 70^\circ \frac{\lambda}{D} \quad (3.14)$$

Combining (3.13) and (3.14) gives

$$\theta_{3\text{dB}} = 70^\circ \sqrt{\frac{\eta \pi^2}{G}} \quad (3.15)$$

In our case the 3-dB beamwidth is about  $27^\circ$ . At a distance of 500m this means that the oven can be anywhere up to more than 100m from the boresight direction and still stay within the main beam. In urban propagation conditions this is quite sure to happen. This also gives an indication about severe multipath problems, because the beam is also wide in the elevation plane and easily reaches the ground between buildings Tx and Rx, although the antennas were about 35m above it. Thus, the received signal from the TX highly depends on the mounting height of the antennas.

The large number of already existing ISM devices, particularly widespread microwave ovens, prevents any straightforward actions to reduce the unsatisfactory interference levels. Besides this, some existing military radar systems may be classified items or belong to foreign nations, making attempts to change their spectrum characteristics out of the question. The spatial distribution of all these unintentional ISM TXs is very wide and varies considerably with time. Ovens may be used in offices in the daytime and in households in the evenings. Radars can move onboard ships, for example, and temporary remote controllers may appear and disappear just with one transport truck or van. On the other hand, current highly integrated commercial access link semiconductors already include sophisticated modulation features that are hard to modify later for better survivability. Obviously the designers have tried to optimize, for example, the spreading parameters to reduce the adverse effects of external interference as far as practical.

The best way to improve the situation seems to be the antenna, but even this is not so simple. If the original equipment design permits, we can leave the TX as it is so as not to violate the national EIRP limit and just have a narrow receiving beam. If, on the other hand, the link design includes a built-in TX/RX combiner (see Chapter 6 for further details) and we have to live with just one antenna, we should consider reducing the output power of the electronics. After that we can again utilize the

interference canceling features of narrow beamwidths. A noteworthy drawback is, however, the sidelobe level, which tends to rise in conjunction with narrower beams. Adaptive arrays or beamforming would cut the cost benefit presently given by the ISM approach despite recent advances in component technology.

### 3.3.3 Reception of Weak Geostationary Satellite Signals

Let us finally consider a small test satellite in the geostationary orbit transmitting at 11.64 GHz an frequency-modulated television signal, which requires a processing bandwidth of 30 MHz in the ground station RX. Our task is to find out the required antenna gain for our ground station. There is a small 11.5-dBW TWT amplifier up in the satellite, feeding an antenna that has a gain of 29.5 dB toward our reception site. The satellite is positioned at 10° East above the equator. If the receiving ground station is located approximately 60° North and 24° East, the elevation angle will be about 20° and we are speaking about true LOS propagation. Our RX on ground has a rather poor NF of 1.8 dB, and the loss between antenna feed and RX input is about 0.5 dB.

The noise floor of our ground station system is partly defined by the RX. However, because at 12 GHz the contribution from the sky is about 20 to 30K [35] and the attenuation due to absorption in the slightly rainy atmosphere is at 12 GHz roughly 2.3 dB [36], the antenna temperature is

$$T_a = 10^{-0.23} (20) + (1 - 10^{-0.23}) 290 = 130 \text{ K} \quad (3.16)$$

as was discussed in Chapter 2. The entire receiving system noise temperature  $T_{\text{sys}}$  is obtained as

$$T_{\text{sys}} = 10^{-0.05} T_a + (1 - 10^{-0.05}) T_{\text{LOSS}} + (10^{0.18} - 1) T_0 \quad (3.17)$$

where  $T_0$  is 290K and  $T_{\text{LOSS}}$  is the physical temperature of the coupling loss between the antenna and the RX. In this case we can assume it to be 290K as well. Substituting these values into (3.17) gives a system temperature of 296K. Combining this and the required processing bandwidth of 30 MHz we can estimate the equivalent noise power in the system input as

$$P_n = 138 \cdot 10^{-23} \cdot 296 \cdot 30 \cdot 10^6 \text{ W} \quad (3.18)$$

which gives about -129 dBW (or -99 dBm). The remaining task is to determine the free-space attenuation from the satellite to the ground station. This is obtained from (3.2) by setting the path distance equal to 39,500 km, which is obtained from simple spherical geometry [37]. Obviously, the wavelength is now 26 mm. Combining the 2.3-dB absorption loss yields a total loss of 208 dB. Thus, if we want a C/N ratio of 20 dB before FM demodulation, the gain of our ground station antenna should be

$$G_a = (208 - (11.5 + 29.5) + 20 - 129) \text{ dB} \quad (3.19)$$

which yields roughly 58 dB. It is quite apparent that we should choose a paraboloid-type reflector antenna to conveniently achieve this high gain. Taking

(3.13) and assuming an efficiency of 0.55, we find out that the diameter of our ground station antenna should be 8.9m. As the 3-dB beamwidth from (3.14) for such a paraboloid at 12 GHz is only  $0.2^\circ$ , additional losses due to pointing errors may appear unless some kind of tracking is installed. A further look at (3.19) suggests that there are two ways to reduce the antenna size. Either the TWT power up in the satellite should be considerably larger, or our C/N requirement (related to the detected video signal quality, which in this case is better than 65 dB after subjective weighting [38]) should be less than 20 dB. Other remaining parameters (e.g., the RX NF) do not have such drastic effects.

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# Circuits and Components for System Evaluations and Design

A real RF system or a piece of RF equipment is made of real components or circuits. This chapter tries to introduce some of the fundamental building blocks, which the designer can use, for example, to construct a demonstrator or even sometimes the final system as well. Both passive and active circuits will be highlighted with the exception of antennas and related hardware, which have a chapter of their own later. In this context, a demonstrator is a special piece of equipment or an entire system that does not necessarily have all the external characteristics of the prototype to come and that might also lack some of the software features. The demonstrator's main purpose is to be a test bed for evaluating the key problems and their solutions. Its main benefits when compared to a real industrial prototype are cost and time savings and the possibility to focus at the essential issues.

## 4.1 Standard or Custom Design?

Sometimes already at the starting phase of a project we face the question of whether the entire system or at least some parts of it should be tailored to the specific task. Could we base the realization on ordinary *commercial-off-the-shelf* (COTS) technology? Circuit designers might want to rush to the workshop to pick some breadboard and wire and immediately switch on the soldering iron. Alternatively, if those who are more theory-oriented could click MATLAB or MathCAD or Maple to the screen or perhaps go straight to some electromagnetic simulation package. The real solution for systems engineers is to simplify.

Custom designs, either homemade or ordered from a vendor, tend to have far more risks than benefits. If performance criteria can be met with existing hardware and software, that is the path to take. Often, we can even adjust the overall design so that system specifications will be met, even though the original configuration indicated severe constraints. For example, if the initial plan calls for previously unavailable TX output power in an end-to-end system, we can sometimes compensate for this by taking more out of antenna gain or by selecting a little bit better RX NF—doing so, of course, assuming that such LNAs are readily available.

Custom designs cannot be avoided, though. Particularly novel military and scientific systems are forced to use such designs to be able to comply with their mission requirements. Typical—and often neglected—difficulties appearing with such modules include the following:

- Severe delays in the schedule and often within the critical path;
- Budget collapses due to labor and hardware costs;
- Unexpected technical side effects (e.g., power supply, temperature, and availability of semiconductors);
- Maintainability problems (only the specialist who designed the module knows it thoroughly);
- Documentation challenges.

Practical experience indicates that even if the financial and timing estimates are made with the best available professionalism, final conclusions after completing a project show a two- to threefold increase in the use of funding and other resources. Seldom can the designed special component or module be applied in, for example, industrial production without considerable refinements and sometimes even redesign.

## 4.2 Passive Modules

Devices that do not contain semiconductors in their RF paths are classified here as passive. Typical items falling into this category include coaxial or waveguide terminations, attenuators, power splitters and combiners, isolators, and filters [1]. Although completely mechanical RF switches do exist and are frequently used as well, that topic is postponed to Section 4.3, where active modules are discussed in detail.

### 4.2.1 Terminations

The purpose of an RF termination is to provide a well-known and stable impedance to a transmission line port, which might belong, for example, to an amplifier or to a power divider. The key performance figures are the impedance mismatch, the bandwidth across which that value is maintained, and the power-handling capability. Very high power terminations are often called dummy loads, because their main usage is in PA and TX testing. Naturally, we must find a termination having the suitable mating interface (e.g., an appropriate coaxial connector or a waveguide flange). Coaxial terminations are available for 50- and 75-ohm systems, but only exceptional designs use the latter variant. Most commercial types can withstand CW power up to 100 mW or slightly more and have SWRs better than 1.2 to 26 GHz. Some waveguide terminations are specified below 1.1 but are usable only within the respective guide bandwidth [2]. Microstrip and stripline systems generally do not have ready-made termination modules available but the designer has to use *surface-mount device* (SMD) resistors having the correct resistance and negligible inductance or capacitance. Precision terminations are mainly used as calibration standards for network analyzers and related equipment. For these metrology-grade applications sliding terminations are also manufactured. They enable one to adjust the phase of the residual reflection so that its effect on the overall measurement uncertainty can be estimated.

Generally, terminations do not cause too much trouble in systems design, if devices of adequate quality and bandwidth are purchased. On some occasions

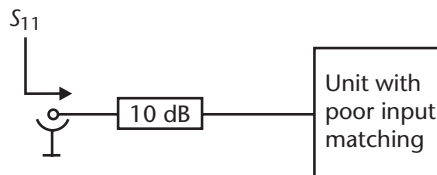
voltage transients may destroy a termination even though we do not exceed the maximum power. The main reason for damage is improper handling. This is particularly true of small coaxial 1.8-mm, 2.4-mm, 3.5-mm, SMA, or K-type connector devices, which do not withstand excessive force or misalignment during mating. Coaxial modules come as male and female alternatives for the best impedance matching.

### 4.2.2 Attenuators

The main goal of an RF attenuator is—as the name implies—attenuation. However, we must take into account, the frequency range, the unavoidable mismatch at both of the ports, the phase or group delay response, and the power-handling capability. Commercially manufactured devices are often processed in a series of 1–2–3–5–10 dB and from there on, in steps of 10 dB up to about 100 or 120 dB. Both fixed and step attenuators are available, and their control can be either mechanical or fully electronic. Some applications require a continuously variable attenuator, the range of which is typically from 0 to about 50 or 60 dB [3]. The actual lossy elements can be simple high-quality resistors, pin diodes, or lossy fin-like designs in waveguides. Fast pin-diode attenuators act like amplitude modulators, if needed, and one of their additional parameters is switching time, which can be further divided into settling time and active time. Where mechanical step attenuators require 10 to 20 ms to change state, all-electronic counterparts operate within 10  $\mu$ s or less and pin switches even in the nanosecond class. All step and variable attenuators have the same kind of settling uncertainty and residual attenuation, which is present even if we select 0 dB. Mechanical devices do quite well with a typical residual term around 0.2 to 0.4 dB, whereas pin-diode devices may have 3 to 4 dB “permanent” attenuation. Part of the uncertainty is stable over time and usage and means a constant difference between each individual attenuation value. Some random variation is there, too. This typically grows in mechanical attenuators over their years in use.

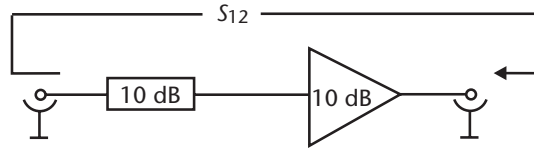
An attenuator is first thought of as a device with which we can reduce the level of a signal to better suit further handling. They are frequently used as measuring standards for metrology-grade tasks. Besides these functions, attenuators are very suitable for impedance matching in those cases where nothing else can be done within the available time and in which the apparent reduction in signal amplitude can be tolerated by our system. This is illustrated in Figure 4.1.

Another trick is to combine an attenuator and an amplifier and thus create an RF isolator as indicated in Figure 4.2. The idea here is first to define the amount of isolation needed and select the attenuator so that this figure is satisfied. Then, we



**Figure 4.1** An attenuator provides an easy way of improving return loss—at the cost of signal amplitude. If the device to be matched shows an almost infinite SWR, here we get twice the attenuator value, making 20 dB, often suitable for any feeding amplifier.





**Figure 4.2** Increasing isolation between two ports can be done with this simple series connection of an attenuator and an amplifier.

compensate for the attenuation by adding an amplifier of similar gain. Actually, this scheme often gives somewhat better isolation values due to the fact that the amplifier, too, has some reverse isolation characteristics, and, if necessary, we can measure the performance and adjust the parameters accordingly. Special high isolation amplifiers are available as well. If we have to improve matching as well, we can try to divide the attenuation into two parts, one before and the other after our amplifier.

### 4.2.3 Power Dividers and Combiners

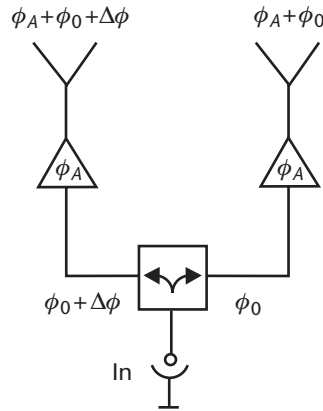
In terms of theory, power dividers also act as combiners, but this is not necessarily the case in real life. The name describes the wanted function very well. We may need to give the same signal to several different processing elements or we perhaps want to feed several signals to the same antenna. This is just a perfect place for a good RF combiner or divider. Different constructions are available having a sum port and something up to 32 or 64 individual channels to be summed up or divided into. The division itself will cause a respective decrease in the power level; for example, a two-way divider has output signals at  $-3$  dB, but additional losses are unavoidable—typically 1 to 2 dB per division depending on the frequency range and vendor. Resistive power dividers have further attenuation due to their operating principle but often give wider bandwidth and better matching.

When selecting a power-dividing or power-combining element for a system, we naturally have to look at the number of ports needed and take care of the frequency range. The power-handling capability is limited, too. Summing devices have to withstand much more than the single input signal [3]. Phase-coherent systems behave in this respect differently to noncoherent designs because voltages may add up in-phase.

In many cases, the phase imbalance between the ports is important. This figure depends on the frequency and may be one of the limiting factors (e.g., in adaptive antenna arrays); see Figure 4.3. Although there might be no reason to ask for isolation between ports when considering the initial input signal, the overall system performance surely benefits from it. Wilkinson dividers found in the coaxial and microstrip worlds give easily more than 20 dB; waveguide structures, on the other hand, do not necessarily yield very much. Isolation and matching of individual ports also depends on the impedance conditions of the remaining inputs or outputs.

### 4.2.4 Filters

We use filters to select or reject certain signals. A very important task of filters is the reduction of input noise, because the noise bandwidth is rather close to a 3-dB

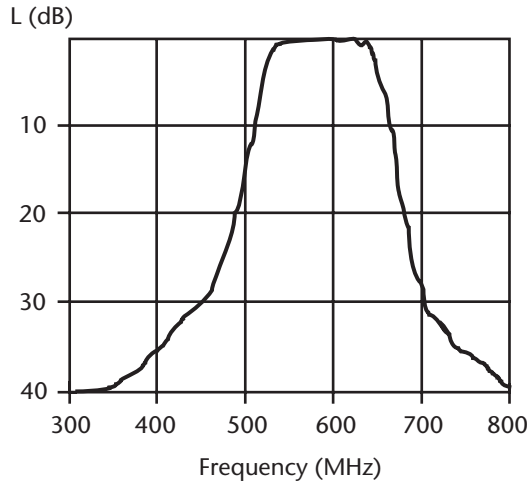


**Figure 4.3** This is how or why power divider phase characteristics affect antenna array performance. Any phase differences of the divider will be directly added to the phasing network values.

bandwidth in most RF equipment. The main characteristics of filters are their frequency response (both attenuation and group delay) and impedance matching [4]. Often manufacturers specify the passband attenuation and stopband attenuation separately. Many filters are reflective in nature, which means that there is large mismatch for frequencies within the stopband [5]. The losses in the passband should normally be a lot below 1 dB, but at very high millimeter-wave frequencies this may turn out to be difficult. Some simple filters do not give much more than 20 to 30 dB of attenuation in their stopband, but most are capable of 50 to 60 dB or even more.

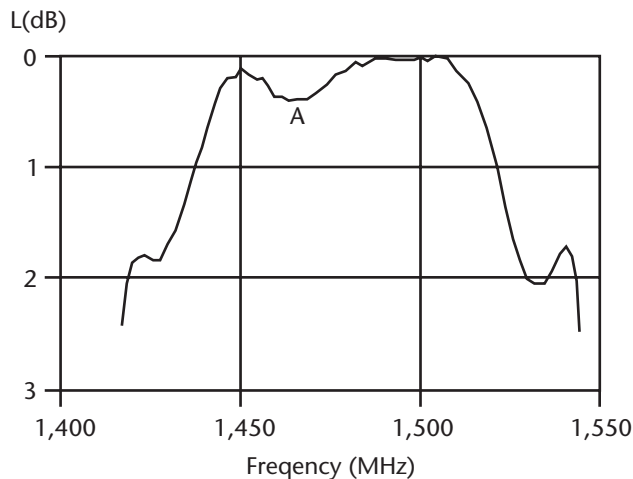
Commercially available filters are found in all four main categories: lowpass, highpass, bandpass, and band reject [3]. Normally, the frequency response is fixed. We can also purchase tunable components, which are based on pure mechanical adjustments. Sometimes the adjusting mechanism is driven by a stepper motor. Faster and durable devices rely on electronics, which can be found as varactor diode designs and as *yttrium iron garnet* (YIG) blocks. Tuning typically compromises other performance figures. Mechanically tuned filters show generally better attenuation responses but suffer from slow speed and wear in use. YIG and varactor filters can be very fast—several gigahertz per second—but this is achieved only when a couple of decibels of additional passband attenuation and not so steep slopes can be tolerated. The systems designer can obtain additional degrees of freedom by suitably combining low and high pass designs in order to get tailored pass bands. One such result is illustrated in Figure 4.4. Physical constructions include coaxial (“tubular”) filters, microstrip and stripline designs (also using air as the dielectric), and waveguides. Narrowband filters are made in the surface acoustic wave (SAW) scheme or as piezoelectric crystals. Although filters are normally very low-loss devices also in the stopband, they still have an upper limit of signal levels that they can handle.

Initially, we have to find a filter having the correct frequency range. This is not particularly complicated if we are dealing with a predefined system (e.g., SSR radar). Then we have to check the interface (i.e., coaxial or waveguide) and the passband loss. This might be surprisingly high in a tunable device even at moderate

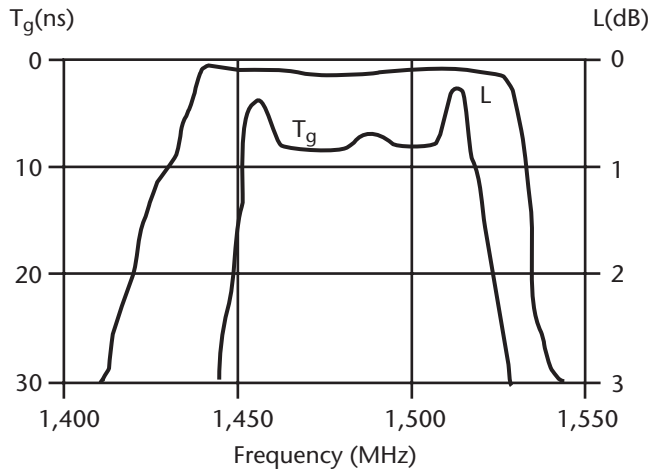


**Figure 4.4** Suitably selected commercial lowpass and highpass filters in cascade can provide a semicustom passband for our system. This shows the response when a highpass filter with a cutoff at 530 MHz is used in series with a lowpass filter with a 630-MHz cutoff. Actually, both filters are sold as 600-MHz devices.

RF frequencies. Many filters have considerable attenuation ripple in the passband (see Figure 4.5), which in FM systems may cause unwanted FM-to-AM conversion. The real struggle in filter selection is often in getting the wanted low passband attenuation in conjunction with sufficient rejection capabilities quite close to the passband. Steep filters are often sought. Modern systems present a further difficulty through the group delay requirement. If a filter has steep slopes in attenuation, we may find severe fluctuations in the delay curve close to the passband edges as indicated in Figure 4.6, and some suggested design procedures unfortunately yield to less satisfactory results [6]. So-called constant-delay filters are special designs intended to overcome this problem [7]. Unfortunately, they are not broadly available as ready-made units for arbitrary frequencies.



**Figure 4.5** Excessive attenuation ripple in the filter passband (around A) may cause unwanted FM-to-AM conversion in frequency-modulated systems.



**Figure 4.6** If a filter has a reasonably steep amplitude response as desired in many systems, its phase characteristics might be far from linear. Here we show the group delay performance of a seven-stage stripline filter.

One of the cases in which custom designs are easily justified is a tailored system-specific filter. This partly comes from the fact that we—and the whole system—can benefit from a suitably narrow bandwidth—just tuned to our needs whereby the noise input will be lowest. Military radio systems require specific filters also due to enhanced jamming resistance [8]. Of course, exact filter characteristics are kept as classified information due to their importance in *electronic countermeasures* (ECM) and *antijamming* (AJ) tasks. Examples of design equations and related data can be found in [2]. Recent trials in the author’s team with selected commercial simulation software packages have indicated that a fully functional RF filter still requires at least one physical manufacturing iteration cycle for optimum performance [9]. For example, the time needed from the announcement of system specifications to produce the first-in-series stripline bandpass filter (prototype illustrated in Figure 4.7) for an L-band radar was about 3 months.

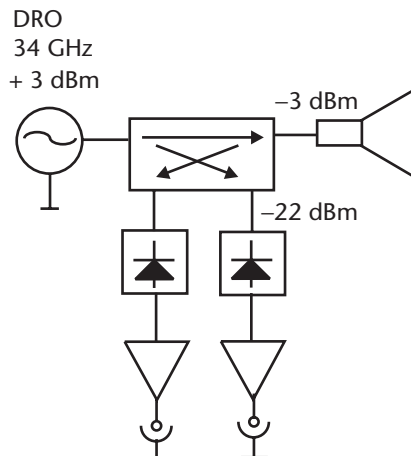


**Figure 4.7** A high performance stripline bandpass filter for L-band radar. Completing this all-milled design from the announcement of system specifications took about 3 months.

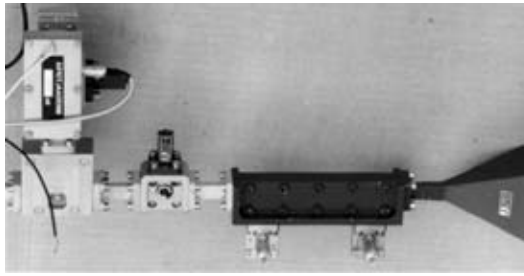
### 4.2.5 Directional Couplers

Many RF systems have blocks or functions that need to know how much energy propagates to one specific direction in a transmission line. Such data is necessary in defining the impedance mismatch at a junction or when adjusting the amplifier output in a TX (source leveling) so as not to exceed a specific voltage limit in the following waveguide. Directional couplers are used as a “transducer” in these cases. One example of scientific use is shown in Figure 4.8. Basically, two distinct types exist, one-way and two-way. The slightly more complex two-way directional coupler can be used to measure both the forward and reflected signal levels. The main parameters used to describe this block are the directivity, coupling, and main line loss, of course, as a function of frequency. As with any transmission line elements, matching is very important for proper operation. Additionally, we may need respective phase responses as well. Special high-power couplers are needed in real TX measurements due to the unavoidable losses.

Suitable commercial modules are available for coaxial and waveguide systems [2], and some very low-frequency surface mount components can be found in the market [3]. The loss in the “through” mode (main line) is typically less than 1 dB. Coupling values are available as a series from 6 to 40 dB, the most typical being probably 20 dB. The more difficult thing is the directivity, because theoretically and in the system designer’s eyes it should be infinite, but 40 to 50 dB is a practical maximum and one has to pay a fair amount already for that. When aiming to achieve the best possible directivity in a real system, we must also take care of the mismatch at each of the coupler ports [10]. This is obvious, because once reflected from the coupler, a signal surely finds another suitable mismatch to get reflected once again and come back to the coupler—perhaps slightly dephased. Very wideband waveguide couplers are complicated and cannot be purchased easily for higher microwave frequencies. There, typical products are limited to respective waveguide frequency ranges or even less. An example of such two-way construction is shown in Figure 4.9. If possible, use coaxial versions instead, but note that their directivity will be compromised below 20 dB or so.



**Figure 4.8** An example of the usage of a directional coupler. Here we are interested in the output power into the test sample (a log) and in the amount of signal reflected back from it.



**Figure 4.9** A two-way directional coupler (in the middle) can be used, for example, to monitor the forward and reflected power of a horn antenna. The feeding dielectric resonator oscillator (DRO) oscillator with its ferrite isolator is shown to the left.

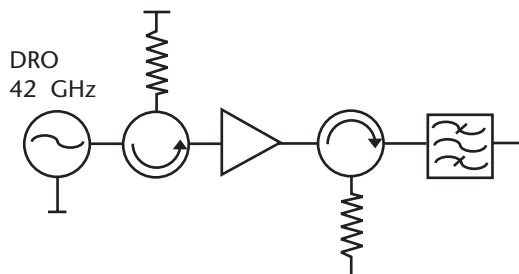
#### 4.2.6 Isolators

Ferromagnetic processes, particularly Faraday rotation, allow us to create nonreciprocal modules in which the electromagnetic wave behaves differently depending on its direction of propagation. Circulators and isolators are components that make use of this phenomenon. Commercially available devices typically provide directivity values around 20 dB, again varying with frequency and show something around 1 to 2 dB as insertion loss in the wanted (forward) direction. Three-port circulators have similar performance figures but applied to all three connections and naturally following the appropriate direction of rotation (clockwise or counter-clockwise). The practical bandwidths available often match those of standardized waveguide ranges, but coaxial modules are also available for the lower microwave bands.

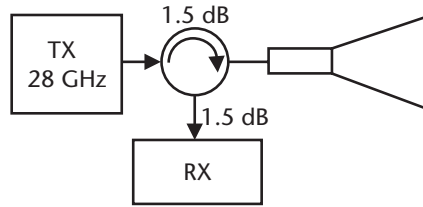
Many millimeter-wave amplifiers and oscillators require the use of an isolator at their output to reduce load-pulling effects and to make the output stage withstand possible RF open/short circuit; see Figure 4.10. Of course this means that we deliberately accept losing a bit more than 1 dB of power. A ferrite circulator can be effectively used as a part of an RX/TX switch in a radar or communication system having one antenna interface as illustrated in Figure 4.11. The loss and NF calculations must be performed accordingly.

### 4.3 Active Modules

The definition of “active” devices is not extremely logical, but most RF engineers consider all elements containing semiconductor devices to be active, although



**Figure 4.10** An isolator can be used to protect PAs from destruction due to severe mismatch and to reduce load-pulling effects in oscillators.

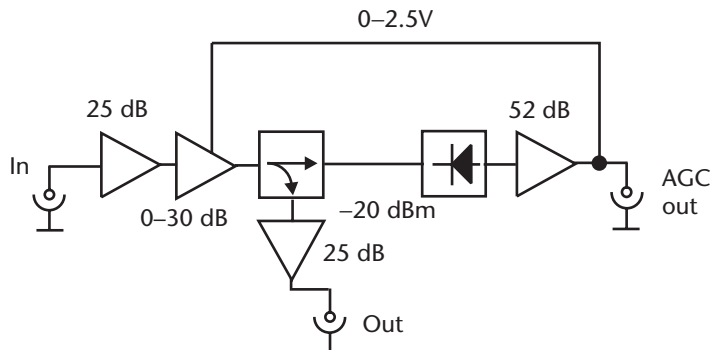


**Figure 4.11** A ferrite circulator can be used as the key element of a TX/RX switch in a single antenna system (radar or communication equipment). The output power is reduced by the insertion loss, and there is additional attenuation in the RX path, making the equivalent NF larger.

diodes, for instance, can be operated in an entirely passive fashion. They do not always need a dc bias and they do not convert dc energy into RF. An RF switch is an interesting example. If made of mechanically operated parts, many of us would say it is passive. However, if the same functional task is handled by a set of PIN diodes, manufacturers list it as active. Modules containing bipolar devices or FETs are easier to classify, which is also true of all microwave tube amplifier and oscillator circuits.

### 4.3.1 Detectors

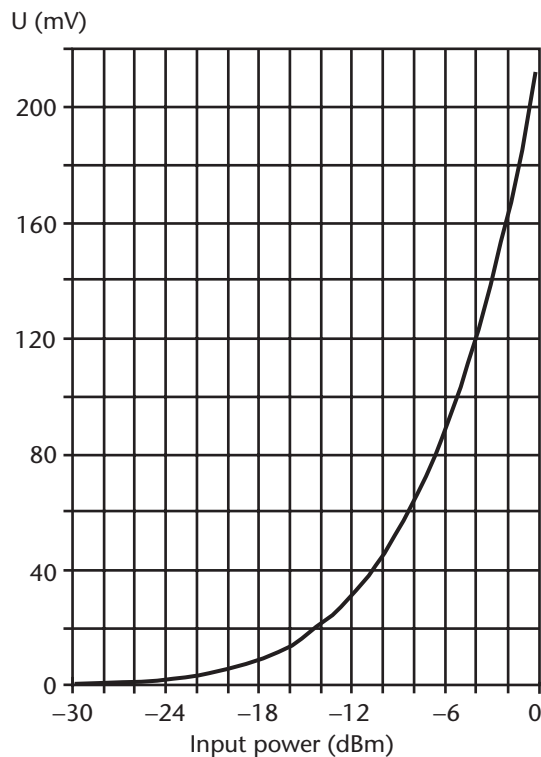
Diode detectors provide a measurable low frequency or dc voltage that has a direct and monotonic relationship with the original RF signal's amplitude. Commercial detector blocks are available both as waveguide and as coaxial hardware. They can be used for a variety of tasks (e.g., power monitoring, leveling, pulsed radar video detection, and diverse laboratory measurements). An example of a leveling arrangement is shown in Figure 4.12. When selecting a detector, its frequency range, RF port matching, and speed are the main parameters. Naturally, we are also interested in the sensitivity—understood here as the conversion coefficient between RF and dc amplitude—and in the nature of this unavoidably nonlinear characteristic [11]. Some applications may require a very large input signal power range, or temperature variations must be taken into account. The upper end of the dynamic scale can—and must—normally be adjusted through attenuators, and this is thus not of major concern.



**Figure 4.12** A simple leveling setup that uses a diode detector, a power splitter, and an adjustable RF amplifier. The entire block performs the desired AGC function.

Many commercial detectors have a very broad frequency range, easily up to several tens of gigahertz from 10 MHz. Very wide ranges cause some fluctuations to the sensitivity curve, possibly up to a 1.5-dB maximum [2]. Also matching is slightly worse going up to an SWR of 3 when similar narrowband designs generally stay below 2 or less. The output voltage varies between about 0.22V for a 0-dBm input down to 1 mV at  $-28$  dBm. A measured detector response (without load) is shown in Figure 4.13. Speed, as measured at the video output, depends both on the detector and on the following circuitry. If a 50-ohm load is used there, the fastest off-the-shelf diode detectors can reliably track RF pulses down to 10 ns, but if typical high impedance oscilloscope probes without any parallel load are utilized, we may reach only 10 ms. Diode and circuit capacitance should be much below 100 pF for fast detection. Maximum allowable CW input power is typically around +20 dBm. Detectors normally come with negative output polarity, but a positive option is available for many blocks.

Putting a detector into our system is a relatively straightforward task. There are two main design factors to manage. First, although the dynamic range is, in principle, quite large, down to about  $-60$  dBm, one can seldom make use of the lowest values due to noise or interference in the low-frequency parts of the system. If other system parameters permit, it might be wise to design a detector operation starting from  $-40$  dBm, for example. Second, if we want speed, we lose sensitivity. For example, a 50-ohm load at the video port of a sample diode means about 0.1 mV dc



**Figure 4.13** A high-quality diode detector has a dynamic range of about 60 dB but making real use of it all can be tedious due to noise and interference, so only 30 dB is shown here. Maximum output voltage is around 0.22V for a 0-dBm RF input, if no load is connected.



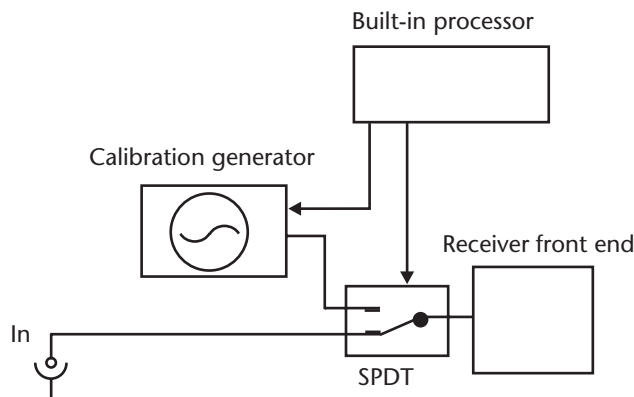
output for  $-20$  dBm RF, but if the load is 50 kilo-ohms or more, the same dc voltage is obtained already at  $-40$  dBm input. Also temperature characteristics are deteriorated if the load impedance is low. High dc impedances yield a total variation of 2 to 3 dB in our sample, but 10 to 12 dB is possible at the 50-ohm point for a temperature range from  $-50$  to  $+55^\circ\text{C}$ . Generally, the power measurement accuracy of diode detectors is not extremely good, mainly because of calibration problems. In many cases the matching of the detector is vital for a low uncertainty value. If severe mismatch exists, one-way and multiple reflections will add up to several tenths of a decibel to the error.

### 4.3.2 Switches

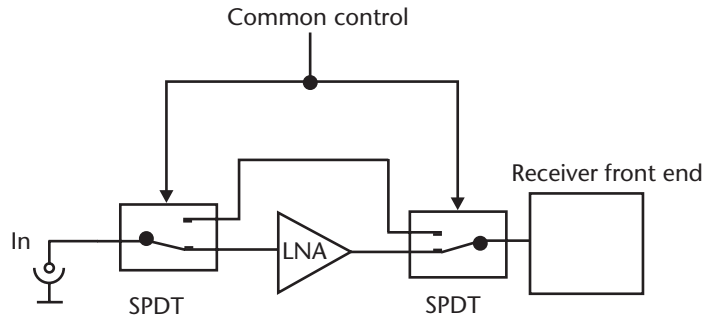
Actually, RF switches are available as all-mechanical blocks with nothing “active” in them, but today, partly due to computer control, most switches have at least their driving system converted into an electronic scheme. Just like ordinary low-frequency electronics hardware, a number of switching configurations exist, ranging from the simple *single-pole single-throw* (SPST) to complicated multiple-pole monsters.

The main applications of RF switches are in signal routing, of course. For example, we want to have an internal calibration generator in our new RX, and this can be connected into the RF input through a *single-pole dual-throw* (SPDT) switch as indicated in Figure 4.14. Alternatively, we might want to be able to temporarily insert some functional module into our system flowchart (for example, a low-noise preamplifier, as illustrated in Figure 4.15). The main parameters of switches are insertion loss, isolation, and impedance matching. Switching speed is important, too, but it is a relevant parameter in solid-state devices only. Also, many mechanical switches must have a specification of the minimum number of switching actions and the respective repeatability of the characteristics along their entire lifetime. Most RF parameters are a function of the operating frequency, and many switches have a relatively limited bandwidth. Power-handling capability is not infinite either.

Electromechanical waveguide or coaxial switches typically have the lowest insertion loss—below 0.1 dB for VHF and increasing up to 2 dB or so for the low millimeter-wave region. Pin-diode devices can have losses up to 6 dB already at 3 to



**Figure 4.14** A simple RF switch application is the connection of a built-in test generator into the RF input of a test RX.

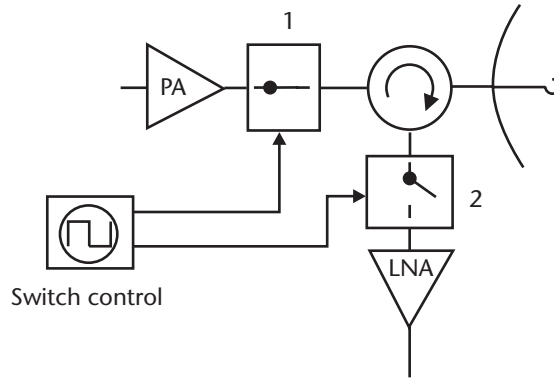


**Figure 4.15** Switches can be also used to insert a functional block, here a low noise preamplifier, into the system flowchart.

4 GHz. The isolation of solid-state elements is often around 30 dB at 2 GHz, for example [12], when coaxial devices provide about 100 dB to 18 GHz and 50 dB to 40 GHz. While wear, contamination, and corrosion can severely degrade the performance of mechanical RF switches, however, semiconductor devices maintain their specs throughout their entire lifetime. The ability of a switch to handle RF power depends very much on the switching condition. “Hot” switching with RF power going across the terminals is more difficult than “cold” switching, where the RF is not present. Normal laboratory-type mechanical switches and associated hardware usually handle 1 to 10W, but very robust hardware for broadcasting and radar TXs is manufactured, too, with specifications up to several hundreds of kilowatts. Semiconductor switches have the inherent problem of maintaining their operating point within acceptable limits also with the RF signal applied. Many manufacturers recommend keeping the signal level below +20 dBm.

The matching of an RF switch is actually not a simple thing at all, because the momentary switching state should also be taken into account. First, check if the switch has an internal 50-ohm termination for the port that is not selected. What happens at the time of transition (break-before-make)? Often switching transients cause unwanted oscillator instability and amplifier overshooting, because RF devices tend to be wideband in nature, when compared to mechanical processes. Thus, 10 ms of unknown SWR at the moment of switching may be disastrous. Also the static SWR figures must be observed. Mechanical switches are often specified from 1.1 to 1.5 at all ports whereas some diode switches may have as high as 30 in the temporarily unused connector!

Electromechanical RF switches may serve 5 or 10 million cycles or they may—according to author’s own experience—fail after the twenty-first action. They are precious pieces of fine mechanics, and sometimes the manufacturing process has perhaps been less perfect. Anyhow, when needing the best RF performance and if speed is not a major limitation, choose mechanical devices. Diode switches are fast and quite reliable, excluding, of course, signal-related destruction. They are the fastest choice, being able to react in less than 10 ns, if necessary. This high speed means that the designer has to carefully analyze arrangements where he or she needs several simultaneous or precisely timed switching operations. One such case is demonstrated in Figure 4.16. A further benefit of diode switches is their superior shielding against dust and dirt; waveguide systems, in particular, can benefit from their smaller size and weight.



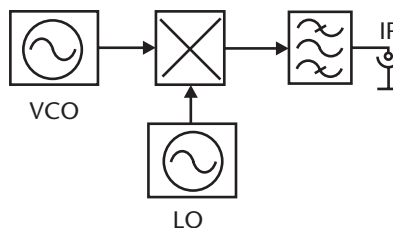
**Figure 4.16** Two high-speed diode switches are here used in conjunction with a ferrite circulator to form a TX/RX switch for a millimeter-wave radar. The mutual timing of switch 1 and switch 2 must be precisely controlled to avoid RX destruction and to get the shortest possible dead time.

### 4.3.3 Mixers

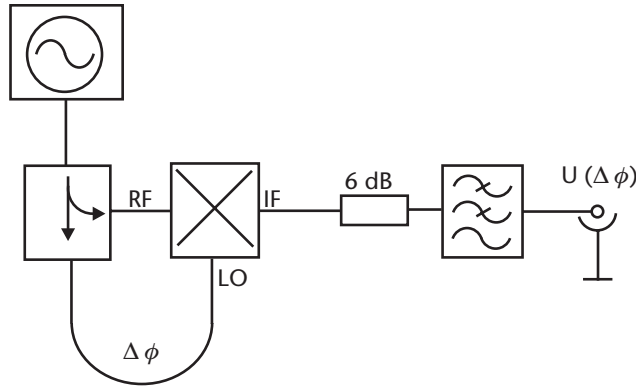
RF mixers are used in systems to perform coordinated frequency changes. In addition to this, a suitable mixer can be used as a phase detector or as a demodulator. Commercial products are available through the entire spectrum starting from 10 kHz and exceeding 100 GHz. Internal circuit topologies include one-, two-, and four-diode assemblies [13]. Four diodes are normally connected as a double-balanced mixer, two as a balanced device.

A very basic mixer application is illustrated in Figure 4.17. We have some initial RF signal, which we want to transpose to an IF. The mixer needs a second input, the *local oscillator* (LO) power, to perform this conversion. Critical performance figures of a mixer include its conversion loss, port matching, isolation between various ports, and naturally the frequency range of all three ports. Often the ports do not have equal characteristics. Practical applications must often be designed for a certain LO power level, for example +7 dBm [3]. Sometimes—mainly in phase detection applications as indicated in Figure 4.18—we have to know the dc polarity and dc offset of the IF port.

One of the first unfortunate characteristics of mixers is the real IF spectrum, which tends to contain an overwhelming combination of mutual sums and differences as predicted by the common mixer equation [14]. The user has to provide suitable filtering at the IF port to select the signal of interest, but this is not enough. We must also take care of the purity of the LO signal and reduce the spectrum coming to the RF input due to similar reasons. An enhanced block diagram is illustrated in



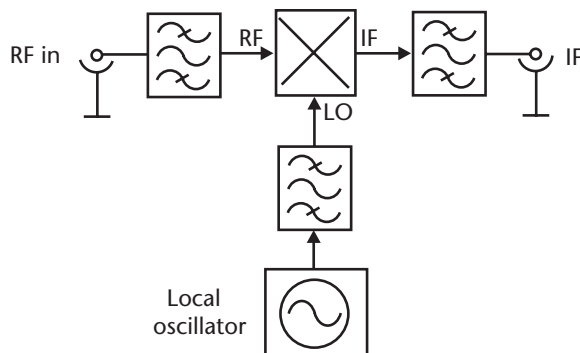
**Figure 4.17** A simple RF mixer application. The first VCO produces an RF signal, and the LO produces the mixer LO input; and the mixer generates the respective IF spectrum, which is filtered.



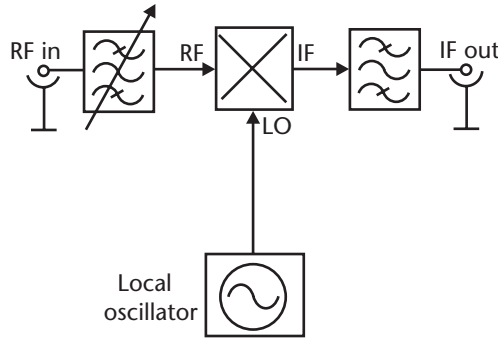
**Figure 4.18** If two signals of precisely equal frequency are fed to the double-balanced mixer's RF and LO ports, the IF signal will be a dc voltage proportional to their phase difference, plus a set of higher sum frequencies. Note the sharp lowpass filter, which is appropriate as we are only interested in the dc output.

Figure 4.19. Such an arrangement is quite easy if our system uses only one predefined carrier frequency, but far from simple when tunability is needed. Besides, putting normal filters in mixer ports involves the risk of perfect mismatch because bandpass filters typically show very poor matching in their stopbands. Multiple reflections may cause out-of-spec conversion loss variations and yield to degraded intermodulation performance. However, filters better in this respect generally provide much less selectivity in the frequency domain.

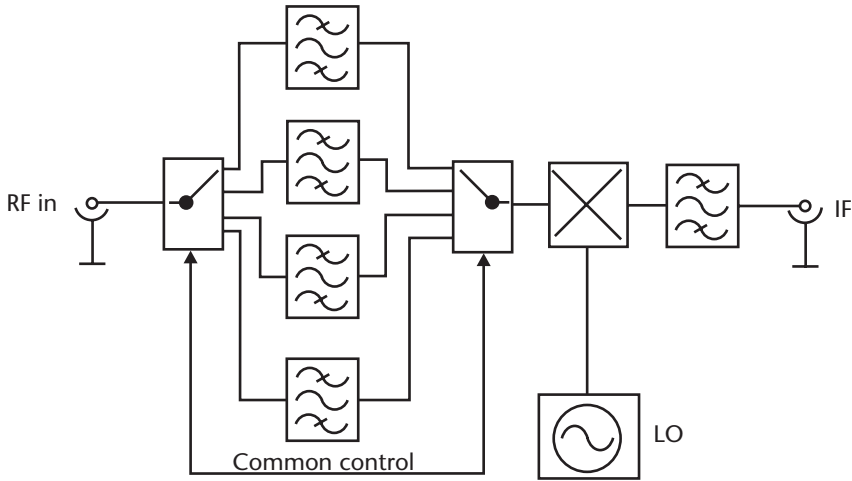
Assume that we want to construct a wideband RX. This makes sharp RF filtering impossible, and a fixed LO frequency cannot be used. The only way of circumventing the problem is to have tunable filters, which are often called tracking filters, because their center frequency is assumed to follow that of the system tuning. We can alternatively choose a filter bank in front of the mixer and use switches to select the appropriate unit for the specific frequency range at hand. These two configurations are highlighted in Figures 4.20 and 4.21. Electronically tuned filters provide speed and may so seem attractive for example in an *electronic support*



**Figure 4.19** Adding suitable filters at all three mixer ports is quite mandatory in many real systems to get a desired IF output. The RF input is filtered to prevent the arrival of possible image frequencies, the LO is sharply filtered for best possible purity, and finally we select from the raw IF spectrum only the band of interest for further processing.



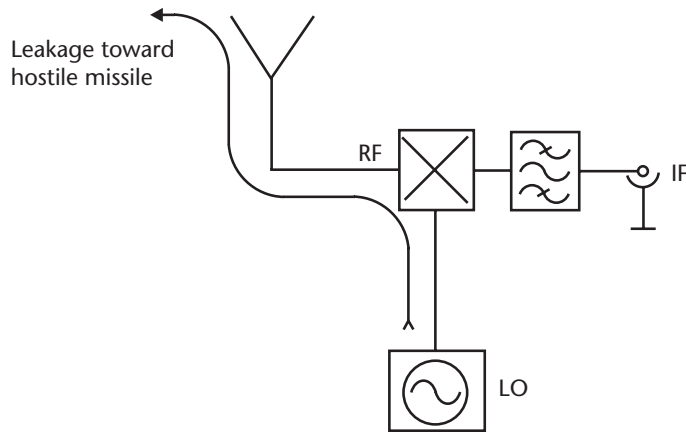
**Figure 4.20** Tunable filters before the RF port can solve the problem of wide input bandwidth, but selectivity will generally be degraded.



**Figure 4.21** A filter bank can be a solution if we need wide bandwidth and good selectivity. However, tracking speed is typically much worse than in a tunable filter RX.

*measures* (ESMs) system, but their drawback is the poor stopband attenuation and sluggish rejection slopes.

Different mixer topologies yield varying conversion losses, but generally 7 to 8 dB is commercially available up to the high microwave frequencies. This figure depends heavily on the LO power level—if lower than suggested, the conversion loss increases drastically. The frequency range of the RF and LO ports is from about 1 to 2,000 MHz in VHF and UHF mixers and above that often covers one to two octaves in one unit. IF bandwidths come in some relation to the two other ports so that the low-frequency mixers have dc to 1,000 MHz and microwave devices start from 1 GHz and cover to about one-third or one-fourth of the upper RF limit. The higher the frequencies involved, the worse the isolation figures. We must often accept 20 to 30 dB as a good result either for LO–RF or for LO–IF even though manufacturers like to indicate respective figures (40–50 dB) for very low frequencies. This rather unavoidable feature may be a serious limitation on the system level. Consider the case presented in Figure 4.22. We have a millimeter-wave oscillator and want to use that to form a very simple low-IF radar warning RX. The oscillator feeds a mixer



**Figure 4.22** An example of a real situation where the poor LO–RF isolation of a mixer may prevent the operation of a proposed system. Here, the feed-through of the LO signal can be strong enough to reveal the RX and cause hostile ARM activity.

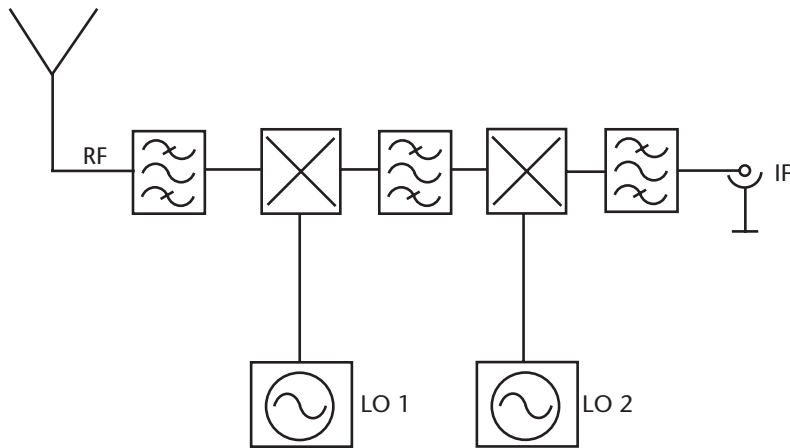
LO port. If an external radar is active, its signal goes from the antenna to the mixer, and the respective IF will be detected. However, the leakage of the LO from our RX may be strong enough to alert the enemy of the presence of a warning device and to be used as an ARM guidance signal.

Some means exist to enhance the performance of our simple RX. First, we may redesign the IF part to allow a larger separation between RF and LO frequencies. In the original layout the IF frequency was 80 MHz, which is nice for direct detection or amplification but will cause the millimeter-wave frequencies (for example the 35–40 GHz range) to be so close together that any conventional filtering will fail. If, on the other hand, we choose a considerably higher first IF, for example, around 2 GHz, we can insert a bandpass filter before the RF port of the mixer and so reduce the amplitude of the LO frequency at the antenna interface. Another possibility is to put an isolator between the antenna and the mixer. However, this would give perhaps only some 20 dB of reduction and unfortunately would completely destroy the NF due to the 1- to 2-dB insertion loss. Now that we have a higher IF, we might run into processing problems. These can be overcome by adding a second mixer to convert the 2-GHz signal into our original 80 MHz, if desired. Adequate filtering must be installed as indicated in Figure 4.23.

A special form of RF mixer is the diode multiplier, into which only one RF signal is fed. They generate normally harmonic multiples of their input and can thus be used as frequency-extension devices (e.g., for microwave oscillators). Similar limitations are valid here, too. The initial power level must be high enough and considerable filtering is mandatory. Available output levels tend to be fractional particularly, if very high multiplying factors are needed. An example of multiplier usage is shown later in Section 6.2.

#### 4.3.4 Amplifiers

The main purpose of RF amplifiers in our system is to enhance the signal level. Initially, two distinct cases existed, where we might have been interested in the



**Figure 4.23** An improved radar warning RX. By using two IF stages and suitable RF filtering we are able to reduce the LO leakage to the antenna port. However, this piece of equipment is not very simple anymore. For example, the amount of filters increases.

capabilities of the module to dig up the weak signal from noise or to give our TX output the full power needed. Recently, however, *monolithic microwave ICs* (MMICs) have appeared with sufficient performance for both tasks—of course, with reasonable limits. The most important parameters [15] describing an RF amplifier are listed as follows:

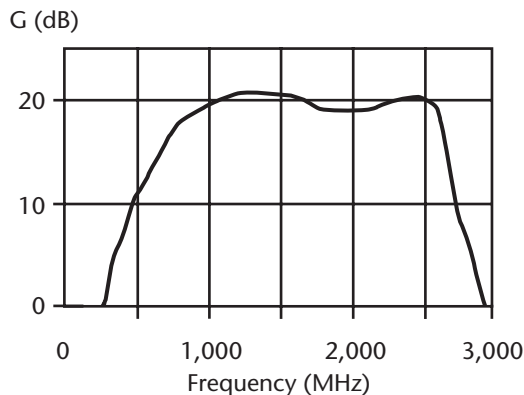
- Frequency range;
- Amplitude transfer function (gain as a function of frequency);
- Phase transfer function (also for stability analysis);
- Input and output matching;
- NF (mainly of preamplifiers);
- Maximum output power (often at 1-dB compression, mainly of power stages);
- IP3 or third-order intercept point (distortion behavior);
- The dc power consumption and voltage (often also of preamplifiers due to the cellular devices);
- Gain control range (when applicable);
- Cooling (when applicable).

The majority (the quantity in use) of today's amplifiers are based on semiconductors but the highest powers still come from tubes. Particularly TWTs are indispensable in radar systems and certain satellite system TXs. These tubes are based first on an electron beam that is accelerated with a high dc voltage between the electrodes and additionally on a multiple resonator structure in which the interaction between the original input signal and the beam takes place. Exotic lower frequency applications can also make use of classic tetrodes and sometimes also klystron amplifiers turn out to be feasible. Most higher microwave and lower millimeter-wave amplifiers rely on GaAs MESFETs [16], and above about 60 GHz also IMPATT and Gunn-diode reflection-type designs become practical.

Often the first question in system or equipment design related to an amplifier is “how much gain.” Commercial devices start from about 10 dB, and the more expensive TWTs, for example, give above 50 dB. Many MMIC building blocks make a nice compromise around 20 to 30 dB with a variation of about 2 to 3 dB across their entire useful frequency range [17]. Figure 4.24 illustrates one less typical measured result. The higher microwave range is naturally more complicated, and we easily end with a cascade of four to five modules just for the same 30-dB net gain due to additional losses in connectors and transitions to and from the microstrip MMIC board. NFs have practically achieved the man-made noise limit so that a further reduction seldom makes sense. Of course, the millimeter-wave devices still have some progress to show. Typical commercially available figures range from 0.3 to 3 dB, depending on frequency. AGC blocks tend to have inherently poor noise performance.

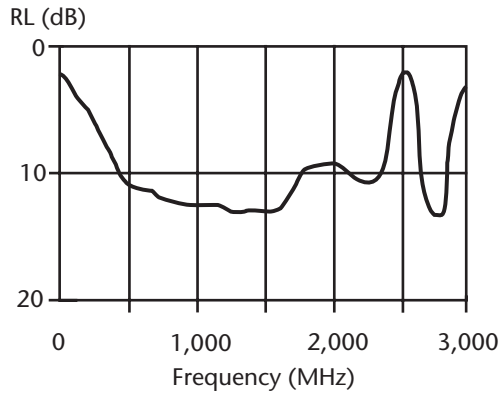
The impedance matching of amplifier blocks has evolved considerably and thus quite easy-to-use modules have appeared requiring just one or two external components. Wideband units have SWR values generally below 2 at their inputs, but 4 or even 5 occasionally appears as a respective output parameter. The best NF is normally not obtained simultaneously with optimum matching. Figure 4.25 shows the actual measured input return loss of our sample amplifier. The maximum output power of normal laboratory-grade blocks at the 1-dB compression point lays somewhere between +10 dBm and +40 dBm, again depending on frequency. Up to 40 GHz we can rather easily get about 200 mW, but from there on semiconductors tend to exhaust. Commercial VHF/UHF transistor amplifiers are available up to 20 to 40 kW of CW power, but these devices are actually rather complicated parallel amplifier matrixes used mainly for broadcasting and radar work.

Good spectral characteristics, which are mainly indicated as low spurious levels, are obtained at the expense of dc power consumption. Most transistor amplifier blocks work below 30% efficiency, but unfortunately very many cannot even achieve the 5% limit. The good thing in semiconductor blocks is that we normally rely on dc voltages less than or equal to 24V. Tubes are known to require huge anode voltages, up to and above 50 kV, which makes system prototyping interesting and sometimes also colorful. Actually TWT power supplies tend to be as



**Figure 4.24** The measured gain of a MMIC amplifier as a function of frequency. The useful range extends from about 500 MHz up to 2.5 GHz. However, in this case the variation is too high to be neglected in systems design.



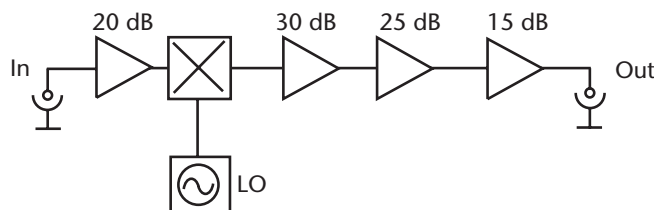


**Figure 4.25** The measured input return loss of our sample amplifier. Despite the careful matching network design, this result is not particularly good. Actually, this amplifier type is known to oscillate if, for example, it is cascaded.

complicated as the tubes themselves and, according to recent experience, have more faults.

A typical example of a set of amplifiers in a system is illustrated in Figure 4.26. There is one millimeter-wave LNA, and after the mixer we have a number of IF amplifiers, which have been arranged according to the best overall noise performance. This means that both gain and NF are taken into account. The total amplification in the chain is about 80 dB when we add the conversion loss in the mixer.

Cooling may be necessary both in LNAs and in final power stages. Cryogenic front ends often use liquid nitrogen. Extreme needs are fulfilled by helium, which provides an operating temperature of about 20K. If no other means exist to satisfy the over all noise floor requirement, this is the way to go, but operating complexity and costs tend to be considerable. Remember, that not all active modules can withstand such low temperatures either—the whole design must often be reconsidered. Even some conventional materials may suffer and become brittle. PAs from the 100-W class upward typically cannot rely solely on convection cooling through fins. Forced-air cooling is the most common choice, but its efficiency is limited. Systems involving liquids, mainly water, are again complicated and expensive and cause reliability problems. Two main variants are employed. Systems in which normal tap water runs isolated from the blocks to be cooled is easier to maintain but the cooling capability is limited. If we use electrically purified water, which is produced through



**Figure 4.26** A typical amplifier arrangement in a receiving system. First, a millimeter-wave LNA is used to make some distance to noise, and after the mixer several IF stages increase the level for detection. Filters have been omitted from the drawing for increased clarity.

ion-exchanging, for example, we can push the water directly in to the electrodes of the TX tube, but already the frequent change of the liquid may be too much of a burden. The highest cooling performance is obtained if we let the water vaporize in the tube and later circulate this water through heat exchangers.

Many amplifier problems are related either to neglected cooling of seemingly low-power units or to poor connection arrangements, which cause oscillations and spurious emissions. First of all, even amplifiers operating at, say +10-dBm power levels, need proper cooling due to their extremely low efficiency. We often have to dissipate 1 to 2W of heat from a small module. Take care to prevent any unintentional RF coupling into an amplifier block through its power supply lines. Of course, any direct coupling between the input and the output may be disastrous. Sometimes the shielding as supplied from the manufacturer is inadequate and allows a coupling through the electromagnetic field. Oscillation problems tend to be more severe, if we have to cascade modules for higher gain. Often manufacturers indicate whether a certain device is not recommended to be use in a series connection (cascaded). A RF amplifier can oscillate totally outside its nominal frequency range, and this characteristic may not come out until complaints from other users of the spectrum start to arrive.

#### 4.3.5 Oscillators

Oscillators are the primary source of RF power. They define the frequency of our setup and have an important effect on many other parameters such as spectral purity. Several oscillator types are manufactured. Fixed-frequency devices rely on some nontunable resonator, which often is a quartz crystal [18]. However, as the practical upper limit of crystals only extends up to about 200 to 300 MHz, *phase-locked loops* (PLLs) and frequency multipliers are needed to achieve the higher UHF and microwave bands. DROs are less stable but find extensive use as microwave sources and, with a suitable multiplier, can be used in millimeter-wave systems as well. In a DRO, the operating frequency is defined by the mechanical size and shape of a small dielectric part, often having the shape of a flat cylinder and placed close to the feedback line of the oscillator's active component (e.g., an FET). Tunable oscillators, on the other hand, must have a resonator with an adjustable center frequency. Varactor-based transistor oscillators are practical up to some gigahertz, and above that, for example, YIG devices are a suitable choice. Power oscillators, like magnetrons, are a special group, because they generate besides the frequency the final output power level of our TX.

Vital parameters when selecting an RF oscillator are the following:

- Center frequency;
- Tuning range (if applicable);
- Tuning speed (if applicable);
- Output power;
- Frequency stability (defined separately for different observation intervals);
- Frequency pulling and pushing;
- Frequency setting resolution (if applicable);

- Temperature characteristics;
- Spectral purity (e.g., phase noise, spurious, and harmonics).

Normal commercial building blocks are available through the entire technically interesting frequency range from less than 10 kHz up to 900 GHz and more. Often the design power level is somewhere between +20 and -10 dBm. Higher millimeter-wave frequencies must accept lower power, of course. Direct crystal oscillators end around 250 MHz. Most high-stability RF oscillators employ therefore some kind of a PLL, which needs a second oscillator for the actual output frequency. The operational idea is to feed suitably divided samples of the crystal oscillator's output and that of the VCO to a phase comparator, which then generates a suitable steering signal for the VCO. Commercial VHF and UHF modules without a crystal are available up to about 3.5 GHz. They give typically +10 dBm of output power but the user has to construct the phase-locked circuit. An example of PLL usage is given later in Section 6.2. DROs are available up to 22 GHz or so. Above that units with a built-in multiplier, for example, 3x, which is nice for 35- to 40-GHz operation, are sold.

Tunable oscillators are often called *voltage-controlled oscillators* (VCOs), because the parameter defining their output frequency is a voltage supplied to a specific connector. Infinite tuning ranges are not possible due to restrictions in resonator Q value and in the capabilities of the active component. Commercial modules yield, for example, about 1 GHz of frequency range around 1.5 GHz and have a power level flatness figure of 3 dB. Microwave YIG devices often cover typical radar bands or derivations thereof—for example, the common 8- to 12-GHz range. Very wideband types are also available, but they tend to cost a lot more. If the oscillator has some kind of tuning possibility, its stability is often not very good.

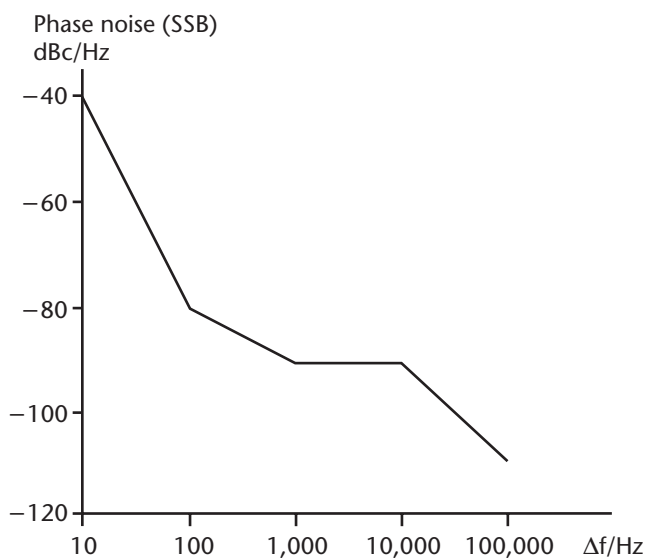
Sweeping and frequency agile radio systems need tuning speed or modulation bandwidth. Various measuring instruments like spectrum and network analyzers and chirping radars are good examples of this. Jammers and surveillance RXs naturally require as high a stepping speed as possible. The previous expression of FM is related to the fact that a VCO is actually used as an FM TX, when we rapidly tune its frequency. Some commercially available cheap UHF blocks give about 1 to 2 MHz of video bandwidth, which indicates that we could in principle achieve a sweep speed of about 1,000 MHz/ $\mu$ s. However, this is not practical, because already the linearity of such a sweep is hard to maintain. Many microwave YIG oscillators operate up to 100 MHz/ $\mu$ s. The indicated maximum tuning speeds assume all-analog control whereby very little information about the momentary frequency is available. On the other hand, if we select digital synthesizer-like tuning, we sacrifice the speed but know the actual frequency more precisely.

The signal quality of an RF oscillator is defined partly by its stability and partly by measuring the unwanted output across the entire spectrum of interest [19]. Far from the actual operating frequency we normally find the odd and even harmonics (integer multiples) of the initial frequency. They can be filtered out, but in a wideband system this is tedious. Also nonharmonic spurious signals may appear, depending on the components and circuit design. Close to the carrier we are typically interested in the phase and amplitude noise of the oscillator, defined as decibels with reference to the carrier (dBc or dBc/Hz) [20]. Here, both the actual figure (for example, -90 dBc) and the distance from the nominal frequency (for example, 10 kHz)

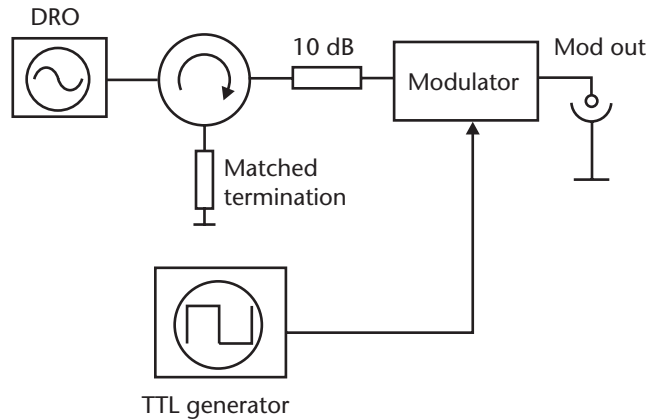
matter [21]. Naturally, the lower the noise and the closer we measure to the carrier, the better the result. One illustrative example is shown in Figure 4.27.

Time domain stability is often analyzed over time spans ranging from 1 second to 1 year. Typically the fundamental oscillator (often a crystal) defines almost everything above 10 seconds. Very short time frequency fluctuations are mainly caused by the PLL (if applicable). Aging is an unavoidable feature of crystal oscillators. This is so, because their operating principle is in fact of a mechanical nature. Sometimes we are able to achieve a reasonable stability only after 1 year of continuous operation without any power interruptions. Drift values around 10 *parts per million* (ppm) are common. Very high-quality designs are either *temperature-compensated crystal oscillators* (TCXOs), *digitally compensated crystal oscillators* (DCXOs), or *oven-controlled crystal oscillators* (OCXOs). DCXO and OCXO concepts can also be applied simultaneously for better performance. As the names imply, the main effort is targeted to reduce the effects of operating temperature on the frequency, but at the same time the general stability tends to be improved as well. Good TCXOs show frequency changes of 0.1 ppm, and oven units go down to 1 *part per billion* (ppb). The latter figure means that aging becomes the dominating factor.

Even high-quality oscillators are quite vulnerable to poor system-level design. The supply voltages of RF oscillators must be properly filtered and regulated. PLLs are very sensitive to supply effects, because there is a VCO in them. Isolate oscillators from their loads. Typical DRO units may experience frequency jumps of 100 ppm due to a return loss change from 10 to 20 dB. Ferrite isolators and amplifier/attenuator combinations work well (see Section 4.2), but often 20 dB of isolation is not enough. Similar problems may occur in commercial UHF VCO blocks. As the oscillators are inherently high-speed devices, they will notice even the shortest load “spikes”—for example, due to modulation or switching. Figure 4.28 illustrates one suitable way of DRO usage.



**Figure 4.27** A graphical presentation of *single sideband* (SSB) phase noise acceptance limits for a specific RF oscillator. Measuring difficulties tend to further limit the results very close to the carrier.



**Figure 4.28** If we expect the oscillator load to meet a varying mismatch, it is better to provide proper RF isolation. Here we use a ferrite circulator and a 10-dB attenuator before a high-speed pulse modulator.

A number of dedicated systems cannot work without even better frequency stability than that available with OCXOs. Fortunately, atomic frequency standards are also available. Miniaturized rubidium oscillators have reached the cost level of the highest performance quartz devices and typically provide at least 10 times better stability. However, they wear in use, and the expected service life may well be less than 10 years. Navigation systems, some network base stations, and radar can make use of the better frequency characteristics. For example, Doppler processing can be extended or integration time constants adjusted accordingly. Cesium oscillators have long been the cornerstone (primary reference) in the time and frequency world. Recent fountain-type units are believed to be more transportable than their classic counterparts. Often, the uncertainty level of cesiums is described as being somewhere between  $10^{-12}$  and  $10^{-14}$ . Very special systems could use such a master clock (e.g., to maintain coherence over extend periods of time in a spread spectrum communication network). The designer must be aware of the cost, which is around 100 times that of good crystals, and about the fragility of these units. Finally, if extreme short-term stability must be obtained, hydrogen masers may be an alternative. They are claimed to outperform cesiums for time spans shorter than one day or so. However, even active hydrogen masers seem still to have some very long-term drift. Maser oscillators find use in very sophisticated scientific instrumentation, space research, and some ultracoherent radar networks. Many specialists agree that a hydrogen maser should be purchased together with its maintenance engineer.

In Doppler radar systems, particularly if we are dealing with CW designs, the most important frequency stability region is that where the target echoes appear. Often this is something from a few kilohertz to some tens of kilohertz around the carrier. Here we have to guarantee the phase noise performance of our oscillators. This means, that atomic clocks generally cannot give substantial improvement, because they also have to rely on their internal crystal units for the very short observation times. One of the major tricks in the past was to use cavity-stabilized klystron oscillators. The practical utilization of oscillators in RF systems will be described in Section 6.9.

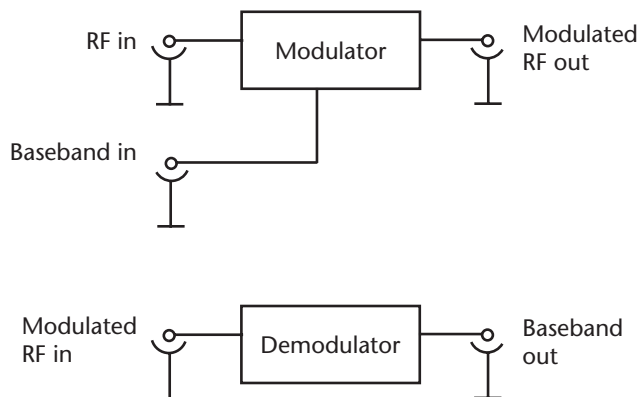
### 4.3.6 Modulators and Demodulators

Some very elementary radio systems, like the NDB/ADF equipment used for air navigation could in principle operate just with the plain carrier (the CW signal as generated by a perfect oscillator) [22]. Practically almost every RF system makes, however, use of one or several kinds of modulation and demodulation schemes. The purpose of modulation is either to put our information as a part of the original wave or to enhance the processing performance of signal detection during reception, particularly in radar applications. Two resources should be monitored when selecting the modulation type and associated technologies. Again, RF bandwidth should not be wasted. Additionally, we should try to get the most transfer performance at the lowest possible TX output power level. This implies that complicated modulation technologies may be advantageous but surely challenge the demodulator designers. The definition of transfer performance is not so straightforward, though. For the engineer, it might be the BER value; for the manager it might be something having a closer relation to value-added services, for example.

Basically, we have three features of the carrier wave in our use, namely its amplitude, frequency, and phase. Naturally, we can use combinations as well, but they may appear as unwanted modulations, too. The nature of the modulating, or the baseband signal can be analog or digital [23], a pulsed square wave found in typical radar systems being somewhere in between these two. In most cases, the idea of modulation is to cause sensible variations to one or more of these signal parameters and then compare the momentary signal with a “reference” in the demodulator. Depending on selected technology, this reference can be a part of the originally transmitted RF signal or it can be regenerated in the demodulator.

If we think of a modulator or a demodulator as a two-port or three-port device indicated in Figure 4.29, its main characteristics are listed as follows:

- Usable RF bandwidth;
- Usable baseband bandwidth;
- Linearity (conversion from baseband amplitude or value to the modulated RF parameter);



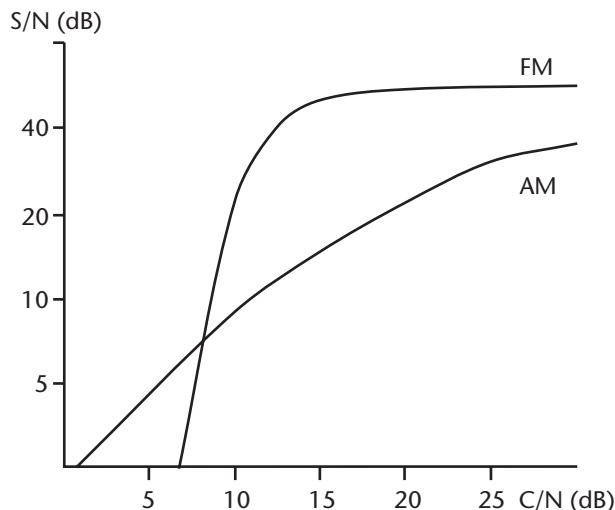
**Figure 4.29** In its simplest form, a modulator can be discussed as a three-port having connectors for the incoming CW and baseband signals and a third one for the modulated output. The demodulator can basically have just the input for the modulated wave and the baseband output.

- Distortion;
- Modulation “range” (depth in AM, deviation in FM);
- Conversion sensitivity (in communication demodulators, the required C/N for a specific S/N).

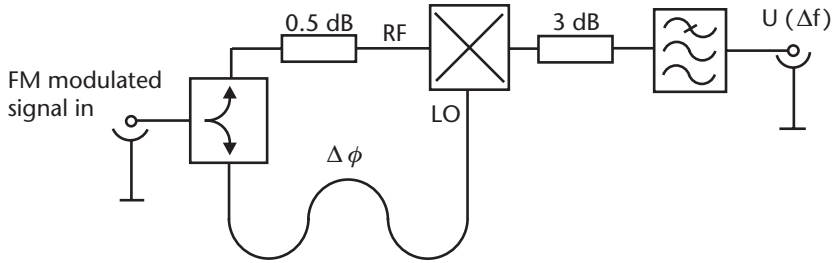
People working outside the telecommunication sector may be confused about the abbreviation C/N, which stands for carrier-to-noise ratio. It is mainly used to overcome the difficulty of simultaneously dealing with two noise ratios—one before demodulation or detection and the other immediately after that process. Figure 4.30 tries to highlight this issue by showing how the S/N behaves differently as a function of input C/N in FM and AM systems.

Amplitude modulators can be assembled for example from AGC blocks, from pin-diodes (either in analog configuration or as switches) or from double balanced mixers. The detection of an AM signal is often done with the ubiquitous diode circuit found in every textbook. FM typically requires some kind of a controllable oscillator. Normally we use a VCO or a XVCO for better stability. Demodulation is accomplished through a discriminator, a PLL, or a mixer arrangement, an example of which is shown in Figure 4.31. Phase modulation is similar in nature and thus the technologies applied do not differ very much.

When configuring a new system concept, the difficulties associated with modulators or demodulators often start with the improper RF bandwidth. Even the center frequency of commercially available components may be totally wrong for our purpose. This is partly caused by the past standardization of hardware, which has led to the use of fixed IFs of 70, 21.4, 10.7, and 0.455 MHz. These frequencies used to be the normal IF bands of all radio hardware, and still many building blocks stick to them. Other commercial devices are available for the modern cellular phone technologies, mainly in SMD chip form. If, however, our application calls for something different, we might consider either mixing, which enables us to shift the wanted frequency to be within the band of building blocks at hand, or we can construct the



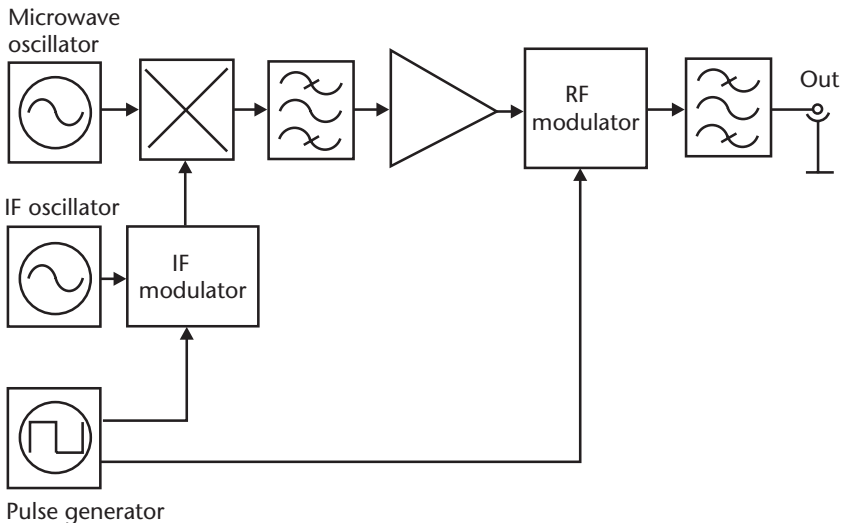
**Figure 4.30** AM and FM demodulators yield very different transfer performance when converting their input C/N to output S/N. The FM threshold is clearly visible at around 8 dB.



**Figure 4.31** A simple wideband FM demodulator can be assembled from a Wilkinson-type power divider, some transmission line, a double balanced mixer, and an IF filter. The attenuator is optional. The difference in electrical length as a function of momentary frequency in the cable first converts FM to phase modulation, which is detected by the mixer.

modulator from existing general-purpose blocks. In the latter case, however, we must perform extensive studies related to the spectral characteristics of the resulting waveform.

The second problem might be modulation bandwidth, which becomes particularly important in high-speed systems and in radar use. Partly parallel to it is the question of FM deviation, which could be inadequate as found in standard off-the-shelf building blocks. Some designers circumvent this by putting the modulated oscillator at a much lower center frequency and then using a frequency multiplier, whereby the deviation will increase respectively. Unfortunately, also all the unwanted defects of the frequency domain will be larger. Pulsed AM can suffer from poor ON/OFF ratio. Cascading identical blocks is possible and will improve this figure but at the cost of longer rise and fall times. If these are vital, we can try to find even faster individual modules. Figure 4.32 illustrates a “distributed” pulse modulator in which part of the ON/OFF ratio is obtained at the IF and the rest at the final operating frequency. The benefit here might be, for example, a lower total cost.



**Figure 4.32** Adequate modulation depth can be assembled as shown here from a combination of IF and RF modulators (actually pin-diode switches). The ON/OFF ratio exceeds 100 dB.



### 4.3.7 Upconverters/Downconverters

Commercial, ready-made blocks for frequency conversion are available particularly for the well-established but sometimes slightly old-fashioned communication and radar bands. Basically they are simple modules to use. We put our own spectrum in and get the same spectrum out but transposed around a different center frequency. Because these converters usually employ active components, we can assume some gain, too. Normally, the key parameters are listed as follows:

- Frequency range of the RF interface;
- Frequency range of the IF interface;
- Frequency stability (similar parameters apply as for oscillators; see Section 4.3.5);
- Level of harmonic and nonharmonic spurious responses (RF port for upconversion, IF for downconversion);
- Required IF level (in upconverters);
- Sensitivity or NF (in downconverters).

Frequently encountered problems in microwave upconverters are related to poor purity of the output spectrum. The internal LO signal, which has to be of very high power, is often inadequately attenuated. Also the level of the image frequency might be too high. Upconverters usually have extremely low dc efficiency. Sometimes our system has to use a very high IF input level to make the upconverter work properly. Typical levels can be up to 20 dBm. The frequency stability of the internal LOs is not always very good, because compact units seldom have enough space for oversized crystal oscillators. This can be solved through AFC in the RX but can in some cases cause additional problems in staying within the allocated bandwidth.

### 4.3.8 Power Supplies

Practically all RF systems need power supplies, which can be based on batteries (either rechargeable or dry cells), rotating generators, or normal mains voltages. The inherently low efficiency of most RF hardware makes the load on power supplies high. The analog nature of operation in such devices as oscillators, amplifiers, and modulators sets very high requirements for the stability and purity of the voltages. Additionally, various biasing arrangements call for several dc levels, which must sometimes even track each other. Split supplies are quite common, particularly during prototyping. The cost of the complete power supply assembly can be about one-third of the overall list price of a complicated RF system.

Commercial dc supplies come in all-linear and switched configurations [24]. Where interference requirements permit, a switched-mode supply runs cooler and thus improves the overall reliability. Even very well-known types and brands have, however, occasionally shown unexpectedly high levels of ripple or RF emissions, which may seriously hamper the operation of our system. Unwanted FM in oscillators and AM in amplifiers may turn up synchronized to the residual 50- or 60-Hz-based voltage (actually often 100 or 120 Hz due to full-wave rectifiers) or at one of the switching frequency harmonics. Particularly PLLs and their VCOs tend to

be sensitive to supply variations. Sometimes block-level tests are easier to carry out by using heavy-duty rechargeable batteries equivalent to the size used in cars instead of a main supply. Very difficult situations may arise if the switching frequency harmonics fall into our primary processing band. This is quite possible, because the striving toward smaller power supply modules has caused a continuous increase in the switching frequencies.

Filtering of individual supply lines when they are about to enter a separate RF block is almost mandatory. High-quality RF feed-through capacitors and entire pi-type filters are commercially available and often come as integral parts of functional modules (e.g., amplifiers). The trick in selecting the most suitable one is in finding the vulnerable part of the spectrum, because filters tend to cover only a limited bandwidth. Sometimes a filter is not suitable, if we want to have for example a wide-band gain control input. In these cases a protected coaxial cable is often an effective alternative. A final remark on supply filtering of RF building blocks: If you do not use all metallic RF-tight enclosures, there is no reason to put expensive filters into the circuit either.

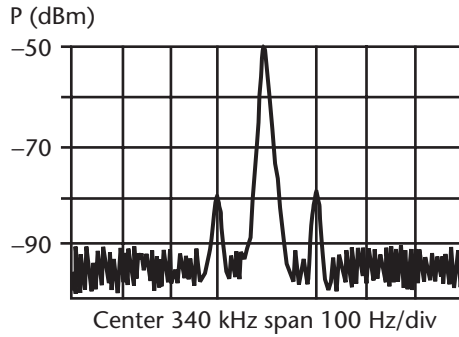
Some RF systems or blocks set specific requirements for their power supply. If pulse modulation is used, we have to make sure that the voltage drop does not have access to the critical parameters and that we are able to arrange all the needed energy to the RF output stage at the very moment of the pulse. A recommended design practice accepts a certain voltage drop and adapts all hardware to it rather than trying to construct a “monster” supply that would keep the voltage constant. Even such a supply would not do it, anyhow. Interesting phenomena may appear if we are working with a combination of a pulsed RF waveform and a switched-mode power supply. This could be the case in a radar tube *high-voltage* (HV) system. If the mutual timing of the RF pulses and the switching waveform is not appropriate, or we do not use intelligent filtering, we can blow both the supply and the tube.

Finally, RF devices employing static or low-frequency magnetic fields (for example, YIG-tuned oscillators and filters) need special protection against stray coupling from the possible mains transformer. The intensity of the magnetic field next to the transformer windings is proportional to the dc loading. In extreme cases, very unfortunate equipment layout combined to a straight forward circuit design can lead to similar problems just through simple magnetic field coupling to PCB strips. One result is illustrated in Figure 4.33.

## 4.4 Mechanics

Besides antennas and antenna towers, many RF devices and systems need a considerable amount of mechanical assemblies, which do not have a direct primary RF function but without which the whole arrangement could not work properly. The majority of these items are some type of enclosures or cabinets or supporting parts for precious RF blocks. Common desires or design parameters related to mechanical accessories are listed as follows:

- Stiffness and rigidity;
- RF shielding;



**Figure 4.33** When everything goes wrong, we can get mains transformer-based FM just through coupling to PCB traces. This is a measured spectrum of an aviation NDB, which suffers from too high levels of unwanted FM. The transformer was located beneath the synthesizer unit in the same 19-inch equipment rack.

- Thermal transfer characteristics;
- Weight;
- Ease of prototyping or assembly.

Although we are used to thinking of cellular phones as almost all-plastic devices, protection against RF leakage into or out of our construction is one of the first topics when defining the mechanics. Light alloy boxes of various shapes and sizes are commercially available, and a skilled workshop team can easily make arbitrary shapes according to our wishes. If adequate care is exercised when selecting the distance between galvanic junctions along the edges of our enclosure, performance is normally guaranteed in this respect. Low-frequency shielding calls for ferrous materials, which tend to be more tedious to shape and have typically inferior corrosion resistance. A single light alloy sheet enclosure (2–3-mm wall thickness) gives about 60 to 90 dB of RF attenuation from about 30 MHz upward. Similar levels are difficult to obtain at HF. Sometimes, especially when working with millimeter-wave modules, we may want to add “internal” attenuation by covering the interior of a metallic enclosure with absorbing material. This can effectively reduce unwanted coupling between units next to each other.

One overlooked characteristic of RF enclosures and equipment racks is their mechanical stability against deformations caused by vibration, shocks, or temperature gradients. Many RF blocks have at least some individual components, the performance of which partly relies on their physical dimensions or shape. A simple wire-wound inductor is an obvious example, and similar features can be observed in passive microstrip elements (couplers and dividers) and in YIG devices. If our external mounting arrangement is not stiff enough, we may encounter unwanted modulation sidebands at vibration-related frequencies. Sometimes this process is referred to as “microphonics.” Alternatively, our waveguide-mounted components may simply get loose, and the whole block will cease to work. In particular, mobile systems found in numerous military applications belong to this group. The spectrum of vibrations is characteristic of each platform and can, among other things, give the potential enemy a nice way of target recognition. Figure 4.34 highlights the mechanical support assembly of a millimeter-wave laboratory scanner. It is made of



**Figure 4.34** An example of a stable positioner for millimeter-wave laboratory work. The span is about 4m and positioning uncertainty less than 1 mm.

commercial aluminum profiles and tailored fasteners. Positioning accuracy is maintained better than 1 mm on all three axes and throughout the entire span (4m) of the system.

Maintaining the required dimensional tolerances—when dictated by our electronic design and layout—seems to be a challenge [25]. Improper impedance matching is one of the obvious consequences of careless mounting or assembly of coaxial connectors or waveguide transitions. Microwave and millimeter-wave building blocks typically need milling accuracy, but VHF and UHF circuits can sometimes be put into simple cut-and-bend-type sheet metal boxes. After writing this, we have to emphasize that overengineering is no good either. Polished brass surfaces look nice and professional indeed, but if there is no direct influence on performance characteristics, why to pay that extra? Painting is another issue. It is sometimes mandatory (e.g., in order to prevent corrosion or overheating), but if done, it should be done properly. Very often paint gets into wrong places and impairs galvanic contacts. It adds weight and may turn out to be a considerable financial burden, if large surfaces are involved.

## 4.5 Purchasing Modules for Equipment Development

Depending on the size of the project and the institution in charge of it, the system designers may well be in the position of carrying out the actual commercial negotiations related to purchasing the desired modules. Larger enterprises and government offices naturally have separate departments to take care of these functions, but also there it is beneficial to know some of the rather brutal backstage matters.

Manufacturers do not produce RF building blocks for charity activities—they or their shareholders have a distinct commercial interest. Further on, dealers are in a similar position, but normally work in smaller scale. However, this means that the terms of purchasing must be carefully checked and negotiated. Take into account the following points:

- The prices you see (e.g., on the vendor's WWW pages) are not the ones that you will find on the dealer's quotation. Typically, the added cost is something from 50% up to 200% or more, being generally more severe in overseas products.
- If advertising indicates list prices as "starting from \$1.5," you are surely not able to use such a module in your project. Those cheap items are normally valid for chips, which are intended for large-scale cellular manufacturing and related mass-market tasks.
- Even if you find a product that has the appropriate technical specifications and fits within the budget, you may not be able to purchase it. It simply is not for sale in quantities below 1,000, for example.
- Some sensitive RF products, such as ultra LNAs, high-speed millimeter-wave switches, tunable wideband oscillators, or compact TWTs, are not sold to everyone. Normally manufacturers "forget" in these cases to respond to your request of proposal, or, if they give a quotation, the delivery time and price have been put outside any sensible limits. There once was—and still may be—a group of manufacturers who sent lower quality components to customers having inadequate acceptance testing. Once caught, these vendors also rapidly changed their habits.
- The local dealer may be in reality just a man and a suitcase. The expertise area of this individual can be quite far from RF engineering. Technical support may or may not be available. It is advisable to check this before ordering, if you expect to run into trouble. Alternatively, you can try to contact the manufacturer directly, but they are often reluctant to bypass their dealers.
- Delivery times are normally exceeded. If advertising promises weeks, expect months. Shipping can really mean ocean traveling, which in fact takes quite long.
- Products shown on data sheets may become obsolete rather suddenly. One reason to put a certain item on such a list is the termination of a large long-term ordering contract. This, of course, has nothing to do with the validity of the module itself. Alternative types simply may not be available. Consider this if your project is expected to run over several years. Perhaps you can buy the spares now?
- The country of origin may not be the one where the manufacturer's headquarters sit. If you expect hardware from Switzerland and get items made on the other side of the globe, there can be a difference in quality as well.
- If you desperately need a specific module or component, for example during the refurbishment of equipment manufactured in the former Soviet Union, do not panic. Items are still available; you just have to know where to ask for them!

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# Antennas and Associated Hardware

RF antennas range from simple whips used in low-cost walkie-talkie toy phones and FM RXs up to huge arrays formed from thousands of individual elements as is the case in the Patriot missile system's radar or even from tens of thousands like the *theater high-altitude area defense* (THAAD) antenna system [1]. Many antennas are still completely passive, just pieces of various metals and some dielectric parts in between, but more and more often also the antenna itself has very sophisticated semiconductor circuits in it. Two distinct parameters roughly define how an antenna looks. The frequency range has its effect just like with any other RF building block. The second issue is the purpose for which an antenna has been designed. Is it supposed to be stationary, or shall it turn and point its pattern toward any target in the hemisphere?

Antennas are perhaps the most flexible, efficient, and vulnerable elements of any RF system [2]. This is caused by the fact that an antenna has to be the one and only interface between our remaining RF building blocks and the propagating wavefront in the space surrounding it. Intelligent antenna design and construction can yield superb performance whereas overlooking or neglecting its possibilities will seriously hamper performance characteristics. Unfortunately, in many cases the latter has been the practice. It is obvious that many lower HF antennas unavoidably are large in size and may well need a couple of hectares for assembly, but today's high microwave and millimeter-wave systems can make use of small-size and light-weight antenna constructions [3]. If started from the theory of electromagnetic waves, antenna engineering is a very sophisticated branch of science. Good results on the systems level—with considerably less scientific output—can be obtained also by employing cookbook-like solutions and by understanding a set of elementary rules for antenna or antenna array construction.

An antenna is connected to the RX or TX through appropriate feeding lines, which can be waveguides or transmission lines. Dedicated RF connectors are also necessary. A brief commented presentation of them has been included, because their simple appearance tends to cause unnecessary design faults on the system level. Finally, many special systems must allow some type of controlled mechanical movement of associated antennas. This may be arranged through flexible waveguide sections, cables, or by using rotary joints. These all are introduced later in this chapter. Four examples are included as well. They deal with a simple vehicle-mounted AJ antenna system, show in detail a complete but still elementary adaptive antenna design, highlight problems related to mobile measuring antenna installations on vans or ships, and introduce a small tracking satellite ground station antenna arrangement and difficulties observed during its commissioning.



## 5.1 Antenna Selection Criteria

Three main themes must be discussed when trying to define the process of choosing an antenna for a specific application [4]. They are described as follows:

- Characteristics toward the surrounding space: radiation properties;
- Characteristics toward the feeding system: impedance properties;
- Characteristics toward the operating environment: mechanical performance.

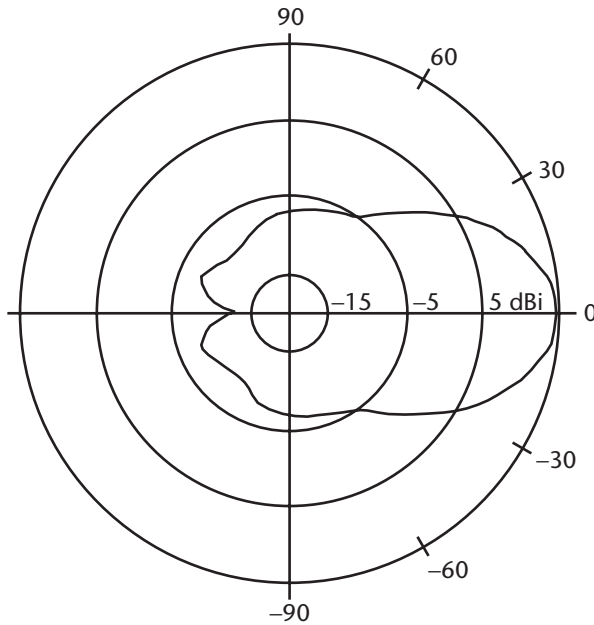
We must first define some descriptive parameters and terms frequently used in the field of antenna engineering. Let us first think of a simple flashlight that has a bulb, a mirror and a lens (and exhausted batteries). If we remove the mirror and the lens, the small bulb will illuminate a considerable space around it, but the luminous intensity will be very low all over. When we put back the optical components, we get a sharper beam of light, but it can only illuminate a rather narrow sector and the rest of our surroundings will be dark. If we take our flashlight out when it is misty, or we use it in smoke, we can see the shape of the light beam. If we ask somebody to hold the lamp, we can have a look from a direction perpendicular to the beam and observe how it becomes wider when going away from the bulb. During this experiment, the electrical and optical power input to our system have been constant; we just focus it in the desired direction.

Antennas work much the same way as the flashlight lens and mirror, and higher microwaves and millimeter waves can actually make use of analogous focusing elements. Antennas have directivity and gain, which means that they are often capable of focusing our power toward a desired direction. Numerical values may be expressed in decibels referred to an isotropical radiating element (dBi) (think about the bulb in our example). Alternatively, we sometimes see decibels referred to a dipole (dBd). In practical antenna engineering the difference between these two is roughly 2 dB. For example, if gain is specified as 6 dBi, it will be about 4 dBd. The term “gain” is not that simple either. We actually have two parameters, which are called directive gain, or sometimes directivity and power gain. In imprecise conversation these two are unfortunately mixed. The analytical expression combining these two is simply

$$G = \eta G_D \quad (5.1)$$

where  $G$  is the power gain,  $G_D$  the directivity, and  $\eta$  the radiation efficiency of the particular antenna. Often the symbol for directivity is just  $D$  but we use in this book  $G_D$  in order to avoid confusion with antenna diameter  $D$ , which is frequently needed in this and following chapters. Theoretically  $\eta$  could be 1, but practical antennas often show efficiencies less than 0.6 or so.

Besides directive gain and power gain, an antenna’s radiation pattern is often described as a set of numerical data, which may or may not be supplemented by measured pattern plots. They often appear as slices of the actual three-dimensional situation as illustrated in Figure 5.1. The situation is analogous to looking perpendicular to the beam of the flashlight in smoke or mist. Parameters of interest include, very often as a function of operating frequency, the following:



**Figure 5.1** A typical measured (scalar) antenna radiation pattern in polar coordinates, based on observed electric field strength  $E$  as a function of turntable angle.

- Main beamwidth [at 3-dB or, sometimes, at 1-dB points (e.g., for tracking antennas)];
- Sidelobe level;
- Front-to-back ratio;
- Polarization and polarization purity;
- Steering angle (if applicable);
- Complete phasor pattern (magnitude and angle) for array use.

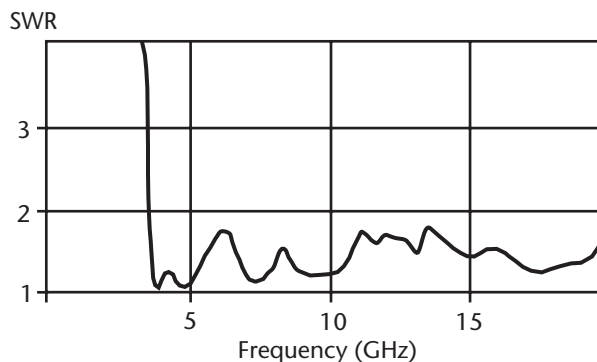
The beamwidths of antenna patterns are typically defined at 1- and 3-dB amplitude points and sometimes between first-amplitude nulls as well. Here we normally speak about the main beam, but practical antennas have sidelobes as well. Their relative amplitude is expressed as negative decibel values; for example, an antenna sidelobe level is  $-25$  dB. This tells us that the observed electric field magnitude in the direction of the largest sidelobe is 25 dB less than the field strength at the main beam maximum. It is intuitive that higher gain means narrower beam. Unfortunately, very high gain can in some antenna types cause higher sidelobes, too.

The radiation characteristics of an antenna are “scaled” by the wavelength in use [3]. A particular antenna can be scaled to smaller size for use with similar performance at higher frequencies, respectively. This means also that a comparison of typical performance figures such as gain or beamwidth can be misleading if antennas for totally different frequency bands are discussed. A 40-dB reflector antenna is very handy at the Ka-band but would be monstrous in size at VHF. For certain computational purposes or for very wideband cases as illustrated later we may want to use antenna gains below 0 dBi. Normally commercial devices range from about 2 to about 60 dBi and even more, but only if we are working at sufficiently high

frequencies. The higher gains are mainly obtained with parabolic reflector antennas. Many communication applications use beamwidths from  $5^\circ$  up to full  $360^\circ$ , whereas radar and scientific measurements go well below  $0.1^\circ$  in one or two dimensions. Sidelobe levels are normally below  $-20$  dB, and often the design target lays somewhere around  $-40$  dB. Very high performance antennas—or ULSAs—can achieve even  $-60$  dB of sidelobe attenuation, but this often happens as a compromise between the frequency range and maximum gain. Polarization characteristics include vertical, horizontal, and two circular ones. A simple analogy to everyday life can be found in sunglasses. They tend to limit the entrance of one linear polarization, but if you bend your head at about  $90^\circ$ , you suddenly start to see the annoying reflections. This happens because the plane of the polarization filter in the glass now allows light to enter your eyes. In antennas, combinations of several polarizations are also possible with special feed arrangements.

For example, point-to-point microwave links and terrestrial television broadcasting can make use of fixed and stationary radiation patterns, after sufficient beamwidths and directions have been established. Mobile systems, space applications, and radar often require that we are able to adjust the direction of our beam according to the situation. This is possible either through electromechanical rotation or scanning of the entire antenna (sometimes even hydraulic) or by electronic beamsteering. Antenna arrays make a good candidate for the latter. In this case, the physical antenna is fixed. The signals coming from (or going to) individual elements are given dedicated phase shifts and controlled amplitudes so as to have the resultant phasor field outside the array pointing toward the desired direction. This process will be highlighted through a practical example in Section 5.2.2. More complicated systems allow simultaneous creation of pattern minima toward hostile directions as well. Electronic beamsteering may be limited to a range of  $20^\circ$ , but it can be extended across the entire circle.

The transmission line interface of an antenna is characterized by its complex impedance, but we are often satisfied just with the scalar return loss or SWR; see Figure 5.2. If an antenna is matched to our transmission line, both have the same impedance—for example, 50 ohms. Poor matching at such a junction will cause much of the input (or incident) power to be reflected back toward the line or back into space. The return loss and SWR definitions of antennas are basically analogous to other RF hardware [4]. Thus, an SWR value of 1 means that we have perfect



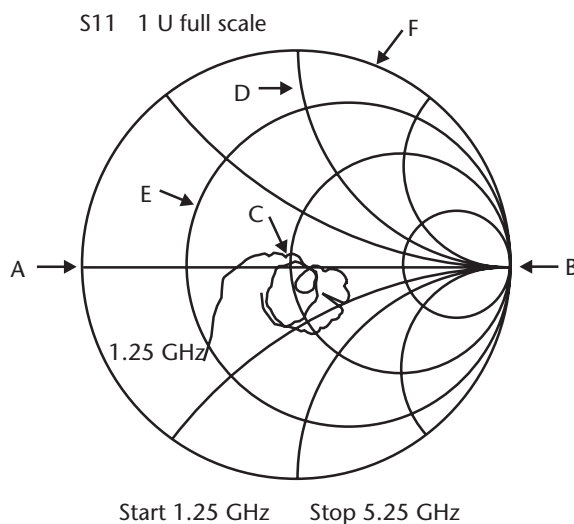
**Figure 5.2** Often it is sufficient to measure the scalar return loss or SWR of an antenna. In this case it is of particular help, because we avoid confusion due to the very wide frequency span.

matching. The term return loss is related to the amount of power coming back from a transmission line junction (e.g., at an antenna interface). Theoretically perfect match is available only if the return loss value in decibels approaches negative infinity. Many commercial antennas show SWR values from 1.1 to 2, and some low-cost examples show up to 4 or even 5 within their operating frequency range. Reasonable return loss results range from about 10 to 40 dB.

Single antennas, which are not used as a part of a larger array, can be judged by their SWR. A SWR very close to unity, perhaps 1.05 or even less, means that we can often ignore the phase of the reflection whatever it is, just because its amplitude is so small. If, however, we intend to do some intelligent beamforming with a typical antenna array, we have to know the phase values as well, particularly when the matching is not very good [5]. The matching of a similar wideband antenna as was used for Figure 5.2 is shown as a Smith chart plot in Figure 5.3. The basic features of the Smith chart are quite straightforward. First, the left-hand extreme of the real axis at A is the short circuit and the right-hand end at B an open circuit. Perfect matching is obtained in the chart center, marked as C. The curved lines D indicate the reactive part of the impedance and the circles E the resistive part. Additionally, there is a wavelength scale on the outer circumference F.

Depending on the specific application, we may have to set a very strict return loss limit. Many receiving antennas do not have to show better than a 20-dB return loss; some lower “commercial grade” devices can hardly meet 10 dB. On the other hand, if we are supposed to put 2 mW of TX peak power into the feedline, we may want to have a return loss value better than 40 dB, which is not easy to achieve particularly, if we want a very wide frequency range. Regardless of the actual return loss value, we have to be aware of the maximum power-handling capability of our transmitting antenna, which might be limited (e.g., by dielectric losses or by electric field strength in transmission lines).

Many real-world antennas are mounted outdoors, in high towers. It is thus obvious that they have to be able to survive in the extreme wind conditions, which



**Figure 5.3** The complex impedance Smith chart plot of a similar antenna, which was used in Figure 5.2. Letter A indicates short circuit, B open, C matched, D reactance axis, E real part “axis,” and F distance scale in wavelengths.

may include severe icing as well, and beyond the arctic tundra area [2]. Similarly, large parabolic directional antennas used for radars and radio astronomy have to be protected against wind forces—not only to make them survive but to enable precision angular recordings under normal prevailing wind conditions of up to 30 m/s or so [5]. Environmental problems may also come up in vehicular installations, particularly onboard aircraft. Historical reasons have caused both civilian and military aviation to use VHF communication frequencies, where wavelengths typically range from about 1 to 3m. To be electrically efficient, a passive antenna should be a considerable fraction of the wavelength in size, which might be too much in this case if flying at a speed of Mach 2—at least as a nonconformal construction (a definition is in Section 5.2). Some ubiquitous MW or SW loops with active preamplifiers are an exception to this size-scaling rule, but unfortunately they are suitable for receiving only.

When starting the search for suitable antennas for a specific new system, we may first want to perform a rough analysis of the following things [5]:

- The approximate center frequency of operation;
- The relative bandwidth (measured from the center frequency);
- Gain as calculated from the link budget or radar equation (as applicable);
- Main beamwidth as defined by the wanted angular resolution (as applicable) or coverage;
- Polarization;
- Direction of energy flow (TX or RX or both);
- Physical size limitations [as applicable (e.g., onboard vehicles)];
- Beamsteering possibilities (if needed).

Generally, some kind of a compromise must often be made between the various desires and taking into account the project budget. The quite arbitrary RF impedance of, say, a bent two-inch steel nail as the radiating element of a monopole, can be matched to 50 ohms at almost any discrete frequency, but this will not make the nail a reasonable antenna. Its radiation efficiency might be minimal. The pattern might have several, almost equal-amplitude lobes pointing to various directions and deep minima all over, of course depending on the frequency of interest. Very high gain is not easily combined with very wide bandwidth, for example. Parabolic reflector antennas with suitable feeds tend to yield the best results here. Extremely high gain also means that the beamwidth will be narrow—at least for one planar cut of the radiation plane. This implies that we have to accept very strict antenna alignment requirements, which is the case in many stationary microwave point-to-point links. Let us assume, for example, that we happen to have two 40-dB antennas at the extremes of our link path looking toward each other and their main beam width is around  $1^\circ$  to  $2^\circ$  as can be deduced from the equations in Section 3.3.2. We can expect this maximum gain in our system calculations only when both antennas are adjusted in their azimuth and elevation angles within  $0.1^\circ$  to  $0.3^\circ$  of the true line. For a 0.4-m Ka-band reflector such an uncertainty means less than 1 mm at the outer rim. Alternatively we must use very different vertical and horizontal patterns. This gives some relief for one angle but sets even tighter requirements for the other, if we

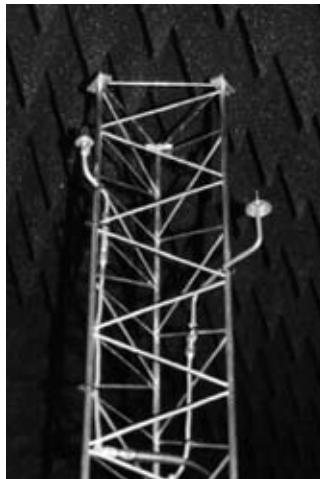
cannot reduce gain. Having unequal beamwidths can be also a useful feature (i.e., in surveillance radars, where a considerable variety of elevation angles caused by target altitudes and distances as seen from the radar station can be covered with one and single scan). It is also rather challenging to create a complete circular coverage around one antenna, if we can not use just one monopole-like antenna or a vertical array of such elements and if this arrangement is not mounted on top of a high antenna tower. Figure 5.4 shows a typical but less efficient monopole antenna mounting on one side of a tower. This installation practice is often only possible if the top position has already been reserved, but it will create a deep minimum approximately to the direction of the tower “shadow,” depending on actual geometry and wavelength, of course. Numerous aviation and maritime stations as well as cellular sites seem to be forced to this. Fortunately, some of the area covered by the deep minimum will be handled by terrain multipath characteristics.

## 5.2 Some Antenna Types

In the view of system design, antenna applications can be roughly divided into three categories. They are partially overlapping, but it might still be useful to speak about them:

1. Standalone single antennas that do not rely on their mounting environment for pattern shaping;
2. Antenna arrays made of elements;
3. Integral antennas that make extensive use of their mounting platforms for pattern characteristics.

The first class naturally contains the vast number of individual antenna types and devices, like a simple dipole, loop, Yagi, or horn. Many systems make extensive



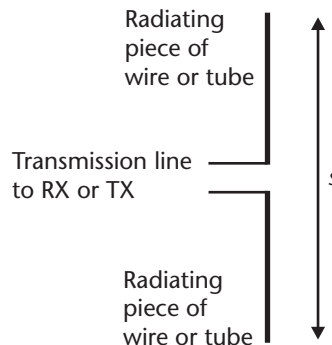
**Figure 5.4** A monopole antenna mounted on one side of the antenna tower will not create circular coverage but there often is a deep minimum toward one sector. The width and direction depend on the actual layout.

use of antenna arrays, partly because they give interesting possibilities for “live” pattern adjustment. Arrays can be either standalone units or more closely integrated to their platform. In the latter case, we often speak of conformal antennas although this is not necessarily always the case. The term “conformal” is used to indicate that the antenna or an entire array is designed and assembled in such a way that its three-dimensional shape follows that of the supporting platform. The aim is often to get simultaneously optimized aerodynamics and RF performance. A very simple case would be to have a planar array as a part of a sheet skin of a vessel, for example. More design effort and manufacturing skill may be required when we try to construct antennas in the conformal way on top of a missile’s nose cone. In the best case such conformalism is obtained by designing the antenna or individual elements as a physical part of the outer skin (e.g., as slots, which are described later in this chapter). If this is not possible, thin and small elements such as PCB patches might be mounted on top of a surface. For example cars, trucks, tanks, ships, and aircraft typically have single antennas or arrays of antenna elements, the performance of which highly depends on the mounting environment. In this case we must take the vehicle into account as a vital “part” of our antenna arrangement.

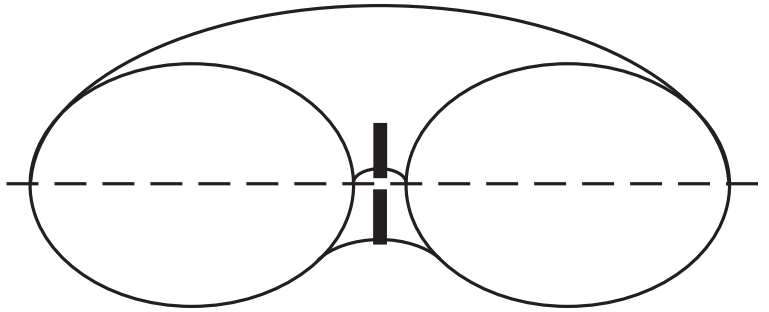
### 5.2.1 Individual Antenna Elements

Monopole and dipole antennas have for long been the choice from the lowest HF up to UHF frequencies and above [2]. A simple ground-plane monopole has been successfully used at 15 GHz, for example. Figure 5.5 shows the classical dipole antenna. Its radiation pattern resembles a doughnut as sketched in the three-dimensional plot of Figure 5.6. Due to the mechanical construction illustrated in Figure 5.7, the planar cut of the monopole radiation pattern is circular. As explained in Figure 5.8, the monopole antenna utilizes the very important concept of “mirror effect.” If we place a radiating element close to a conducting surface (often called a ground plane), the resulting field is the same if we had no surface but an additional “image” element.

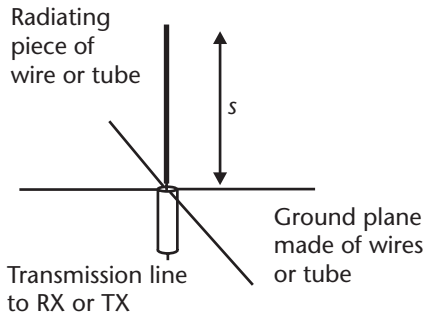
The plots indicate that we might consider selecting one of these simple antennas if there are no special desires of sharply pointing the RF energy toward a certain direction and we want to have a relatively simple and low-cost system. If properly dimensioned, both can be matched over a substantial frequency range to a suitable transmission line. Commercial devices are available from 30 MHz upward. Many



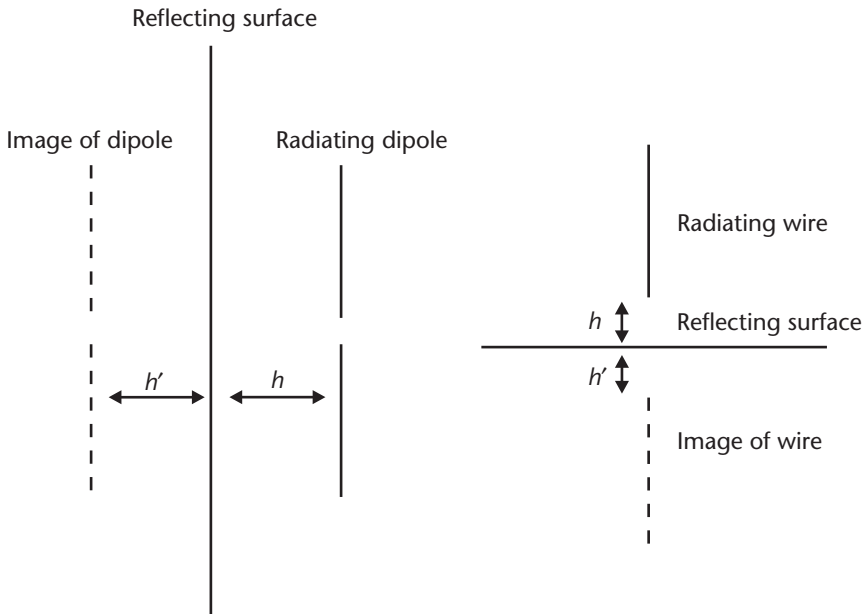
**Figure 5.5** A dipole antenna is made of two linear wires that are connected in the middle to the balanced feeder cable. The total length  $s$  of the design is often one-half of the wavelength.



**Figure 5.6** Half-wave dipoles produce doughnut-like radiation patterns. There is a steep null toward both ends of the radiating elements (up and down) and the azimuth cut is circular.



**Figure 5.7** The monopole antenna utilizes a ground plane whereby the radiating element is only about one-fourth of the wavelength. Depending on frequency and application, the ground can be made of wires or of metal sheet. Feeding is possible with coaxial cable.



**Figure 5.8** If a radiating element is placed close to a conducting surface (ground plane), the effect is similar as if we had no surface but additionally an image of the original element at an equal distance ( $h = h'$ ).

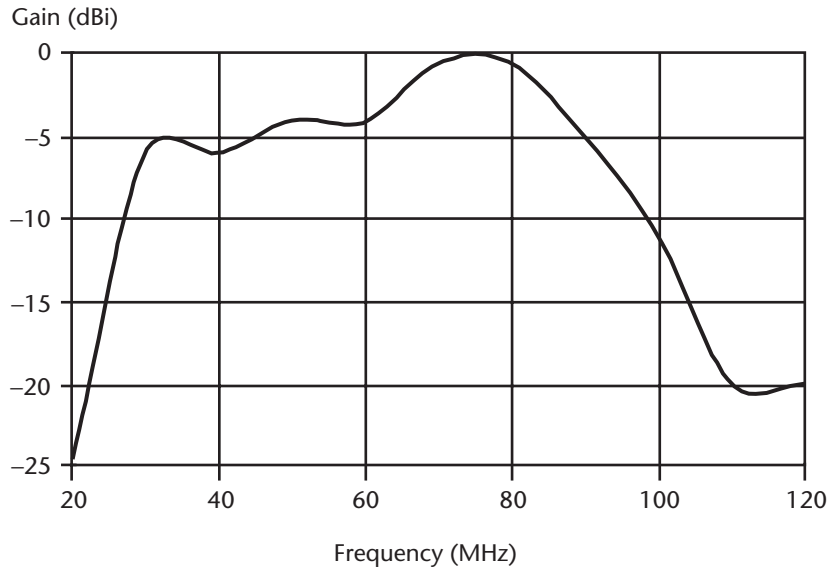


BS systems utilize either monopoles or dipoles or small dipole arrays. They are also suitable for vehicle installations, because the momentary attitude of a vehicle does not have a severe effect on the apparent signal level.

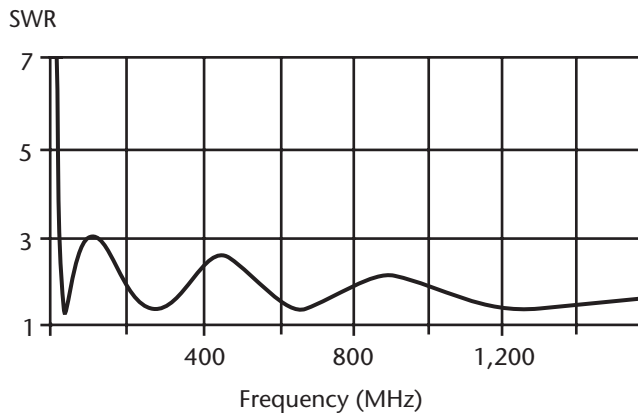
Monopoles are also used below their efficient frequency range because no other reasonable construction would do any better. For example many aviation HF ground beacons have such antenna systems. This situation implies that we have to use external lumped components for matching and thus the tuning will be of narrow bandwidth [2]. If a monopole antenna has to be matched over a very wide frequency range that is several times the apparent center frequency at hand severe problems will be encountered. Because the antenna's radiating part will be of totally different electrical lengths at the two extreme ends of its proposed frequency range, adequate reactive matching turns out to be impossible. Figure 5.9 shows one rather unfortunate commercial product intended for mobile tactical military VHF radios. Because there is no solution for a lossless matching network, the manufacturer has decided to do the matching by putting a sufficient amount of resistors into the monopole base. Of course, in this way the transceiver always "sees" about 50 ohms as calculated from the parallel connection of individual resistors, but the actual power to the radiating whip is minimal. The measured real gain is indicated in Figure 5.10. Because isotropic antennas are not available, this test was actually performed as a comparison to a tuned narrowband monopole, and the results were postprocessed to show gain in decibels referred to an isotropic antenna. Similar results have been obtained with commercially available aircraft VHF communication antennas. Because aerodynamics dictates a small physical size if we assume a nonconformal layout, good impedance matching over a frequency range from 118 to 150 MHz, for example, turns out difficult if no losses are added. Figure 5.11 illustrates the matching of a commercial passive collinear dipole antenna, advertised as a broadband device. The SWR values are reasonable from about 20 MHz upward. A schematic view of the design is given in Figure 5.12. However, the measured gain shown in Figure 5.13 suggests major problems in systems use. First, the matching and the arrangement of



**Figure 5.9** A very wideband matching of a monopole antenna is impossible with a lossless network. One commercial military VHF antenna includes a set of resistors as a questionable solution to the problem.



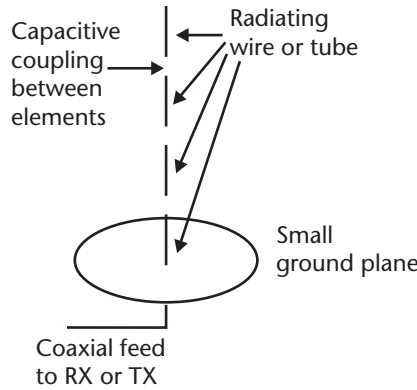
**Figure 5.10** The radiating efficiency of the resistor-matched monopole antenna is far less than that of a tuned counterpart at the same frequency. Although the antenna is matched, there is not very much power output.



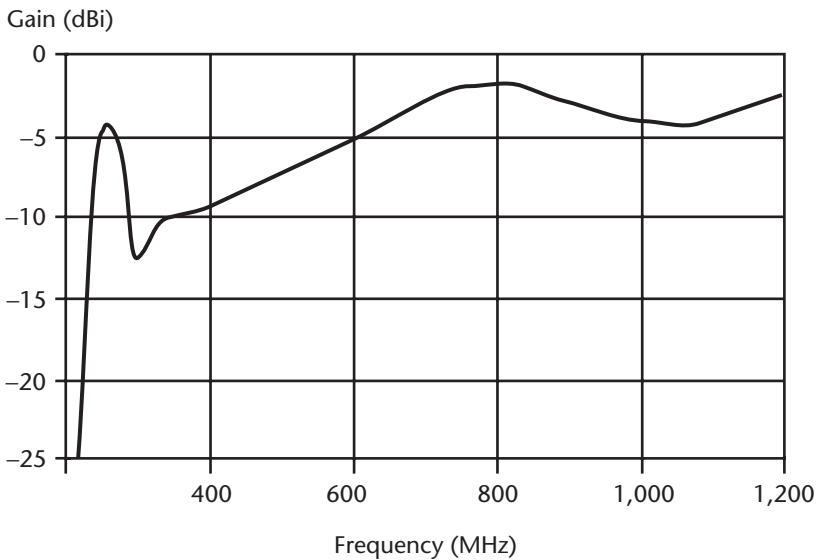
**Figure 5.11** A passive dipole setup (of collinear character) can be matched reasonably well (e.g., from 20–1,300 MHz, as illustrated here).

radiating elements inside the structure have caused the net gain to be negative. Besides this, many systems might not tolerate  $-10$  dBi. In fact, the gain stays fairly stable only from about 600 to 1,300 MHz—an octave band! However, although the antennas of Figures 5.9 to 5.12 have poor radiation performance, they can be used as system elements if the entire higher-level design process is adapted accordingly. Serious conflicts may appear if we leave this task to the end user or if such items are sold separately without proper and open specifications.

Different wire antennas, in which the dimensions are on the order of several or several tens of wavelengths still find extensive use below 30 MHz. This is natural, because they are the only practical constructions capable of providing reasonable directivity and matching combined to radiation efficiency at such low frequencies.

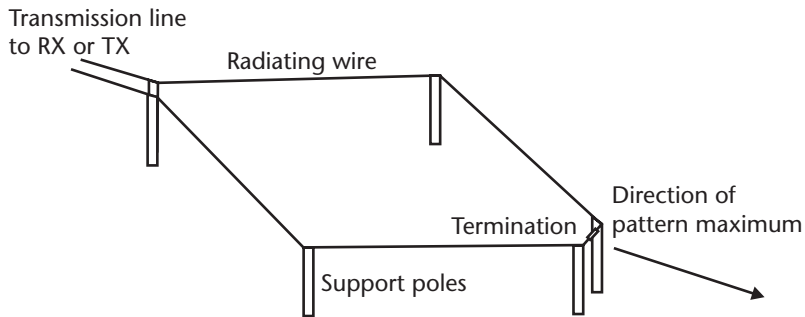


**Figure 5.12** Several dipoles can be assembled as a collinear “array,” whereby the vertical beam-width becomes narrower. Different coupling arrangements between individual elements are used in commercial products.

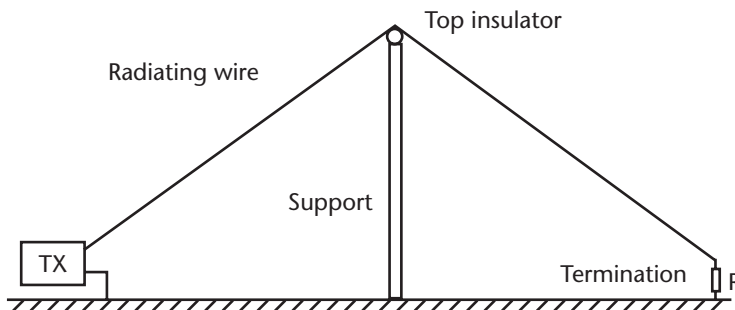


**Figure 5.13** Tricks to widen the bandwidth of inherently narrowband antennas tend to fail. Although the SWR is good, we have to accept a gain below unity (or negative in decibels referred to an isotropic antenna) and values down to  $-10$  dBi, which might be too little for our system.

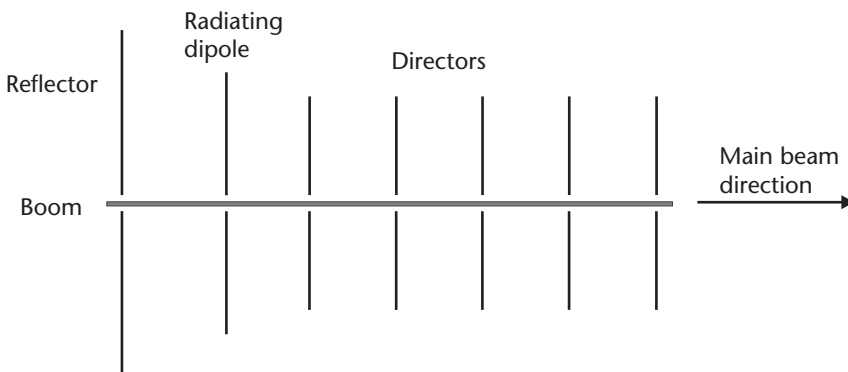
Some examples of these types are the rhombic (Figure 5.14) and the half-rhombic or inverted V antenna illustrated in Figure 5.15. It is basically possible to use them at higher frequencies, too, but their benefits start to lose meaning at around 300 MHz due to the more favorable dipole derivations, particularly the Yagi antenna, which is shown in Figure 5.16, and the log-periodic design. A typical commercially available VHF/UHF log-periodic antenna has a frequency range from 80 to 1,300 MHz, and its reported average gain is 7 dBi although in many conservative system calculations we might use 6 dBi instead. The matching is also appropriate but varies periodically between about 1.4 and 2.5. This is due to the mechanical construction illustrated schematically in Figure 5.17 and the operating principle. At a given frequency, that particular wire closest to resonance will act as the radiating element and the slightly



**Figure 5.14** For example, SW systems can make use of rhombic antennas that are made of wires or ropes supported on poles as illustrated here. The termination reduces backward radiation.

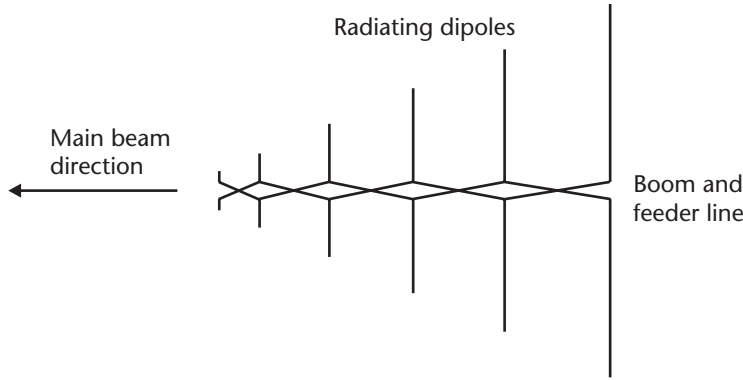


**Figure 5.15** An inverted V antenna is a low-cost directional design (e.g., for frequencies from about 20–80 MHz). The height of the supporting pole is critical to obtain the desired main beam elevation angle. Again, there is a terminating resistor suppressing the backward beam.



**Figure 5.16** If narrowband operation is acceptable, a simple Yagi antenna formed of a feeding dipole, reflector, and directors is a very lightweight and cost-effective choice.

longer one next to it is a reflector. However, as these elements of the log-periodic design cannot form a continuously varying set of lengths but have discrete values instead, both the radiation pattern and impedance matching tend to show periodic fluctuations as a function of frequency. Thus, in this way the periodic mechanical structure is reflected to the SWR and—to lesser extent—to the gain as well. This antenna type is an example of true wideband operation, but gives only linear polarization. Its 3-dB beamwidth is about  $60^\circ$  in the plane of the elements. A crossed

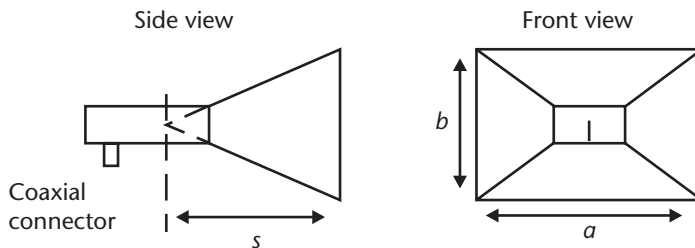


**Figure 5.17** Log-periodic antennas are truly wideband in nature. One of the elements closest to resonance (at the operating frequency) acts as the radiator, and the next longer one is a reflector.

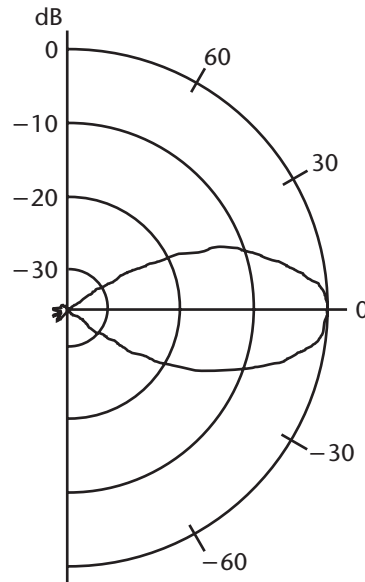
log-periodic design can be equipped with a suitable hybrid to work with circular polarizations, and we can further compile a set of suitably positioned elements to create, for example, a full 360° azimuth coverage. However, depending of course on the specific frequency range, such an installation is not more compact and might turn out to be pretty expensive. Other periodic structures like the spiral antenna provide similar benefits.

The horn antenna, an example of which is illustrated schematically in Figure 5.18, may well be considered in its various versions as the basic antenna at microwave frequencies and above [5]. Their easy connection to typical rectangular waveguides makes them particularly practical because in this way also the transition losses will be avoided. Sometimes horns appear combined with dielectric lenses for even better pattern characteristics. Already the basic rectangular horn has very low sidelobe performance as indicated in Figure 5.19. The maximum gain can be calculated from the dimensions quite accurately, which has led to the definition of “standard gain horns.” In addition to the rectangular type, circular and elliptical versions are extensively used. Particularly, the corrugated circular horn is employed in systems utilizing circular polarizations.

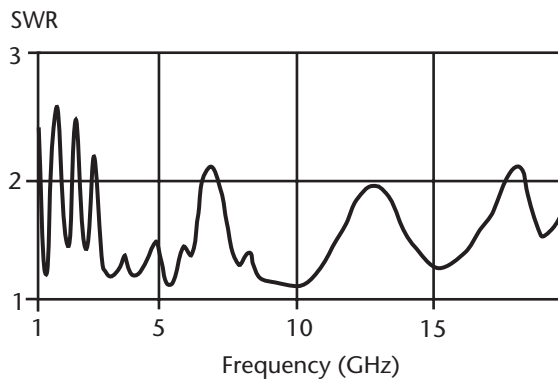
An interesting wideband construction is the so-called double-ridged horn, which can be designed to cover, for example, frequencies from 1 up to 18 GHz. Of course, some compromises must be allowed. Reasonable impedance matching is possible across the entire range as indicated by the SWR plot of Figure 5.20. However, the



**Figure 5.18** A pyramidal horn antenna has an aperture size  $a \times b$ , and the length of the pyramidal section is  $s$ .



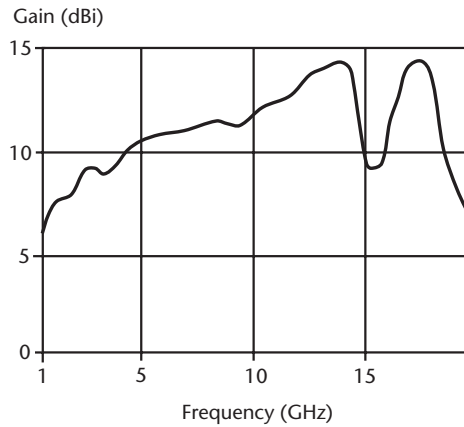
**Figure 5.19** Horn antennas are characterized by their smooth radiation patterns. This means that sidelobe levels below  $-30$  dB are not rare, depending on the selected polarization. This plot presents the H-plane (defined by the magnetic field vector) pattern of a Ku-band pyramidal horn shown earlier in Figure 5.18.



**Figure 5.20** When carefully dimensioned, a double-ridged horn antenna can be nicely matched across the entire frequency range of 1 to 8 GHz.

variation of overall size expressed in wavelengths unavoidably causes a gain change as a function of frequency. The measured performance of one typical commercial product is illustrated in Figure 5.21, showing a peak gain value of about 14 dBi whereas the lowest gain is near 7 dBi. Because the ridge spacing is very limited near the coaxial-to-waveguide transition, the maximum power-handling capability seldom exceeds 1 kW. Only linear polarizations are possible. One of the main applications of this design could be in various monitoring and surveillance systems, particularly for defense-related tasks and EMI measurements.

Slot antennas give an opportunity in those cases where we want a “flush” mount of the radiating element along a predefined surface such as the skin of a



**Figure 5.21** The physical size dictates some gain alteration as a function of operating frequency, but, in any case, the double-ridged waveguide horn is an interesting system element for wideband applications.

vehicle. They, too, can be easily fed with waveguides and are very often parts of larger arrays. Figure 5.22 shows one very basic configuration of a slot antenna array, formed of eight individual slots. The obtained performance is equal to the corresponding dipole array as predicted by theory.

Very many conformal antennas are based on the patch element configuration, which has become popular in conjunction with the tiny cellular handsets but finds extensive use in various radars and, for example, as surface-mounted parts on ammunition projectiles. Patches can give reasonable radiating efficiency and proper impedance matching over more than 20% of relative bandwidth. Their essential feature is the planar mechanical layout, which is readily implemented as a PCB design.



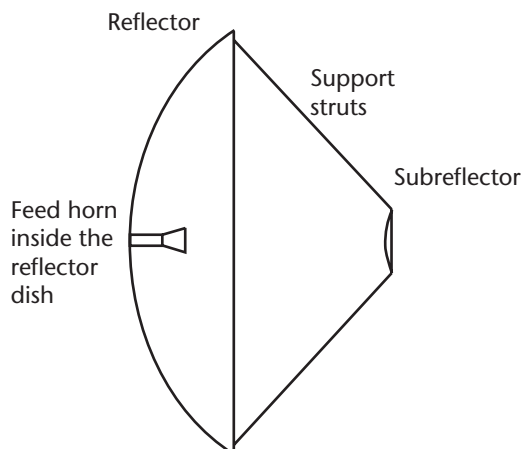
**Figure 5.22** If we have to align our radiating elements along a surface of predefined shape (e.g., an aircraft fuselage) we may find slot antennas interesting. Here, an eight-element array is shown for the military L-band links.

Further extensions of the frequency range become possible if slight deviations from the planar construction can be allowed. Dedicated materials are often needed, though, in order to achieve the desired bandwidth. Patches are very suitable elements of array antennas and benefit from relatively low mutual coupling, which typically stays below  $-20$  dB. Mutual coupling describes the interaction of two (or more) antennas—for example, elements in an array. It can be measured by feeding a known signal to one element and measuring the power available in the other. Relative phase can be of interest as well. An analytical approach relies sometimes on the concept of mutual impedance  $Z_{ij}$  between elements  $i$  and  $j$ . If the coupling is strong and there is only, say, 5- to 8-dB attenuation between the elements, we cannot treat the element excitations individually. Thus we very often desire minimal coupling.

The highest gain is often obtained through the use of parabolic reflectors or derivations thereof [5]. They normally have a horn feed, which can either be of Cassegrain type or the simpler focal mounted version. Figure 5.23 illustrates the principal layout of a Cassegrain system, and Figure 5.24 shows the arrangement for focal point feed. If very low losses are needed, we can also use a quasi-optical beam feed, but they become really practical only in the millimeter-wave bands. Gain values exceeding 60 dB are very feasible at higher microwave frequencies. The vital parameters defining the actual reflector performance are—besides its diameter—the surface accuracy and the feed illumination. Accuracy is studied both in terms of the *root mean square* (rms) error and as defined by shape deflections. Intelligent design schemes enable the mutual optimization of sidelobe level and maximum gain, but they are always conflict with each other. Larger reflectors require very sturdy pedestals, and their rotating mechanisms must be dimensioned according to the main beamwidth. Physical defects can have quite adverse effects on the radiation performance. An example of gain loss due to rough surface quality is highlighted in Figure 5.25.

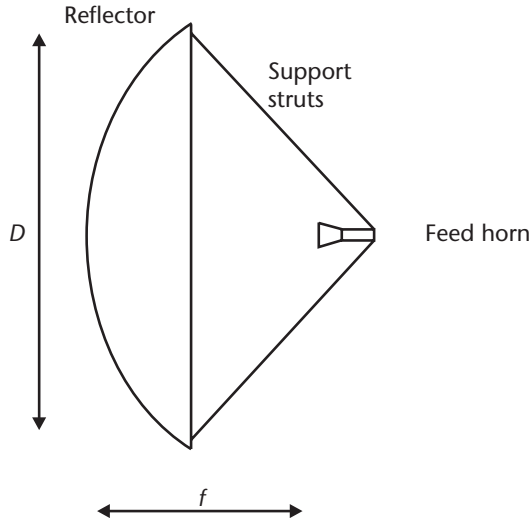
### 5.2.2 Antenna Arrays

As the name suggests, an antenna array is a combination of two or more individual antennas, arranged into some specific geometrical configuration and connected through active or passive circuits to a common RX or TX. Very often the element

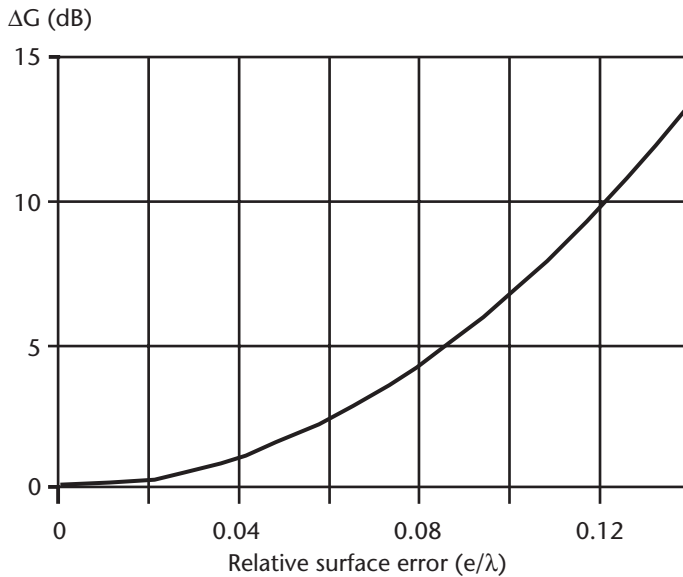


**Figure 5.23** A Cassegrain-type paraboloid antenna has actually two reflectors, because there is a subreflector in front of the main surface. The feed horn is mounted in an opening at the focal axis.



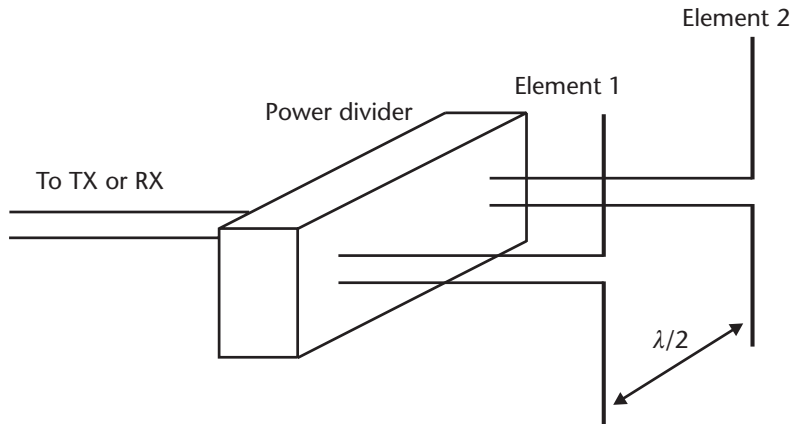


**Figure 5.24** The most straightforward way of feeding a reflector surface is to put the horn at the focal point and have it illuminate the surface.



**Figure 5.25** Both faults during assembly and deformations due to such factors as gravity can spoil the radiation pattern of a reflector antenna. The influence of rms surface roughness on the net gain is shown here.

antennas are identical but this is not mandatory. Diverse antenna types are suitable as array elements ranging from dipoles to paraboloids. Simple arrays employ equal distances between elements. An array can be one-, two-, or three-dimensional, depending on application. Fundamental antenna arrays produce fixed radiation patterns, and there is nothing adjustable in them. A very primitive but fully functional array can be formed of just two half-wave dipoles parallel to each other and mounted at a suitable distance—for example,  $\lambda/2$ , as is illustrated in Figure 5.26. They can be fed through a two-way Wilkinson-type power divider (see Chapter 4 for



**Figure 5.26** Two dipoles mounted parallel to each other at a distance of  $\lambda/2$  and fed through a power divider make a simple antenna array.

further details about power divider characteristics). The radiation pattern  $G_a(\theta)$  of such simple constructions is of the form

$$G_a(\theta) = F(\theta) \cdot G(\theta) \quad (5.2)$$

where  $F(\theta)$  is the array factor and  $G(\theta)$  the element pattern as a function of observation angle  $\theta$ . In this simple case we have assumed operation only in one plane, but general array patterns are functions of azimuth and elevation angles.

Many antenna arrays are used in the VHF/UHF frequency bands for FM or television broadcasting or in radars. Also a number of mobile phone BSs can make use of such arrays. The main reason for their popularity within these frequencies is the relative easiness of getting gain compared to reflector antennas that tend to show much larger wind and ice loading. It is also possible to define to a certain extent the final radiation pattern during array assembly by adjusting the number or position of elements, whereas a reflector antenna comes “as such” from the factory. In higher microwave bands this is the main reason to use the array concept, because reflector antennas would generally give here better gain and beamwidth values for the same physical size.

Modern cellular communication systems, including the European GSM 1800 and UMTS, defense communication networks, and numerous radar applications for military and civilian purposes can make extensive use of phased array antennas [6], which provide, for example, a rather convenient way of steering the radiation pattern toward wanted directions or creating deep minima against unwanted, interfering regions [7]. A more flexible—and today realistic—application will focus the beam to the area of mobile phone subscribers and at the same time create several minima in order to suppress some nearby harmful transmissions or RF energy spillover to adjacent cells. Note that digital processing in the IF RX is not adequate for the latter purpose, we must have processing in the TX part as well. In an RX, the basic operating principle relies on vector summation of voltages from available antenna elements. This task can be either a straightforward analog function with power combiners, phase shifters, and adjustable amplifiers or a DSP session.

Because A/D conversion requires many samples from each period of the incoming waveform, digital processing is done, for example, at the IF interface due to the easier sampling action. The tricks used for pattern shaping can thus be thought as adjustments of amplitudes and phases although the three-dimensional multipath nature, particularly in an urban environment, tends to cause severe complications. Similar challenges are met in a multiple target-multiple jammer military scenario. If time is added as a parameter, signal processing people normally speak about *space-time adaptive processing* (STAP).

Regardless of the application, most antenna arrays should today be capable of an almost real-time radiation pattern adjustment, which means that all electronic and numerical tasks must be performed within a few milliseconds or even faster. This may be based in the case of radars on the location of moving targets and ECM units or in the case of cellular systems on the respective location of individual mobile phones. This is accomplished through mathematical algorithms, generally implemented as considerable amounts of, for example, assembly- or C-code, typical examples of which are described in [8–11]. When time constants far shorter than 10 ms are anticipated the phrases “adaptive antennas” or “smart antennas” are often used. This implies that the host system will not observe pattern adjustments as interruptions or delays in its performance.

The following section uses one real but elementary software-controlled phased antenna array designed by the author’s team, described in detail in [12, 13], to demonstrate some of the techniques and problems involved. Here we assume that the whole array is intended to be used as a standalone device risen to a tower, which effectively limits the adverse effects of the mounting platform. For a less fancy situation the reader is encouraged to wait until Section 5.2.3. The system was chosen to utilize analog RF processing of signals. This enables retrofitting existing units with new antenna configurations and makes the design independent of manufacturers of cellular or military equipment, thus possibly giving it a larger market potential. Initially the aim of the project was just to give the hardware people a tool with which they could evaluate the performance of various antenna modules. During the project—partly based on customer initiative—the task was extended first to study the feasibility of a software-based correction of some or perhaps most of the defects or imperfections found in the analog RF parts. Later the research was to give some insight to an automatic pattern-shaping action based on simple network operator decisions. For example, if cellular customers are in the city area in the daytime but move to residential districts for the evening, the coverage should be tunable too. The prototype system was configured for BS reception, and thus its active modules were not suitable for transmitted pattern shaping. However, the physical array configuration itself is reciprocal. If appropriate PAs are used instead of the LNAs, a transmitting antenna could be assembled. The phase shifter electronics presented later can handle both situations, but of course not at the same time.

Although chaotic designs have been recently introduced in the laboratory level, it is still quite customary to assemble a phased array antenna from a number of identical subarrays, which in turn can be just a row of almost equally spaced simple elements like dipoles or planar microstrip radiators. They both form a so-called *uniform linear array* (ULA). Depending on the nature of the specific application we may assume that the subarrays have a fixed excitation, or they too can be phase-steered,

which has been put into practice (e.g., in the Aegis radar) [7]. In the frame of the past project for which all the software was developed, we need to consider only the one-dimensional case, the pattern of which can be presented as shown in [6] in the electrically familiar form of

$$G_a(\theta) = \sum_{i=0}^{N-1} G_i(\theta) U_i e^{j(\phi_i + i k d_i \sin(\theta))} \quad (5.3)$$

where  $U_i$  is the element's voltage,  $d_i$  is the mechanical spacing,  $\phi_i$  is the electrical phase shift of an element, and  $\theta$  is the angle of observation, measured from the normal of the element plane. The individual element's own radiation characteristics are given by its pattern  $G_i(\theta)$ . The factor  $k$  is the common propagation constant and is defined as

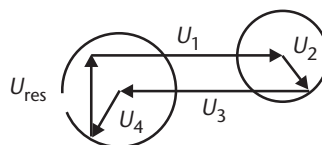
$$k = 2\pi/\lambda \quad (5.4)$$

Although we here assume the antenna to transmit signals from the surrounding space, the general theorem of reciprocity states that the results obtained are valid for reception as well. The total number of elements in the summation ( $N$ ) is typically tens or hundreds but may in the case of a large radar system be tens of thousands. A considerable relief in the computational task, generated by (5.3), is available if the element patterns and amplitudes were identical and can be moved in front of the summation. This yields to a simplified pattern as

$$G_a(\theta) = G(\theta) U \sum_{i=0}^{N-1} e^{j(\phi_i + i k d \sin(\theta))} \quad (5.5)$$

where also the mutual distances between individual elements are assumed to be constant. The similarity to (5.2) is recognizable. Despite their electrical appearance, most of the parameters in (5.3) and (5.5) are of mechanical origin and thus based on dimensions, shapes, and materials.

We will next discuss a couple of problematic case examples, which were found with the initial prototype antennas and later became targets for elimination by software actions. In a real phased array system both the amplitude and phase of each element contribute to the final result, which is schematically illustrated in Figure 5.27. The lengths of the vectors are defined by gain or attenuation and mismatch whereas the relative angles are those coming from various wanted and unwanted phase shifts, typically affected by mechanical and electronic features of the array



**Figure 5.27** Unsuitable relative amplitudes of the individual element voltages  $U_1$  to  $U_4$ , caused, for example, by severe impedance mismatch may prevent their phasor sum  $U_{res}$  from going to zero although  $U_2$  or  $U_4$  is given a phase change (rotated) if  $U_2$  and  $U_4$  are kept fixed.

assembly. The main causes of difficulties are careless positioning of subarrays and amplitude differences either from amplifiers, feeder lines, or possibly from impedance variations.

An antenna array operating below 3 GHz is typically a metal construction with either welded or bolted joints between sheet metal or machined parts. Its overall dimensions range from about half a square meter to above one hundred. Because of weight reduction, sturdy machined supporting plates are generally not attractive; nor is the exaggerated use of steel. The mathematical formula in (5.5) assumes a constant spacing between various radiating elements and subarrays and—although not explicitly shown—requires that a full symmetry exists along the main axes of the assembly. Thermal deformations during welding [14] and the often careless handling of sheet metal parts during various manufacturing phases tend to cause bends, kinks, and misalignments to the array, which disturb the symmetry and notably alter the interelement or intersubarray spacing. Let us consider the illustrative case shown schematically in Figure 5.28.

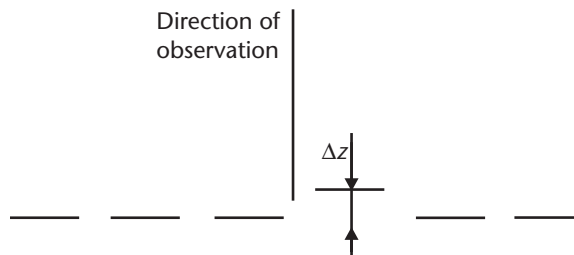
In the very simple case of one single fault the phase error can be roughly estimated as

$$\Delta\phi = 2\pi \frac{\Delta z}{\lambda} \quad (5.6)$$

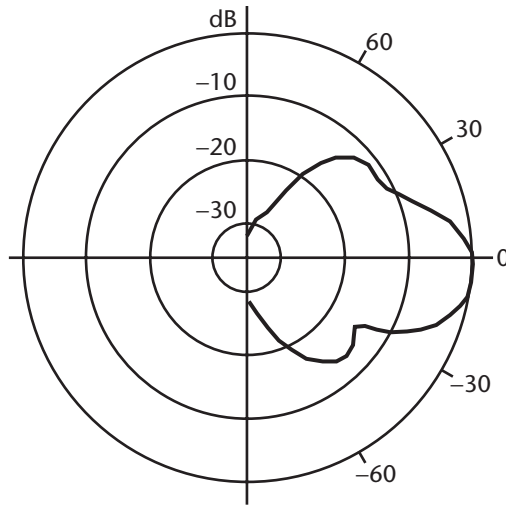
in which  $\Delta z$  is the radial positioning error due to factors such as assembly tolerances as measured from the point of observation, and  $\lambda$  is the wavelength in free space. We normally expect that measurements are performed at a point in the so-called far-field region making  $r$  sufficiently large. The observed result is a deterioration of the radiation pattern, which, too, will no longer be symmetrical as can be seen in the measurement result of Figure 5.29 obtained from one of the early prototypes. Not only the position of individual elements is crucial but also the alignment of every tiny radiating stub, because, if incorrect, both will cause a distortion to the phase pattern of the complete array.

If a sharp pattern minimum is desired, the phasing error  $\Delta\phi$  from (5.6) will in the primitive case of two identical subarrays in a planar configuration cause it to be filled up to a level of

$$L = 20 \log_{10} \left( \sin \left( \frac{\Delta\phi}{2} \right) \right) \quad (5.7)$$



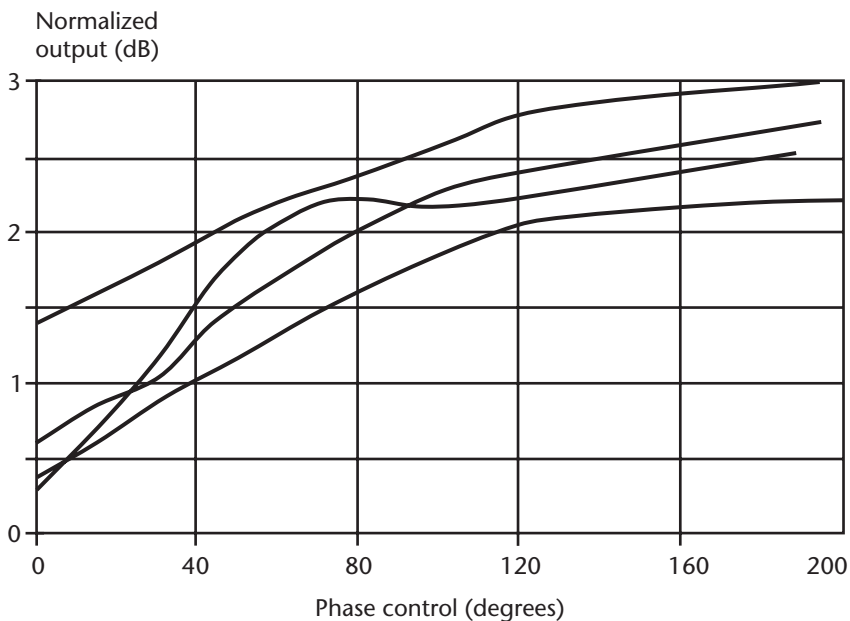
**Figure 5.28** A radial displacement of  $\Delta z$ , if small compared with the distance to the point of observation will cause a phase error between individual elements. Here the specific type of elements (i.e., patch or dipole) is of minor importance.



**Figure 5.29** Errors in the mechanical layout create nonsymmetrical radiation patterns (e.g., different sidelobe levels).

in decibels, which is readily obtained from simple geometry. In practice, a dephasing of  $11^\circ$  reduces the minima to be not deeper than  $-20$  dB, which is a limit that can seldom be allowed as the design value is often below  $-40$  dB.

Figure 5.30 shows test results from several identical preamplifier modules to be controlled through the developed software. A mechanically inaccurate connector mounting—see, for example, [15, 16]—has caused a gain spread of more than 1 dB. This was due to excessive return loss degradation, as the center pin of the connector



**Figure 5.30** Mechanical uncertainties in the mounting of coaxial connectors have caused a gain difference of more than 1 dB between front-end units of our phased array demonstrator. The curved nature depends on the characteristics of the phase shifters.

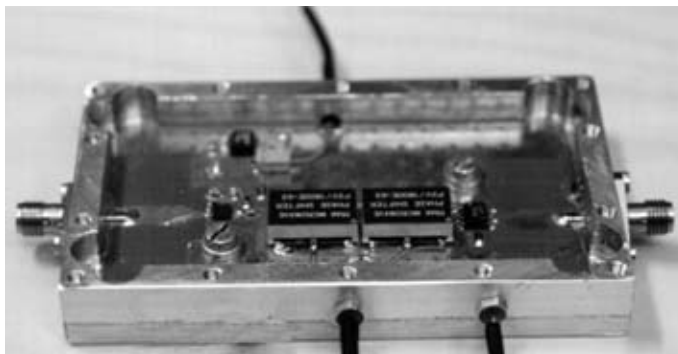
hardly met its stripline counterpart on the PCB. A phase unbalance was observed as well. The curved nature of the plots in Figure 5.30 does not come from the physical setup but from the operating principle of the electronic phase shifter. When the milling accuracy was enhanced to 0.02 mm, the difference between individual gains disappeared. This forced us to use quite expensive milled front-end mechanics as observed in Figure 5.31.

The amplitude errors are especially annoying when the beam-forming algorithm tries to create a sharp minimum toward an interfering TX. For a primitive case with just two subarrays brought to opposite phase with each other the obtainable depth of the minimum in decibels can be estimated as

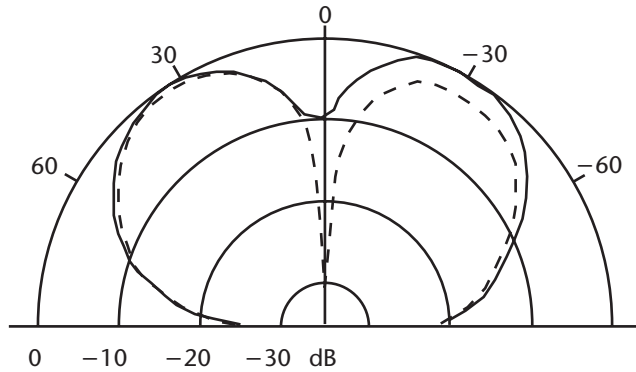
$$L = 20 \log_{10} \frac{1 - 10^{-\frac{A}{20}}}{1 + 10^{-\frac{A}{20}}} \quad (5.8)$$

where  $A$  is the difference (error) in the feeding amplitudes of the two arrays, respectively in decibels. It is seen that already a bias of 1 dB prevents us from getting more than 25 dB of interference rejection. The effect is further clarified in the measured pattern of Figure 5.32 where a wanted  $-30$  dB minimum has been filled just to reach 10 dB. Unfortunately this error type cannot be corrected by a simple phase adjustment in the software. We must find the smallest resultant value, which indeed is the difference in the amplitudes of the respective components. Therefore, we must implement an amplitude adjustment as well.

The definitions for array and software development pointed out three partly separate situations, with which the system should be able to dynamically cope. A most natural wish is to have the pattern maximum to a selected direction, or perhaps more generally sufficient coverage within a certain rather narrow angular sector, say a suburban highway strip or a battalion operating area. Following this, the second case might be the creation of a minimum. Finally, if the physical characteristics of the array itself permit, it could be desirable to have a flexible choice of coverage in a full  $360^\circ$  area. Such a feature might be attractive when mobile units move according to the tactical situation. However, we typically have two distinct directions into which a tactical UHF link is supposed to communicate. All other azimuth angles are



**Figure 5.31** When the connector mounting was milled with an uncertainty less than 0.02 mm, the electrical data spread was easily controlled. However, this is not the cheapest enclosure for a front end.



**Figure 5.32** The black line does not reach the wanted minimum (dashed) because there is a 4-dB amplitude difference due to hardware defects.

potential DOA angles of hostile interference or may have surveillance RXs hidden below the horizon. Thus, several minima are desired into these regions.

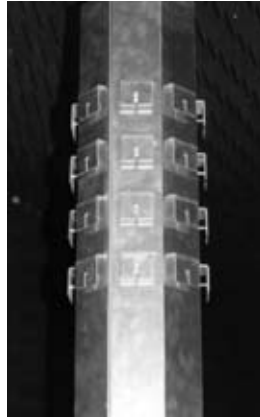
Three physically different antenna arrays, all for reception at a BS site, have been used for the evaluation of the developed hardware concept and software modules. Two are planar constructions with either dipoles or modified patches (see Figure 5.33) as radiating elements, and the third one is a full-coverage conformal design (shown in Figure 5.34) of an octagonal cross-section. Actually two separate types of the planar patch antenna have been available with either four or eight elements per subarray to meet absolute gain requirements. From the software's point of view, all planar patch arrays are equal except for the slightly increased mechanical tolerances of the largest construction. The conformal device has different element patterns. Due to the reduced size, the electrical coupling between adjacent subarrays in the azimuth plane is stronger, thereby complicating the pattern-shaping task.

The simplified block diagram of the RF electronics, common to all tested versions, is shown in Figure 5.35. There are two control variables in each subarray: the electrical phase angle produced by varactor phase shifters and the amplitude,

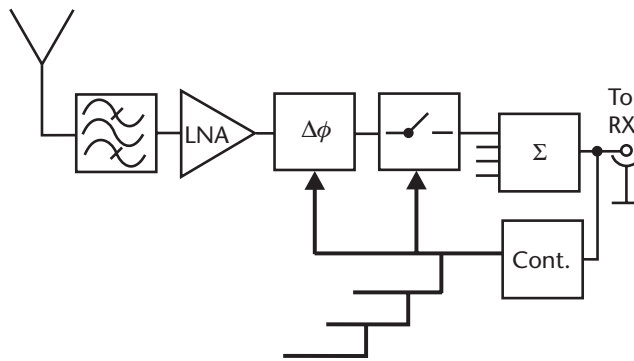


**Figure 5.33** A part of the dipole array used for tests. The unfortunate feature of this construction is the high mutual coupling between adjacent rows, not better than  $-20$  dB.





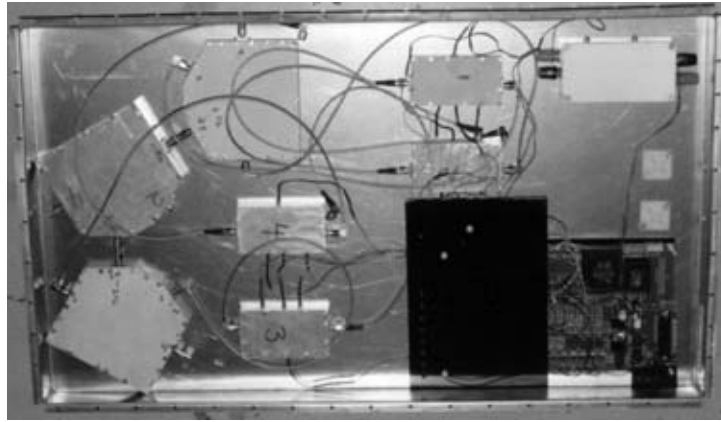
**Figure 5.34** For a complete 360° coverage the hardware people suggested an octagon-shaped antenna construction for which a special piece of software was also developed.



**Figure 5.35** A simplified block diagram of the analog electronics showing both the RF phase shifters and switches, which were later used as variable attenuators, too. Thick lines indicate the routing of the parallel digital control bus that serves also adjacent array elements (not shown).

initially thought of as an ON/OFF-information, formed by a commercial gallium arsenide RF switch. The phase adjustment happens by changing an analog control voltage, and the switches require a digital but bipolar input. After this RF processing all signals are summed in, for example, a Wilkinson power divider or in our case in a special ring hybrid. It soon turned out that the selected electronic phase shifters had a 3-dB change in attenuation as a function of their control input. Therefore the concept was edited accordingly to include programmable gain correction through the analog switches.

The computer interface of the test antennas was based on a commercial I<sup>2</sup>C bus implementation and compatible *digital-to-analog* (D/A) converters and optically isolated digital switches. One such card can simultaneously handle eight subarrays, and four cards can be tied together if necessary. Data transfer from the controlling computer is arranged through its printer port and has bidirectional optoisolators, which provide the necessary interference rejection in tower environments. Two of the prototype antennas also have an internal single-chip processor, which can be used for controlling *built-in test equipment* (BITE). An overall view into one of the earlier prototypes is illustrated in Figure 5.36.



**Figure 5.36** An interior view of one of the prototype antennas. The computer interface card is in the lower right corner, and the RF electronics are housed in the milled boxes in the middle.

The generation of software code proceeded in a stepwise fashion and partly parallel to the ingenious achievements made by the hardware designers of the team. This resulted in an iterative cycle where successive improvements of the code were carried out “on-the-fly.” Initially, a simple manual but GUI only for the adjustment of various phase shifter voltages and switch states was created both for the first planar arrays and for the conformal construction. It turned out to be very impractical to aim at ultimate manufacturing and assembly precision of the passive RF components. When the nonidealities described above were found almost at the same time, it was questioned whether a completely software-based correction could be created. In such a way, the physical or electrical inaccuracies might be correctable through the same piece of software, which performs the initial pattern-shaping action. The first fully automatic pattern-shaping algorithm was soon added featuring a combined wanted maximum/wanted minimum search—at that time having the unavoidable restrictions coming from the planar physics package. The algorithm was to find a combination of element amplitudes and phases that yields the best possible ratio of signals coming from the desired TX and the interfering one. Because the array was not truly three-dimensional, only a limited sector, about  $60^\circ$  in azimuth, could be handled.

As an example of similar studies, the work in [17] was examined. This Swedish approach is based on the commercially available (former) Hewlett Packard *visual engineering environment* (VEE) platform and—as the original authors also found—suffers from apparent speed restrictions. Besides this, [17] was of minimal help because some fundamental questionable simplifications have been made there (e.g., ignoring the effects of varying test distance on the measured phase patterns of individual subarrays). This involves the case where the passive array is being rotated around its axis of symmetry. Apparently, the individual subarrays follow a circular path, and the maximum error across the test range will be  $D$ —the array diameter.

All the software for this project was created in C++. The elementary versions, including the one developed for the octagon-shaped conformal array, only provided the hardware designers a means to use their electronics and measure antenna performance. Thus the main components of these programs are the following:

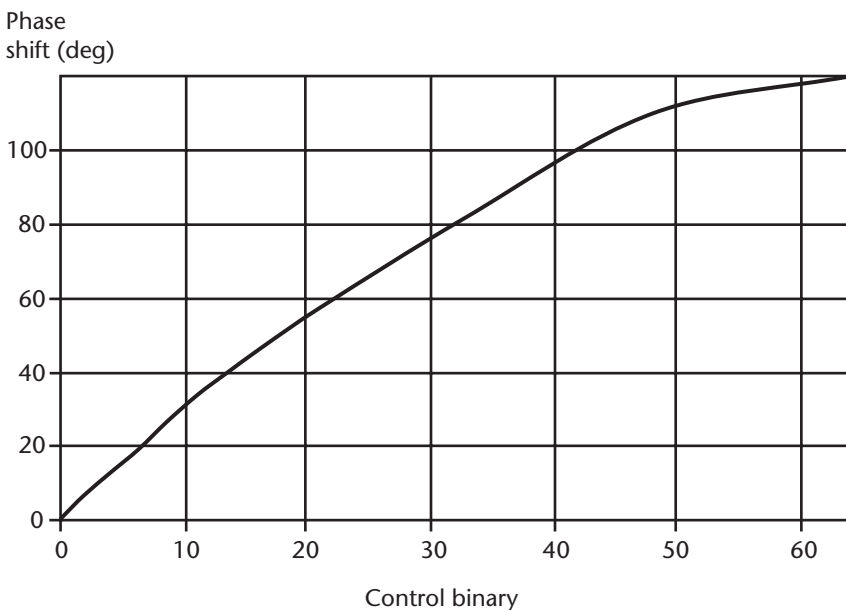
- A friendly GUI;
- Required controlling of the intelligent industrial communication standard (I<sup>2</sup>C) bus interface card;
- Conversion between the actual phase shifter angle and control binary (see Figure 5.37).

A special feature of the octagon variant is the possibility to virtually adjust the diameter of the cylinder and so simulate the amount of lobing, as shown in Figure 5.38. This would be impossible with real hardware. Naturally, when the octagon radius gets larger, the distances between individual array elements and the array phase center expressed in wavelengths are longer as well and therefore possibilities for deep minima increase.

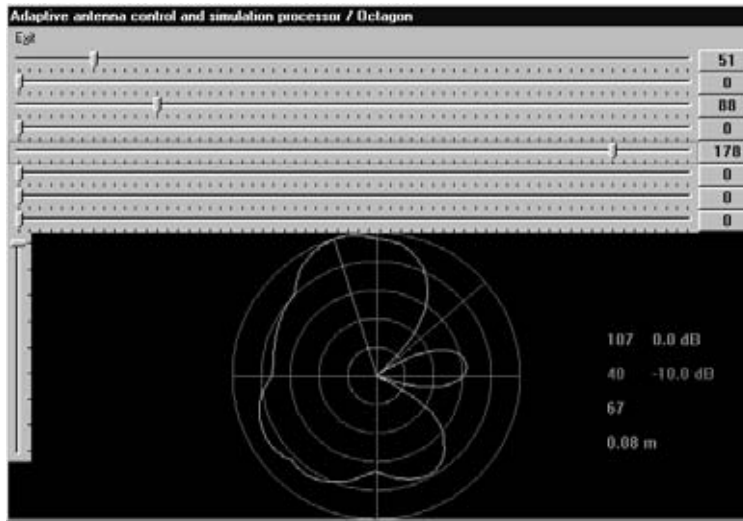
Using the very basic vector summing principle from (5.3) as a background, a program capable of reducing the effects of hardware imperfections was introduced. The additional parameters, which are taken into account, are listed as follows:

- Individual subarray amplitude patterns;
- Individual subarray phase patterns;
- Front-end gain variations due to phase control.

Because the mutual coupling in a patch arrangement stays below  $-20$  dB, there was no real need to try and eliminate that effect. In a dipole array, this would be, however, vital. All corrections are implemented as look-up tables, which were created for the prototype antennas manually from the measured data. Only the conversion between control voltage and phase shifter angle is maintained as an internal piece of code, but all other correctional data has been arranged as individual files



**Figure 5.37** The highly nonlinear characteristics of phase shifters were taken into account already in the very first prototype software version. However, the programming task was simplified due to the similarity of the four subarrays.



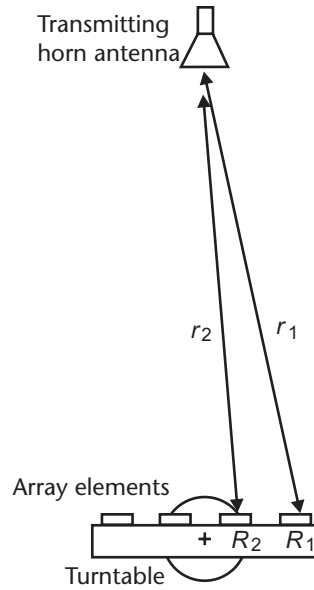
**Figure 5.38** The octagon-array program has a feature (vertical slider to the left) for changing the radius  $R$  of the octagon. This helps in estimating the amount of adverse lobing in thick antenna towers.

that can be easily updated (e.g., for following antenna samples). Because the pattern gain information was available only at  $5^\circ$  intervals the software used linear interpolation to create the missing data for each  $1^\circ$  step. The numerical resolution is 8 bits, which gave an uncertainty of 0.03 dB near the beam maximum but increased up to 3 dB when the overall element gain had dropped to its minimum.

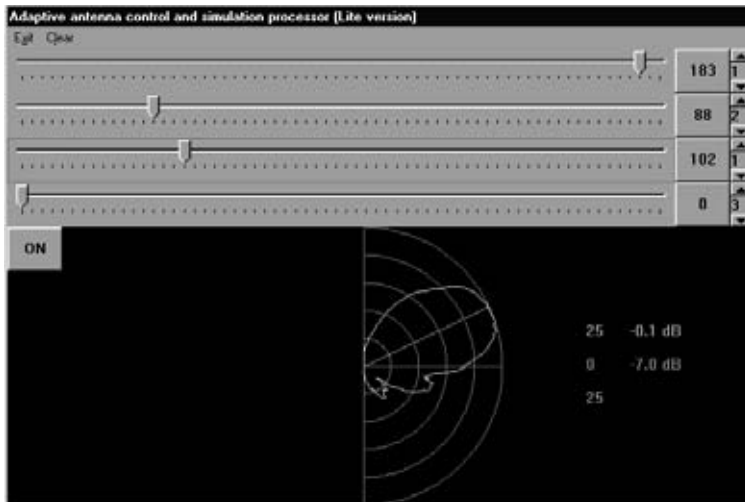
A specific problem was the correct processing of phase recordings, because the test installation in the anechoic chamber prevented a truly centered positioning of each subarray. This was the problem, which was neglected by the parallel research of [17]. In our case it was solved by creating a short program in C. It directly took optically scanned phase results, converted them to a numerical form, and subtracted the respective correction due to the apparent free-space propagation distance; see Figure 5.39 for a graphical presentation of the problem. Considerable effort was needed to cover the ambiguities encountered with the rotation of phase angle, particularly because the functions of C had their own convention. With properly scanned phase recordings the measured error due to the data conversion here was below  $3^\circ$  of electrical phase angle.

Although this program eliminated the amplitude errors caused by all known reasons, there was a manual “button” for fine-level adjustment of each subarray as well. The phase shifter sliders are similar to those used for the eight-element array but instead of the dimensional adjustment there was a button for switching on or off the element-based phase correction. One version of the GUI is shown in Figure 5.40.

As suggested by the industrial partners of the main project, the latest (at that time) software set was designed to automatically optimize the planar array’s pattern to produce a minimum to one certain direction and a good maximum to another selected angle. Actually, however, this software was aimed at finding the best possible *carrier-to-interference* (C/I) ratio, because it turned out that for most of the cases the differences between an absolute maximum and a local one within the  $180^\circ$  sector (theoretically) covered by the planar construction are just 3 to 6 dB.

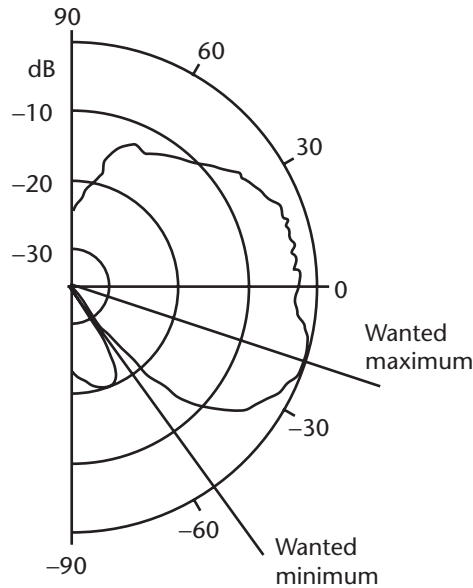


**Figure 5.39** The physical distance  $r_i$  of each subarray behaves differently during the rotation of the measurement turntable because  $R_1$  and  $R_2$  are not equal.



**Figure 5.40** The GUI of the prototype planar antenna array. This display was developed for measurement and evaluation purposes, not for the end user.

Because computing power is relatively cheap compared to the rest of one phased array assembly, the initial choice has been to use “brute force” and check each different combination available through the phase shifter and switch combinations. To be feasible, though, this really means *central processing unit* (CPU) frequencies around 1,000 MHz, otherwise one complete pattern estimation action will take minutes. Compared to the simple beam-shaping software described above we here do not care about anything else except the two specified directions, which naturally yields sometimes quite astonishing patterns with, for example, multiple lobes however satisfying the preset requirement. One example is shown in Figure 5.41.



**Figure 5.41** The automatic search algorithm produces the desired combination of one minimum and one absolute beam maximum, but otherwise the pattern may be “unexpected.”

Phased array tests require considerable investment in the laboratory infrastructure. Most of the numerous experiments with the developed hardware configurations and associated software were performed with a full anechoic RF test range which, in addition to signal generators and RXs or spectrum analyzers, has absorbing internal surfaces all around for the cancellation of unwanted reflections and a rotating mechanism onto which the antenna to be measured was put. All phase-related recordings were taken with a VNA. Some of the test gear is shown in Figure 5.42 together with the designer wondering at some of the curious results.

Pure software evaluations, which did not need the complete antenna hardware, followed the typical benchmark practice. Four platforms were available for



**Figure 5.42** The author’s son is seen here tediously measuring the phase and amplitude patterns of a full array. Two microwave generators, a spectrum analyzer (the pile on the left), and a VNA (in the foreground and thus not visible) were used.

estimations of computational speed ranging from a low-end 133-MHz laptop to a 450-MHz Pentium III with 128 Mb of RAM.

The application requiring much processor power is obviously the one developed for automatic pattern optimization. Table 5.1 shows how the time for one complete pattern search got smaller through various editing actions and demonstrates the measured difference in processing time needed with some typical platforms. We observed that the effect of some nonelementary mathematical operations is not as large as expected whereas some innocent-looking pieces of unnecessary code may cause several tens of percent of additional wait time. For example a 10-base log, which was used to convert the *C/I* ratio to decibels, had a total effect of only 300 ms for the whole pattern cycle.

Another performance figure for the search process is its dependency on the complexity of the task. We evaluated this by giving it successively an angular difference from  $5^\circ$  to  $80^\circ$  as measured between the desired maximum and minimum. Figure 5.43 illustrates that only a change of about 10% was observed which is to be compared to something like the effect of battery condition on the same platform where an improvement of 25% to 30% was available just by charging. The various software-based correction functions, when switched off on purpose, gave a relief of only 1%.

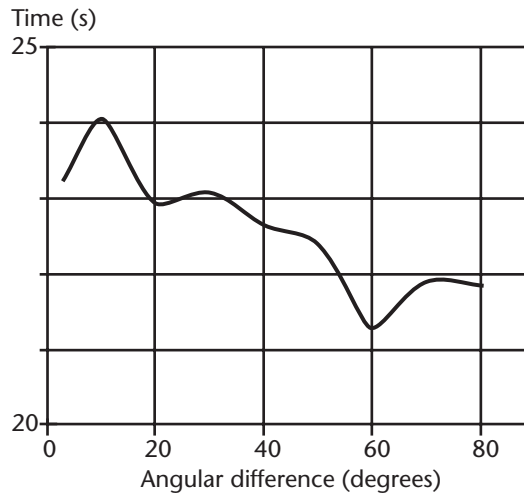
The obtained radiation patterns were used as a tool for evaluating the RF performance of the developed software. Depending on the application either the width, direction, and shape of the so-called main beam are of the greatest interest, or occasionally the direction and depth of minima are the most interesting factors. Table 5.2 summarizes some key findings. Illustrative examples of the influence of various corrections are further clarified in Figures 5.44 and 5.45. It seems that the simulated patterns have an angular integrity better than  $5^\circ$  with the measured ones whereas the amplitude discrepancies—excluding the areas of sharp minima—generally stay below 5 dB. The accurate prediction of a sharp minimum would require still more precise measurement data.

### 5.2.3 Vehicle-Mounted Arrays

Unlike arrays mounted on high towers, their counterparts on top of vehicles must be designed to live with their reflecting nearby environment. Two approaches have

**Table 5.1** The Progress in Computing Time of the MINMAX Routine on Different Platforms

<i>Software State</i>	<i>Time (133)</i>	<i>Time (333)</i>	<i>Time (450)</i>
Straightforward computing	—	140 seconds	—
Removal of redundant cycles	—	25 seconds	—
Cleaning of code, no log or sqrt	36 seconds	10 seconds	—
Precomputed sines, and more	25 seconds	7.5 seconds	5.3 seconds
Removal of ON/OFF switches	—	4 seconds	—



**Figure 5.43** The computational time of the search algorithm is highly stable. A reduction of only 10% was observable when the task was made easier by widening the angular distance from wanted maximum to wanted deep minimum.

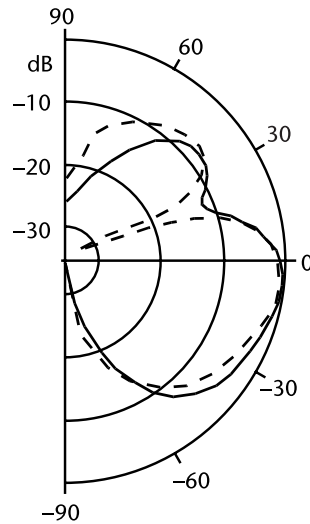
been tried in the past. The first attempts were motivated by a possibility to create a “universal” concept, which would cancel out all or most of the adverse effects due to the vehicle body. More recently, constructions accepting and taking advantage of the ship’s deck or a fighter’s airframe have been favored. Artificial impedance surfaces are expected to change the situation, because they enable close mounting of radiating elements on otherwise short-circuiting conducting bodies. Our “normal” design rules state that the real part of permittivity or permeability in a media must be positive. We also expect that metal surfaces exhibit very small resistivity. Artificial impedance means here that these rules are not strictly valid. For example, by using special geometry resembling porcupine’s skin we may create exceptional conditions for some direction on wave propagation. It is also possible to use tiny active elements that, for example, cancel reflections from a surface within a limited frequency range. Such an approach might be feasible for example in anechoic coatings if the wavelength is too large for pyramidal absorbers.

Regardless of principal path selection, one of the major differences of vehicle installations is caused by the type and performance of individual antennas, which tend to follow more closely factors like aerodynamic requirements than RF performance criteria. As a demonstrative example of this type of a task, we consider next a simple airborne phase controlled antenna system for AJ purposes. The construction is very straightforward and is not to be understood as a milestone of AJ

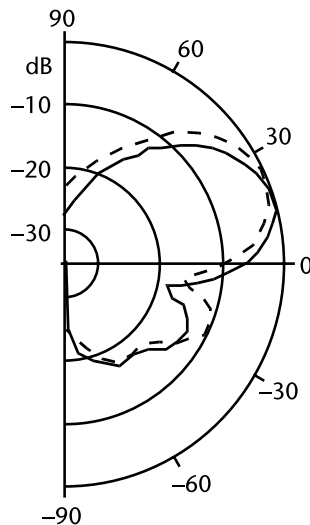
**Table 5.2** A Comparison of Some Vital RF Results from Measured and Predicted Patterns

<i>Parameter of Interest</i>	<i>Simulated</i>	<i>Measured</i>
Depth of wanted minimum	16–23 dB	20–25 dB
Half-power beamwidth	30°	28°
Beam-steering range	23°	27°
Sidelobe level	–8 dB	–10 dB





**Figure 5.44** If the phase patterns of individual subarrays are ignored, there is a poor correlation between measured (black) and simulated (dashed) patterns although the average dephasing is only about  $20^\circ$ .



**Figure 5.45** These two patterns show how, without subarray amplitude correction, conventional beam-steering for one maximum works as predicted by simulation (dashed).

electronics but more as a way to introduce a number of detailed problems and ways to circumvent them.

Many of the present systems have been designed by using numerous radiating elements, separate RF switches, and complex driver software. For land battle electronic warfare (EW), various approaches have been available through sources like [18], but the military aviation community has been a bit more conservative.

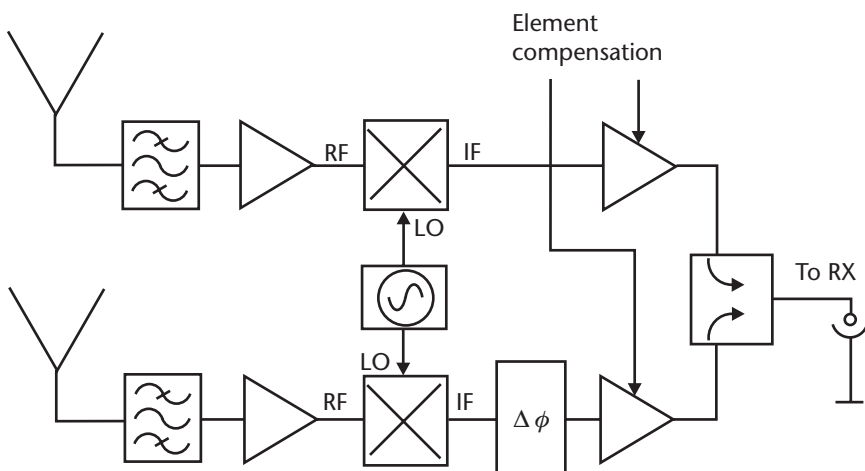
The purpose of an AJ antenna system is to produce a desired minimum to the direction of the jamming TX and, at the same time, to optimize the C/N or S/N ratio

of the communications chain. Normally the situation is more critical up in the air, where possibilities also exist for long-distance jammers. Much concern is caused by the complex airframe structure, which is dictated by aerodynamics, RCS, and weapons payload. Very little is left for the antenna system designer and besides, the typical, nonconformal, radiating elements used for air-to-ground or air-to-air communications are only tiny vertical whips.

In this case a primitive AJ system was designed for the F-18 A/B fighter aircraft. The VHF communication antenna element is basically a monopole operating around 120 MHz, and two of these are mounted on top of the airframe just behind the cockpit canopy. The typical radiation pattern of two coupled monopoles is the familiar “number eight,” where a deep minimum in one direction is achieved, when the distance between the two elements is  $\lambda/2$ . Of course, because the communication system must be able to utilize the entire allocated spectrum, this requirement cannot be continuously met. Further limitations are set by the airframe construction and items inside it. The adjacent radiating element unavoidably deforms the initial single element pattern, which is no longer circular.

Besides the two radiating elements, the designed system had a two-channel coherent downconverter with adjustable phase difference between the channels, a control electronics unit, and some software. The operating principle relied on summing of the two antenna signals but in such a way that one of them is phase-shifted to produce a deep minimum toward the jamming TX. The actual phase shift happened at the selected IF. A phase-locked transistor oscillator was used as LO to feed the two mixers via a resistive power splitter. An external attenuator was installed to reduce load-pulling effects. Two gain-controlled amplifiers were further needed to cancel differences in element pattern amplitudes. The block diagram of the total system is shown in Figure 5.46.

The phase shifter had four identical, adjustable amplifiers in parallel with associated fixed-delay lines, which had nominal phase shifts of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$ , and  $270^\circ$ . Theoretically, three constant delays would have been enough for one carrier frequency, but more were needed to cover any reasonable bandwidth. The phase



**Figure 5.46** The basic operating principle of the simple jammer canceler relies on two aircraft antennas and vector summation of their voltages.

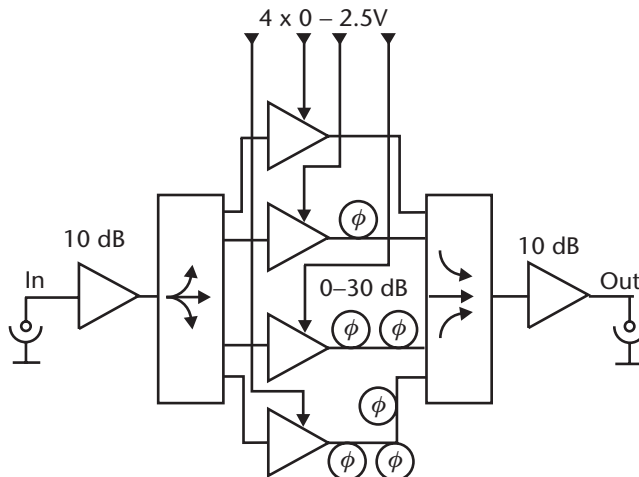
shifter principle can be seen in Figure 5.47, and the antenna mounting on the F-18 A/B scale model can be seen ready for tests in Figure 5.48.

At first, any differences in the original phase or gain behavior of the amplifiers were recorded for future use as a function of the gain-control voltage. Then, the computer generated four vectors for the virtual directions of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$ , and  $270^\circ$ —having no direct connection with actual compass bearings. The lengths of the vectors were presented by the gain of each channel. Furthermore, discrete vectors such as the lengths and the fixed angles of the four channels were calculated for each point of the  $360^\circ$  azimuth circle. The computer controlled the D/A-converters through all possible combinations (262,144, 8-bit D/A) and selected the best one for each angle. This actually took about 35 hours but gave the best accuracy. The long delay was caused by the network analyzer sweep time. Some test results of the phase shifter can be seen in Figure 5.49.

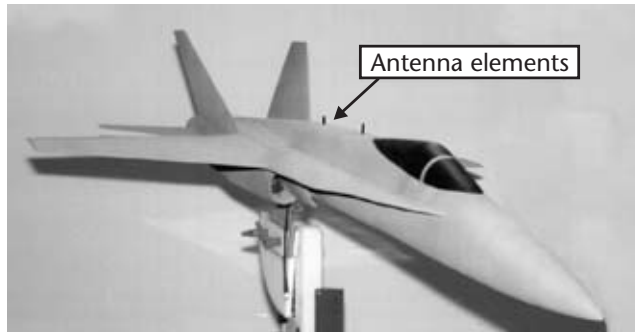
All software, except that for radiation pattern simulation, was coded in BASIC. Separate automatic routines were used for measuring phase shifter and leveling amplifier performance. Theoretically, there were two possible pattern minima at separate directions for each electrical phase angle. Due to this, two level corrections must be available as well. After that, the final phase-shifting data was computed for each phase angle in  $1^\circ$  steps and using 8-bit resolution for D/A conversion. During this process, the individual antenna elements were already mounted on the aircraft fuselage.

Most of the tests were performed with a high-end spectrum analyzer, a network analyzer, and an RF-generator of equal performance. A calibration of zero electrical length was accomplished, as required, by inserting a double balanced mixer mounted instead of the final two-way power combiner. All test equipment was controlled through the IEEE-488 bus. Part of the test was performed with the F-18A/B scale model described in detail in [19] and by using an anechoic test range described in [20]. The model had a wingspan of 1.2m and the fuselage length was about 1.75m.

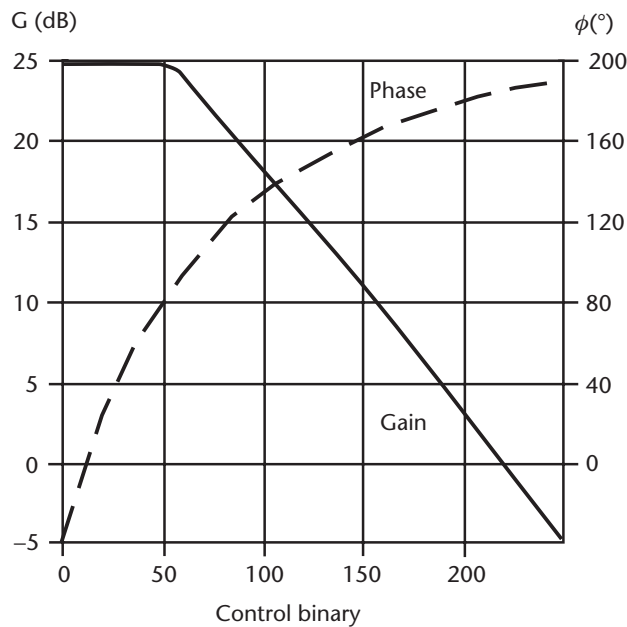
Tests proved, that it is possible to produce minima deeper than 20 dB in all desired compass bearings despite the complex airframe, as shown [21]. Up in the air



**Figure 5.47** The constructed simple A) system uses gain-controlled amplifiers and predefined fixed-phase shifts to produce the desired radiation pattern minima.



**Figure 5.48** The two VHF antenna elements on top of the F-18 A/B model.



**Figure 5.49** Gain and phase of the final phase shifter are controllable within about a tenth of a decibel and about  $3^{\circ}$  of the target values. We can even accept nonmonotonic behavior due to computer memory control.

the situation might be even better as there are fewer reflecting surfaces to fill minima. The improvement of the radiation pattern minima achieved through the leveling scheme was typically 10 dB compared to the original, noncompensated situation. This result is well in line with the observations described earlier in the case of patch antenna arrays. The stability of the phase shifter could be maintained within  $3^{\circ}$ . Sharp excessive minima caused by the airframe structural items (e.g., fin, elevator, and canopy) could, however, not be totally avoided. The effects of the F-18A/B tail structure with its tilted double stabilizer are especially hard to overcome. This means, that although the system was capable of producing a wanted minimum almost anywhere in the  $360^{\circ}$  sector, a maximum cannot be always guaranteed. With only two antenna elements it is difficult to totally eliminate jammers of nonzero vertical incident angles. A benefit was the capability of the designed system

to automatically adapt itself to the operational platform—taking into account the unavoidable effects of the antenna ground planes. The growth of computational workload and the attack time of the AJ function decrease, if the actual “live” algorithm is kept simple enough.

## 5.3 Antennas as Mechanical Elements

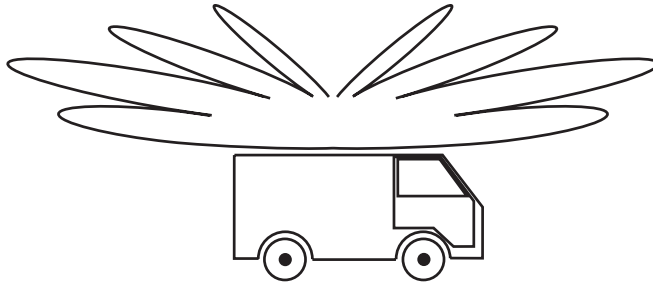
The question of antennas as physical pieces can be understood in two different ways. First, we can speak about the effects of the physical mounting place of an antenna—for example, onboard a ship—on the radiation characteristics or impedance [5]. Alternatively, the theme can be used to describe the different arrangements of mechanically turning the antenna so that its radiation pattern maximum usually points toward a wanted direction. In both cases the RF system designer should continuously monitor the status of things to be able to intervene if a potential loss of performance is foreseen. Requirements conflicting with the radio engineer’s desires might push the precious antennas somewhere behind 50-mm-thick armored plates or make the mounting of our reflector array unable to move freely toward the target direction. We use two examples to highlight some of the difficulties and possible solutions related to both interpretations.

### 5.3.1 Antenna Mounting on Test Vehicles

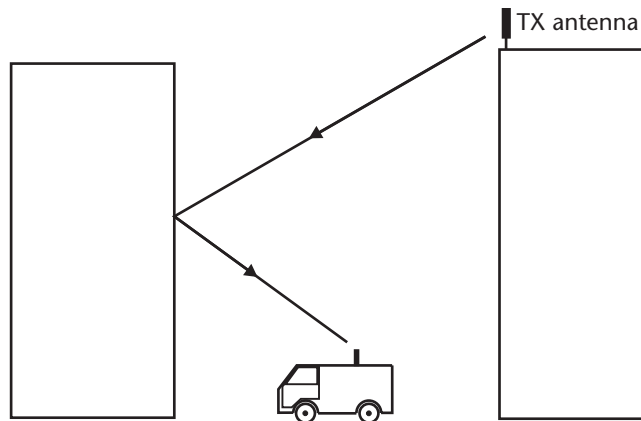
Most commercial cellular operators and telecommunications authorities carry out extensive coverage measurements both for network planning and later tuning. Several alternative mobile test systems are in use, but most of them apply some kind of calibrated test antenna to be able to measure the absolute field strength. Serious attention has been paid to the accuracy of the instrumentation (e.g., the test RX [22]), where an absolute level error of less than 2 dB is called for, but very little has been said about the antenna installation itself.

To reveal different multipath and fading problems within the coverage area, very high sampling rates are to be used giving a spatial resolution around some tens of millimeters or even less. Many methods have also been suggested for the testing of GSM-like TDMA/WCDMA-practices, where different time-of-arrival due to multipath reflections from the environment can be disastrous. The new UMTS operates in the 2-GHz band and uses TDMA/WCDMA and sophisticated power-controlling algorithms. At UMTS frequencies the radiating element is less than 50 mm high. If the antenna is mounted slightly above the real car roof surface for example to prevent blocking by signal lamps, a vertical radiation pattern like that sketched in Figure 5.50 appears.

In heavy urban areas, tall buildings surround the main streets and from the test van, look angles above 30° toward the incident wavefront are not rare. A possible situation from the test vehicles’ point-of-view is outlined in Figure 5.51. Although several reflections do exist, the dominating one (largest or first or the one having best BER) can come from the least expected direction, and the result is questionable. The system accuracy is further impaired by the typical multiple test antennas not to mention the necessary FM antenna for the amusement of the driver and additional



**Figure 5.50** The vertical pattern of a measuring antenna on top of a van roof (height  $> \lambda/4$ ).

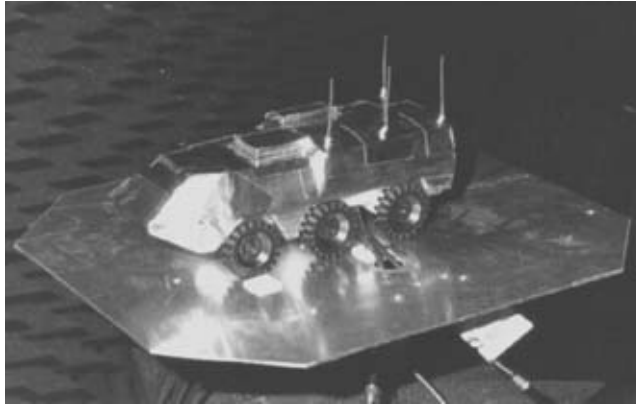


**Figure 5.51** Multiple reflections from building walls have a look angle of  $5^\circ$  to  $40^\circ$  as seen from the test vehicle.

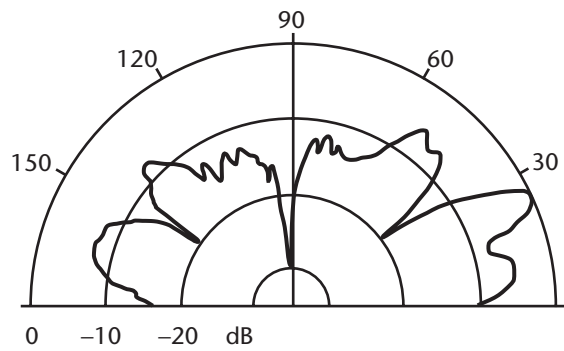
obstacles on the vehicle roof, where a spare tire, the air conditioner, and other necessary items are located.

A 1:7.5 scale model van was constructed for the measurement of the typical radiation patterns following the principles stated in [23]. Polyurethane foam, wood, aluminum foil, and conductive copper tape were the main raw materials. Figure 5.52 shows another similar scale construction, but in this case of a military high mobility vehicle. If working with a 7.5-scale car, UMTS-frequencies are scaled to around 15 GHz. Two lateral mounting positions were tried. Also the effect of a whip-type FM car radio antenna was considered. The BS antenna was highly directional to avoid external multipath problems. Elevation and azimuth patterns were measured with different mounting heights above the van roof.

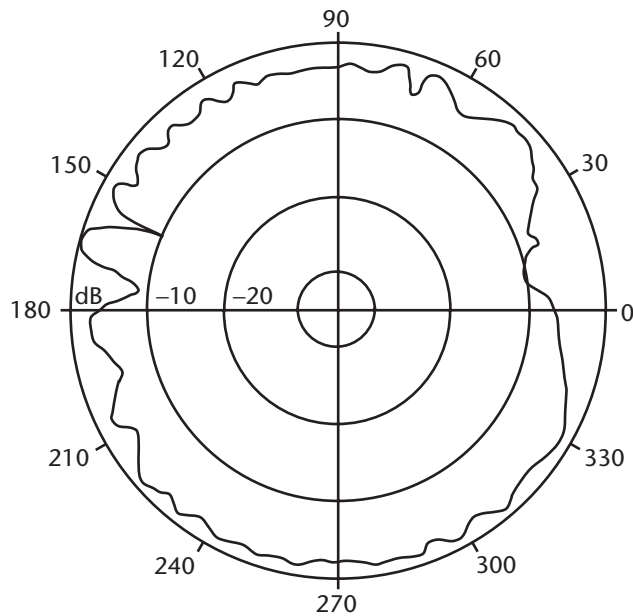
As predicted, the vertical pattern gets badly split as soon as the test antenna is raised from the roof's metal surface. This can be clearly seen in Figure 5.53, where the maximum points to about  $24^\circ$  above the local horizon. Also the azimuth results depend on the situation above the roof, as is natural. Even quite deep minima can appear (see Figure 5.54), which suggests a loss of 10 dB depending on the angle of arrival. If, for some reason, the behavior is accepted as is, there seems to be no real need for a roof-center mounting of the test antenna, because the pattern will be far from ideal. For reasons of technical curiosity, another scale model (1:20) was



**Figure 5.52** A scale model vehicle for radiation pattern measurements.



**Figure 5.53** A measured vertical pattern at 2 GHz, with an antenna height of 0.15m (in real life).



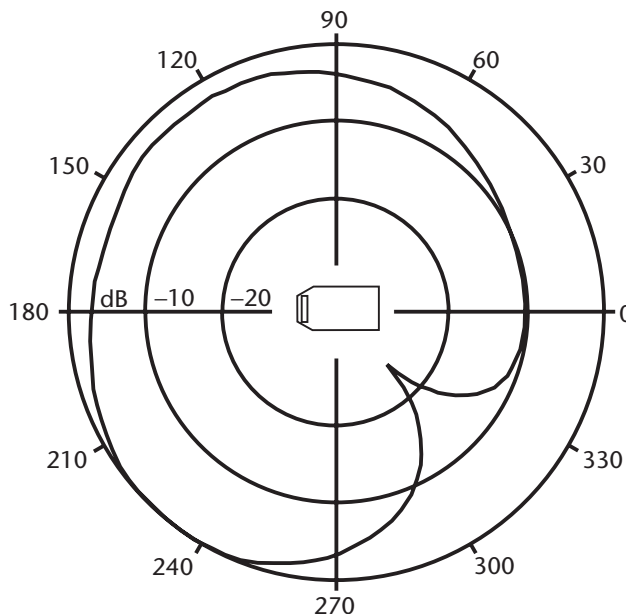
**Figure 5.54** The azimuth pattern is affected by minimal obstacles. An air-conditioner box creates the minimum at 0°.

constructed to see if lower frequencies suffer from the same phenomena, as well. In the common VHF band the results are much like those seen at 2 GHz. A typical pattern can be seen in Figure 5.55. This result might be of interest for those designing for example tactical mobile VHF installations.

In conclusion we observe that vehicle-mounted antennas and related mounting procedures are the most problematic element of the whole coverage measurement system due to the complexity of gain judgment and the formation of antenna pattern nulls in the local multipath environment on top of the vehicle roof. The problem is multiplied at higher frequencies like those of UMTS and can be as deep as or deeper than  $-10$  dB. This exceeds the normal field strength measuring inaccuracies by one order of magnitude. Similar difficulties are met in mobile tactical VHF systems. Modern electromagnetic simulation software packages provide an efficient means to predict the performance and assist in finding suitable mounting positions. Scale-model measurements may be used for further verification or in those cases where the computation time is unacceptable. This can happen if the wavelengths of interest are very short compared to the dimensions of the mounting platform.

### 5.3.2 A Tracking System for a 3-m Reflector Antenna

Polar orbiting satellites are used to monitor the weather and the environmental state of our Earth and to gather classified information for national defense. The spaceborne platform altitudes are typically near 850 to 1,000 km, and one full orbit lasts about 100 minutes. Many of these satellites transmit at L-band, often near 1.7 GHz. Their EIRP is not particularly high, which implies the use of about 3-m paraboloid antennas for the reception together with RX front ends having NFs below 1.5 dB. At 1.7 GHz the antenna 3-dB beamwidth would be roughly  $4^\circ$ . Therefore, there must be a reasonably accurate tracking arrangement in the ground receiving station.



**Figure 5.55** Complex pattern shapes turn up if multiple VHF antennas are mounted on the same roof, which was the case in Figure 5.54. Here the real-life frequency was 60 MHz.

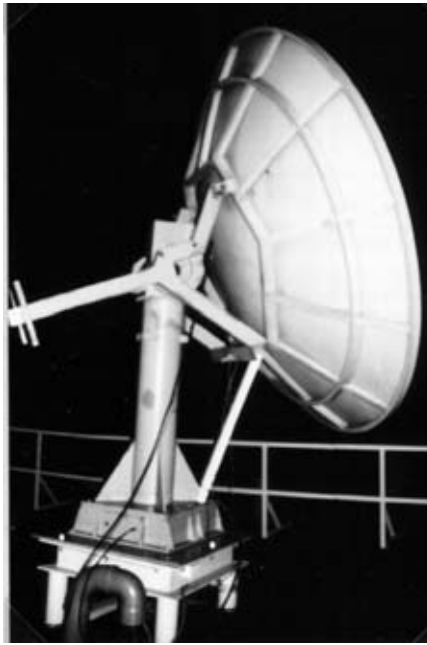


If satellite beacons operating at higher frequencies are to be received, considerably smaller beamwidths must be taken into account (e.g.,  $0.35^\circ$  for 20 GHz). Actually we here face the dilemma of pattern width and tracking accuracy. If we want more input power to our ground station RX, we tend to increase antenna diameter for larger gain. However, when doing this, we set much higher pointing accuracy requirements to the electromechanical rotating system or, in some cases, to the phasing electronics. At some critical stage, the size and associated wind and inertia loads will start to “eat” the true gain more than is brought in by increasing the diameter. In practice this means that although, for example, a 15-m paraboloid antenna surely has very high gain, the main beam might never reach the satellite to be tracked.

The design data for the constructed Earth station were derived as follows. Because the orbital altitude is so low, the apparent angular speed can be more than  $20^\circ/\text{second}$  although we might exclude tracks passing directly over the local zenith. Because orbits start and end near the local horizon, the other angular speed extreme also has to be handled. Naturally, the azimuth and elevation tracking ranges must be able to cover the entire visible hemisphere. The operating concept was made a bit easier by requiring two full azimuth turns of the antenna both clockwise and counter clockwise whereby there was no special need to check, due to hung cabling, which way the antenna was brought to the initial position before new tracking starts. The tracking system resolution was defined to be one-tenth of the worst-case beamwidth giving  $0.035^\circ$ , and the mechanical uncertainties were initially assumed to be near  $0.01^\circ$ . Maximum operational wind speed was selected to be 25 m/s; above this speed the antenna was supposed to be brought into stow position. The main issue to be “monitored” by the RF system designer was the tracking performance.

The final tracking station is illustrated in Figure 5.56. It consists of a steel pedestal inside which both azimuth and elevation motors and reduction gears were hidden. The moving parts are brought to static balance through counterweights (visible in the front), but this happens unfortunately at the cost of increased inertia. Backlash of the elevation mechanism can be nulled by allowing a slight unbalance (“nose down”). The maximum turning torque needed by a 3-m dish is roughly 40 Nm, if we want to keep the thing in steady motion also at 25 m/s wind. When this result is combined with the necessary turning speed, we end up at 400W of power on the primary axes. If desired, two motors working against each other could solve the azimuth backlash.

Several drive possibilities exist and were indeed tried during the system evaluation phase. Stepper motors were the first choice. It was thought that they would be easily connected to the controlling computer. However, the required tracking accuracy could not be met due to the excessive vibrations caused by the stepping actions. This was mainly caused by the displacement of the feed horn, which made the main beam direction wobble around the reflector’s focal axis. Actually, this defect had a direct connection to the antenna design philosophy as well. Because there was a desire to have as little blocking due to the feed and RF front end in front of the reflector, the whole horn and LNA assembly was deliberately designed as a long but shallow cylinder. Its outer diameter was just that required by the circularly polarized horn, whereby the total length became quite impressive—to about 600 mm—to be able to fit in all the necessary downconverter hardware. When this assembly was placed to the focal point of the reflector, it created considerable bending torque



**Figure 5.56** A rather typical 3-m tracking station for polar orbiting satellites. Maximum speed is  $20^\circ/\text{second}$  and the tracking uncertainty is below  $0.03^\circ$ .

already just due to gravitation. A further mistake was to use very thin aluminum tubes as feed supporting struts—again when fearing excessive blocking. However, these three tubes acted like torsional springs. Uneven motion caused the whole set to vibrate.

Additionally, it was soon found that the main gear started to wear very rapidly. In the end, after a test period of about half a year, the combination of these two drawbacks made the antenna feed tremble violently and break its welded mounting into pieces. After considerable trial and error the system was modified to work on variable-frequency ac converters and associated ac motors. Modern frequency converters have rather straightforward serial-bus control and their speed range is about 1:100. Still, the lower speed was not slow enough but a separate mechanical reduction gear had to be added. After drive system modification the desired tracking accuracy has been achieved and thus the faults in the antenna design path have been circumvented through better mechanics.

Measuring the azimuth and elevation angles was also a tricky task. Radar people are familiar with synchros—actually precision three-phase generators—which have been after the advent of the digital age supplemented by synchro-to-digital converters. The primary synchro transducers are mechanical precision instruments but evolved over the years into quite robust designs. However, at the time of the project, all-digital encoders had appeared and they were chosen as the method to use. Unfortunately, due to cost limitations from the customer's side, the team was forced to start with so-called incremental transducers. Although they provide a very high resolution in terms of pulses per full revolution, they have two quite serious drawbacks. First, they “do not know where they are” unless a special zero pulse is generated every time the power is switched on. Second, they have a separate

indication for the direction of rotation, and this must be decoded from the output pulse train. The result was a very complex arrangement of wired logic gates, mechanical microswitches on the azimuth axis, and stand-by batteries to maintain the directional “database” of one bit. Moreover, despite all these precautions the vital piece of information was abruptly lost during some mains failures. Only after a couple of reflector runaways and broken IF cables was it possible to demonstrate the benefits of optical absolute encoders. The final solution was a 14-bit design, which gave the wanted resolution and gave as a byproduct some relief in the computational burden as well.

Because the antenna system has an active feed with a low-noise downconverter, it must be equipped with dc supply as well. Although a really primitive task in itself, also here the team initially failed. About 30m of IF cable was needed from the actual RX to the front end, and this was correctly computed to be very possible with just RG-214-type hardware with out excessive loss or degradation of noise performance. The idea was to use the same cable to carry the dc power to the front end, but nobody checked the current needed, or the resistance of the relatively thin center conductor. Finally we were struggling with 15-V supply variations of several volts and measuring for weeks, based on wrong assumptions, power line interference, and other obvious defects. When the real reason was found out, the whole idea of remote dc had to be discarded, and a separate switched mode supply was installed in the pedestal.

True interference problems were also encountered after about half a year of continuous operation. Received satellite pictures started to show occasional white lines, which extended across the entire plotting area. These lines were more often encountered at low southern elevation angles. Simple measurements indicated that the ground station RX was getting interference, and a more detailed spectrum analysis showed very typical pulsed radar spectra. Because no 1.7-GHz radars were known to operate in that region, the situation seemed rather confusing. Almost by accident I came to think about a possibility of harmonics, which later on turned out to be the case here. A couple of foreign radar stations using the UHF band and located south of the satellite receiving site created sufficient second harmonic power to mask the weak satellite signal, but only when the antenna was more or less pointing directly toward the radar.

## 5.4 RF Transmission Lines

Basically there are four main RF transmission line types to choose from. They are listed as follows:

- Metallic waveguides;
- Microstrip lines;
- Striplines;
- Coaxial lines.

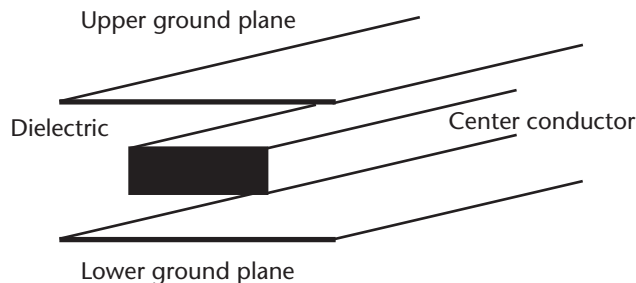
The main use of stripline and microstrip layouts is on circuit boards and inside monolithic ICs. This is quite obvious, as the manufacturing process can often rely on

similar board etching, for example. Figure 5.57 illustrates the cross-section of a stripline design. A thin center conductor is placed between two parallel ground planes and is supported by a suitable dielectric [24]. Because the construction is completely closed, we expect minimal radiation losses. The more common microstrip layout has a cross-section similar to that shown in Figure 5.58. It is similar to the stripline, but the second ground plane has been removed [25]. We can easily add lumped components or semiconductors to our circuit, and the manufacturing cost will be lower. Also interfacing toward coaxial connectors is fluent. Unfortunately we must accept some radiation losses.

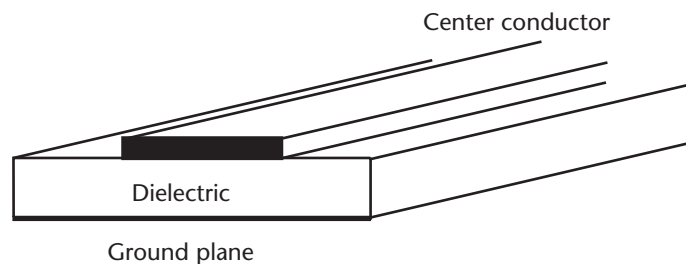
However, a system designer might less often use microstrip or stripline. Other circuit-level solutions such as coplanar [26] or dielectric waveguides [27] are suitable for millimeter-wave work, although there are certain physically large-scale applications, where stripline, for example, is the only possibility. Anyhow, the two main line types in system use are coaxial cables and metallic waveguides. Cables provide very wide relative bandwidths and at least some flexibility for the mechanical layout but at the cost of high attenuation. Waveguides, which mainly have rectangular or circular cross-sections, have only about octave bandwidths and tend to be physically rigid (exceptions will be mentioned later) but have typically very low attenuation. Sometimes these parameters allow one to make a decision, but often more specific studies are needed.

#### 5.4.1 Coaxial Cables

Coaxial RF cables have been in use since World War II. They are characterized by the *transverse electric and magnetic field* (TEM) propagating mode whereby their



**Figure 5.57** A simplified presentation of a stripline-type transmission line. The center conductor is positioned between two parallel ground planes and is supported by suitable dielectric material.



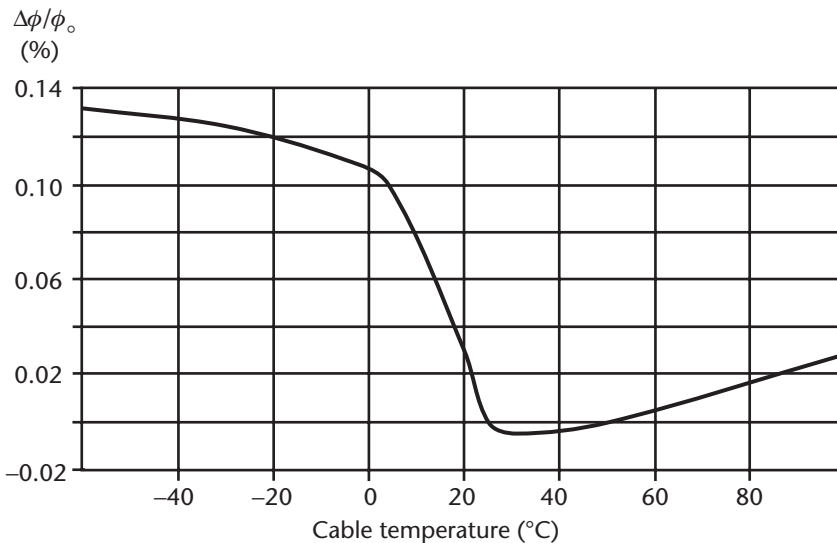
**Figure 5.58** Microstrip transmission lines have just one ground plane and the center conductor with suitable dielectric material in between.

relative frequency range is very wide. Some form of dielectric as supports between the center and outer conductors is mandatory, which implies high losses particularly in the microwave and millimeter-wave bands. Almost all RF cables today have 50-ohm characteristic impedance, but some special types are available at 75, 90, and 100 ohms, for example [28]. There is an unavoidable tolerance for the impedance value, often guaranteed to be within, for example, 2 ohms of the nominal, but this is normally not the main factor due to more severe connector mismatch. Because of the geometry, the capacitance of a cable is rather fixed, being around 100 pF/m, even though so-called low-capacitance types are available. This parameter may be of key importance (e.g., in certain HF systems, where the transmission line length as such would not cause any noticeable attenuation but rather acts like a capacitor loading the respective feeding amplifier output). A real-life example of this problem will be pointed out later in Chapter 7.

Other important parameters of coaxial cables include the following:

- Velocity of signal propagation (typically around 60% to 70% of  $c_0$ );
- Attenuation;
- Power-handling capability;
- Maximum operating voltage;
- Shielding;
- Static and dynamic bending radii;
- Type of dielectric in terms of environmental durability.

As indicated by field theory, the speed of wave propagation in a coaxial cable is defined by the dielectric constant of the insulator [29]. The real value is important in phased arrays and similar applications. There may be a strong temperature effect. This is illustrated in Figure 5.59. Many cable manufacturers use a relatively low-frequency value as a typical attenuation figure for their products. The two main

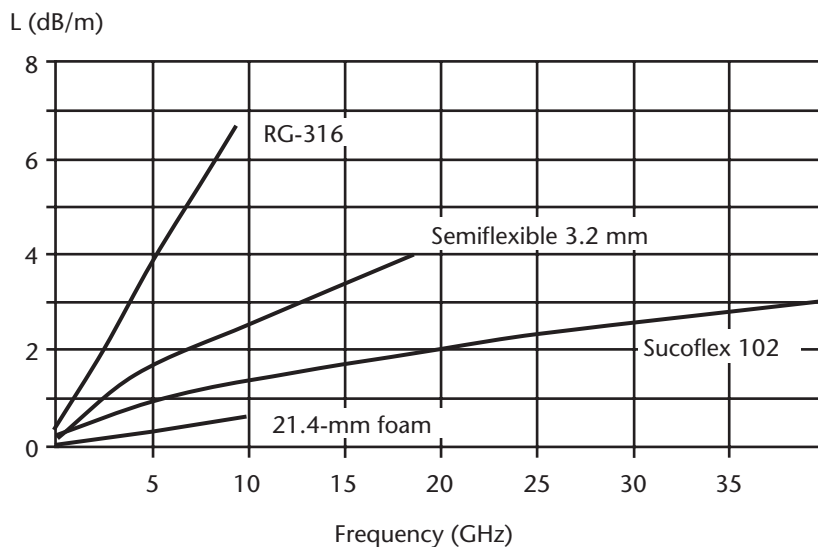


**Figure 5.59** The relative change in phase  $\phi$  as a function of temperature in a typical high-grade microwave coaxial cable assembly.

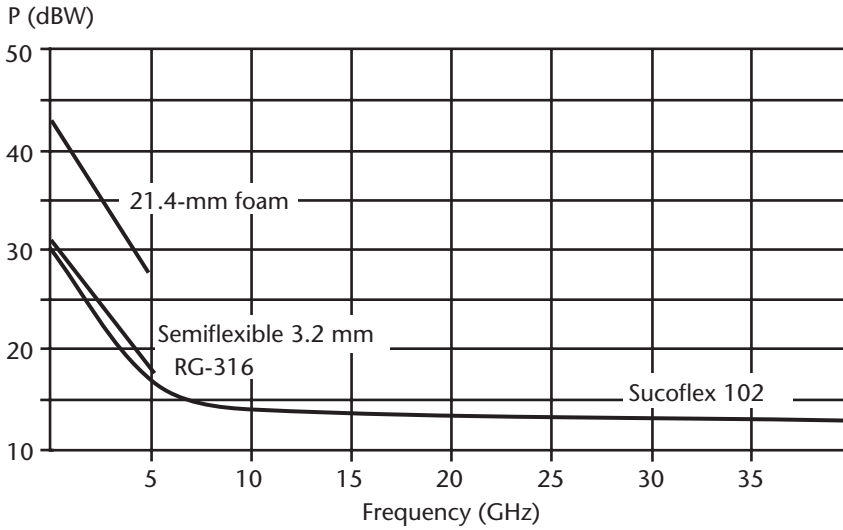
parameters defining the attenuation are the radius of the insulating layer and the type of dielectric. Very thick, practically rigid coaxial “tubes” of the size of 100 mm or more in diameter and having just thin disks of dielectric as supporting devices for the center conductor may have attenuation values as low as 0.01 dB/m even at 1 GHz, but normally used flexible types provide, for example, something between 0.1 and 5 dB/m. However, the actual attenuation depends on the frequency. This is illustrated in Figure 5.60 for some cable types. Another factor influencing the attenuation performance is the type of dielectric. The more air it contains, the lower the attenuation. This feature is utilized in so-called foam cables. The drawback is the possibility of water being absorbed by the air cells whereby the cable gradually gets spoiled. In particular, outdoor installations are considered to be vulnerable in this respect.

Also the RF power-handling capability depends on the operating frequency and cable diameter. At around 10 MHz we could put about 20 kW into a 15-mm cable but at 5 GHz the same coaxial can handle about 200 to 300W. Naturally, normal laboratory-size PTFE cables work at even lower levels, somewhere around 10 to 100W. Solid coaxial tubes used in broadcasting stations take for example 100 kW at 1 GHz. Figure 5.61 shows the power transmission performance of some typical cable sizes. In applications involving power peaks—such as pulsed radar—the main phenomenon encountered as a limiting factor is the voltage breakdown. The cable itself, the transition to the connector, or the connector will be irreversibly damaged. Normally, the most sensitive spot is the transition, partly because there the SWR tends to be worst. Figure 5.62 shows the voltage-limited power-handling performance of typical microwave cables.

The screening or shielding effectiveness of a transmission line works in a reciprocal way, of course. The RF power traveling in the cable will be attenuated when trying to escape, and external interference will face a similar reduction when attempting to enter our system. The key factors defining the shielding performance are the type and material of the cable outer conductor and possibly the number of

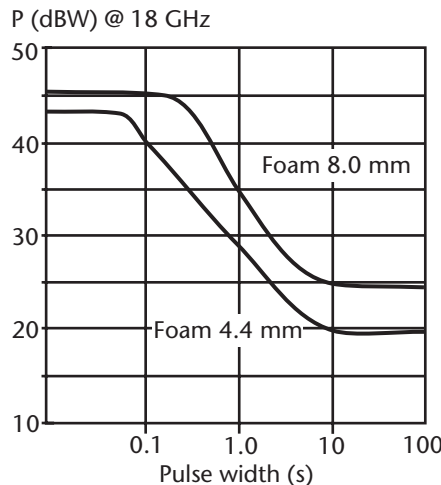


**Figure 5.60** Coaxial cable attenuation  $L$  as a function of operating frequency. Cable diameter and type of dielectric are used as parameters.

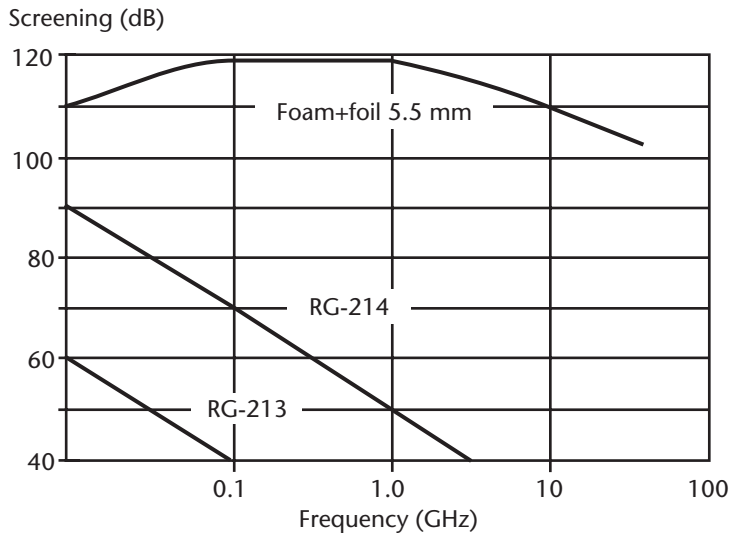


**Figure 5.61** The power-handling capability  $P$  of coaxial cables depends on their diameter and dielectric and on the operating frequency.

screening layers. Of course, this performance figure is also a function of frequency as indicated in Figure 5.63. Numerical data given by manufacturers tells us by how many decibels a signal will be attenuated if trying to cross the border formed by the outer conductor. The standardized measuring length is one meter. There is a wide gap between the best and the worst values ranging from about 30 to more than 120 dB at 1,000 MHz, for example. Special measures and materials can provide very good shielding (e.g., at HF bands) (down to 150 dB) but microwave frequencies above 10 GHz have to live with 100 dB regardless of how much money is used. An alternate way of expressing screening loss is to use the transfer impedance concept, but it does not include capacitive coupling effects. Multiple screening layers may cause PIM problems, if their mutual interface gets corroded.

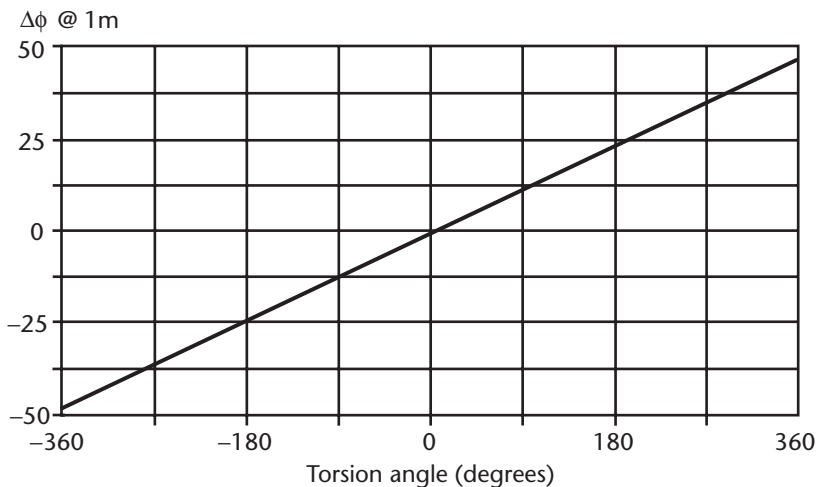


**Figure 5.62** In pulsed systems the power-handling capability  $P$  of coaxial cables is restricted by voltage breakdown. Here the characteristics of selected commercial types are shown.



**Figure 5.63** Multiple screening and solid foil-type outer conductors often give the best shielding performance in coaxial cables particularly at microwave frequencies.

Finally, we want to write a few words about the bending of RF cables during use or assembly. Because the characteristic impedance is bound to the cable cross-section, any bending will have a negative effect on the SWR value. Repeated bending will additionally cause wear in the screening and in the dielectric and even in the center conductor. Thick cables having foil-type screens cannot withstand a radius much below half a meter even though we are working at, say, 100 MHz. Near the millimeter-wave edge we first observe the bending as a change in the electrical length of our line, which means phase instabilities. They may come up already if we try a radius of 10m! Figure 5.64 highlights the phase change measured in a high-grade millimeter-wave cable at 33 GHz when a torsional force is applied. Such a force tries to rotate the end of the cable around the primary axis.



**Figure 5.64** Phase change  $\phi$  in a millimeter-wave cable at 46 GHz as a function of torsional angle. In this case, the manufacturer's application limit is set at  $10^\circ/\text{m}$ .



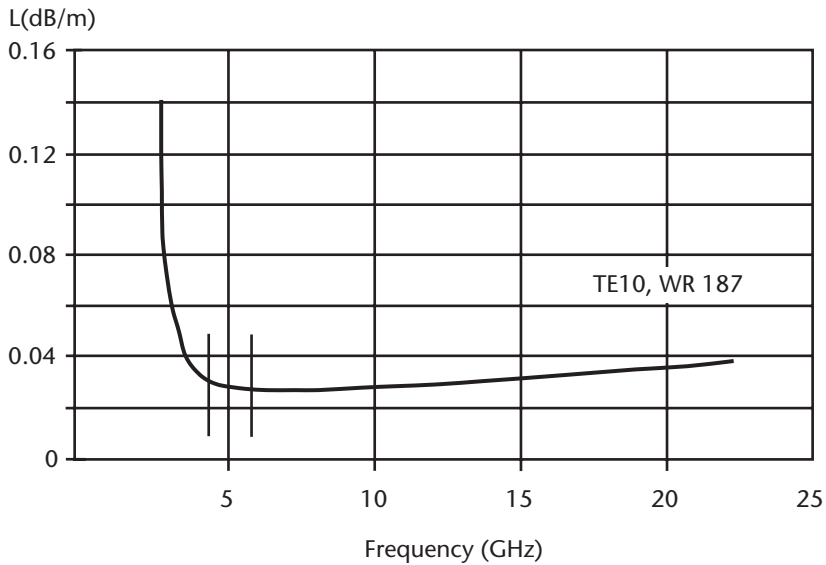
### 5.4.2 Waveguides

Rigid waveguides are still the way to proceed in case we need very low transmission loss or particularly if we are working at high powers and high frequencies [30]. Both circular and rectangular cross-sections are in use, but more standard hardware seems to be available for the latter. Some kind of a sensible limit is 900 MHz, below which the guide opening is simply huge in size. As mentioned earlier, one of the restrictions in using waveguides is the relative bandwidth per selected size, which is typically around one octave. Of course, higher-than-specified frequencies will propagate, but the mode is no more pure  $TE_{01}$ . Commercial products extend currently above 1 THz. Besides straight portions of waveguide, twists, bends, and angles are available. Aluminum and copper alloys are used as materials. Copper gives lower attenuation at high microwave bands, but aluminum enables lightweight and corrosion-resistant designs. The internal guide surface can be plated with silver or gold. Table 5.3 gives some examples of standardized waveguide sizes following the *Electronic Industries Association* (EIA) nomenclature. Theoretical maximum attenuation and allowable peak power occurring at the lowest operating frequency are indicated. Attenuation will be less when working closer to the upper limit of the frequency range. Larger peak powers are possible there, too. Other existing standards include *International Electrotechnical Commission* (IEC) and *Joint Army Navy* (JAN).

All deformations of the guide geometry from the ideal one will have some effect on the transmission properties—attenuation, phase characteristics, or mode purity. This implies that sharp-angled bends will become problematic particularly above 40 GHz or so and only soft “arcs” can be used. As a rule of thumb, about five guide wavelengths is suitable for a soft gradual change in physical direction or polarization. The low attenuation in rigid waveguides is demonstrated in Figure 5.65 for a

**Table 5.3** Examples of Standardized Rectangular Metal Waveguides

Frequency Range (GHz)	IEC Type	Inside Dimensions (mm)	Maximum Attenuation (dB/100m)	Maximum Power Peak (kW)
2.6 to 3.95	WR-284	72.14 × 34.04	2.9	7,600
3.95 to 5.85	WR-187	47.55 × 22.15	5.3	3,300
5.85 to 8.2	WR-137	34.85 × 15.80	10.5	2,000
7.05 to 10	WR-112	28.50 × 12.62	12.4	1,300
8.2 to 12.4	WR-90	22.86 × 10.16	19.4	760
10 to 15	WR-75	19.05 × 9.53	23	620
12.4 to 18	WR-62	15.80 × 7.90	29.1	450
15 to 22	WR-51	12.95 × 6.48	39.5	310
18 to 26.5	WR-42	10.67 × 4.32	62.1	170
26.5 to 40	WR-28	7.11 × 3.36	70	90
33 to 50	WR-22	5.69 × 2.84	97	60
40 to 60	WR-19	4.78 × 2.39	120	45
50 to 75	WR-15	3.76 × 1.88	180	30
75 to 110	WR-10	2.54 × 1.27	320	14



**Figure 5.65** A typical C-band rectangular waveguide attenuation characteristic  $L$  as a function of frequency. The two vertical lines indicate the typical operating bandwidth around 5 GHz.

C-band copper sample. Naturally operation near the cutoff frequency will drastically increase attenuation, but this is not always a problem if the total line length is only a few meters. The clear benefit of using slightly “undersized” waveguides is the 20% reduction in size and weight.

Additional benefits of waveguides include their relatively easy feeding arrangements when connected to horn or certain slot antennas. Because no transitions are needed, the overall efficiency will be improved. However, interfacing with many semiconductor modules tends to be more cumbersome especially below 15 GHz, as a relatively large guide will not mate with normal PCB microstrip electronics. Special transitions are available both directly from microstrip to waveguide or through an intermediate coaxial connector. Individual waveguide sections are connected together by dedicated flanges that are either rectangular or circular in shape. Note that rectangular guide cross-sections may employ circular flanges. Special flange designs allow pressurization whereby moisture defects to the guide interior can be reduced. The solid mechanical nature of waveguides gives them one of the best phase stabilities against vibration. However, fine-tuning the electrical length might not always be so easy.

## 5.5 Connectors

Coaxial transmission lines or cables are seldom soldered directly (e.g., to circuit boards) or fastened by other irreversible means to mechanical structures like antenna elements, filters, or enclosures containing functional modules. Instead, we normally use special microwave connectors, which are good examples of precision mechanics in high-volume production. Depending on application, environment, and frequency range, these components are available from the tiny SMB-types up to

relatively robust 7/16 constructions [31]. Even if a disassembly action of the final product in the field is not anticipated, many modules require a connection arrangement due to tuning or performance checks during its manufacturing process.

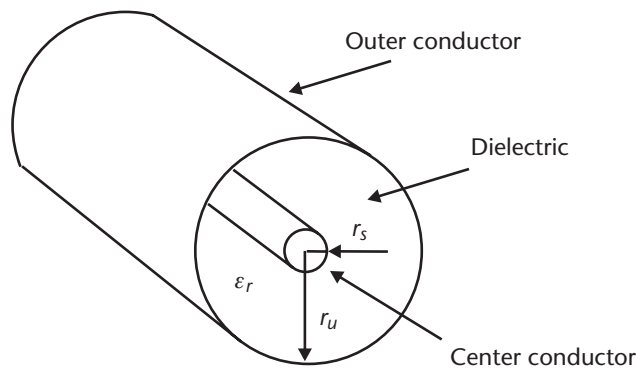
Another extreme might be to say that the fewer the connectors, the better the microwave system. This may be the case, if, for example, very high reliability is desired or if we want to minimize the total weight. Many connector bodies are made of brass, which has a rather high specific weight. When special procedures are employed, a zero-connector approach can give a slightly better RF performance, too, but at the expense of tedious mounting and fine-tuning of physical layouts. Corrosion, contamination, and dust are not such frightening problems if we do not have to worry about connector-mating surfaces.

### 5.5.1 General Performance Requirements

As such, a coaxial connector, which is inserted to the end of a cable or as an interface to an instrument, should be as invisible as possible. In RF terms this means a perfect impedance match and zero attenuation plus an operating bandwidth from dc to several hundred gigahertz. The phase response should be linear or the group delay constant and the power- or voltage-handling capability should be unrestricted. Mechanically we would like to have an infinite number of mating cycles (connect-disconnect) without any deterioration of characteristics, and often there are strict requirements concerning water or dust trying to enter the connector interior. Additional wishes are easy mating—even a blind process—and perhaps a solderless assembly scheme. Some applications call for resistance against ionizing radiation or aggressive chemicals.

### 5.5.2 Fundamental Construction

Similar to the coaxial cable itself, the respective connector has both an outer conductor and an inner conductor as illustrated in Figure 5.66. Due to physical restrictions, an air-insulated component is not possible. Different solid dielectric materials are used instead to support the center conductor in its proper position and to create, together with the dimensions of the cross-section, the desired impedance. For RF



**Figure 5.66** The basic layout of a coaxial structure. The inner conductor (radius  $r_s$ ) is supported by a dielectric material ( $\epsilon_r$ ) in the middle of the outer cylinder (radius  $r_u$ ). If the dielectric is air or some other gaseous mixture, only thin supporting beads are used.

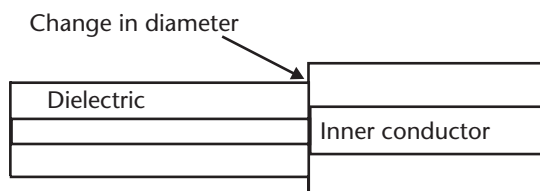
and microwave devices, only 50W is used. A variety of mechanical solutions exist for the locking of the joint. Alternatively, the outer conductor is at the same time the mechanical locking device, or there may be a third level of cylindrical symmetry just for providing a place to put the fastening threads. Most of the coaxial connectors are either of male or female type but some so-called sexless devices like the 7-mm connector simply have one, intermating style. Normally a flange is available at the enclosure or chassis side of a coaxial connector. Both rectangular and round versions are manufactured, the former requiring two or four fastening screws to be inserted to the enclosure wall. Special devices are available for a hermetic seal.

The two connector-mounting schemes—bulkhead or screws—have clear differences. If the bulkhead version is utilized, only one hole in the panel is needed. All the mounting hardware comes with the connector, and the method can withstand some vibration without getting loose. However, the hole must be D-shaped or possibly a double-D, which calls for special punching tools or tedious manual preparations. Besides this, the connector's outer conductor with its nut and washer protrudes deep inside the instrument, making it hard to create a small-sized joint with minimal stray capacitances or inductances. On the other hand, four screws means four more holes, and these are normally threaded. Locking compound should be applied if the end use is subject to vibration. The main benefit of the screw versions is the possibility to have an extended dielectric, which enables a completely invisible fastening of the connector as observed inside the panel. All mounting screws are chosen so as not to exceed the wall thickness of the enclosure. This compact joint gives the best possible impedance characteristics. For many purposes, a very attractive and dimensionally nice alternative is the two-screw SMA-connector.

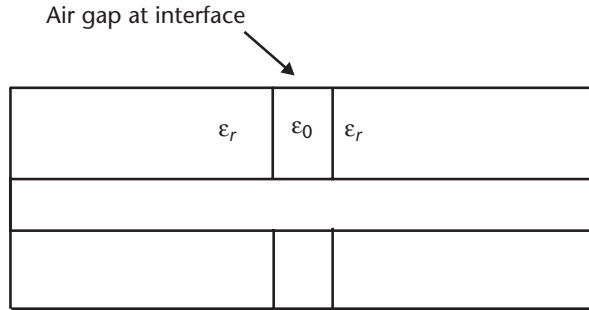
A perfect geometry both as a cross-section and along the direction of wave propagation is hard to obtain. Very often the radii of the connector's inner and outer conductor are not exactly the same as those of the cable as shown in Figure 5.67. Although a correct ratio is maintained, the discontinuities introduce parasitic capacitances, which disturb the impedance matching of the joint. On the other hand, there may be a small air gap between the two mating surfaces of the facing connectors (see Figure 5.68). Here the impedance is totally wrong but fortunately only for a short distance of propagation. The higher the frequency in use, the more severe the effect will be. The situation may be much worse inside the enclosure where we typically have just the center conductor either as a rod or as a thin strip.

### 5.5.3 Common RF Connector Types for Mechanical Modules

The evolution in coaxial connection principles has been based on the trend toward higher and higher operating frequencies and on the desire for smaller equipment



**Figure 5.67** A change in the cross-section at the cable/connector-interface will usually cause a stray capacitance, which deteriorates the impedance matching.



**Figure 5.68** Although the cross-sectional dimensions are similar, a small air gap or a layer of dirt or grease between the two dielectric surfaces can disturb the electrical behavior of a coaxial connector interface.

size. Naturally, the cellular explosion brought also the question of lowest cost in volume production into the front line. The following is not a chronological presentation but rather tries to highlight the problems in using the specific connector types as an interface to passive microwave modules.

#### 5.5.3.1 MCX Connector

These small connectors having a maximum external dimension around 4 mm are suitable for operation up to 3 to 4 GHz. A quick snap-on mating process is used. For applications in mechanical RF modules (i.e., filters and power dividers) only female bulkhead mounted versions with a short (1.5-mm) soldered center conductor are generally available. An extended dielectric is not feasible. If low-cost PCBs in large volume production are the main targets, this type should be considered. However, high microwave tasks or tight performance requirements should be carefully analyzed.

#### 5.5.3.2 BNC Connector

If a simple connector below 1 GHz is needed, this might be a good choice. The well-known bayonet locking is quite practical but prevents a repetitive mating process if stable scattering parameters are necessary. Both male and female chassis connectors are available, and either bulkhead or four-screw panel flanges are used but here, too, no extension of the internal dielectric is available. The relatively large size (center conductor  $\phi = 2$  mm) and poor performance above 1 to 2 GHz limit the usage.

#### 5.5.3.3 TNC Connector

This is almost like the BNC in its appearance but with a threaded mating mechanism. Reliable connections are possible up to and even above 10 GHz. The main technical limitation when considering mechanical modules is again the lack of an extended dielectric version whereby only very thin chassis walls could be accepted. All center conductors are of soldered type as seen from the equipment side. They might be used as an interface to the outside world particularly in aggressive environments and as a slightly smaller ( $\phi = 1.5$  mm) alternative to N-type devices.

#### 5.5.3.4 N Connector

The N-type coaxial connector was originally developed during World War II for the U.S. Navy. Since then several dozens of variants have appeared. Today, commercially available designs include both bulkhead- and panel-mounted designs having either a conventional or an extended dielectric, and they are naturally produced both as male and female versions. Also special microstrip launchers are manufactured in which the equipment side center conductor is only 0.15 mm thick. The female connector, which is the typical choice for an instrument panel or a module chassis, is relatively robust, and only a potential destruction of the external portion of the coupling thread needs special attention. Of course, the center conductor is easily broken by improper mating actions but normally not during module assembly stages. The power-handling capability at 1 GHz is around 500W but goes down to 100W at 10 GHz. Certain N-type designs can be used up 18 GHz, but generally 12 GHz is regarded as practical maximum. Above 1 to 2 GHz the relatively large overall dimensions ( $\phi = 20$  mm) may turn impractical because a well-defined joint geometry in all three dimensions is hard to obtain. Unfortunately there is also a 75-ohm N-type connector available. If mated with a 50-ohm version by force, the center conductor of the female part will be spoiled.

#### 5.5.3.5 7/16 Connector

High-power devices up to and above 10 kW—if the external dimensions above 30 mm are not preventive—could utilize this sturdy connector. Only female units are available for panel or chassis mounting, but both bulkhead as well as four-screw designs are produced. Normally the electrical characteristics are specified to 5 GHz but special arrangements are typically needed for the connection of the center conductor to the instrument's interior.

#### 5.5.3.6 SMA Connector

A very wide variety of SMA-type connectors are currently in production. The list includes bulkhead- and panel-mounted versions, those with and extended dielectric (up to 15 mm) and either straight or 90° angles toward the cable receptacle. The largest cross-sectional dimension—excluding a possible mounting flange—is around 6 mm whereby also relatively small constructions become feasible. Mounting flanges with just two screw positions are also available for the smallest possible footprint. Toward the instrument we can either have a conventional 1.28-mm center conductor or alternatively a 0.15-mm stripline/microstrip launcher. Maximum allowable RF power is—depending on frequency—approximately 100W, and the highest specified frequency for selected SMA-types is 26 GHz. Normally, however, operation above 18 GHz is not recommended. Mated connectors are tightened to the specified torque with a high-quality wrench.

#### 5.5.3.7 K and PC 3.5 Connectors

Although they look very similar to SMA and even allow an intermating action, these two microwave connector types are internally quite different. They can be used to

33 or 40 GHz (K) and give the highest possible repeatability and reliability in this size. The main limitation is the small number of available panel or bulkhead connectors. A special center pin-coupling geometry is necessary, and no dielectric extension is available. For the highest microwave frequencies the user may want to choose the 2.4-mm connector type, but the main benefit is just the avoidance of unwanted propagation modes in the associated cable.

#### 5.5.3.8 SMB Connector

This is a small ( $\phi = 4$  mm) snap-on connector, which can be used up to 2 to 3 GHz. For instruments and passive modules, a couple of bulkhead-mounted versions are produced. No extended dielectric is available, and the inherent impedance mismatch is relatively poor.

#### 5.5.3.9 SMC Connector

This is similar to SMB but uses a threaded coupling mechanism, and thus the upper frequency limit is around 8 GHz. Bulkhead-mounted female receptacles for instruments are available.

#### 5.5.3.10 SMS Connector

This is similar to SMB but with slightly reduced external dimensions. Both male and female connectors for bulkhead mounting are available. Mating connectors are joined by a snap-on process and thus the upper operating frequency should be kept below 2 to 3 GHz.

#### 5.5.3.11 APC-7 and 7-mm Connector

These are very high quality sexless connectors for the upper microwave range. Their external dimensions (around 15 mm) preclude small-sized applications. Extended dielectric versions and both bulkhead and panel mounted screw types are produced. The mating of two connectors requires the use of a special torque wrench. Normally these types are found in high-end test instruments where a large number of successive mating cycles must not disturb the impedance matching.

### 5.5.4 Connectors as Components in Milled or Sheet Assemblies

The most notable problems appearing in the connector selection or mounting process are either caused by difficulties in the center conductor fastening method or by the too large physical discontinuity at the transition. If the connector does not have an extended dielectric, additional pain may arise because of physical impossibilities in arranging a connection between the connector's center conductor and that of the equipment interior. Too short is too short.

Irrespective of selected connector type some general guidelines can be applied [32]. First, a coaxial connector is intended to support only its own weight and proper forces appearing during the mating action. This implies that a connector

should not be left to carry the full weight of a long cable, and more importantly a connector should not be used as a device for strengthening the overall mechanical construction. Thus we must allow the mating connectors couple first and after that see that no bending or twisting happens when individual units are joined together. For example, regarding lowest attenuation, an SMA-series female adapter between two blocks may well seem attractive. Its own mechanical tolerances in conjunction with those of the panel-mounted counterparts can, however, easily yield to a non-matable physical arrangement.

As long as we have a microstrip PCB inside our instrument or building block, it is feasible to try and use a respective launcher-type connector. Many application examples in this book show designs with air as the dielectric. This necessitates more sturdy center pins, which can withstand the weight and possible slight bending of the hanging center conductor arrangement. Only 1.28-mm rod-type constructions are currently available for these purposes (e.g., in the SMA- or N-series). Soft soldering can seldom give the required reliability, and its application is restricted to copper and brass. Experiments with press-fitting in aluminum alloys have yielded promising results, but here the axial force during assembly may disturb the primary protruding distance within the connector.

An ultimate size limit for waveguide feeds and similar “open” designs is set by the dimensions of the connector’s dielectric and those of the center pin. This is due to the fact that the discontinuity caused by the dielectric gap in the guide wall must be small compared to the wavelength. Also, the feed diameter—yet having a positive influence on the bandwidth—should not be excessive. Similar challenges appear if the connector is to be used as a filter section. Our experiments indicate reasonable performance up to 18 GHz, but the tuning process of such probes gets very tedious above 15 GHz. As a practical alternative, a combination of commercial semirigid cable and coaxial connectors could be tried. Here, the probe is formed by the somewhat smaller center conductor of the cable, and the respective outer copper is directly soldered or press-fitted to the ground plane.

Different materials at the interfaces between connectors and equipment interior are sure to cause intermodulation. The severity of this problem is proportional to the power levels involved but also depends on the choice of materials and joining technologies. The question is not straightforward because requirements for successive numerous mating cycles without performance degradation suggest material pairs for the facing connector surfaces. On the other hand, different principles may have guided the selection of alloys inside the module. Poor mechanical contacts, material migration in soft solders, and corrosion will further impair the situation.

## 5.6 Rotary Joints and Flexible Waveguides

Coaxial cables are the natural solution to situations where two or more RF units must be allowed to move with respect to each other. This might be the case in a radar system or in an industrial sensor assembly. However, if very low attenuation or high power- or voltage-handling capability is desired or if we need, for example, continuous unidirectional rotation, the normal way to go is to use a rotary joint. Flexible waveguides give similar benefits but do not naturally allow unlimited



rotation. Less typical solutions have been developed as well. If we are working at millimeter-wave frequencies, we could consider quasi-optical systems [33]. In the case of millimeter-wave reflector antennas we might consider an arrangement where the reflector rotates around a stationary feed, which has a circular azimuth pattern [34].

### 5.6.1 Rotary Joints

Typically a microwave rotary joint forms the RF connection between the stationary and circularly (a complete circle or an arc of it) movable parts within a microwave transmission line system. Such an arrangement may be a part of a radar or possibly also in a tracking communication or telemetry antenna setup. This type of joint could be assembled (e.g., between the TX and/or RX and the rotating antenna) for the azimuth and elevation positioning of antennas, in foldable masts, for example. A practical example of a commercially available rotary joint is shown in Figure 5.69.

A rotary joint has either one or often several parallel microwave channels around the same concentric axis. Normally the frequency ranges and thus the waveguide sizes of the different channels are unequal. In general there are five types of rotary joints, which differ from each other either because of their functional or physical properties:

- Basic waveguide rotary joints;
- Swivel joints;
- Coaxial rotary joints;
- Hollow shaft rotary joints;
- Contacting rotary joints;
- Dual-channel rotary joints;
- Multichannel rotary joints.

The basic waveguide rotary joint construction consists of two parallel waveguides, which have coaxial transitions and a short coaxial line in between. The



**Figure 5.69** An example of a rotary joint designed for a missile. This Siverts RJ 6947 model has two axes of freedom to facilitate both azimuth and elevation scanning.

coaxial part is circularly symmetric allowing free rotation without having disturbing effects on the performance. The position and polarization of the probes remain constant. In the rotating part electrical RF continuity is achieved typically by using  $\lambda/4$ -chokes, which eliminate the need for metal contacts. The waveguide ports can be placed in various positions depending of the application in which the joint is used:

- Both ports at a right angle to the rotational axis;
- One waveguide port at a right angle and one in line;
- Both waveguide ports in line.

As the name describes, swivel joints were developed to twist only about  $60^\circ$  around their neutral position. This is often quite enough for such devices as tracking antennas. The main advantages of a swivel joint are its small dimensions and good peak power (voltage) capacity. Coaxial rotary joints can be classified into hollow shaft rotary joints and contacting rotary joints. In a hollow shaft rotary joint the inner conductor of the coaxial line is hollow, which allows coaxial cables to be put through the waveguide part. There are several geometrical variations on this joint type. In contacting rotary joint models the electrical contact is maintained by utilizing precious metallic sliding contacts. The main reason for this is to realize a very wide operating frequency range down to dc. Mechanical wear is considerably larger and thus the total lifetime will be much shorter than in the noncontacting types.

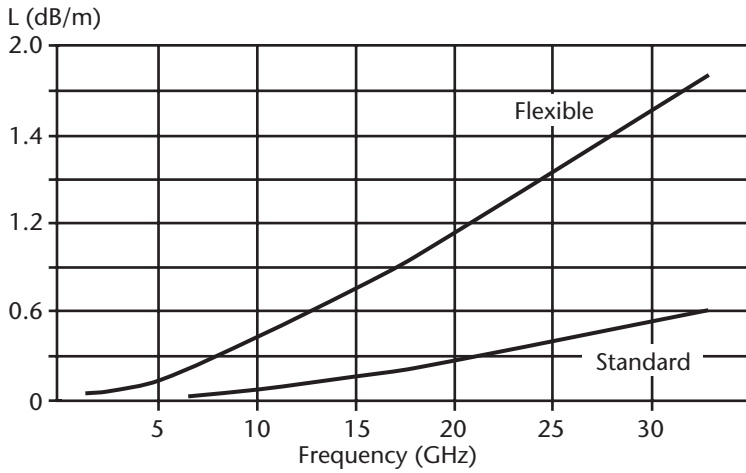
Dual-channel and multichannel rotary joints combine in principal a waveguide rotary joint module with additional coaxial or waveguide sections. For example, the transmission line center conductor will be used as the outer conductor for the next module. By using the hollow shaft of a waveguide module for coaxial cables a number of coaxial modules can be stacked to form a complete assembly with several low-power channels. In many rotary joint constructions additional high precision slip-rings are necessary to transport, for example, low-frequency (or dc) antenna control signals or LNA power supply voltages.

An operating rotary joint will increase the RF noise level of the transmission line because of small changes in its insertion loss and electrical phase as a function of physical angle(s). Also temperature effects can impair the matching at the joint interfaces. The lifetime depends on operating temperature, rotation speed (in typical applications from 1 to 200 rpm), internal barometric pressure, and other mechanical loading conditions. For most sophisticated rotary joints even 50 million revolutions are guaranteed without maintenance actions. Still, the selection of a rotary joint as a replacement for flexible transmission lines should be considered accurately in systems where continuous unidirectional rotation is not mandatory.

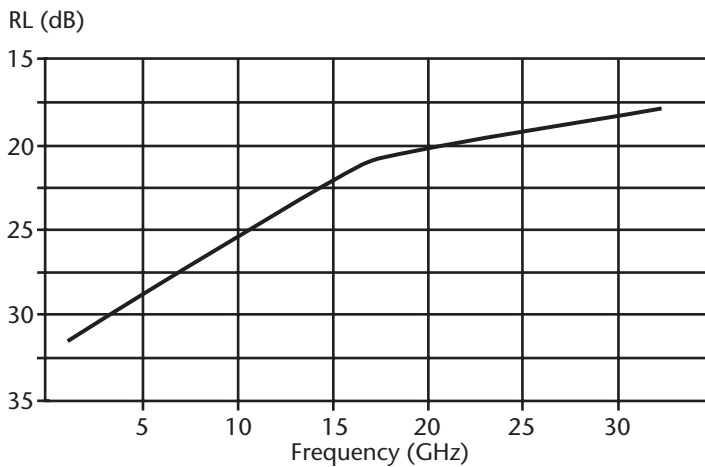
### 5.6.2 Flexible Waveguides

A portion of a flexible waveguide is often used as an intermediate solution between coaxial cable and a rotary joint—naturally only if continuous unidirectional motion is not required. The RF performance is typically a compromise showing somewhat less attenuation than the best cables (at the same frequency) but generally worse phase characteristics than well-designed rotary joints. The mechanical durability is

not easily analyzed. Of course, there are no similar moving parts as in a rotary joint, but tiny internal displacements take place continuously as the waveguide part is being flexed. Figure 5.70 shows the attenuation performance of a flexible waveguide sample and Figure 5.71 the return loss characteristics of the same design. Please observe that although the graphs here cover a considerable frequency range, you have to select the proper guide size for your application. One physical setup can never cover such a wide bandwidth. If only small physical displacements are needed or the flexible portion is only used as an assembly aid, the performance figures are quite close to similar low-grade rigid guides but at higher microwave frequencies the difference is quite remarkable. If aggressive bending is needed, both the expected lifetime and the RF parameters will collapse compared to a completely rigid design.



**Figure 5.70** The attenuation  $L$  of a sample flexible waveguide construction. Naturally, the most appropriate guide size must be selected for each operating band.



**Figure 5.71** The return loss (RL) performance of a sample flexible waveguide construction. As in Figure 5.69, one has to choose a specific guide size according to the operating band.

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# TXs, RXs, and Transceivers

In its simplest form an RF TX is nothing more than a fixed-frequency oscillator. When assembling a TX, we are normally dealing with rather high-level signals, which do not disappear into thermal noise immediately when we turn our back. Microwave RXs, on the other hand, are typically very sensitive and must handle signals whose power is close to that generated by the random motion of electrons due to their nonzero temperature—thus the words “thermal noise.” A more detailed description of this was given in Sections 2.1 and 2.4. Many TXs have only one specific output signal and its modulation to handle at a time whereas RXs must live in a “deep jungle of RF garbage.” This is formed of numerous emissions that can be of man-made or natural origin—thunderstorms being one reason for the latter. Unfortunately, lots of devices that are not intended to generate RF power outside their enclosures behave so in practice. In that way, many of the peaks seen, for example, on a spectrum analyzer screen, if the analyzer is connected to an antenna, are a pure nuisance in the eyes of an RF engineer. Nearby electric motors and dirty high-voltage insulators of overhead power lines are examples of annoying sources of interference below a few hundred megahertz. Of course, broadcasting stations, radars, and cellular phone systems give their own contribution. Thus, the opinion expressed by a number of designer colleagues emphasizing the relative easiness of TX design when compared to RXs might be justified. On the other hand, some quite extraordinary challenges have to be met. Our TX system must not radiate anything else than the initial wanted waveform. The output power may be enough to kill a pigeon flying too close into the main beam. Power supplies might have to handle tens of megawatts of pulsed power. Many military systems have to use *fast frequency hopping* (FFH) or noise-like pseudorandom sequences for the spreading of the spectrum [1]. A modern, state-of-the-art TX with a lot of functional features is not that easy anymore.

Radio RXs, on the other hand, work in the crowded spectrum environment described above and handle signals near the thermal noise power level. Particularly military surveillance RXs, used for *electronic support* (ES), may have to process unknown modulations and waveforms and sustain severe intentional jamming [1]. Depending on the discussion forum, RF RXs can be considered to be individual systems, or they are just modules in very large, often geographically dispersed arrangements.

The following sections first discuss some of the most relevant radio TX parameters and then describe a couple of different topologies. Guidelines for selecting building blocks are given. After that we define some of the most important technical requirements related to RX design and selection. Simple block-level schematics are

discussed to highlight the basic functional ideas. Later on, an overview of commercially available radio RX modules is given, and some suggestions related to selecting a complete ready-made device are outlined. Special topics dealing with transceivers—physical entities that combine an RX and a TX—are shown, and finally a set of practical project examples are used to demonstrate the different challenges. The reader is encouraged to turn back to Chapters 4 and 5 for detailed information about individual modules and technologies when needed. Particularly, the description of filters in Section 4.1.4, mixers in Section 4.3.3, amplifiers in Section 4.3.4, and oscillators in Section 4.3.5 may be useful. Also the discussion of up/downconverters in Section 4.3.7 could be helpful. Different transmission line types were treated in Sections 5.4.1 and 5.4.2. General system-level suggestions can be found in Sections 2.1 and 2.2 as well.

## 6.1 Requirements for TX

International standardization, professional practices, and sometimes good luck have guided the development of radio TXs during the first 100 years of their existence. According to current understanding, at least the following performance figures are important:

- Carrier frequency (or frequency range if applicable);
- Frequency stability (defined in several ways);
- Type of modulation and occupied RF bandwidth;
- Level of unwanted emissions (harmonics and nonharmonics);
- Output power at the antenna connector;
- Baseband characteristics.

Carrier frequency and output power are the two things that have the largest impact on how a TX is constructed and how its actual physical layout looks. This is partly due to the requirement of correct impedances (see Section 5.4), and on the other hand, due to voltage breakdown limits or to volumetric power density expressed in  $\text{W}/\text{cm}^3$ , for example. This means that very high RF output powers can only be produced with physically large circuits. Also, very high voltages call for large enough insulation distances between adjacent components. If the dimensions involved are large compared to the wavelength, we must use transmission lines almost everywhere in our transmitting device.

Historical evolution has caused radio TXs to have rather narrow tuning ranges. Their operating frequencies tend to be tied to the expected use. For example, FM broadcasting TXs use the frequency band from about 88 to 108 MHz or a bit lower in the former Warsaw Pact countries. Cellular phone networks have allocations around 900 and 1,800 MHz, for example, and television satellites transmit around 12 GHz. Even tactical military radios have normally rather restricted bandwidths, for example from about 30 to 80 MHz. The development in radar was first bound to components, then to some tactical desires and elementary phenomenology—mainly propagation and target RCS characteristics—and then again on component technology. Therefore, we still might use symbols like C-, X-, and Ku-band—or the newer

standard—to tell the approximate operating frequency although these letters were originally intended as camouflage. All this implies that RF TXs are typically designed for a relatively small frequency changes and even if implemented, such an action might require intervention into the hardware. Of course, not all systems follow these “mutual arrangements.” Military RF devices may be stored for the “bad day” and jammers especially have to be able to cover quite large bandwidths very rapidly. Also some scientific devices utilize similar features, but they are supposed to operate in closed and controlled environments. Modern software radios, or software-defined radios, are a further exception to the rule. In these devices the operating frequency is determined under program control—within certain feasibility limits—“on-the-fly” depending on usage (voice, video, or data), location, and budget.

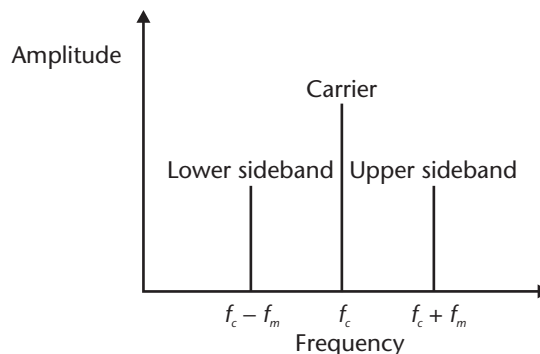
The very first radio TXs had no frequency control at all, but the rapid increase in their number caused the requirement for improved stability. Also a simple technical feature related to the noise performance of RXs encouraged such enhancements. The explanation is as follows: Noise in RXs is proportional to the input bandwidth whereby a narrow frequency range in processing will yield lower noise. Consider the primitive AM bandwidth of phone-quality voice, which is about 6 kHz if both sidebands are taken care of. This is due to the fact that the highest frequency radio signal will appear at a frequency that is the sum of the initial carrier (fixed) and the highest modulation frequency. Also, the lowest frequency will be that of the carrier minus that of the modulating signal. Figure 6.1 illustrates the situation where the modulating audio signal happens to be a pure sine wave of frequency  $f_m$ . The time domain presentation of the obtained voltage in such a case can be expressed as

$$u(t) = (1 + m \sin(2\pi f_m t)) \cdot \cos(2\pi f_c t) \quad (6.1)$$

where  $m$  is the modulation depth from 0 to 1, or if given as a percentage value from 0% to 100%. The mathematical background for the spectrum of Figure 6.1 is obtained through simple manipulation of trigonometric relationships from which we get the same voltage as

$$u(t) = \cos(2\pi f_c t) + \frac{1}{2} m \sin(2\pi(f_c + f_m)t) + \frac{1}{2} m \sin(2\pi(f_c - f_m)t) \quad (6.2)$$

and have the three frequency components clearly visible.



**Figure 6.1** A primitive AM spectrum is obtained if the audio signal happens to be a pure sine wave having a frequency of  $f_m$ .

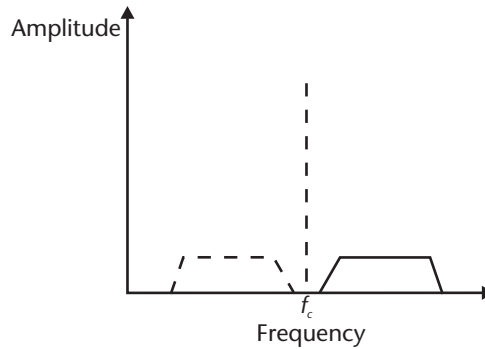


For example, if our carrier is at 400 kHz, the lowest AM output frequency would be  $400 - 3 = 397$  kHz and the highest  $400 + 3 = 403$  kHz. Anything outside these frequencies is useless to the RX (mostly noise and interference) and thus we might design a filter that removes anything below 397 or above 403 kHz. However, if such tailored filters are used in RXs, the potential instability of a free-running carrier may push the signal out of the 6-kHz RX channel already at lower VHF bands.

Because a TX may be considered as an interferer by external systems, the stability requirements are connected to the output power. Practically all modern devices have some kind of quartz crystal oscillator as a backbone. Depending on the specific type and design, the relative uncertainty coming from it may be anything between  $10^{-5}$  and  $10^{-9}$ ; see, for example, [2] and Section 4.3.5. The actual fluctuations of the carrier frequency can be on the same order of magnitude if proper phase locks are utilized, but frequency multipliers (see Section 4.3.3) do not provide this feature. There, instead, the frequency errors are also multiplied. Very high power broadcasting stations and a number of coherent radars use atomic oscillators, mainly rubidium-based units, with which medium- to long-term stabilities on the order of  $10^{-11}$  are feasible.

Many modern radio systems take their real performance from the modulation characteristics. This is obvious, because output power is limited through supply constraints and interference issues. The selected modulation is very much connected to the system in use, and freedom is often available only in some military radar and communication designs and for scientific instrumentation. We can think of modulation as a tool that can be used to put as much information as possible into the allocated portion of spectrum. In other cases, like in the *orthogonal frequency division multiplex* (OFDM) principle used in the new digital terrestrial television, modulation tricks give better quality in the same environment [3]. Here, the initial video information is split onto and transported by a large number of closely spaced carriers in such a way that temporary changes in the propagation path characteristics cannot destroy the quality. As was outlined in Chapter 2, each of the OFDM carriers has its individual multipath behavior. In radar work we might speak about detection and tracking performance instead of the amount of data, although if pondered in more depth, the two things converge. The general principle is anyhow that a more complicated modulation scheme normally gives better performance but sets harder to fill requirements to the TX and respective RX hardware.

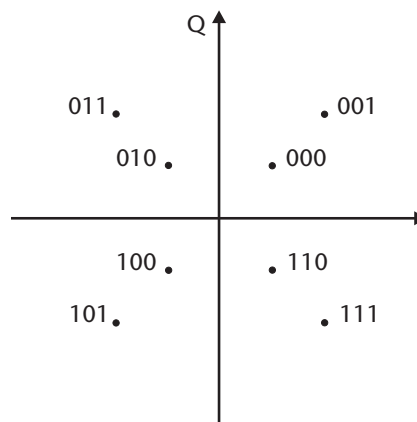
Pure analog AM is still being used. It is simple in terms of equipment and does not use so much bandwidth either. Just consider the difference between seven actual GSM phone channels taking 200 kHz and seven AM phone channels requiring theoretically about 60 kHz. The problem in AM is mainly in poor signal quality in case the C/N is not high; perhaps 40 dB or even more is desirable. Also the necessary TX output power is rather high for comparable distances. Derivations, particularly SSB are more intelligent, because their spectral efficiency is even better and we have a respective relief in TX power, too. In an SSB TX we filter out the carrier and one of the initial AM sidebands leaving just one sideband to be amplified and fed to the antenna as shown in Figure 6.2. This is possible, because the basic AM spectrum unavoidably contains the *upper sidebands* (USBs) and *lower sidebands* (LSBs) plus the carrier, but the fixed carrier has no information in it and the sidebands are duplicates of each other. However, we have to accept a slightly more sophisticated RX,



**Figure 6.2** This is how SSB modulation removes the carrier and one of the sidebands. In this case we speak about USB because the lower sideband is filtered out. Removed spectrum components are shown dashed.

where some means of regenerating the missing carrier must be available to have proper demodulation.

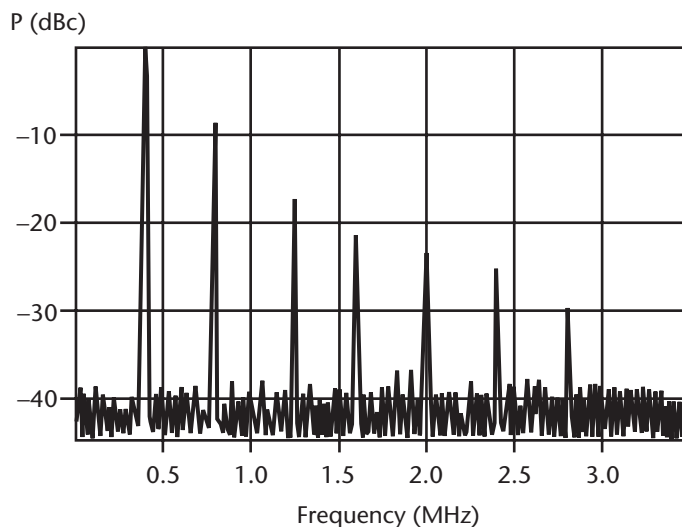
Frequency and phase modulation either with analog or digital baseband signals are common, and their use is expected to continue. Combinations with simultaneous AM are quite effective. The principle is depicted in Figure 6.3, which shows a constellation diagram. It contains eight different carrier vector positions, four of which have lower amplitude values. The numbers indicate an example of the respective digital modulation input bit pattern. In the communications world we often find variants of QAM, which means that different values of the incoming modulating signal are transferred to selected, discrete momentary combinations of the output signal's phase angle and amplitude. For example 64QAM indicates that we have exactly 64 different output vector values available. Radar systems use chirped pulses in which the carrier frequency is rapidly changed (e.g., swept linearly) during the pulse duration. Here the challenge is similar to the difference between AM and SSB. Take an advanced modulation, and you get better transmission performance but have to design an even more complicated TX or RX. A more detailed discussion about various modulation topics can be found in Section 4.3.6.



**Figure 6.3** A constellation diagram shows as a polar coordinate presentation different carrier phase and amplitude values that correspond to the actual modulation input. Numbers show the digital input.

One of the parameters, which used to be the most probable reasons for rejection upon TX type approval, was a too high level of spurious signals. Practices followed by the authorities have changed, but the importance of unwanted signals grows constantly. There is no technical or functional reason to let anything else enter the antenna output except the wanted carrier and its modulation, but because the cost of reducing interfering signals is high, it is often neglected. Integer multiples—that is, harmonics—of the carrier and just arbitrary signals normally pop up at the output and occasionally also elsewhere (e.g., at the power supply interface). Two specifications are used. Very high-power TXs should be judged based on the true unwanted output power at each separate frequency of interest but otherwise relative values (in decibels referred to the wanted carrier amplitude) can be used. Figure 6.4 illustrates as an extreme example the measured output spectrum of an unfortunate commercial aviation NDB, which utilizes a square-wave output stage. Such a beacon is assumed to transmit a single continuous sine wave, which is AM-modulated only momentarily with the station call sign. The azimuth radiation pattern should be omnidirectional and aircraft are using their ADF RXs to define the direction of arrival of this signal. In this case the NDB manufacturer had relied on the filtering performance of the antenna-tuning circuit, but apparently the result was not very successful, because several harmonic multiples of the carrier have high levels. An example of unsuccessful NDB/ADF cooperation was documented in Section 3.3.1.

Naturally, the TX's output power is one of the main parameters defining the entire system performance. Besides the radio link budget, it also has its unavoidable effects on the financial issues, feeding of primary power, weight and size, mobility, and reliability. We can define the smallest TX output as low as we want to, but generally levels around 0 dBm seem to be some kind of sensible limit even at the very high millimeter-wave frequencies. The upper corner, where we might face kilowatts or megawatts either in pulsed or continuous form, is then a completely different story, as is described in Sections 2.3 and 4.3.4. If our operating frequency range is



**Figure 6.4** This is one example of spectrum quality due to poor TX design. An aviation NDB was supposed to work with a square-wave output stage and use the antenna-tuning circuit as a filter. International specifications were not met.

already fixed, we can rather easily figure out the maximum power available from commercial products. Transistor PAs reach even some tens of kilowatts of average power at VHF frequencies, but if we need more or if we have to design a microwave system, we have to go to tubes, which can be, for example, tetrodes, TWTs, or klystrons as was outlined in Section 4.3.4. Naturally, the PA class of operation is very important [4]. Constant power envelope means higher thermal stresses, which happen in conjunction with supply energy wasting but may otherwise be easier to create. In this mode, the final stages of the TX operate at a fixed power as a function of time, and therefore rapid changes do not appear in the supply loading.

One of the not-so-obvious problems of working at very large RF output powers is the required high dc voltage. The energy that we get out of our equipment in the form of an RF signal comes from the power supply circuits. There it is as dc, maybe as a set of rechargeable batteries or as a diode rectifier bridge connected to the mains transformer. The RF transmission line impedance has been often set at 50 ohms, which means, that tens or hundreds of kilowatts of RF require several kilovolts of voltage span even if we use impedance transformers. This voltage span—within which the momentary RF voltage must stay—is just the one given by our dc supply. Additionally, microwave tubes have to have their accelerating voltages, which can be anything between 10 and 100 kV, for example, in typical TWT TXs. This tube technology was already introduced in Section 4.3.4. It is of course thinkable to use power combining, and this is done (e.g., in some modern broadcasting TXs), but combiner losses may prevent a meaningful result. If this scheme works, we may be able to put simultaneously a little bit of redundancy in the output block.

Finally, high power tubes (e.g., TWTs, magnetrons, or triodes) all are quite susceptible to shock, vibration, and humidity. Cooling tends to become a challenge above 100W or so. Klystron output stages in the 20-kW class (average output power) weigh about 100 kg and measure about 1 cubic meter. Mobile installations are surely not easy, and frequent maintenance may be necessary.

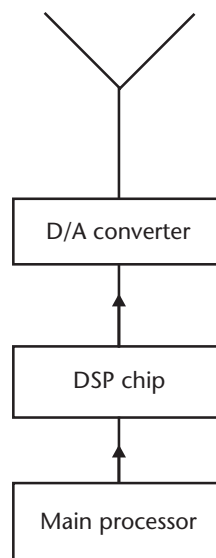
The baseband interface in a TX is mainly defined by the modulator if it is, which is normally the case, a separate functional module. The requirements in this respect are clearly set by the wanted capacity of information transfer, which can be expressed as bits per second or as an analog bandwidth. Simple phone-quality speech typically requires about 3 kHz, modest analog television 2 to 5 MHz, but high-speed data could need 155 Mbps. Generally, there is no strong relation between the RF parts of a TX and its baseband signal, but interesting effects may occasionally come up. For example, let us assume that our baseband signal is uncoded binary data in serial form, such as “1010111100010111010101001....” We have selected multiple-level frequency shifted keying (FSK) as our TX modulation type. It uses several distinct RF frequencies to represent the logical “1s” and “0s.” What happens if our incoming data stream contains just all zeros like “0000000000000....” or all ones “1111111111111111” for a considerable time? The TX output will possibly seize at the one specific frequency although the system concept assumed a rather uniform spreading of energy across the entire allocated bandwidth. Care should be taken also if the receiving site is supposed to perform a carrier locking to the TX—for example, in a link chain. An unfavorable modulation pattern due to unexpected passband signals can seriously hamper such processes.

## 6.2 Block Diagram

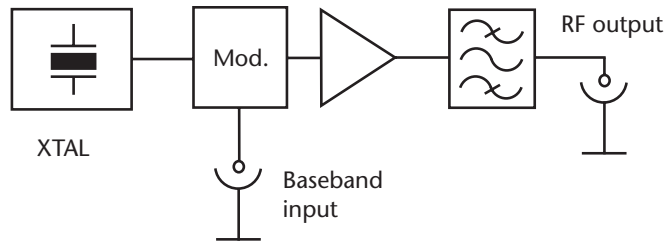
Today's fast DSP chips, which were discussed in Section 2.9, might well enable the construction of an RF TX depicted in Figure 6.5. There is the primary processor containing our information to be sent, a high-speed D/A converter and an antenna. The main idea is that the mathematical algorithm inside the processor can create in real time a numerical equivalent of the desired output signal when the processor input is fed with the baseband data. We here assume that the DSP is also able to take care of all spurious signals and that the D/A block can work at the required frequency and voltage and power levels. Of course, the D/A unit performs the conversion from numerical values to analog antenna input voltages as a function of time. As you see, the only "classical" RF building block is the antenna. Current processing power works already quite fluently in the VHF range, but some time might be still needed for similar microwave/millimeter-wave constructions to appear.

An example of a more conventional and still currently a bit more feasible approach is shown in its most simplified form in Figure 6.6. Our carrier frequency is so low that we can use a direct crystal oscillator, which is connected to a two-level amplitude modulator. Logic "1" may be chosen as the higher envelope power and logic "0" as the lower level. The modulated signal is filtered after which we have an output amplifier [5]. Finally, after some more filtering, we put our signal to the antenna. The described primitive solution can be made to fulfill all specifications related to spectral purity just by selecting or designing suitable filters. Basically, the output power is also freely settable by cascading amplifier stages and putting more and more powerful ones as the last block. As long as the crystal oscillator works, we can also select the operating frequency.

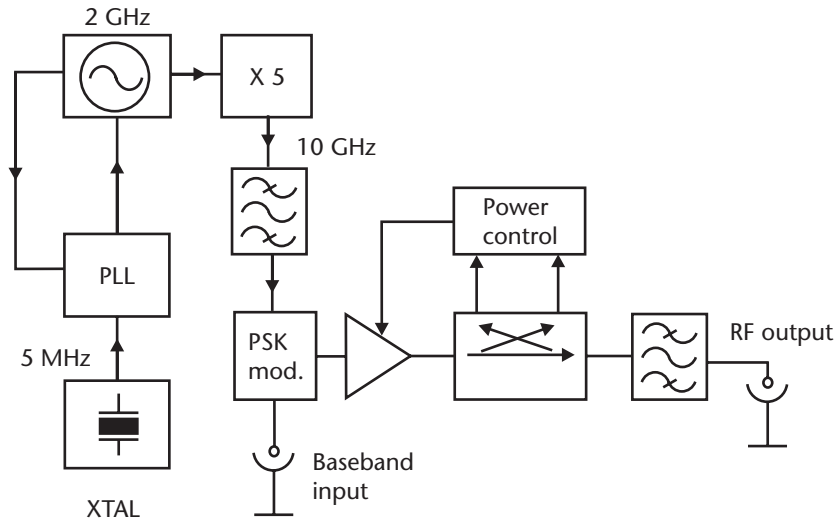
If the carrier frequency happens to be in the microwave or millimeter-wave bands, we can try the concept shown in Figure 6.7. Here we have chosen two-level PSK due to better noise immunity. There is also power leveling installed, which



**Figure 6.5** An all-digital radio TX. Information to be sent sits in the main processor, which feeds the DSP chip. A high-power fast D/A block is directly connected to the antenna.

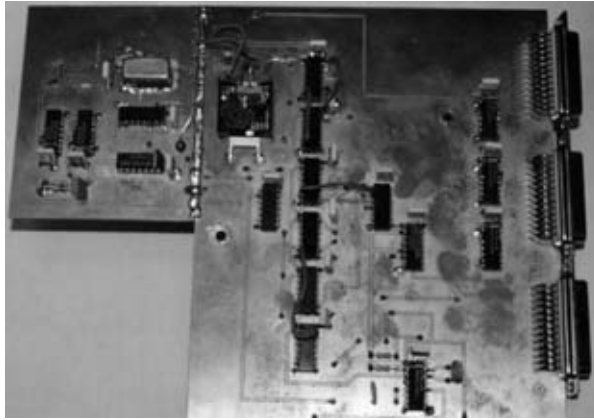


**Figure 6.6** A simple AM TX. The frequency is so low that we can use a direct crystal oscillator. Two-level amplitude modulation is assumed.



**Figure 6.7** Microwave and millimeter-wave TXs often use a combination of frequency multipliers and PLLs to create the final carrier signal. This TX uses PSK modulation and has additionally output power leveling.

means that we have the same output level regardless of slight antenna mismatch. A directional coupler first provides a small sample of the forward and return signals. These samples are detected and processed to set the gain of an AGC block located before the final output stage. In case of antenna failure, this unit can work as an emergency shutdown, which might reduce damage to the TX itself, depending on the time constants. The carrier frequency is based on a chain of a frequency multiplier and a phase locked L-band oscillator. The basic oscillator is still a crystal but now operates at around 5 MHz. Its output signal is fed to a phase comparator together with the output from a voltage-controlled L-band device. The phase comparator produces a suitable control voltage in such a way that the L-band oscillator's output is bound to that of the 5-MHz unit as was described in Section 4.3.5. Because direct phase comparison is not possible if unequal frequencies were used, both the 5-MHz and the L-band signals must be divided by suitable integer numbers before actual comparison. As dividers are not available (or not practical) at higher microwave frequencies, we have to use multiplier diodes that produce harmonics of the L-band signal—for example, the third and the fifth. A suitable filter can extract the desired harmonic for further processing. Figure 6.8 shows a prototype board for



**Figure 6.8** An assembled prototype board for an L-band PLL suitable for the system of Figure 6.4. Observe the large amount of ground plane copper.

an L-band synthesizer, which is capable of 1-Hz resolution. More information about PLLs and oscillators in general is available in Section 4.3.5.

### 6.3 Choosing the Building Blocks

The essential elements needed in a conventional RF TX are the following:

- The carrier frequency generating setup;
- Filters;
- Output amplifier;
- Modulator;
- Power supply;
- Monitoring devices.

A large number of alternative solutions or technologies are either dropped out or become feasible or attractive immediately after the desired carrier frequency or frequency range is known. Similarly, the rough output power class can clarify the situation. Milliwatts or watts, again depending on the frequency, are the task of semiconductors, but kilowatts may require tubes, at least if working at microwave frequencies. Further details can be found in Section 2.3.

If really transmitting into the surrounding space, we should have a controlled frequency. Only some scientific test systems might have completely free-running carrier oscillators. A special class is those magnetron radars where the tasks of the oscillator and the PA are combined. If very stringent stability requirements are not set and the desired carrier of our radio system is below 300 MHz or so, we could try to use direct crystal oscillators as a first alternative [6]. They are simple and we do not have to worry too much about phase instabilities or comb-like harmonics. Most often our choice is, however, a PLL set, either as a single or multiple configuration [7] as shown in Section 6.2. This may become necessary just due to a requirement for frequency stepping during operation or because a crystal unit will not reach that

high. Frequency multipliers are unfortunately needed in the very high microwave and millimeter-wave bands, but proper filtering must be used in conjunction with them.

RF filters should be selected very carefully for TX applications. Unwanted passbands may turn out to be annoying if found only after taking the system in use. Excessive attenuation in the wanted passband degrades output power put may also become a heater, if the absolute power dissipated in it is high enough. Water-cooled antenna combiner filters are perhaps impressive, but they also tell something about the system-level design. Putting a reactive filter at the output of the final power stage may be risky, if the reflected spurious signals are of a sufficient level. The amplifier can then work as a mixer or even be destroyed due to bad SWR.

The choice of modulator type and performance is mainly defined by the initial system specifications whereby there are not so many alternative ways to go. However, the position of the modulator in the consecutive block diagram can be sometimes varied to yield better performance or lower cost. For example very high ON/OFF ratios can be difficult to achieve for pulsed radar work. The task is generally more complicated at higher microwave and millimeter-wave bands. One possible solution is to try a configuration of using an upconverter and have most of the TX-related signal processing in the lower frequency IF unit. This could mean that we have an IF pulse modulator giving 70 dB of the ON/OFF performance and an additional microwave modulator providing another 40 dB or so. Similarly, the frequency sweeping could be performed in the IF section, because it is easier and cheaper, and then transferred to the millimeter-wave band. The additional burden in this scheme is to lock the two frequencies (IF and LO used in the upconversion) effectively.

Poor power supply characteristics are perhaps the most often encountered reason for out-of-spec transmission quality. Selection should be based on the following parameters:

- Adequate surge current capability without unacceptable voltage drop;
- Long-term operation capability at full load in the highest environmental temperature;
- Level of unwanted ac components in the dc;
- Protection against load or supply defects (when applicable).

Special care should be taken to avoid PA current changes from affecting the supply voltages of PLL units or modulators. This coupling can even happen through the mains connection, if this is not of adequate rating. Most AC-type rectifying supplies suffer from relatively high series inductance values, which are particularly difficult in pulsed TX systems. In some cases we may want to use heavy-duty rechargeable batteries (100 Ah or even more) at least for test purposes to make sure that the peak current resources are sufficient. Naturally, this kind of an arrangement involves obvious risks of equipment damage and injury, because efficient protection against short circuits can be rather difficult to assemble.

If size and weight permit, it is advisable to select a power supply unit having about 20% to 30% reserve capacity. When still in development, a system might face a need of further galvanically isolated supplies. Therefore one or two additional



transformer secondaries might well be justified. Regulators and dc capacitors (if applicable) should be specified accordingly. Many conventional “electronics-grade” electrolyte capacitors have huge inductances when tested with RF.

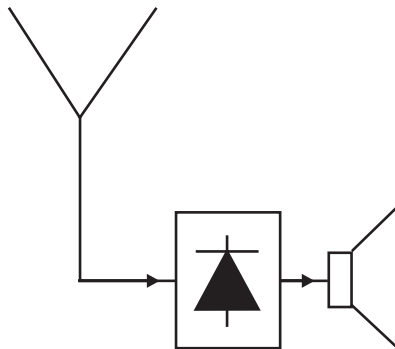
Radar TXs have output powers from about 1 mW up to several MW. The most common modulation types are pulse, *frequency-modulated CW* (FMCW), and a combination of these two. A frequently appearing TX scheme involves pulse compression that can make use of a frequency sweep inside every pulse. The main component of a radar TX is the stage, which produces the output power. This device can be a magnetron, klystron, TWT, or a transistor. In some millimeter-wave systems diodes can also produce the needed low power. Semiconductor TXs are interesting (e.g., for antenna arrays and missile homing heads).

Magnetrons are essentially high-power oscillators. Once they are given dc power, they start to oscillate. Therefore, pulse modulation is accomplished through sophisticated high-power supply systems. The popularity of magnetron TXs in radar work is based on relatively low cost and weight, a good ratio between peak and mean power, and reasonable efficiency. The output frequency of a magnetron is related to its mechanical construction, but slight frequency tuning is possible (e.g., by a motor-driven arrangement). So-called coaxial magnetrons have better frequency stability, but normal types suffer from temperature effects and rather high noise levels.

Coherent, high-stability radar TXs are mainly based on TWT amplifiers that are fed from synthesizers, for example. Output powers from about 10W are commercially available throughout typical radar bands. TWTs have the largest available bandwidths. If we are designing a medium- to high-power radar, we unavoidably face high dc voltages regardless of power stage construction. At the time of writing, Ka-band semiconductor modules suitable for mass production are under development for output powers up to 100 to 300W.

## 6.4 Requirements for RXs

Many colleagues regard the designing process of radio RXs as more complicated than designing TXs. This opinion is perhaps not totally wrong, especially today, although the very simple RX structure of Figure 6.9 was completely satisfying in the



**Figure 6.9** Back in the good old days a radio RX was just a wire antenna, a detector, and headphones. However, selectivity was exceptionally poor, and sensitivity was hard to control.

1920s. During the early years of radio there was less concern about crowded bands and reception quality was not of great importance due to the novelty of the innovation. Most RXs were stationary and could utilize huge antennas. A modern RX has to work in a heavily crowded spectrum and dig up our desired signal from a desperate mess formed of other intelligent transmissions and unintentional interference. People are continuously moving, and most of us require the best detected signal quality.

The most essential criteria in modern RX design are listed as follows:

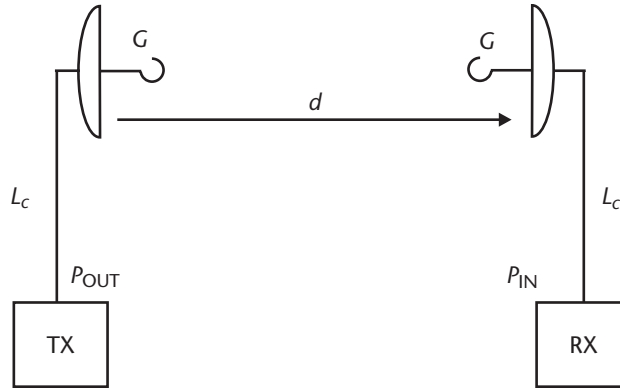
- Sensitivity;
- Dynamic range;
- Selectivity;
- Baseband performance;
- Frequency range (if applicable);
- Tuning speed (if applicable).

Modern RXs having extensive and fast DSP blocks in their IF and baseband circuits can work with a very noisy raw signal coming from the front end, but even they benefit from a proper C/N (in radar S/N) before detection [8]. In more conventional systems of an analog nature this is one of the corner issues. Basically the definition is quite simple. We take our processing bandwidth  $B$ , the front-end noise temperature  $T$ , and Boltzman's constant and multiply them to form the equivalent noise power as was suggested in Chapter 2. This shows that we have only two parameters available. Either the NF of our front end must be low or we have to use a narrow bandwidth. Because the later is typically predefined by other vital characteristics of our system (i.e., data speed, radar range resolution, or rain clutter reduction), only  $T$  remains. A couple of notes must be made, though. The designer must take advantage of all the components involved in defining the noise temperature. Although we often say that only the first amplifier or mixer matters, this is not actually true, particularly when working in the higher microwave or millimeter-wave bands. If our LNA can give only 10 to 15 dB of gain, we must ensure that also the stages immediately after it have sufficiently low noise temperatures. Second, we can have some relief through a sophisticated detector/demodulator design. The conversion from C/N to S/N may allow a couple of decibels more, if we are lucky. Anyhow, a completely uncontrolled system-level design may lead to very impractical RX and TX requirements. For example, let us study in brief how excessive cable attenuation may destroy the noise characteristics of a tactical UHF link. The operating scenario is shown in Figure 6.10.

The power  $P_{IN}$  available at the RX's antenna connector can be expressed as

$$P_{IN} = P_T G_T G_R L_{CT} L_{CR} L_p \quad (6.3)$$

where  $P_T$  is the TX output power at its connector,  $G_T$  is the transmitting antenna gain,  $G_R$  is the receiving antenna gain,  $L_{CT}$  is the cable attenuation between TX and its antenna,  $L_{CR}$  is the cable attenuation between RX and its antenna, and finally  $L_p$  is the propagation path loss. Let us try a numerical example. A commercially available mobile military L-band link uses reflector antennas having a specified gain of



**Figure 6.10** Although actually full-duplex, the tactical link configuration can be effectively analyzed based on this layout.

20 dBi across their entire frequency range (!). Telescopic antenna masts are needed to overcome shadowing caused by trees and vegetation, whereby flexible antenna cables must be used at both sites. This arrangement yields to coaxial feeder cables both for the RX and the TX. Cable attenuation has been measured to be 10 dB. If we assume a separation of 40 km between the two sites, LOS conditions, and the TX output power as +38 dBm, we get at the RX's antenna connector

$$P_{IN} = 38 \text{ dBm} + 40 \text{ dB} - 20 \text{ dB} - 130 \text{ dB} = -72 \text{ dBm} \quad (6.4)$$

Here 40 dB comes from the two antennas ( $2 \times 20 \text{ dB}$ ),  $-20 \text{ dB}$  comes from the two cables ( $2 \times -10 \text{ dB}$ ), and  $-130 \text{ dB}$  comes from the path loss. The noise temperature  $T_R$  at the RX antenna connector (excluding the contribution from the sky; see Chapter 2) is found from

$$T_R = T_{\text{CABLE}} + L_C T_N \quad (6.5)$$

where  $T_{\text{CABLE}}$  is the cable temperature and  $T_N$  is the RX's own noise temperature. Because the specified RX NF is 6 dB, which was caused by a compromise between large-signal characteristics and noise performance, the overall noise temperature is 11,500K. Observe that in (6.5) we have to put  $L_C$  in linear form, not in decibels. If we are able to use a tower-mounted, state-of-the-art LNA having an NF around, say, 1.5 dB and we assume other parameters to be equal, the respective value would be 120K. This means that we are actually able to reduce the TX output power by 20 dB, which has two positive effects. First, the dc drain from the batteries will be considerably lower. Second, and more importantly, hostile surveillance RXs have a much more challenging task in finding the link TX now, because the EIRP also toward the side and backlobe directions will be 20 dB less.

The dynamic range of an RX describes its ability to work in an environment where very weak and large signals appear simultaneously. In principle this definition might be used to analyze signals appearing only at the desired frequency but the real-life scenario often calls for survivability also in those cases where the undesired, out-of-band emission has a very high amplitude. The main elements inside an RX having a direct influence on this parameter are the LNA and—depending on the

topology—the first mixer. Typical commercially available high-grade L-band front ends can handle signal levels roughly between  $-110$  and  $-30$  dBm if we assume, for example, a 200-kHz processing bandwidth. Naturally, if selectable attenuation is provided at the RF input, much larger levels can be tolerated.

Selectivity was originally sometimes defined as the attenuation characteristics of the IF filter, particularly as the ratio of the 3- and 60-dB bandwidths (for example). Good selectivity was something around 1:1.5 and poor perhaps 1:10 or worse. In modern RXs selectivity should be understood in a broader sense, meaning the RX's ability to reject all unwanted signals possibly trying to enter through the antenna interface. Of course, the IF filters still have an important role in this respect, but additional necessary blocks include preselection filters, mixers, and amplifiers. Tracking filters immediately after the antenna connector are today often mandatory in, for example, military surveillance RXs. Their purpose is to reduce the possibilities of unwanted mixing processes in the following stages and to avoid blocking. One of the challenges is to obtain or develop such filters having simultaneously the following characteristics:

- Rapid tuning;
- Low insertion loss;
- Matched or near-matched passband width (with respect to processing);
- Very good passband/stopband ratio;
- High stopband attenuation;
- Low phantom passbands.

The selection of mixers and amplifiers may considerably improve selectivity if types with low intermodulation levels are available for the particular application. This may turn out problematic (e.g., if the RX has to cover a very wide absolute frequency range).

Baseband or demodulation performance is naturally very important, because the end user is anyhow only interested in what he or she gets out of the final connector. Two of the most often used parameters are detection or demodulation performance as a function of the IF C/N and the bandwidth or rise time of the baseband chain. Generally there is not much reason to expand the baseband width much beyond that of the IF circuits, but this is sometimes seen in certain commercially available designs.

Mainly various monitoring and surveillance RXs have to provide a very wide frequency range and possibly also a high tuning speed. Although theoretically an extension of the frequency coverage is just a couple of additional amplifiers and mixers and perhaps some filters, high-quality wideband RXs tend to be complicated, expensive, and sometimes less reliable in use as well. The RF performance is easily compromised if proper practices are not followed. A normal way is to divide the band of interest into reasonable subbands and arrange some multiple RF switches to direct the signal accordingly. This reduces, but does not totally eliminate, problems related to the bandwidth and dynamic range of individual blocks but unavoidably slows down scanning, for example. Also the IF and demodulation stages are often affected, because the prevailing transmission parameters (e.g., deviation) tend to

change with carrier range. Actually very wideband RXs are more or less RX arrays sharing a common power supply, user interface, and enclosure.

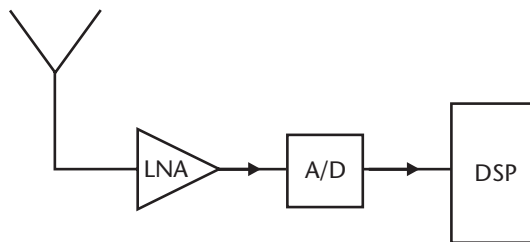
A single mixer RX could be made quite fast. Because the signal must stay within the observation band long enough to allow settling, the main limiting factors are the necessary IF bandwidth and the performance of the tunable oscillator. The better the RF characteristics of our RX become, the slower the tuning will generally be. Tracking preselector filters, if of best quality, are especially tedious to steer at ultrahigh speeds across the band. If speed is one of the major parameters of interest, we may consider an RX bank and arrange for a switching matrix at the demodulated outputs. However, such a system is not one of the cheapest, even though we can share internally some functional blocks.

## 6.5 Block Diagram

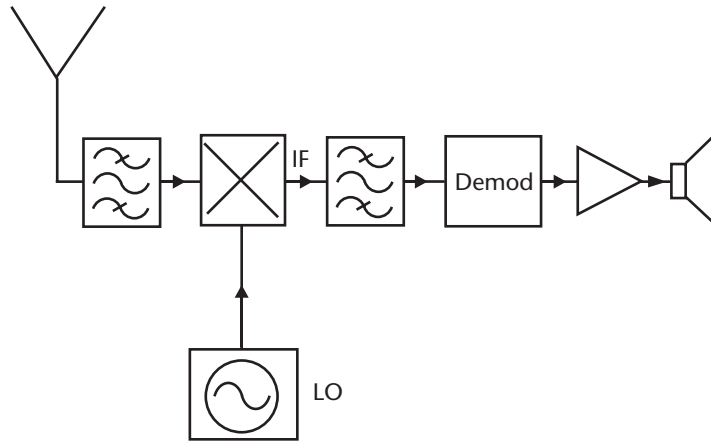
We today very seldom face the possibility of using a direct RX configuration in which the signal captured by the RX antenna is immediately demodulated. Some HF RXs do work according to this principle and the soon-to-come all-digital devices could follow the same scheme, which is schematically shown in Figure 6.11. If we are able to find an A/D converter having three essential features: high enough sampling rate, high enough dynamic range, and low enough power consumption, there it is. However, instead of that, we normally still must use at least one frequency conversion with a mixer. Therefore, one of the simplest practical RX block diagram is like that illustrated in Figures 6.12. The signal goes from the antenna through some kind of a preselection filter to the mixer, which gets its LO frequency from a suitable stable oscillator. The IF is filtered to give the optimized processing bandwidth, and after that we put our demodulator and baseband amplifiers. An audible signal is available from the loudspeaker, for example.

Essentially almost all more complicated RX derivatives are also based on the same functional principles. The basic idea in mixing is here to have a possibility for creating the frequency domain selectivity with a fixed-frequency IF filter. At the same time, demodulation is made easier due to the lower frequencies involved. Actually, this is not always the case, because, for example, classical FM demodulation is easier if the relative deviation is below 10%. This relative peak deviation is simply

$$m = (f_{\max} - f_c) / f_c \quad (6.6)$$



**Figure 6.11** In a way this makes the RF engineer's life easier than ever. An all-digital RX has, after the antenna, a low noise preamplifier after which the whole bundle of signals is brought to an A/D converter. Afterward DSP takes care of the rest.



**Figure 6.12** A simple heterodyne RX. The mixer transfers the original signal and its modulation to the IF, where—after filtering—the actual detection or demodulation takes place.

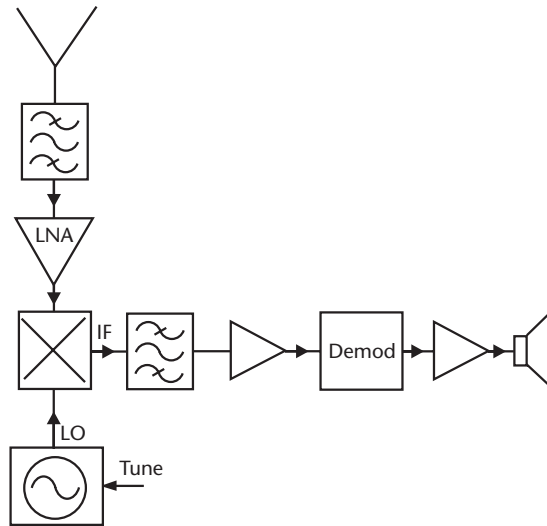
where  $f_{\max}$  is the highest momentary frequency of the modulated RF signal and  $f_c$  is the nominal carrier frequency. If desired, a percentage value is obtained by multiplying (6.6) by 100. For example, if our carrier is at 100 MHz and the modulation causes the TX output to go up to 100.075 MHz,  $m$  will be 0.00075 or 0.075%.

Because no RF or IF amplifiers are present, the entire noise performance relies on the mixer. This RX layout has two strict limitations. First, the input level at the antenna connector must be sufficiently large, because the sensitivity is not good. Also, because there is no real-time tuning possibility, the RX can only handle one single predetermined RF signal.

A lot of conceptual improvements are normally needed. Some possibilities to enhance the characteristics have been collected to Figure 6.13. It illustrates a heterodyne RX, but this time we have an RF preamplifier, an IF amplifier, and a tunable LO. These additional functional modules give us the following features:

- The noise temperature is now lower, defined by the combination of environment, LNA, and mixer (mainly) whereby much weaker transmissions can be demodulated with adequate S/N or BER.
- The demodulator is easier to construct because the IF level will be considerably higher, maybe about 30 to 50 dB.
- We are able to tune our RX to different carrier frequencies as implied by the common mixer equation.

One special problem is created by the layout of Figure 6.13 because now we might not be able to know the spectrum of the LO signal. Due to the tuning action, there is possibly no easy way of filtering, either. This depends on the desired frequency range. Let us assume a LO sweep width from 0.9 to 2 GHz. This means that the first unwanted spectral components will be from 1.8 to 4 GHz and the next between 2.7 to 6 GHz. If the amplitude of the double frequency signal is initially not low enough, we cannot solve the problem with a filter between the mixer and the oscillator. The first multiple would anyhow get through if the wanted frequency were from 0.9 to 1 GHz.



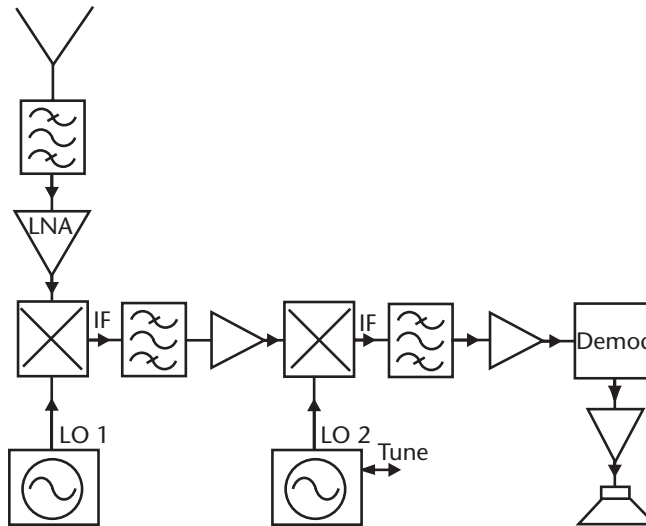
**Figure 6.13** Adding a LNA and an IF amplifier and changing the LO to a tunable unit will enhance the performance of our RX.

As the reader may be well aware of, modern heterodyne RXs are much more complex than the design presented in Figure 6.12. The main reasons for the increased complexity are in the adverse signal spectrum found currently at the antenna interface connector almost everywhere on the Earth or in its vicinity. The reduction of unwanted signals must be extremely high and radio wave propagation characteristics call for a very wide and adjustable dynamic range. These lead to the use of multiple IF frequencies and associated mixers, of course, and to the addition of AGC amplifiers at the IF stages as highlighted in Figure 6.14. Very often AGC blocks are placed at the RF interface as well.

One of the current trends is to use a spatially distributed RX architecture. This means that all the functional blocks are no more found inside the same enclosure or 19-inch rack but parts of the front end, for example, are brought up to the antenna. The driving force behind this has obviously been the introduction of adaptive phased-array antenna systems, but as a byproduct we can get better NFs and have some functional redundancy as well. When configuring the system layout, the designer may consider moving just the LNAs and phase shifters to the antenna or maybe also the first downconversion could take place there as indicated in Figure 6.15. IF signals could be somewhat easier to transport longer distances but this scheme naturally puts more requirements on LO generation and filtering. In particular, the frequency stability of the LO source may be difficult to maintain in the harsh outdoor environment.

## 6.6 Choosing the Building Blocks

One of the very challenging but also most important tasks prior to jumping into the catalogs and Web sites of block vendors is to try to figure out the critical judging parameters of the RX to be built. After that, we should focus the funding and effort



**Figure 6.14** If more capabilities in a hostile spectral environment are needed, the RX should have multiple mixers (called often superheterodyning) and IF amplifiers with adjustable gain.

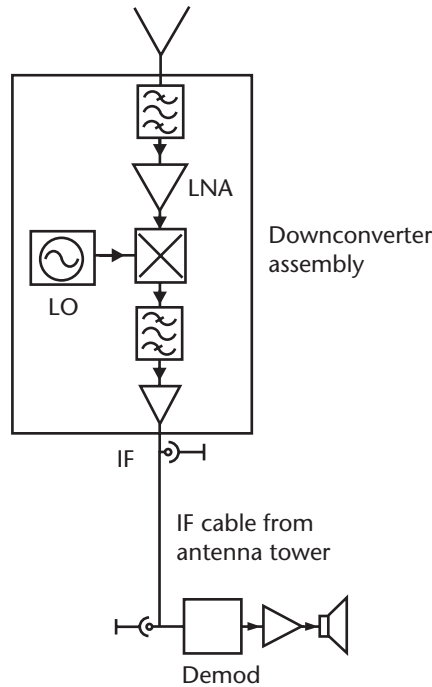
to fulfilling these requirements. The parameters can be nontechnical as well. Particularly cost and schedule may turn out to be such. If, for example we have to have that radio beacon working before D-day, we should perhaps not run around for the lowest possible power consumption. Alternatively, if the RX should detect the weakest return signals from a distant target, could we perhaps put the user interface aside for a while?

Large system projects seldom can afford poor quality of subassemblies or separate functional blocks. After writing this, we must confess that a lot of overengineering happens daily. In many cases “just to make sure” equals “completely in vain.” Some critical notes must be made, anyhow, on the selection of individual modules.

NFs in LNAs and the output power of general-purpose amplifiers are two frequently misleading parameters. Commercial off-the-shelf designs are mainly available because somebody has already had a large enough lot to purchase. Thus the specifications are fixed and based on some existing or near-to-come system, maybe similar to the one with which we are currently working. It means that if the “global” specification for NF, for example, has been 1.4 dB within a certain frequency range, the block manufacturer has been able to satisfy the first customers—but by which margin and what was the percentage of out-of-spec units? In practice we will find out that selecting an amplifier, which is specified just at our design limit (NF,  $P_{OUT}$ ), might lock us or our to-be past friends later into the lab for weeks trying to squeeze the last 0.5 dB out. Maybe the other project’s guys used a bit higher supply voltage or did they cool their preamps? So it is not a bad idea to buy one or two decibels more gain or a little bit lower NF than absolutely needed. Similar trouble may turn up with mixers as well, because their conversion loss highly depends on the exact application circuit, too.

LOs—either VCOs, synthesizers, or fixed-frequency crystal devices—seldom have power problems, and the frequency is normally within specifications. Problems may come up with spurious signals, which can be harmonics or just anything, and with frequency stability. Temperature effects may be surprising especially if we





**Figure 6.15** In a distributed RX architecture we can consider putting the downconverters up to the antenna array, if the easier transportation of the IF signal is of respective benefit.

take the oscillator outdoors. Load pulling, which means unwanted frequency changes due to variations in the oscillator's load impedance, or supply voltage variations may spoil the spectrum in a pulsed system. Generally, if something does not appear on the written specification, there is no way to fight for compensation or for a remedy later.

Filters are very vital elements for RX performance, because they have a role both in sensitivity and selectivity issues. Typically the indicated passband specifications are met but often the stopband is described only qualitatively. If group delay values are not given, anything can come up. Impedance matching is normally very poor outside the passband, and this might create severe problems for mixers and amplifiers. In tunable filters we have to carefully document the possible change in the shape of the amplitude response as the center frequency is being swept. Some commercially available units only have a “typical” insertion loss indicated, for example.

Finally, a few words about radar RXs. The three main requirements for a good device are excellent sensitivity, adequate selectivity, and reasonable dynamic range. Radar RXs must work in a  $1/r^4$  world, which means much less input power compared to the  $1/r^2$  scene prevailing in communications. As many radars use—at least partly—pulse modulation, we must have enough IF bandwidth. Therefore very low NFs might be the only way to reach the desired sensitivity. The dynamic range requirement comes partly from the simple fact that target echoes coming from different distances and different cross-sections have a large amplitude variation. In certain systems we might want to have an RX that is not saturated to close-by clutter, for example. Interference and jamming are important here, too. Logarithmic amplifiers are often used in the IF circuits to enhance the dynamic range before detection.

Modern microwave semiconductors, such as GaAs FETs, can achieve reasonable NFs when properly biased. They are good candidates as radar RX LNAs.

## 6.7 Selecting an RX for the System

In large-scale system projects we might not have the time or the manpower to tackle RX development ourselves and if commercial types are available, the task is simplified to just selecting the most suitable one. The general parameters outlined in Section 6.1 are valid for this process as well, but the attitude might be different, because we now have no later practical possibility to adjust performance once the order has been made. In addition to the primary RF parameters we should also check the following:

- Available IF and baseband outputs;
- Oscillator reradiation toward the antenna (especially in military systems);
- Shielding;
- Computer control (interface or interfaces and commands);
- Supply voltage requirements and connection scheme;
- Environmental limitations (e.g., temperature, humidity, and barometric pressure, vibration);
- Mechanical mounting;
- Size and weight.

Ready-made professional RXs come in three main variants:

- High-grade measuring RXs for EMI tests and research;
- Task-specific RXs for such tasks as satellite ground stations, microwave links, and broadcasting;
- Monitoring and surveillance RXs used by such agents as government authorities and national defense organizations.

If the system under design clearly falls into the second category, we really do not have much to do, because an RX intended for operation in a certain satellite network or as an element of an air navigation system [9] should inherently fill those requirements. Our task is to carefully verify the product's status after delivery. This is not always so easy and might require considerable amounts of dedicated measuring hardware, time, and skill. However, omitting such an inspection might cause serious trouble later when the entire system, in which this particular RX sits, comes to the acceptance phase. Warranty periods seldom exceed 1 year but a large-scale ground station project might take 3 to 5 years to complete before the first-ever signal is captured from space.

Currently available commercial measuring RXs have evolved from pure high-quality radios into entire test setups, which normally include spectrum analysis capabilities and internal processing power. Their frequency range extends from less than 10 kHz up to the millimeter-wave bands and good care has often been taken to

maintain reasonable selectivity and sustainability related to unwanted emissions. However, typically the main performance feature in these devices is the relatively accurate measurement of input level across a very wide dynamic range. Some selectable filters are available, but they are normally not the sharpest ones on the market in order not to spoil amplitude accuracy.

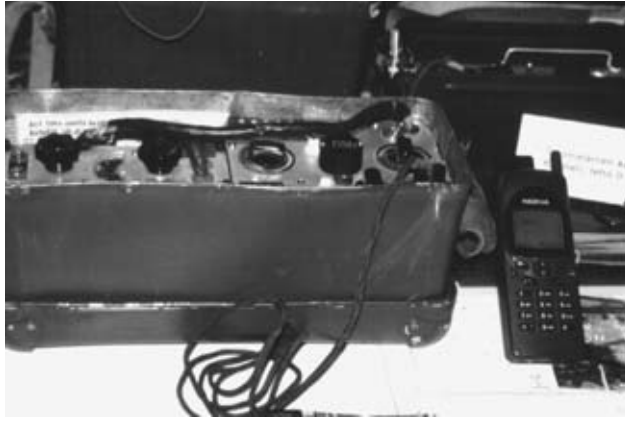
A rather unfortunate thing is that the focus in modern test RXs seems to be in their built-in mathematical and statistical processing capabilities and the user interface. Computer control should be easy indeed through the IEEE-488 and serial-type ports with all that code inserted. Actual RF performance, however, is not much better than it used to be almost two decades ago. Some specialists consider it actually as somewhat worse due to the increased local interference from all that processing power and due to the replacement of very high-grade RF/IF engineering with a set of software-based correction factors. Of course, one of the motivations behind this has been the need to cut production costs.

A military RF system might anyhow most likely take advantage of a typical radio monitoring-type RX. ECMs and counter-countermeasures plus different radar systems can effectively make use of their characteristics. They have some common features with test RXs, but a couple of very essential differences. Most notably, the selectivity, as understood in the broad sense, is superior to all other commercial designs. Also the tuning speed is typically good with figures around 1 to 10 ms being rather normal for an arbitrary step inside the whole available frequency range. Processing bandwidths are provided from 100 Hz to 100 MHz and even beyond. The selection of this kind of an RX seems to end up in a compromise between speed and selectivity. When mechanically tuned tracking filters are used for the highest possible rejection of unnecessary signals, the frequency stepping cannot be as fast as in an all-solid-state device.

Pricing of commercial RXs is not always very clear. Advertising may show remarkably cheap offers but already the first contact to the respective sales people reveals or should at least reveal that the indicated bargain product actually has the worst internal oscillator ever made by that vendor and there is only one demodulation option installed. If you want performance and features, not just a fancy operator interface, be prepared to pay for it.

## 6.8 Transceiver Specialties

Modern cellular phones (or terminals, because much of the information going through them is not voice) are typical examples of highly integrated transceivers. Basically, a transceiver is a radio device and contains both an RX and a TX for the same network or communication connection. So, a box, which has an ordinary FM RX and an RF garage door opener in it, is not a transceiver. Figure 6.16 shows as examples two generations of mobile communication transceivers manufactured in the author's home country. Both utilize state-of-the-art technology and components of their time to achieve smallest size and weight simultaneously with sufficient coverage and achieved unparalleled performance—yet judged on different basis. The older unit from the times of World War II has an interesting approach, because its RX operated on a completely different frequency range than the Morse-code



**Figure 6.16** A portable military HF communication transceiver manufactured in Finland during World War II and sold to Sweden in large quantities and a modern GSM handset from the descendant of the same Finnish vendor 60 years later.

modulated TX. In fact, the RX utilized even different demodulation mode. Besides communications gear, many radar and satellite communication Earth station [10] systems can be considered to contain the transceiver principle as well.

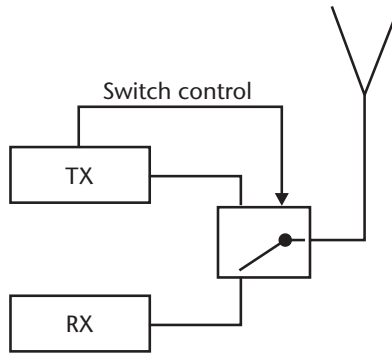
A few noteworthy system aspects come up when configuring a functional transceiver entity. They include the following:

- Mutual coupling, which may happen at the antenna or elsewhere, either at RF or some other frequency;
- Requirement for or benefit of sharing functional modules;
- Flexibility of system layout.

Physically colocated TXs and RXs are often a practical must for true full-duplex communications and for monostatic radar. Normally the antenna arrangements will be so close together that some kind of special solutions must be used. The most obvious one is to use a single antenna and provide suitable means to enhance the isolation between the TX and RX. Frequency separation between the uplink and downlink enables the use of a filter coupler. It can be a really tricky element. The insertion loss must be very low at both bands in order not to spoil the RX NF or to lose too much of the output power. However, rejection from TX to RX might be 60 to 90 dB and the frequency difference is perhaps not very large. The unfriendly impedance characteristics of the RX part at the TX band may also create problems. Sometimes isolators are of help, but they add losses.

If the operational concept is based on some kind of time domain multiplexing, a simple TX/RX switch (SPDT) can handle the whole task as indicated in Figure 6.17. The main concern here is to prevent performance degradation due to mismatch at the moment of switching and in the unused port. Commercial switches for this purpose are therefore available as terminated and reflective versions. However, there is an unavoidable difference in the isolation performance as well.

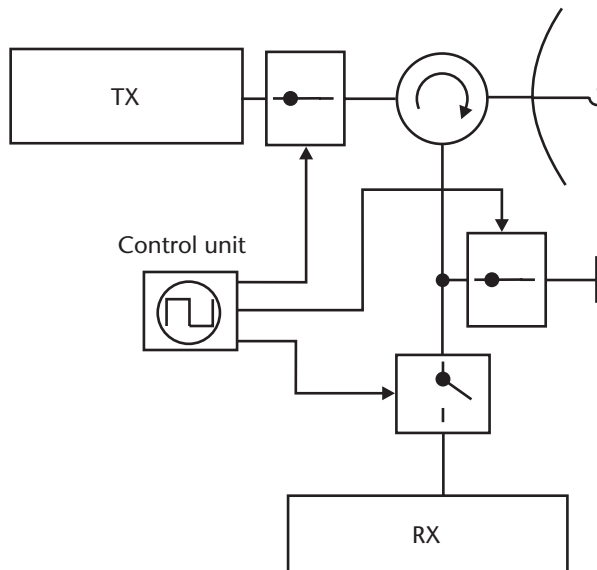
Radars have the problem of same frequency (apart from the possible Doppler shift) during transmission and reception. This is often overcome through switch arrangements as illustrated in Figure 6.18 but it is efficient only for pulsed operation.



**Figure 6.17** Simple TX/RX switching may be possible in a multiplexed system where separate time intervals are used for transmission and reception. Impedance matching may become the main challenge.

The TX port switch is open and the RX switch is closed, respectively, during reception and the two blocks reverse their state for the output pulse period. The RX port has an additional short-circuiting “crowbar”-like device, which becomes active upon receiving some fraction of the TX power. A circulator provides the final link in the combination chain but typically cannot give more than 20 to 30 dB of isolation. Bistatic radars may well circumvent this problem, because their TXs and RXs are not necessarily colocated. However, if we are using monostatic radar devices in a twin-bistatic way, the problem pops up again.

One natural systems-level requirement for RXs and TXs is some degree of coherence in the spectral domain. For physically separated units this is normally handled by distributing the reference frequency to those units needing it. In a transceiver we can use a shared oscillator, or shared oscillators. This removes some of the jitter and noise and may lead to better spectral characteristics. Frequency stability issues have



**Figure 6.18** A TX/RX switching arrangement for a pulsed radar system. The TX switch is open during reception. When the pulse is on, the crowbar absorbs most of the unwanted energy entering the RX branch.

created an entire field of “electronics art” in the classical pulsed radar world with *stable LOs* (STALOs) and *coherent LOs* (COHOs). Their details fall outside the scope of this book, but the reader is encouraged to have a look at [11], for example.

One byproduct of this approach is the reduction of size and production cost. In particular, if we extend the sharing process to other functional modules, like for example power supply units, user interfaces and displays, considerable overlapping can be omitted. Of course, a common enclosure saves lots of mechanical parts. However, some tightening of the specifications of individual blocks may turn mandatory. For example, a modest power supply might have been adequate for the TX, but the RX part requires far better filtering. Internal leakage of signals between functional blocks may also be a nuisance.

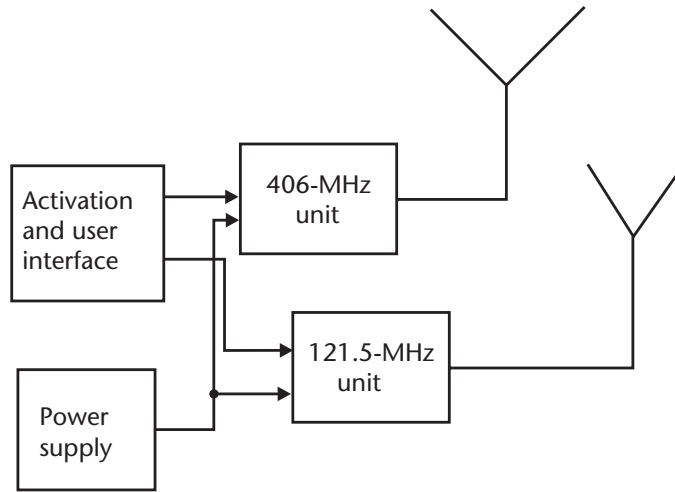
## 6.9 Examples

The diversity of RF TXs and RXs is only partly highlighted by the following five real-life project examples. We have first a look at an emergency satellite system’s portable ground beacon, which has two TXs at the VHF and UHF bands. Next we describe a millimeter-wave system containing both RXs and TXs for the industrial analysis of materials. The following pages contain also material about a millimeter-wave measuring radar and about a short-range unidirectional microwave telemetry system. Finally we show the construction of a UHF TX used for distributing metrology-grade time and frequency references to mobile users.

### 6.9.1 Satellite System Ground Beacon

As an example of a simple TX configuration we briefly consider the design and prototyping of a satellite emergency beacon. The device was intended to be used in conjunction with the SARSAT/COSPAS rescue satellite constellation, which was set up around 1985 as a joint venture between the United States, Canada, the United Kingdom, France, and the former Soviet Union. The principle of operation relies on (1) detecting the existence of radio signals on one of the selected distress frequencies (121.5, 243, or 406.025 MHz), and (2) on measuring and processing the Doppler frequency profiles as seen by the orbiting satellite payloads. Cost and ease of operation caused satellite RXs to be installed as additional payloads (e.g., onboard environmental monitoring satellites using polar orbits and altitudes around 800–900 km). The highest beacon frequency was defined to have an additional identification feature, which should enable the detection of specific type of emergency such as fire onboard a ship, and the identification of the beacon user.

After acquiring the satellite system specifications, the beacon design was initiated as a dual-unit configuration as illustrated in Figure 6.19. Because the satellites were basically compatible with existing aviation *emergency locating TXs* (ELTs) and maritime *emergency positioning indicating radio beacons* (EPIRBs), the initial design incorporated such a unit connected to a separate antenna. Actually this was also mandatory due to the need for a suitable signal for airborne *search-and-rescue* (SAR) units, which do not have equipment for the 406-MHz frequency. The two functional elements were supposed to share a common enclosure and power supply.

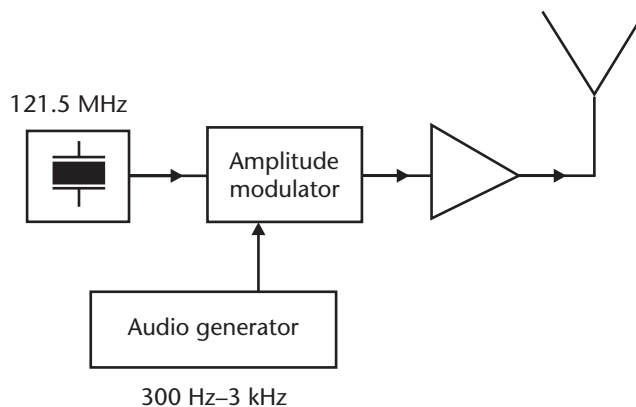


**Figure 6.19** The fundamental layout of the SARSAT/COSPAS emergency beacon configuration. The lower 121.5-MHz signal is produced and transmitted completely separately.

Although certain older commercial devices were operating on 243 MHz only, in this case it was assumed sufficient to allow the harmonic frequency of the lowest carrier to go unattenuated up to the antenna.

The functional block diagram of the 121.5 MHz TX is shown in Figure 6.20. Some technical relief was readily found, because a direct crystal oscillator could be used. A multifunction generator chip produced the specific emergency-vehicle-like repeated audio sweep, which is fed to the amplitude modulator. The wanted output power of about 200-mW PEP was produced in a single transistor unit. Antenna design is somewhat complicated due to the relatively low operating frequency. If an emergency beacon should be portable, and usable also (e.g., in the harsh sea environment), the radiating element cannot follow the quarter-wave rule. Thus a shortened and therefore not very efficient flexible rod antenna was selected.

At 406.025 MHz the satellite system designers were free to start from scratch and this brought some new ideas to the beacon development as well. Phase modulation (two-level PSK) with an uninterrupted carrier and a peak deviation of 1.1 radians was selected. The digital message was to be sent once every 50 seconds and it

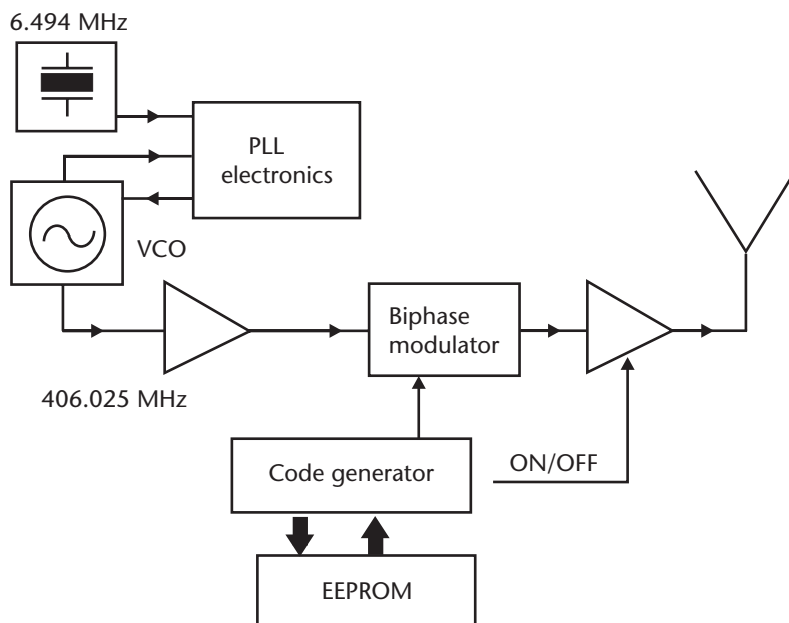


**Figure 6.20** An amplitude modulated 121.5-MHz emergency beacon block diagram. The relatively low carrier enabled us to use direct crystal oscillators.

lasted, with its CW preamble and synchronizing part about 500 ms. The output power into an omnidirectional yet fictitious antenna was specified as +37 dBm. Figure 6.21 shows the prototype block diagram. A phase-locked UHF oscillator produced the carrier, following the principle outlined in Section 6.2. It was fed to an isolation amplifier after which we have the phase modulator. The output amplifier consists of two hybrid stages. An EPROM-based code generator provides the distress message, and the three control signals for the amplifiers and oscillators. In this case the antenna can be “full-size” and a modified folded dipole was chosen. Its mechanical survivability was easily solved, because it fits in the carrying handle of the prototype.

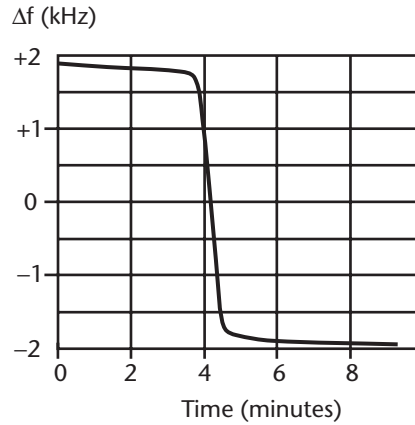
Special concern was caused by the Doppler-related location process. Assuming that the theoretical measured beacon frequency in the satellite would follow a curve like that shown in Figure 6.22, any short or medium term fluctuations or drift in the beacon’s own carrier oscillator would seriously hamper the location process. Theoretically the beacon position was really to be determined just by the accurate shape of the frequency curve and the only remaining problem should be the ambiguity with respect to left and right as seen from the satellite. Much effort was thus put in stabilizing the beacon’s fundamental oscillators.

In a real operating configuration, temperature, short-term variations due to vibrations and medium-term drift were suspected to be the most harmful threats to frequency accuracy. A typical satellite overfly would last up to about 15 minutes so frequency changes occurring within that time period are the most dangerous ones. Two main problems were encountered. At 121.5 MHz the crystal was already near its performance limits due to the high frequency and no real temperature compensation would make sense. The only thinkable solution was to put the whole thing in an oven, but this was not very easy either. First, the TX was assembled on



**Figure 6.21** The higher operating frequency and different modulation scheme together with digital distress message coding require a considerably more complicated TX layout. The carrier frequency is generated with an OCXO-based PLL.





**Figure 6.22** When the SAR satellite passes over the distressed vehicle (not directly over it in most cases, though), the RX on board the satellite will observe a Doppler shift resembling this result. The location of the ELT will be based on the accurate shape of the curve.

one PCB and therefore also the PA would be heated. Besides this, the power needed to heat the oven was at least tenfold compared to the total drain of the 121.5-MHz unit. The second difficulty was faced with the 406-MHz PLL. Although here the fundamental oscillator was a separate unit and could be both compensated against temperature and even put in a tiny oven, the locking circuit had a tendency to wobble during the start of each 500-ms transmission, because the internal impedance of voltage regulators was too high. Finally this defect was temporarily corrected by reconfiguring the supply scheme so that the PLL unit had power connected continuously after activation of the beacon.

Three different prototypes were built to evaluate the performance of the developed TX concepts in real operating environments. Figure 6.23 shows the combined 121.5/406-MHz beacon. The device was about the size of a shoebox and had an overall weight near 2 kg, of which the most came from the lithium batteries. The 406-MHz folded antenna was configured inside the carrying handle. Unfortunately the design in Figure 6.23 was appealing only to the engineers. Sales people considered the unit clumsy and not “stylish” enough. Thus that unit remained a prototype.



**Figure 6.23** The prototype 121.5/406-MHz beacon. Overall size is about that of a shoebox. The UHF antenna is mounted inside the carrying handle, because the whole outer skin is of special ABS plastic. Lithium batteries fill most of the enclosure.

Initial problems in maintaining the short-term stability of the 406-MHz carrier caused the funding partners to request a separate 121.5-MHz version. Two different types were constructed. Figure 6.24 illustrates a civilian marine beacon manufactured into a light-alloy tube with a bright red color. The antenna visible in Figure 6.24 is actually a commercial FM car radio component, which turned out to work very well at slightly higher frequencies. This beacon design was more complicated to assemble, because the electronics and batteries had to be pushed in through the top cover. For the military users a different and slightly more compact version was sketched and actually built as well. It is shown in Figure 6.25. Both beacons have identical electronics except for the activation mechanism. The maritime unit can be switched on manually or automatically through a trigger rope attached to the ship's structures. An interesting detail is the enclosure material used in the military version. After several attempts the TX was assembled into surplus S-band rectangular waveguide, which had been inadvertently cast to slightly wrong cross-sectional dimensions. The different size of units in Figures 6.24 and 6.25 is partly due to the smaller battery capacity required by military users and partly due to the nonbuoyant configuration. Polyurethane foam filled the top part of the maritime beacon for buoyancy.

Real tests in typical or even modest distress environments were quite educational to the system design team, which actively participated in a number of trials. Lots of measuring instrumentation, a bundle of beacons and considerable amounts of other supporting hardware were taken to the test sites, which were selected as far up near the polar region as possible. Figure 6.26 illustrates the isolated test site “in action,” the team trying to identify the correct alignment for the receiving horn antenna. The main reason to go above the polar circle was the higher visibility of satellite overpasses, which meant more test data per used man-hour. However, some rather primitive lessons were learned before a single satellite-based location could be obtained. One of the beacons actually sank into the sea—fortunately close to the shore. The designers had forgotten that the thing should be actually buoyant, which is not so easy if we do not make the enclosure completely watertight. This fault was corrected “on-line” by superfluous amounts of two-component adhesive. A second trial was rather humiliating. Yes, the beacon was now floating, but very well stabilized into a completely horizontal position. The antenna was aligned with



**Figure 6.24** Due to delays in the 406-MHz frequency stabilization, a separate 121-MHz TX for the maritime users was assembled.



**Figure 6.25** Similar in internal electronics to the maritime beacon, this dark gray TX was intended for military users.

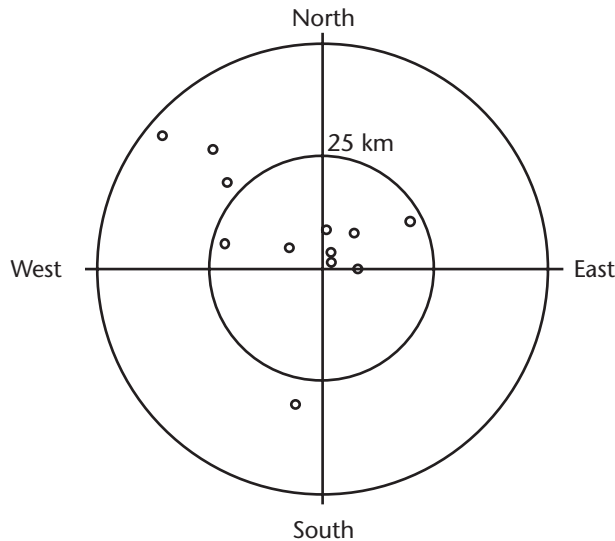


**Figure 6.26** A typical isolated test site. The task at the time of photographing was to find the correct alignment for the receiving horn antenna.

the sea surface and was most of the time covered by waves. This sadly incorrect balancing was adjusted by taking the strainer of the hired cabin's dishwasher and using it as a mechanical counterpoise! The reader should not think that system engineering or system engineers generally fail but unexpected things happen in real life. What is essential instead, is the fact that procedures should be adaptive and we ought to plan for alternative actions in case something goes wrong. The other important issue is the difference between lab tests and work out in the field. Some of these topics are treated further in Chapter 7. Anyhow, after these practical exercises true research could start. One rather successful location result with the corrected beacon is finally shown in Figure 6.27.

### 6.9.2 Material Analysis with Millimeter Waves

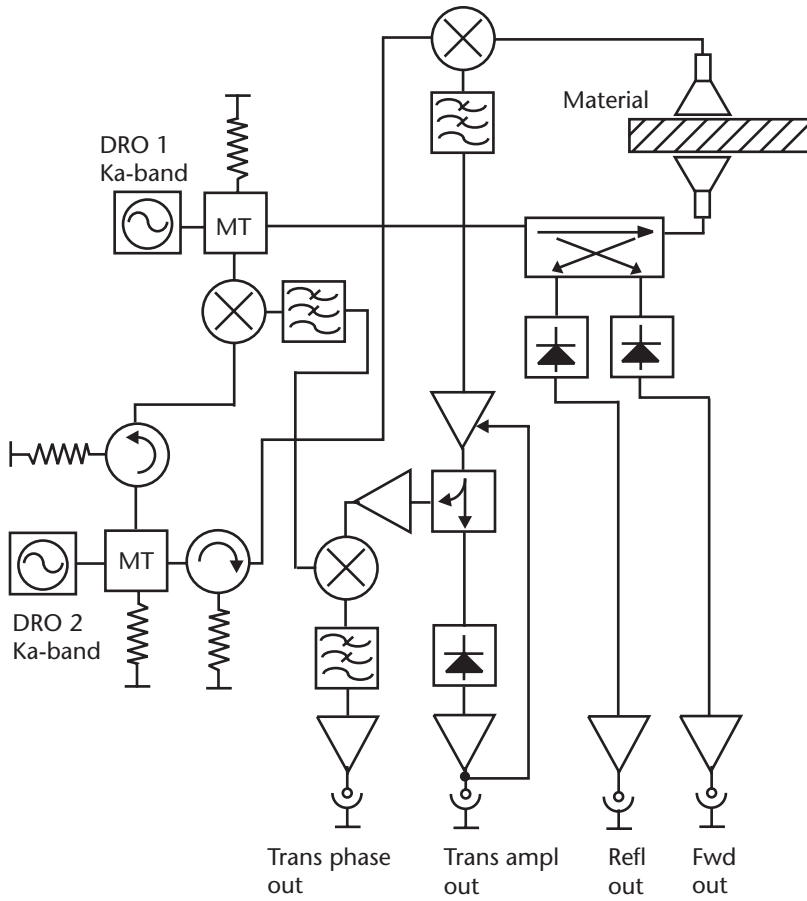
Our second example highlights the main features of a scientific transceiver system, which is used for measuring the properties of logs or related lossy dielectric samples such as concrete or gravel beads. For example, typical parameters which influence



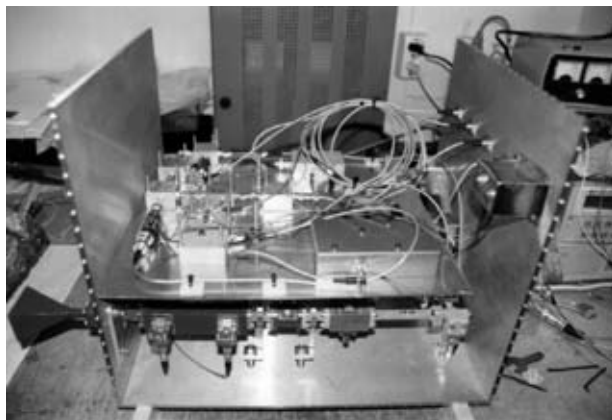
**Figure 6.27** Measured location accuracy from a set of satellite tests. The true beacon position is in the origin. Typically the radial error stayed below 25 km.

the value of a log and its coming use include the number of knots per volume, their size and relative location. Concrete castings may suffer from internal cracks or too large grain size. Many of these defects could be figured out by using RF energy as has been done in, for example, [12, 13] for industrial materials like relatively thin veneer or particle board but require a better spatial resolution than obtained in [13]. Real log-measuring systems operating at L-, S-, and Ku-band microwave frequencies have been developed (e.g., by the author) and are documented in [14, 15]. These systems were shown to be capable of detecting in favorable conditions deformations as small as 3 mm. Normally the size of observable anisotropies was around 10 mm. Besides this, even at the Ku-band the transducer dimensions are too large for an accurate location of a physical defect. As the cost of commercially available millimeter modules is continuously reduced through their extended application in communications, it was recently feasible to try even higher frequencies whereby an idea was born to test the possibilities of related functions at 30 to 40 GHz.

The Ka-band prototype system basically follows the topology of author's previous microwave devices intended for similar tasks, see [14, 15], and so measures the millimeter-wave transmission and reflection in a log or a piece of timber. The complex  $S_{12}$  is measured as a function of distance along the tree. The magnitude of  $S_{11}$  is recorded as well, which helps in evaluating the surface quality of sawed timber and can be used to reduce the effect of bark in log tests. A detailed block diagram of the prototype is shown in Figure 6.28. The system has two Ka-band GHz oscillators, which are actually a combination of X-band DRO devices followed by diode multipliers and waveguide filters. Section 4.3.5 explains the DRO principle in more detail. Ferrite circulators are needed to isolate the oscillators from the high SWR caused by the proximity of the test samples in front of the transducers. All millimeter-wave components are based on waveguide technology, as can be seen in Figure 6.29, which also highlights the internal construction of the unit. As can be seen, waveguide hardware has the drawback of very precise alignment



**Figure 6.28** The block diagram of a Ka-band heterodyne measuring setup for lossy dielectric materials.



**Figure 6.29** The internal construction of the millimeter-wave system.

requirements. These are met by having threaded support struts at suitable locations. The other problem is caused by the numerous flange screws, which need some access

for fastening. Not visible in this illustration is the 1.5-m waveguide line, which connects the RX to the transducer on the other “hidden” side of the test piece. A short piece of millimeter-wave cable is needed to enable adjustments of the gap between the two transducers according to the thickness.

The signal of the first oscillator having a power level of 0 dBm goes through the test sample and is downconverted after that in a balanced mixer. A reference IF is created by mixing the two Ka-band signals. The current construction relies on oscillator power without any amplifiers in the TX part. Similarly, the RX’s only active millimeter-wave device is the mixer. This seems to be an attractive decision in terms of system cost and maintainability as well, because there is no risk of amplifier destruction (e.g., due to mismatch problems). Magic tees are used as isolated power dividers. A typical magic tee is a rectangular waveguide four-port in which three ports (numbered consecutively 1–3) form a planar T-shaped junction and the fourth waveguide arm is connected precisely to this junction straight from the side. In this way, equal power goes from port 2 to ports 1 and 3 but nothing to port 4 or—alternatively—from port 4 to ports 1 and 3 but not to port 2. Three transducer types have been tried. Pyramidal horns are suitable for many coarse recordings and have good polarization purity. Alternatively, small waveguides, which have an aperture of  $2.5 \times 5 \text{ mm}^2$  can be utilized for the best spatial resolution.

The current system has an intermediate frequency of 4.0 MHz and a bandwidth of 2 MHz. This is beneficial because much of the IF hardware is relatively low-cost and lumped circuit elements can be used. Naturally, the frequency drift of the DROs must be kept reasonable. The IF part consists of a LNA and an AGC chain, which has a dynamic range of 65 dB. Phase information is obtained by feeding the amplified IF and the reference to a third mixer, which gives the dc output proportionally to the momentary phase difference. As is seen in the diagram, dc amplification in the range of 40 to 60 dB is used after AM and phase detection.

The AGC attack time had to be improved compared to the previous microwave system due to the much better resolution along the direction of measuring head movement. Therefore the author selected an all-analog scheme as indicated in Figure 6.28. The diode detector output is directly amplified and fed back to the controlled amplifier that enables reaction speeds up to 30 dB/sec. Otherwise the millimeter-wave construction was designed to be directly interchangeable with the two previous systems of [14, 15] as needed and it can make use of the same PC interface card for data acquisition.

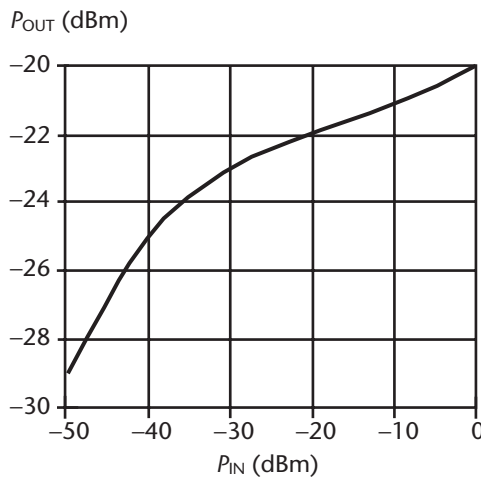
One specific problem is the relatively narrow IF bandwidth, which is required to have the lowest possible noise contribution and to remove some of the spurious mixing products. However, at the same time we set a strict frequency stability target. It turned out that simple DROs require additional thermal stabilization. This relies on the relatively constant temperature ( $8^\circ\text{--}10^\circ$  centigrade) of tap water, which is forced to circulate around both DRO enclosures. The DRO frequency can be further adjusted within 7 MHz by varying its supply voltage.

First measurements with sawed timber of varying quality have indicated that the spatial resolution has been improved to 1 mm or even less and the knot location accuracy is around 3 mm. Both show an improvement by a factor of three compared to the previous Ku-band system. As anticipated, the available output power is not enough to cover thick logs having high internal water content and we are currently

restricted to about 200 to 300 mm of penetration. Typically, the attenuation in dry timber (pine or spruce) is 5 dB/cm but varies from 3 up to 7 dB/cm. Actually, the measured lower limit of  $-92$  dBm at the Ka-band is not set by noise but by the signal leakage through the phase reference channel between the two magic-Ts. This causes the IF signal to appear even in the case of indefinite attenuation. To reduce the effects of unwanted coupling, a 20-dB waveguide attenuator was added between the TX magic-T port and the RX as indicated in the schematic of Figure 6.28. Although we so get a lower output power for the phase detection reference (around  $-20$  dBm), this approach works and improves performance, because we still after attenuation have about 60 dB of margin in the reference path. It looks also possible to enhance the dynamic range further by an active canceler—similar to that used in modern radars against unwanted sidelobes—which feeds a phase-shifted copy of the IF to a summing power combiner at precisely the same amplitude. However, the frequency stability of the DRO devices is challenging for a phasing action.

The AGC range is sufficient as indicated in Figure 6.30. It defines the practical difference between the minimum and maximum input signal levels. The highest practical IF output of  $-5$  dBm is still within the linear region of the control loop, the design of which was also highlighted in Section 4.3.1. The two millimeter-wave mixers do behave as real double-balanced devices and have a lower IF cutoff at dc (0 Hz) but their biasing gets easily disturbed if a dc blocking feed-through is not used between them and the IF amplifier inputs.

Similarly to the Ku-band experiments we here, too, find good correlation between rapid phase changes and knots. Due to the smaller wavelength, a large knot (20–50 mm) can cause a phase shift of  $200^\circ$  or more. Amplitude changes (here from 10–30 dB typically) are a practical detection parameter as well, but they do not show such sharp response times and again, they may turn out to be suitable for coarse analysis only. The amount of water in the tree sample has naturally a severe effect on the attenuation and very wet samples cannot be measured at Ka frequencies without additional amplifiers. Anyhow, the new system cannot be used as a sole means for log analysis but it must be seen as a necessary supplement to the previous L- and



**Figure 6.30** The AGC control loop performance as a function of the raw IF power. The total deviation from a perfect linear fit is 9 dB over the entire AGC range.

Ku-band arrangements. As a joint entity, they are capable of providing a reasonable RF image of the log or timber under test.

### 6.9.3 Mobile Millimeter-Wave Radar

Although approaching its seventieth birthday, radar is still one of the key sensors for national defense and antiterrorist actions. The requirement for mobility pushes the design effort toward a reduced physical size and thus toward millimeter waves due to favorable antenna apertures, even though [16], for example, does not encourage such constructions. For example, the operation of short-range antiaircraft guns and missiles could be supported in bad visibility with suitably scaled autonomous sensor systems, which match the typical kill radius of 2 to 3 km. This example shows some preliminary setups related to evaluations of a small measuring radar operating at some millimeter-wave frequency.

The first known mobile millimeter radar to enter large-scale military service is possibly the U.S. Long Bow system for the Apache AH-64 helicopter. This device operates in the Ka-band in a look-down fashion and is intended against such targets as tanks. No real performance data is available. Russian industry has demonstrated a ground-based millimeter-wave antitank missile radar. Several civilian 60- to 90-GHz systems for vehicular anticollision monitoring have been developed. Both pulse-Doppler and FMCW setups based on laboratory equipment or integrated components are shown in [17–19]. Output power varies from 2 to 13 dBm, NF is typically 4 to 7 dB and processing bandwidth around 150 MHz. Electronic tuning through 1 GHz has been achieved at 94 GHz. A range resolution of 0.75m and a measuring range above 150m have been obtained. The angular resolution was  $1.5^\circ$  and the processing of the whole field of view in [18] took 13 ms (10,000 cells). The work in [20] describes a millimeter-wave add-on to an existing C-band radar. Here at 36 GHz the output power has been 2W, the RX NF below 4 dB, and the parabolic antenna main beam covered  $6^\circ \times 12^\circ$ . Remarkable achievements in millimeter-wave modules are demonstrated also in the Roi-Namur (Marshall Islands) radars, designed by the MIT Lincoln Laboratory. For example TWT performance up to 50-kW peak power and bandwidth exceeding 2 GHz was achieved at 35 GHz. However, the mobility of such constructions is quite limited indeed. The same laboratory has been involved—together with Goodyear Aerospace—in the development of the *advanced detection technology sensor* (ADTS), which is a 33-GHz *synthetic aperture radar* (SAR) not too heavy (around 60 kg) to be put on board *unmanned aerial vehicles* (UAVs), for example, [21]. Neither of these two Ka-band systems is available outside the United States. Lincoln Laboratory has also been active in the 94-GHz frequency range.

For a lightweight jeep trailer or a patrol boat, the radar system mass should be kept below, say, 500 kg and the antenna diameter smaller than 1 to 2m. Man-portable devices must be considerably smaller, of course. Operation on battery power is highly desirable as is the capability to withstand shocks and vibration, which does not encourage using high-voltage tube amplifiers. One of the fundamental questions is if operationally sufficient TX power levels and RX NFs are technically feasible in such configurations and if so, how to select the frequency. Particularly ground and weather clutter data at millimeter-wave bands seems to be



hard to find and even if such were available, they might not fit the real operational environment. Thus, there is need for region-specific measurements.

The well-known radar equation from [22, 23] can be combined with target RCS information, such as that in [24, 25] to find a rough estimate for the needed power or to judge the range characteristics under free-space conditions as a function of operating frequency. One very simple form of the radar equation is

$$P_{\text{IN}} = (P_L G_T / 4\pi r^2) \cdot A_T \cdot (1/4\pi r^2) \cdot A_{\text{eff}} \quad (6.7)$$

where  $P_L G_T$  gives the equivalent power toward the target,  $1/4\pi r^2$  is the free-space loss (appearing twice as the signal propagates to and from the target),  $A_T$  is the target's radar cross-section and  $A_{\text{eff}}$  is the effective area of the radar antenna. All additional losses have been excluded here. If we assume a fictitious, perfectly perpendicular (as seen from the radar) rectangular planar target having dimensions much larger than the wavelength and a circular parabolic radar antenna [26], we have first

$$A_{\text{eff}} = (\lambda^2 / 4\pi) \cdot G_T \quad (6.8)$$

and

$$G_T = \eta(\pi d / \lambda)^2 \quad (6.9)$$

and

$$A_T = 4\pi a^4 / \lambda^2 \quad (6.10)$$

and then obtain through elementary manipulation the received power as

$$P_{\text{IN}} = \frac{P_L \pi^2 d^4 a^4}{16r^4 \lambda^4} \quad (6.11)$$

where  $r$  is the range to the target,  $\lambda$  the operating wavelength,  $P_L$  the TX output power,  $a$  the width of the square and  $d$  is the radar antenna's aperture diameter. In (6.11) we have assumed that the aperture efficiency is 100%, which means that  $\eta = 1$ . If we instead of the square plate assume a spherical target having a radius of  $a$  (again considerably larger than the wavelength), our target's radar cross-section would be

$$A_T = \pi a^2 \quad (6.12)$$

and we get as input power

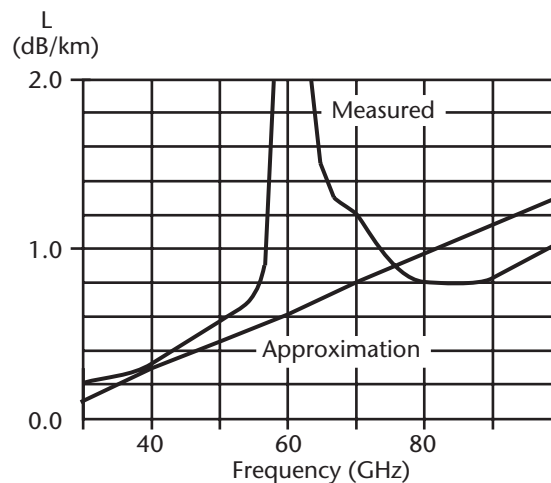
$$P_{\text{IN}} = \frac{P_L \pi^2 d^4 a^2}{64r^4 \lambda^2} \quad (6.13)$$

The additional attenuation due to absorption (excluding direct rain and resonances due to water and oxygen molecules in the vicinity of 22.2 GHz and within 50–70 GHz) in the atmosphere can be coarsely approximated as

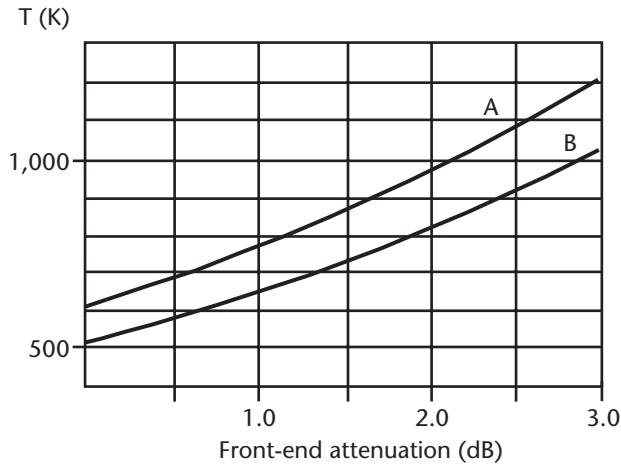
$$L_{atm} = \left[ 0.1 + \frac{(f - 30)}{30} 0.15 \right] \frac{\text{dB}}{\text{km}} \quad (6.14)$$

where  $f$  is to be given in gigahertz and should be between 30 and 50 GHz or from 70 to 100 GHz. It is obvious, that (6.14) will give far too low attenuation values within the oxygen absorption range at 55 to 65 GHz, where sea-level values up to 15 dB/km have been measured at 60 GHz, for example. This fact is highlighted in Figure 6.31 where measured data is plotted for comparison together with (6.14). Combining the characteristics from (6.11) and (6.14) we find out that for some elementary targets an increase in radar frequency might be justified as the radar cross-section of aligned planar sheets is proportional to the fourth power of  $f$ . In many real-life radar applications, though, there's no clear benefit. The reader can have a look at Figure 2.4 for a graphical presentation where also the effects of typically available TX output power have been included. Besides, it is noteworthy that [27] reports the attenuation of already one single fir tree at 60 GHz to be 20 to 25 dB. Anyhow, pure absorption will not prevent clear-air operation at frequencies from 30 to 45 GHz within 10- to 20-km distances, because the two-way attenuation would be about 0.2 to 0.4 dB/km. Thus the selection of frequency can be based on other parameters, such as on tactical and operational requirements (scanning speed, tracking resolution), and on the performance of commercially available components, which are described for example in [28–32]. Figure 6.32 indicates as an example the effects of two off-the-shelf front-end configurations on the obtainable system temperature.

Assuming a shaped parabolic antenna of 1-m diameter for mobility and mostly MMIC-based electronics similar to those documented above, we can estimate for example at Ka-band a detection distance up to 8 km for basic targets. These figures will be drastically reduced against stealth targets (like the F-117) for which [33] gives an RCS of 0.025 m<sup>2</sup> although no frequency is indicated. Modern adaptive processing algorithms are assumed to give a gain of 15 to 20 dB, which enables an



**Figure 6.31** A comparison of measured two-way attenuation at sea-level in clear sky conditions and the approximate equation (see text). The large discrepancy due to oxygen absorption from 55 to 65 GHz is evident.

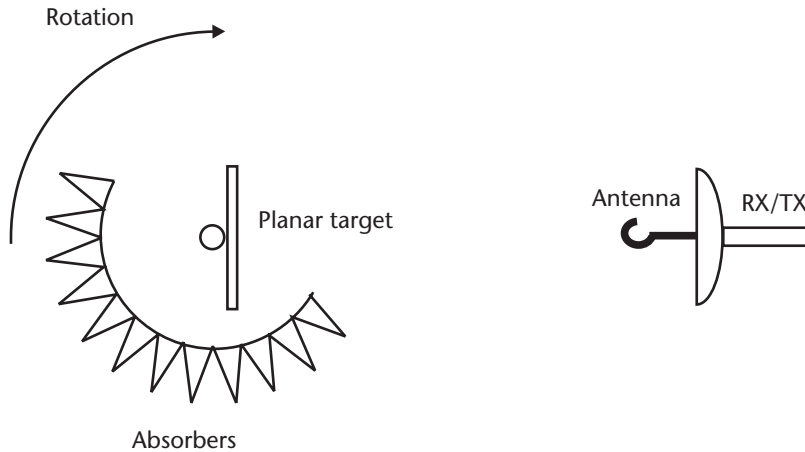


**Figure 6.32** The effect of attenuation (before LNA) in two selected radar front-end configurations on the overall system temperature. Examples are A (LNA 300K/20 dB, mixer 500K/3.5 dB, IF amplifier 100K) and B (LNA 100 K / 20 dB, mixer 500K/3.5 dB, IF amplifier 100K).

$S/N$  near 0 dB for the raw IF port. If we want to use short pulses, even on the order of 10 ns, the detection bandwidth might be above 150 MHz. Of course, such wide IF interfaces are not a general requirement in millimeter-wave radars. To have some frequency agility, also millimeter-wave oscillator tuning is desirable.

Very important are the clutter characteristics of the typical operating environment. A compilation of unclassified data for millimeter-wave frequencies, including further references, can be found in [34]. The secondary “targets” caused by clutter inside the main beam are most prominent in heavy rain or snowfall. For example, [16, 17] show some empirical equations and graphs for the estimation of rain-based clutter but both have been obviously created much before the era of commercially available and cost-effective millimeter-wave technology. Thus their results and test campaigns have been focused at S- to X-band systems. On the other hand, the more recent [28, 29, 35], which indeed discuss millimeter-wave systems, all put an emphasis on such very short propagation distances where rain cannot be a serious practical problem.

In order to be able to create a suitable database, which contains both attenuation and rain clutter data at interesting millimeter-wave frequencies, a simple two-way test system has been constructed, see Figure 6.33. Here the idea is to have a rotating cylindrical absorber temporarily mask the fixed reflector. In this way the TX/RX system can see, once per every revolution, the reflector for calibration but it is also able to measure the true background without the strong backscattering from the reflector. The target is typically a conducting rectangle or sphere (physical size about  $0.1 \text{ m}^2$ ) and it contains a flush-mounted millimeter-wave waveguide antenna together with a band pass-filtered diode detector. This simple RX gives a straightforward way of estimating the one-way path loss. There is an obvious trade-off related to the test path distance. If a large rain clutter volume is desired while utilizing short pulses, the distance should be large. However, a large distance means higher two-way attenuation. Another challenge is the path elevation. In many operational scenarios the antennas will be pointing to very low angles, but if a test track is constructed like this, ground

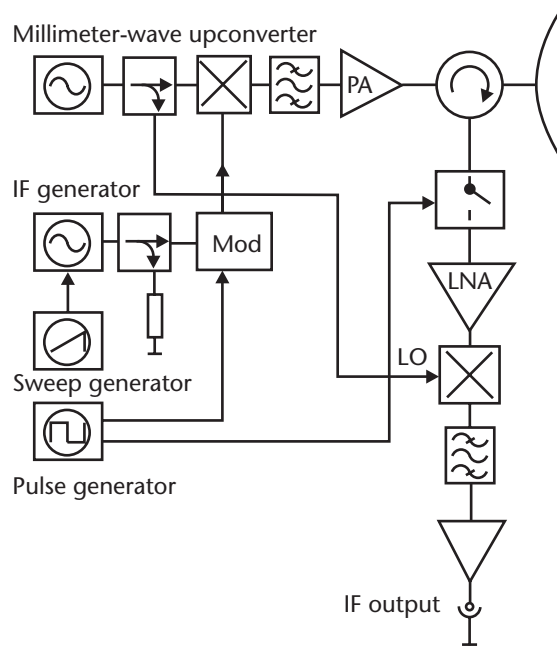


**Figure 6.33** Principal layout of a simple millimeter-wave backscattering test range. A fixed metal reflector is temporarily masked by a rotating cylindrical absorber that has an opening on one side. Thus the TX/RX unit “sees” both the reflector and the background situation without it.

and sidelobe clutter might mask rain effects completely. Besides, the rain profile unavoidably depends on observation angle. If the test angle is high enough to remove such factors as ground effects, protecting the antennas from being covered by snow may be difficult.

The transmitting site has also a coherent RX, which detects the returning signal, but because of pulsed operation, can distinguish between target and clutter echoes. Naturally, the pulse width is very short (30–100 ns) to enable an unambiguous separation of the spatial clutter elements. The dynamic range of the system exceeds 100 dB and the sensitivity of the diode RX at the target is about  $-30$  dBm. A simplified block diagram of one of the test TXs is illustrated in Figure 6.34.

One of the interesting features is the use of switchable beamwidths (from  $1.4^\circ$  down to  $0.2^\circ$ ) together with different pulse lengths (30–300 ns) whereby we are able to “adjust” the apparent three-dimensional clutter volume [36]. Additionally, the hardware supports a pulsed or continuous frequency modulated scheme for a Doppler measurement of particle velocities. A simultaneous recording of relevant weather data supports the test sessions [37]. The preliminary hardware configuration used in the first field trials is illustrated in Figure 6.35. Its measuring range is about 500m when using the horn antennas shown in the photograph. Due to operational safety (outdoors!), the millimeter-wave unit was fed with a separate 15-V dc supply shown in the background. As die-cast aluminum enclosures tend to have inaccurate sealing, some conductive copper tape was required to prevent RF leakage. Figure 6.36 shows the elementary test target, which was used for calibration purposes. Different schemes were tried including completely stationary and various rotating arrangements. After some trials we selected a fixed zinc-bronze plate supported by a plastic tube and surrounded by a slowly rotating microwave absorber cylinder. The upper and lower bearings, just visible in Figure 6.36, caused much trouble due to improper mechanical alignment. As can be seen, there is plenty of empty space behind the calibration target (about 200m) that enables a straightforward cancellation of spurious returns based on their much longer propagation delays. One demonstrative recording obtained with the prototype is shown in



**Figure 6.34** One of the test radar TXs utilizes one intermediate frequency and a combination of linear FM and pulse modulation.



**Figure 6.35** Prototype millimeter-wave radar hardware ready for first field trials. The unit uses a pulse width of 70 ns and has a measuring range up to 500m. Separate 20-dBi pyramidal horn antennas are used for transmission and reception.

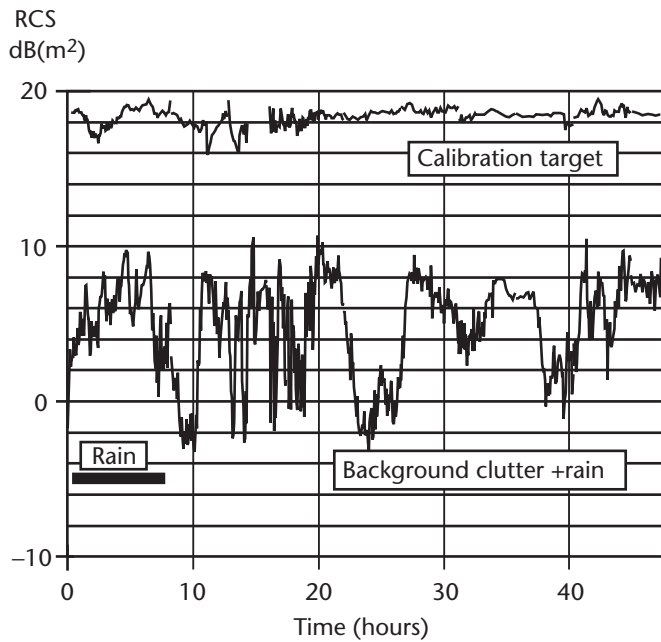
Figure 6.37. However, detailed measuring data and further hardware developments remain classified here, too.

#### 6.9.4 Microwave Telemetry System

Traditionally, measurements from rotating shafts in the industry have been based on, for example, slip rings or an inductive coupling, both of which suffer from mechanical wear and substantial electromagnetic compatibility and reliability problems. Typical parameters of interest within these industrial measurements are shaft movement as a function of time, currents, temperatures, flux densities in rotor area,



**Figure 6.36** Calibration test target in the author's wife's garden. In this case an absorber plate was rotating around the rectangular zinc-bronze plate. Plastic bags provided a temporary weather shield.



**Figure 6.37** An illustrative example of weather results obtained with the first prototype sounding radar. Upper trace indicates effects of attenuation (in decibels), lower backscattering due to rain and ground clutter.

and naturally torque characteristics of electric drive systems. Application areas include but are not limited to paper mills, steel plant drives and brushless synchronous machine tests. A totally contactless transfer of test information from a rotating body is possible either by light, such as laser, or by microwave energy [38], which is further discussed in this example.

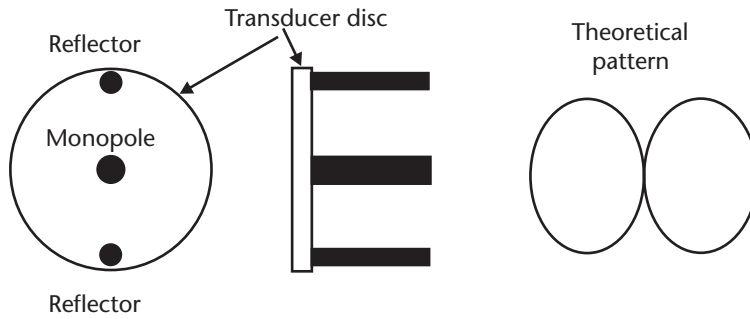
Just like every telemetry system utilizing RF transmission, the proposed equipment includes a small TX with its dedicated antenna located on or in the rotating shaft or rotor, associated task-specific transducers and some interfacing electronics, an autonomous power supply and—located in any convenient nearby place—the RX and signal processing circuitry. The antenna aperture must be visible as seen from the RX but all other TX system components can be installed in an optimum way as required by the parameter under test and as defined by mechanical limitations. The key design feature to look after is the survivability of the whole TX assembly and transducers in the presence apparent g-forces.

The choice of radio link frequency depends, besides regulations set by telecommunication authorities, on such factors as existing interference levels and the preferred size of the TX. Naturally, the higher the microwave frequency the smaller individual elements are usable. If a very low output power is desired we probably can select any frequency we want to but special ISM bands should be used where possible—taking into account limitations shown earlier in Chapter 3. Of particular interest is the antenna aperture area, which can be estimated to be roughly of the size of the wavelength squared—which means, for example, at 15 GHz typically about  $2 \times 2 \text{ cm}^2$ , fitting well inside a typical shaft diameter. However, many drive mechanics can easily accommodate external antennas of tenfold dimensions, which gives some relief in the TX electronics and the actual shaft and rotor structures can be left untouched.

Because the propagation path is typically very short and unobstructed, the TX output power and modulation type are not generally critical. If the RX can be located within some meters we certainly do not need more than 0 dBm—often a fraction of a milliwatt is enough. For worst cross-polarization and very heavy interference a 10- to 20-dB margin is nominally sufficient. To some extent the performance can be improved with an intelligent modulation scheme but very complicated digital TX designs might get physically too large and consume a lot of dc power. Wideband frequency modulation is suitable for many applications and together with a matched RX deemphasis, it can operate reliably far below the 8-dB C/N ratio.

Suitable TX antennas include various waveguide and horn designs, microstrip elements and conventional ground plane systems for the lower frequencies [39]. Circular waveguides can be milled inside a shaft whereas a planar microstrip element should be laminated on the shaft perimeter. The latter choice will create a rotating pattern as seen from the shaft axis normal. Shaft speed or to some extent even rotor position are measurable also from the radiation pattern of a suitable element as spatial amplitude or phase modulation. The very simple case is schematically demonstrated in Figure 6.38 where pattern minima are caused by the two passive reflectors, which were needed due to mechanical balancing. The RX must obviously have an extra AM detector but the conventional tachometer may be left totally out from the test installation. At higher microwaves, a more detailed pattern with also sensible polarization characteristics and a resolution of around  $2^\circ$  can be realized and then measured with an *orthogonal mode transducer* (OMT) feed.

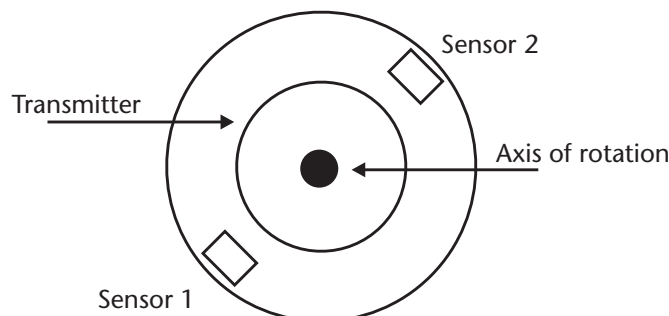
Survivable miniature electronics for the TX instrumentation can be constructed from SMD components and suitable adhesives. Because of small size also the acceleration forces can be kept reasonable. Up to 3-GHz frequency, ready-made



**Figure 6.38** Direct measurement of shaft position with a double cardioid radiation pattern.

commercial tiny TX modules are available at affordable prices. The circuit board layout must be tuned for the specific application in order to avoid any vibration-induced cracks or deformations. If possible, an optimum location for the heavy components is in the middle of the board, which in turn should be mounted in the middle of the shaft or rotor where by radial forces will be minimized. The RX as such has nothing very dramatic in it but must be designed to filter away any out-of-band emissions and have a bandwidth corresponding to the phenomena under test [40].

The torque of an electric machine is normally defined by using measured motor currents and estimated motor flux linkages. Possible sources of error here are saturation in the inductance parameter vector  $L$  and errors in the unmeasurable rotor currents. Specially when load transients occur the estimated motor flux linkage and thus the torque value may have large errors and several high-frequency flux components causing electromechanical vibrations may appear. Other sources of vibration may be, for example, motor asymmetry in the three-phase system, or the mechanical setup may have such a critical frequency that vibrations occur at sudden load changes. The common method for torque measurement is so-called torque shaft where the torsion angle is proportional to the applied torque. During a fast torque transient this shaft acts like an elastic spring and high frequency transient effects are difficult to capture. An improved method measures angular acceleration for example with a wheel containing two acceleration transducers, mounted as shown in Figure 6.39. A “pure” angular acceleration value is obtained if two sensors are used at opposite locations related to the Earth’s gravitational field because most real electric drives operate in a horizontal position.



**Figure 6.39** The measuring wheel layout. The Earth’s gravitational effects have been compensated with two identical but opposite acceleration sensors.

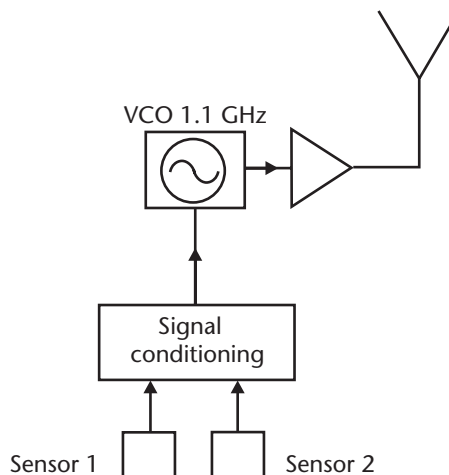


A key feature of this system is its capability to operate without any mechanical contact between the rotating shaft and the stationary laboratory or factory environment. This is of great benefit in torque transient measurements because now the whole motor assembly can be installed in a normal fixed mount with no need for flexible joints or spring loaded bolts. In fact, measurements can be made in the normal operating configuration of the particular drive.

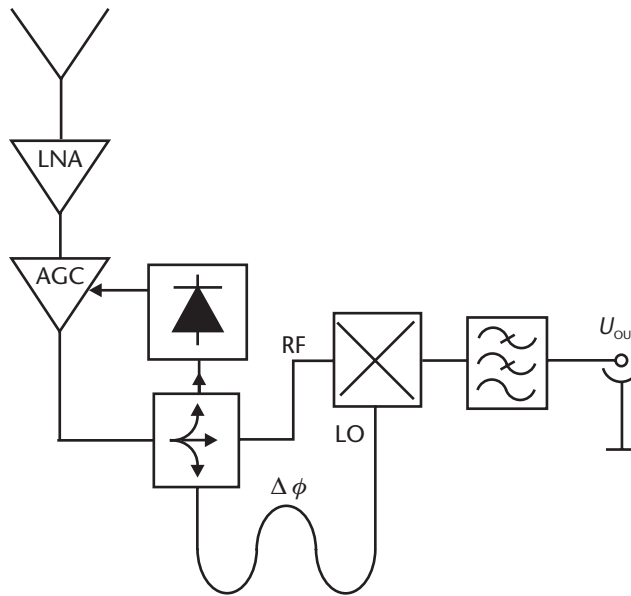
The very simple analog electronics of the TX are shown in Figure 6.40. Wide-band frequency modulation with a deviation of 15 MHz is used at around 1.1 GHz in order to get improved noise reduction. TX power, fed to a radome-covered ground plane antenna, is below 10 dBm. The dc power supply inside the TX is a single miniature 9-V dry cell battery. The respective RX block diagram is in Figure 6.41. After an interference limiting bandpass filter and an AGC amplifier we have an FM-detector, which actually operates as a phase demodulator where the fixed-length delay line “converts” frequency deviation to phase alteration; see Section 4.3.3 for specific operational details. Also the RX had to be battery-powered for operational safety in a power electronics environment.

All but the very low-frequency part of the mixer output spectrum are filtered out with the lowpass filter before feeding the result to a data logger, which was replaced in preliminary tests with a digital oscilloscope. Typical RF input power varied between  $-40$  and  $-10$  dBm and the low frequency output voltage sensitivity was  $4.5 \text{ mV/ m/s}^2$  without any additional amplification. The total torque conversion factor was  $0.32 \text{ mV/s}^2$ , which equaled  $0.15 \text{ Nm/mV}$  in that particular physical installation.

System performance was briefly evaluated in a case where an ac motor was started with a load, which was formed by a dc generator feeding some resistors. Torque transients were caused by the frequency converter and measured correctly through the developed microwave link. Already preliminary tests indicated that the performance of the speed-adjusting algorithm in the converter was far from optimum and severe mechanical shaft oscillations lasting about 3 seconds were recorded. The developed telemetry system proved to be fully functional and it was thus put into operational use.



**Figure 6.40** The TX included two transducers, a preamplifier, and a voltage-controlled 1.1-GHz oscillator followed by an output stage and an antenna tuner.



**Figure 6.41** A delay line fed mixer was used to convert frequency modulation to phase changes and further to a low frequency in the RX, which had normal AGC and filtering circuitry.

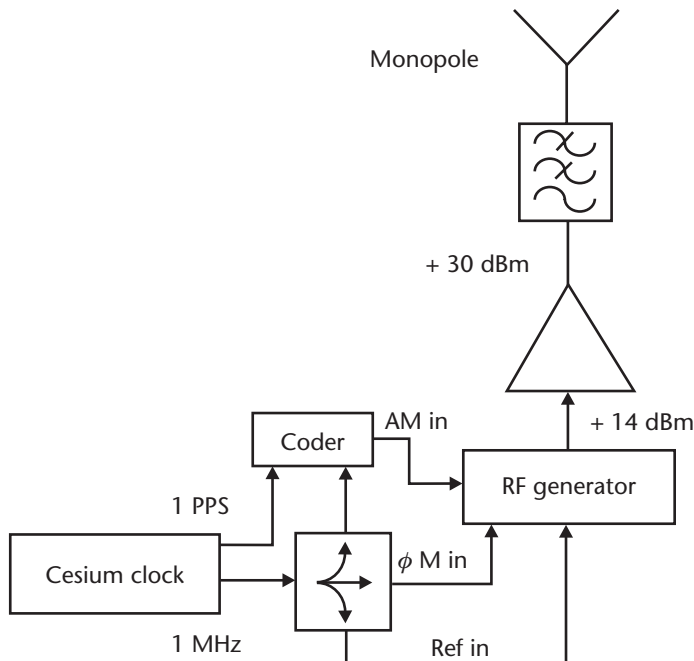
### 6.9.5 UHF Time and Frequency Reference

Our final example of TXs and RXs discusses a system for precise frequency and time dissemination through microwaves. The motivation for such a design is as follows. Many modern mobile communication systems require extremely accurate oscillators forming either their transmitting frequencies or synchronizing their bit patterns or both. Typical uncertainties include the  $10^{-9}$  level of GSM networks and the super performance frequency agile transceivers in military aircraft approaching  $10^{-11}$  for about 2 to 3 hours. The obvious problem is how to verify the performance of individual TXs or RXs without being limited to the artificial laboratory environment or relying on the highly filtered data available through the respective network management computer. Particularly systems where the base station tries to adjust the mobile to follow its frequency or synchronization pattern, may provide misleading interpretations due to the “rubber band” effect. As long as the geographically large control loop is closed and does not reach its adjustment limit the system will work but we do not know the individual offsets or delays. Some observed related problems are documented in [41, 42].

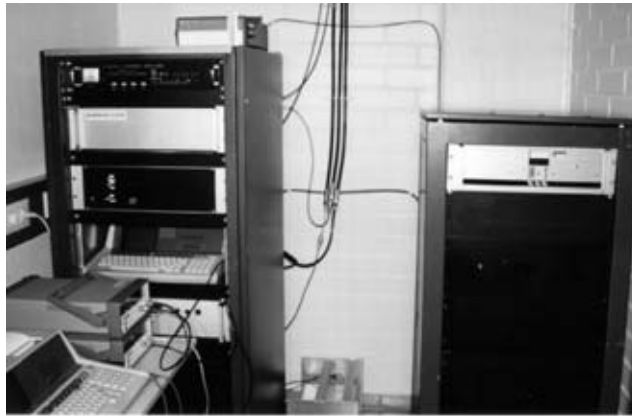
Although a number of task-specific commercial time and frequency transceiver systems exist, they are either at least partly based on the utilization of satellites, which are generally out of the direct control of the user segment, or on very dedicated expensive hardware. The attempt here was to evaluate the possibilities of using less costly COTS devices and to find both a system architecture and a simple modulation or coding scheme providing a straightforward way to test mobile oscillators. Typical applications would be analyzing the frequency control characteristics in a GSM1800-system or measurements of *frequency hopping* (FH) radios installed in tactical vehicles. Both of these radio designs produce a signal that is hard to track from, for example, a stationary reference site.

The idea has been tested with a cesium-locked TX–RX system designed by the author. The frequency standard locks the carriers and simultaneously modulates them with, for example, a combination of a sine wave and a 1 *pulse per second* (PPS) pulse train. Thus we can in principle measure frequency deviations, estimate the Doppler shift and test the synchronization as well. The enhancement in accuracy is mainly based on the quasi-continuous characteristic of the test signal when compared to the complicated hopping or spread spectrum principles employed in the systems under test and being the only thing to track if a measurement were tried from the stationary site. Besides, a very high momentary resolution can be obtained from the UHF carrier itself, which has a roughly 1-ns cycle time. The mobile instrumentation includes a cheap commercial communications RX, a time interval counter and a phase comparator. If accessible in the transceiver under test, its own reference can be compared (e.g., with the regenerated 1-MHz signal), but a fractional measurement is naturally also possible.

The TX system is presented schematically in Figure 6.42 and a view to the equipment room is given in Figure 6.43. The left rack contains time-code RXs, a rubidium frequency standard, a computer-controlled precision delay line, and the associated computer. Two time interval counters are positioned sideways on the small table. The rack to the right contains standby batteries, battery chargers, and one old but very reliable cesium oscillator. An unfortunate second cesium clock [43] was used as the main reference in an attempt to be compliant with the high stability requirements of military radios. Quite soon after the very first field trials the stationary brand new cesium in Figure 6.44 failed due to tube collapse. The manufacturer needed 4 months to repair the unit and it was lost in transport by the local dealer for several weeks. Many of the threats outlined in Chapter 1 and Section 4.5 suddenly popped up.



**Figure 6.42** The test TX includes a cesium clock, a UHF wideband generator and a coder, which can be omitted if the 1 PPS signal can be used on the mobile platform as such.



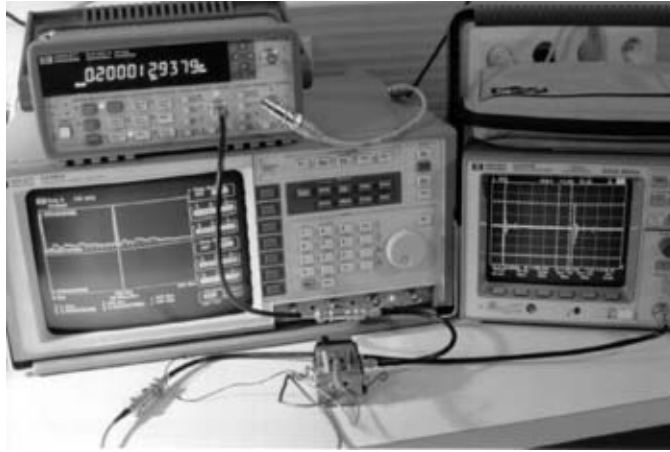
**Figure 6.43** A view into the TX room. An older backup cesium unit is visible to the right of the main rack. See text for more details on the equipment.



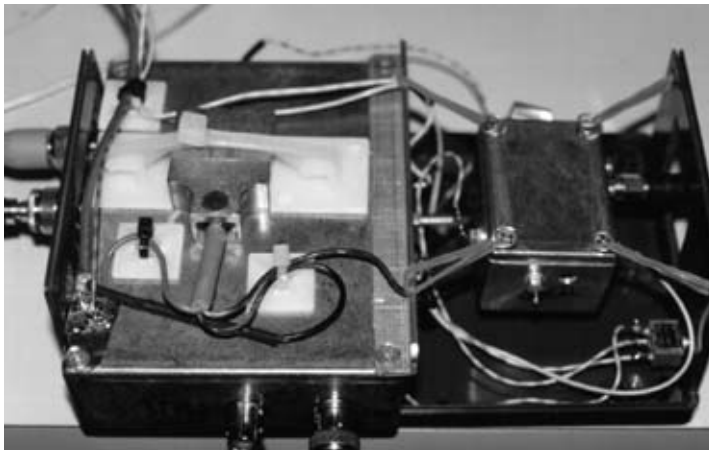
**Figure 6.44** The most unreliable part of the test installation was the new commercial cesium clock, the prolonged repair of which caused a 4-month delay to field trials.

The measuring equipment on the mobile platform uses a communication RX; an ovenized crystal oscillator, which was documented in [44]; a counter and a phase comparator. The oscillator was initially tested to find out its performance under shocks and vibration. That arrangement is shown in Figure 6.45, where the oscillator has an acceleration transducer on its back. The idea was to synchronize the oscilloscope (right in Figure 6.45) with the frequency analyzer (left) in such a way that the two traces—one for relative frequency jumps and the other for acceleration—could be compared. After these tests a spring-loaded suspension was found necessary. One of the first prototype assemblies is illustrated in Figure 6.46. The reader can note that the rubberband suspension could not react in similar way regardless of shock direction (i.e., left-right and up-down).

The challenges of instrumentation layout in the test van are demonstrated in Figure 6.47 whereas the most vital elements are better visible in Figure 6.48. In field experiments, lengthy and tangled temporary cabling is not only a nuisance but can cause difficult-to-locate interference problems. Note also the PC on top of the communications RX. That position was perhaps the worst thinkable and caused severe disturbances. Most of the tests performed relied on a comparison of 500 kHz, 1 kHz, or the second tic signals. The latter is simply a train of pulses that appear



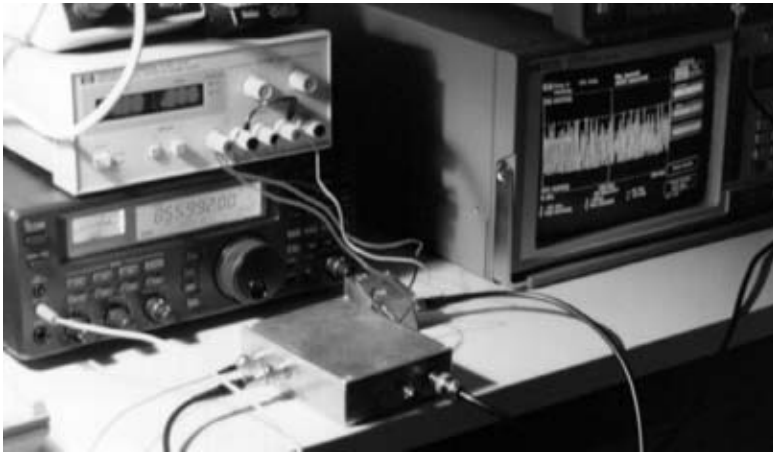
**Figure 6.45** The mobile crystal oscillator was put in shock tests. Here the unit is seen with an acceleration transducer on its back.



**Figure 6.46** Elementary shock absorbing was built from flexible rubber bands and foam. The phase lock circuit is in the left box, under the acceleration transducer.



**Figure 6.47** A view of the test vehicle interior with the time interval RX and the analyzer in the upper corner.



**Figure 6.48** A close-up of the receiving installation in the lab environment. The phase lock unit is in front of the communications RX.

exactly once per second and have some predefined duration, for example,  $1 \mu\text{s}$ . They are often formed in logic gate circuits and therefore assume *transistor-transistor logic* (TTL) or *complementary metal oxide semiconductor* (CMOS) voltage levels. Because the environment is usually very harsh, the key requirements for the “production version” of the RX instrumentation are small size and weight plus a very reliable construction, which is often a synonym for simple as was indicated in Section 1.6. These were partly not met by the prototype, mainly due to the continuous need of reassembling the equipment into the van for each separate test case.

Numerous tests with both airborne and terrain vehicle mounted RXs indicated a very satisfactory service volume with only  $+30 \text{ dBm}$  of EIRP. This is partly due to the limited 20-kHz IF bandwidth found suitable for applications where epoch information is not needed. It is also assisted by the low NF of the radio amateur RX. Figure 6.49 is a radar plot from northern Finland showing a test flight with a jet trainer. The top and bottom turning points indicate the sites, about 150 km from



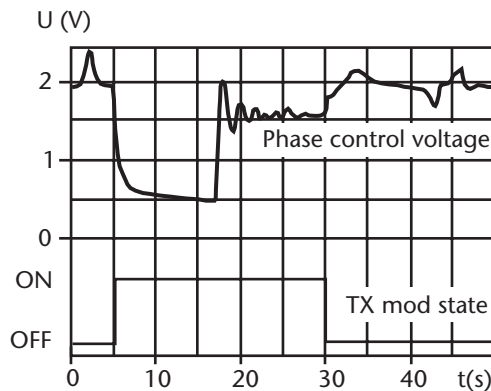
**Figure 6.49** This radar plot shows the expected operating range obtained at 3,000m with just a simple ground plane antenna at the TX site.

the TX, where the signal went unusable. The altitude was 3,000m and the aircraft antenna gain was  $-10$  dBi. For more aggressive flying, needed, for example, when evaluating equipment characteristics important in ground support tasks (high G values, turbulence), tests have been conducted around 150m above average terrain level. Typically the measuring equipment could be used up to 10 km but over hilly areas only to roughly 3 to 5 km. From an RF engineer's point of view, these true field tests with operational aircraft were very educational. Particularly it became clear, that aerodynamics and mechanical airworthiness requirements must be followed in detail when configuring any temporary physical items, for example antennas or simple fasteners. On board military jet aircraft, only precertified hardware is acceptable. This yields to using antennas, which have very low efficiency and often unknown radiation patterns, for example.

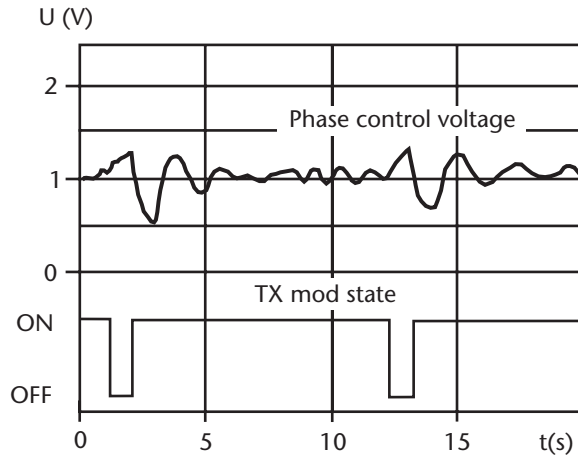
The first simple configuration tried was to use just pure sine wave modulation and from that to extract at the RX site both a frequency reference and time interval information. For this system the most obvious problem would be a momentary loss of carrier due to, for example, heavy shadowing or interference. Because we have to regenerate all the timing information after the RX, a phase lock seems to be an attractive alternative. Its time constants are crucial for the operation under dynamic conditions but are more or less a compromise between the steady state noise characteristics and partly drift-like phenomena as well.

To start with, Figure 6.50 shows the rough test where the carrier was kept silent for 1 minute, which should simulate a reasonably severe propagation problem or even equipment turn-on. As is seen, the design, which was optimized for steady state performance, requires about 17 seconds to relock, and what is still worse, the frequency error increases immediately after the switching of carrier power. Apparently this is avoidable with the addition of a fast-hold diode switch.

In a more realistic scenario we have to accept a continuous stream of interruptions occurring at random intervals. This has been tested with a waveform like the one shown in Figure 6.51, where the carrier appears and disappears consecutively. The state of the lock just before the carrier turn-on is almost as critical as the accumulated frequency error with reference to the ground oscillator. The typical pattern of 1-second dropouts occurring approximately every 10 seconds indicates that the



**Figure 6.50** The high stability phase lock unit unfortunately requires almost 17 seconds to stabilize itself after the carrier has been momentarily lost.  $U$  is the oscillator control voltage of the PLL circuit, expressed in volts.

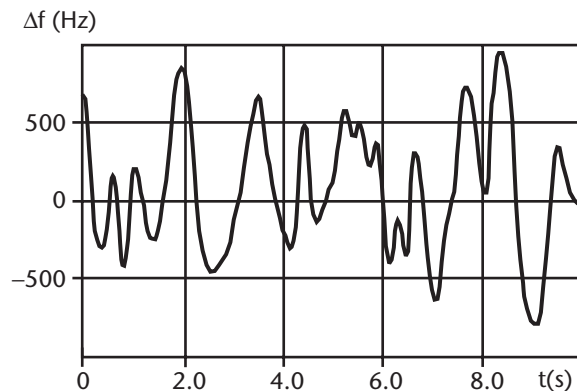


**Figure 6.51** If the carrier is frequently interrupted, both the ON (9 seconds) and OFF (1 second) time constants become important. It seems adequate to allow 3.4 seconds for stabilization.  $U$  is the control voltage for the PLL oscillator.

selected circuit would be more than 30% of the time in an unlocked state. Very obviously the PLL's comparator and the associated VCO were not properly aligned.

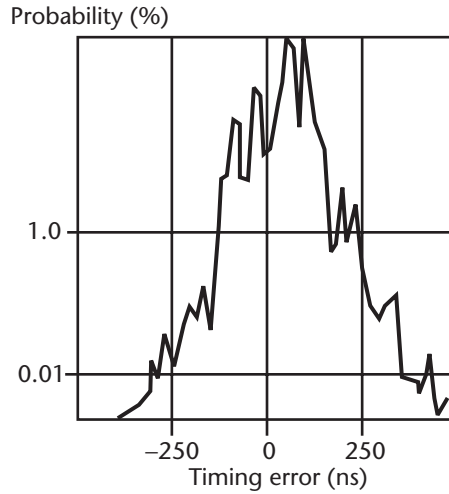
If the PLL time constants were adjusted or the RF path could be improved (e.g., by using space diversity reception), we would be able to do much better, as can be seen in the frequency profile in Figure 6.52. The system no more noticed the absence of carrier if the duration is less than 130 ms. Measured at an average RX input level of  $-100$  dBm the momentary, single sample frequency error was here less than 1 kHz (no averaging).

When the UHF path satisfies the prescribed conditions, quite nice performance figures can be obtained by using just 20 kHz of bandwidth. The momentary timing error as obtained from the phase locked crystal oscillator stays at around 500 ns (no averaging), which is clarified in the probabilistic timing histogram Figure 6.53. The test RX input level was kept at  $-90$  dBm but the signal came via a realistic propagation path having conventional industrial interference and temporary multi-path problems due to adjacent moving objects.



**Figure 6.52** This plot shows PLL oscillator frequency difference as a function of time. A reduction of the carrier OFF time improves the performance because the dropout is not visible when it is shorter than 134 ms.

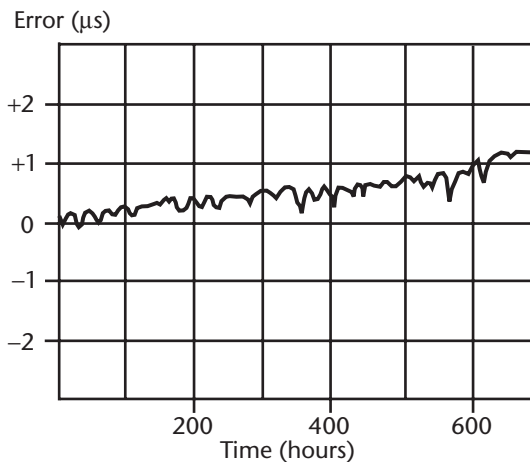




**Figure 6.53** Phase-locked 1 PPS error histogram at  $-80$ -dBm RX level. The receiving bandwidth was only 5.5-kHz, FM modulation.

Long-term tests are not very probable with moving oscillators staying within reasonable geographical limits. However, as the necessary instrumentation was there, a several 100-hour recording was taken, this time with the RX stationary and using the system of [45, 46] as a reference. Obviously we see in the plot of Figure 6.54 the error both in the TX phase lock and additionally the drifting of the time interval counter’s trigger level as a function of elapsed time.

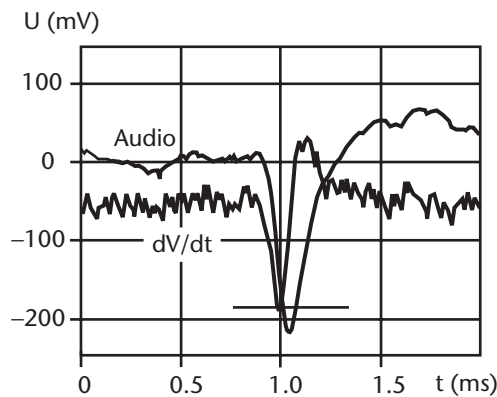
After a satisfactory level was achieved with the continuous FM scheme, a trial was made to include absolute timing information to the transmitted signal as well in an easy form for the RX. The main problem was to keep the signal within the bandwidth allocated by the telecommunication authorities and at the same time maintain the highest possible timing resolution. It turned out to be possible just to pulse-modulate the test generator so that the carrier was suppressed by  $-80$  dB for  $50 \mu\text{s}$



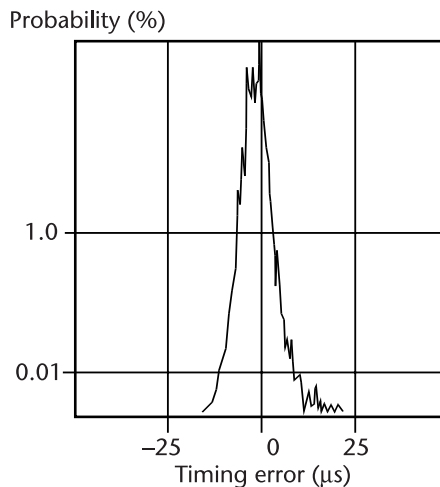
**Figure 6.54** In the long run, with averaged samples over 100 seconds and a plotted sample spacing of 10,000s we observe the TX system drift in this 1 PPS comparison recording at  $-90$ -dBm RX level, FM demodulation, and 100-kHz IF bandwidth.

and to detect this from the same audio output where the previous phase locking information came out. To have the sharpest turnover point for triggering, a careful evaluation of the RX's AGC performance was carried out. The optimum result is shown in Figure 6.55 as the first derivative of the input waveform. Naturally, due to the switching nature of the pulse-modulated signal, we had to use the widest available bandwidth of the RX, which caused a 10-dB increase in the predetection noise floor.

The communication RX was able to simultaneously detect both the quasi-continuous FM signal and the 50-ms absence of the carrier to form a separate 1-MHz frequency output and true 1 PPS information. However, partly due to the required increase in the IF bandwidth giving more noise power and partly due to occasional spikes in the audio waveform, the uncertainty became a compromise showing a momentary worst-case value (99% of cases) value near  $10\ \mu\text{s}$ , as is clarified in Figure 6.56.



**Figure 6.55** If using simple carrier keying, an optimum time interval reference point was found at the point of highest rate of change of the RX audio output. RX level was  $-80\ \text{dBm}$ , FM detection, and 100-kHz IF bandwidth.



**Figure 6.56** The cost of epoch information is the increased time interval uncertainty.

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# RF Measuring Instrumentation

Many system engineers are at least a bit experimental in their inherited character, and thus severe objections are seldom heard related to testing of created concepts—either in the lab or on the field. Although this work closely follows the practices and principles of development done on the lower hierarchical layers, some specialties are frequently encountered. This chapter tries to give an overall impression of how to define a system test setup, which features of test equipment might be vital, and how to decide between commercial ready-made test systems versus one's own compilations. A few fords about computers in system-level tests were considered necessary as well. Finally we try to highlight practical test arrangements for a variety of RF systems and tasks.

## 7.1 Defining a Test Setup

The designing of required verification and performance tests should start briefly after the initialization of the main project or effort. Last-minute actions tend to be hasty just as in everyday life, and also confused and less well-prepared, which will easily yield to omissions of important factors or expensive mistakes [1]. Particularly annoying is the fact that very often unique RF systems need test equipment performance, which is not readily available even in advanced laboratories. Instead, these capabilities must be developed either in-house or by subcontractors or project partners. Such activities need considerable time, often measured in years rather than months and may be faced by problems of equal difficulty when compared to the main path. Just as an example we can think of a highly precise measuring system for defining the modulation depth of certain TXs. How can we make sure, and make the final user or customer believe, that the performance target is really met, if we do not have a signal source capable of providing us an even better reference?

Although many of the individual instruments used for system testing are similar to those in the component lab, large-scale testing involves a couple of additional challenges [2]. Maybe the main difference might come from the geographical dispersion of the system under test. TXs may be on board a ship sailing 1,000 miles away or perhaps currently on their way to the outer skirts of our universe. Already a separation of some hundreds may provide enough problems to solve. Galvanic potential differences, electromagnetic interference, or the microclimate in the workshop can make our test instruments crazy. Of course, just the plain field test environment may hamper our attempts. The tundra scenery in Figure 7.1 is far from the most adverse, but already sets specific requirements. What is the phase noise performance of our



**Figure 7.1** Making RF measurements far out in the tundra is exciting but sets high requirements for the instruments. Here some satellite field strength recordings should start.

spectrum analyzer at  $-30^{\circ}$  centigrade? Do we get frost in the waveguide when it is brought out on top of the roof?

Interoperability of commercial test instruments may look self-evident, but unfortunately it is not. Assume the relative simple and straightforward case of creating a coherent test environment for a jammer/surveillance RX-pair. Individual RF and baseband signal sources and respective analyzers have all their internal oscillators running at their own pace. Physical connection interfaces are normally provided for the input, output or combined in/out of the reference, but quite certainly at least one instrument does not have the suitable frequency or there is only an output. Alternatively, perhaps the accepted or generated waveform happens to be TTL style, which typically means a square wave between about 0 and 4V, when all other devices use sine. Another hassle may be the differences in absolute power levels even if measured very carefully and following appropriate practices. In a setup of three instruments, say a spectrum analyzer, an RF signal generator, and a power meter we will very obviously have three opinions about 1 mW at 20 GHz. The task is further complicated by nonequal nominal reference output levels chosen by various vendors. Where one device gives 0 dBm the other wants  $-10$  dBm. More specialized quantities, for example modulation depths, tend to show even larger discrepancies.

One of the main tasks of the system engineer is to define and describe the entire test environment in a “closed form.” First we have to know which parameters have to be measured and what is the required uncertainty [3]. An environmental test facility is required if, for example, the operating temperature range is vital. RF performance measurements may call for an anechoic room, a shielded room, or an open test range. Size and weight of system elements must be carefully evaluated so that the test facilities can handle their coming task. Comprehensive and long-term tests are usually supported with computer automation. Interfaces and software requirements must be outlined early enough. Particularly important may be the interfacing of the actual system under development into the measurement controllers. Although commercial test instruments may or may not have such strengths as IEEE-488 or serial interfaces, our missile tracker does not. Means to overcome this have to be defined so that the work on the test field can later progress fluently.

## 7.2 Typical Test Instruments for Systems

The most often required RF test instruments in system development and testing are the following:

- Spectrum analyzers;
- Vector analyzers or VNAs;
- Measuring RXs;
- High-speed oscilloscopes;
- Signal generators (baseband and RF);
- Modulation analyzers;
- Frequency counters or frequency domain analyzers;
- RF and baseband switches and switch matrixes;
- Dummy loads, terminations, and other transmission line hardware;
- Controlling computers (typically PCs).

Task-specific complete test-sets are available for such tasks as radar simulation [4] or the checking of radio navigation systems. Very often, however, also they incorporate commercial instruments as functional modules. Figure 7.2 illustrates a relatively simple but not so cheap homemade measuring system used for an anechoic antenna test range. The main instruments include a spectrum analyzer, a microwave synthesizer, a dedicated bus-controlled switch matrix, and suitable hardware for turntable functions. An old workstation, which provided the man-machine interface, is barely visible in the background.

When selecting a test instrument either to purchase one or just to pick it from the lab shelf, consider the following topics:



**Figure 7.2** A relatively simple anechoic antenna range instrumentation setup showing a microwave synthesizer, an IEEE-488-type switch matrix, and a high-end spectrum analyzer.



- What is the parameter to be measured and with which instrument do we get the result most fluently but within required accuracy limits?
- Is there a risk of external interfering signals—do we need selectivity?
- Is the measurement of scalar or vector nature—do we need coherence?
- Do we need information in the time or frequency domain or both?
- Do we need multiple input signals simultaneously?

Generally, spectrum analyzers tend to be quite helpful devices, because we can use them to get a continuous overall view about what is happening in our system and additionally a glimpse of the surrounding world as well. If their selectivity or level measuring uncertainty is not adequate, we can supplement the test set with an RX. It is very common that one signal source is not enough. One or several more are needed to simulate jammers or friendly users on adjacent frequencies. Here considerable savings can be earned by selecting some generators from the “basic” class and only one or two with the extreme performance money can buy. If prestige reasons do not prevent it, consider checking the sales lists of refurbished instruments if you have to set up a whole arsenal of simple generators, for example. A general principle might be to look a bit forward, though. Even if we do not need a feature in the current project, what happens next year? Common sense has to be used naturally, because all the available options are surely not necessary for everyone. Maybe the point to consider is those factory-only choices, which you cannot order later.

### 7.3 Ready-Made or Tailored

Major test instrument manufacturers offer complete test setups for certain tasks and additionally offer turnkey products according to customer specification. Using such a solution in supporting own system development may be beneficial, but it is not very cheap and can sometimes cause considerable difficulties.

If a measuring principle or a set of measuring practices has evolved long enough, it may be so mature that it is used in production testing almost on a routine basis. Cellular phone manufacturing is a good example of this. Very comprehensive and totally automatic test systems are available for such purposes but we are not actually talking about totally new development projects—if it is not a new phone according to the same standard. Another example of complete test systems is EMI compatibility checking, for which very expensive but at the same time high-quality instrumentation is for sale. Radar simulators or testers have been on the market since the 1950s, but again mainly incorporate features and functions already implemented somewhere. This is also true of EW simulator systems. If we are really working on something new, tools for the development seldom appear in the catalog.

Tailored test systems, designed from customer wishes, are a business for the instrument manufacturers [5]. They make profit. The reason to use a subcontractor might be lack of manpower or time in our main project or perhaps because a specialist is not at hand. In-house test system compilation work is not cheap and you have to know a lot, but it also gives as a byproduct fruitful information for the main path. Besides, managing is normally but not always a bit easier if we omit one additional gap between two enterprises. If nobody in the main project team knows enough

about test software, then an external vendor might be a better alternative. If going to a vendor for an entire test system, check the following issues:

- Write your own specifications—do not immediately take what is offered;
- Take care that they do not sell you expensive features that you do not need;
- Compare the offer with the vendor's standard list—too close-looking?
- Try to evaluate the vendor's capability in your specific area of activity—ordering a radio navigation tester from a keen company whose main financial interest is in petrol pumps and processing software may not give desired results;
- Require an acceptance test plan from the vendor for the measuring system—and evaluate it;
- Require a complete warranty to start from test system acceptance—otherwise individual instruments may have reached their limit far before startup;
- Require adequate financial sanctions to cover also that very unfortunate and difficult case where no test system is available at the time when the main project needs it;
- Go through the entire acceptance tests, make appropriate notes on noncompliances and do not pay the bill—even a fraction of it—unless faults are corrected.

## 7.4 About Computer Control

Computers and programs in them can be essential elements of a modern RF system. Comprehensive systems testing can also make extensive use of computer-controlled measurements but only if the software and the computer interfaces are properly defined. On the other hand, a poorly outlined and hastily compiled bundle of programs and computer hardware will lead to endless iterations, and the focus is soon transferred from the initial system to be tested to repairing the software bugs and noncompliant interfaces [6]. In the past, test system control was relatively easy, because almost everything was made with BASIC-type languages and data transfer to PCs was not required as there were none available. Two major challenges appeared on the scene around 10 years ago. Modern test system control must run on PC platforms and in practice code must be written in one of C-derivatives for dedicated purposes if drag-and-drop-type building blocks will not work. Test equipment manufacturers agreed on *standardized commands for programmable instruments* (SCPIs), which was perhaps intended to enable better interoperability of devices from different vendors. In fact, however, many test system engineers consider the improvement to be at least partly masked by the increased complexity of very simple operations. For example, when getting a frequency value from a counter required in the past two lines of BASIC code, we are now forced to use at least 10 or even 15 (equivalent) lines for the same action.

Defining a test system software package should not start from what is commercially available or possible or what is offered by a vendor but rather from the fundamental requirements of our own project. We must first find out what is to be measured and how often. Parametric testing is one of initial questions, too. Maybe the following issues are worth considering:

- Data should be gathered at a speed matched to the phenomena under test. If time constants in the physical environment are several seconds, there is no point in sampling at 100 kHz. Test system designers tend to exaggerate here as a rule. Most high frequency “phenomena” in the obtained data are just noise and interference.
- A computer can collect huge amounts of data in a very short time. Think about the analysis before you have all those numbers in your hands.
- A nice test software user interface with fancy graphics and live animations may look attractive. However, time used to develop those features does not help you in solving the main measurement challenges. In addition, delayed software delivery may seriously hamper the actual system testing.
- The more finesses and unnecessary features a computer program has, the more likely it collapses [7]. Often the cycle time between successive measurements is increased as well.
- Most commercial test software environments create measurement routines that surely tilt—sooner or later. Be prepared for that and do not try to collect too long test runs as one file. The ultimate reason may be in the software itself, in the data acquisition cards or in the main PC. Knowing the guilty will not bring the lost data (or time) back.
- One decibel is often good enough resolution and one-tenth can already be in analog noise. Adjust the documentation and presentation resolution accordingly and reject the six-decimal default suggested by the software or coming as output from the instrument’s internal processor.
- Try to provide easy connectivity in the primary system under development (i.e., radar or communication link) so that the test computer can have a direct path to those vital parameters as well. If frequent human interaction is required, the whole idea of computerized testing might be jeopardized.
- Keep the test software designer—if in-house—paid well and have his or her coffee mug filled regularly. If he or she quits, you and the project will be soon in serious trouble, because software documentation, no matter how detailed it may be, never replaces hands-on experience.
- Allow adequate time for software testing before attacking the primary system. Doing both simultaneously will be ineffective and may spoil a lot of the team spirit as well.

Besides the next section, the reader is encouraged to have a look at the adaptive antenna case in Section 5.5.2, where extensive software development was necessary and results very promising. Good team spirit and general management of affairs combined to individual talent yielded to successive task-specific software.

## 7.5 Examples

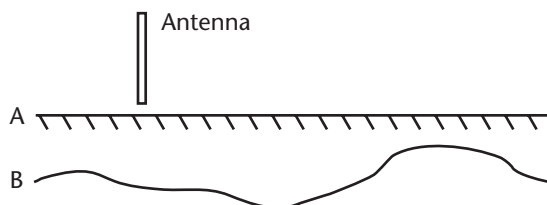
This section describes a couple of real test systems that have been designed for RF measurements. The intention is to show the multitude of application possibilities but also highlight a number of problems that may turn up.

### 7.5.1 Estimating VHF Ground Conductivity

Since the early 1930s a number of RF systems have been proposed, designed, and taken into operation in which the physical ground or the soil is used as the electrical ground plane for their antenna arrangements. This is particularly true for such frequencies where a metallic structure would have been too large or heavy. Comprehensive tests—see [8]—and extensive measurement campaigns have been carried out to get data about the real characteristics, conductivity and dielectric performance of the Earth’s outer cover. First, however, many of the efforts have been focused at the HF bands, and second, very little information is available on the short-term or small-scale spatial variations observed regardless of frequency. This has been partly based on the fact that measurements have been mostly done from low-flying aircraft in the days before GPS and thus the spatial resolution has been poor. A schematic cross-sectional view of the problem, using a single vertical element as a test case, is illustrated in Figure 7.3. The real situation, though, is three-dimensional and depends very much on the frequency and soil parameters. Basically, however, the deviation from predicted performance is caused by the too-large vertical distance of the mirror element(s).

Today’s most critical installations, which fully rely on the physical ground in their pattern-shaping functions, include but are not restricted to typical military or commercial VHF radio communication networks [9], and many of the simple air navigation ground devices. For example, the most widely used precision landing system both for civilian and basic military aviation, the *instrument landing system* (ILS), is known to suffer from ground plane imperfections in its glide path (around 340 MHz) formation process as demonstrated in [10]. This is mainly seen as bends and seasonal changes in the apparent path angle and the deteriorating effects may come from a wide physical area.

Recent, yet preliminary observations made by the author suggest that also the azimuth guidance information, provided by the 30- to 80-m-wide VHF antenna array (localizer in aviation terms) at the other extreme of the runway, may be affected by temporary changes in the nearby spatial ground conductivity [11]. The principal reason is assumed to be the considerably huge size of the “coupled” ground. Normally prevailing constant conductivity as a function of lateral  $x$  or  $y$  coordinates—a situation typical for just one local monopole or similar design—can no longer be applied. Affirmative results have been obtained by nonsymmetrically wetting the foreground of such a system (kind cooperation from the airport fire brigade is deeply acknowledged) and simultaneously recording the navigation signals. These phenomena are of great practical importance for the safety and regularity of air traffic. The possibility to measure the effective or apparent conductivity—though



**Figure 7.3** Many antennas prefer perfect ground conductivity at the boundary (A) but in reality the effective level might be deeper by (B).

on a completely arbitrary scale—at selected VHF frequencies was thus studied. Tedious preparations or expensive installations were not acceptable but still data was desired as a function of the two-dimensional distance along the surface of interest.

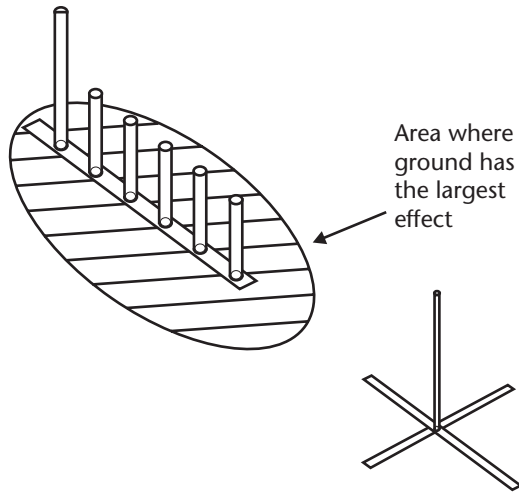
The test system, as shown in Figure 7.4, included a ground plane Yagi antenna, discussed in [12], a tunable oscillator operating as a TX, and a small monopole with a diode detector as an RX. The human interface was just a cheap digital hand-held multimeter having a 0.1-mV resolution but it could have been a portable data logger as well. The TX was equipped with a directional bridge and two additional detectors to monitor the proper impedance match, which nevertheless in preliminary experiments did not turn out to be a particular problem. A view of the test site geometry is in Figure 7.5.

An artificial ground plane, made from three or four half-wave copper ropes was provided for the Yagi's radiating element but the directors did not have anything else except the real ground to be tested. This was the key feature of the principle because the impedance of the transmitting antenna element stayed fairly constant with a return loss always better than 10 dB and thus the real antenna input power was not changing. The gain and main-beam elevation angle were both affected by the nearby environment mostly beneath the Yagi's directors. For calibration purposes, a flexible metallic chicken hatch, extending about one quarter wave to each side of the Yagi's center line was temporarily first mounted and the respective relative field strength and, optionally, beam tilt angle recorded at a distance of about 20 to 30m, as considered suitable in [13]. A comparison to respective values obtained without the hatch gave an indication of the vertical location of the effective conducting soil—how deep and how good (compared to the hatch) it actually was.

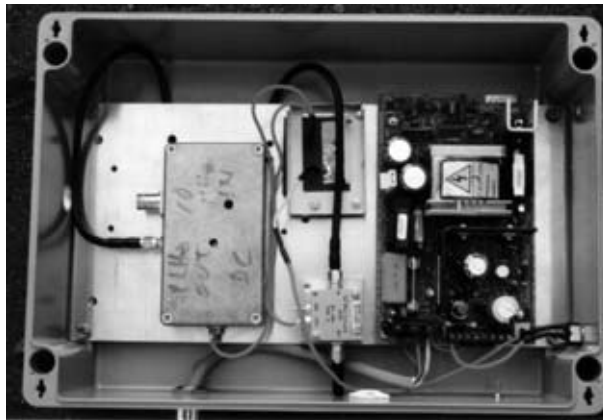
To be able to rapidly evaluate the feasibility of the method, a prototype installation was assembled from off-the-shelf hardware. Wood was initially tried as supporting material due to its favorable strength/weight ratio but an even more flexible solution could be based on nylon rope and fasteners. Standard VHF building blocks were extensively used for the generation of test port power as can be seen in Figure 7.6. The VCO was tunable up to 118 MHz, but alternative similar modules are available up to 3,000 MHz as suggested in [14]. It is highly recommended that battery power be selected instead of mains supply not only due to increased safety but also because of the much easier setting process and the avoidance of pattern-disturbing ac feeder cables. Our choice was a four-arm monopole antenna for the receiving site, as can be noted in Figure 7.7. There was enough dynamic range to allow for a test track length exceeding 30m even under worst-case conditions (lowest conductivity) and the built-in ground plane of the monopole effectively reduced any disturbing phenomena, which might come from the receiving site's spatial ground characteristics.



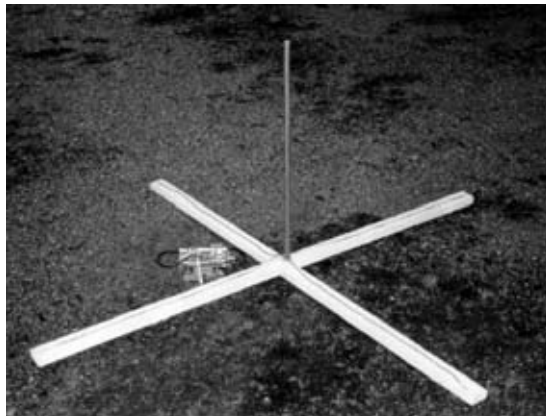
**Figure 7.4** The block diagram of the whole measuring chain includes a VCO, an amplifier, a directional bridge, and a display. The propagation path is formed between a ground plane Yagi and a small monopole.



**Figure 7.5** The area where the Yagi's directors are mostly affected by the ground conductivity is shown hatched.



**Figure 7.6** The TX included a VCO, a medium PA, and two separate power supplies. The directional bridge was not an integral part of the TX.



**Figure 7.7** Just a small monopole antenna and a diode detector (with integral display for dc output) following a 24-dB RF amplifier was enough to record the relative field strength.

One RF amplifier (20 dB) preceded the diode detector, which in turn fed a simple integrating DVM. As such, the complete setup was very compact to be lifted up for the main beam tilt-angle measurement.

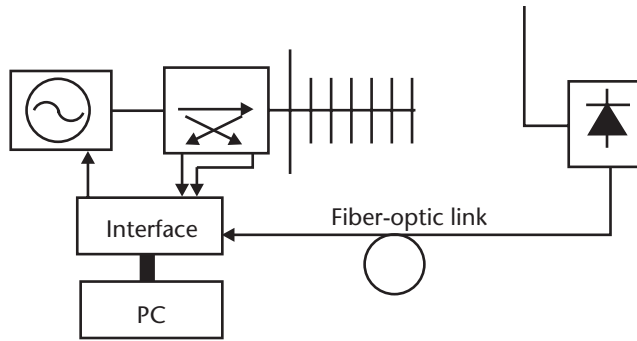
Evaluations with the prototype suggested that a 6- to 8-element Yagi is sufficient to provide a wide enough dynamic range, typically more than 30 dB for the measurements. The reflector, dipole, and all directors can naturally be of telescopic nature to allow a tuning to the specific VHF frequency of interest. With a TX power of 10 mW we were able to observe changes of gain from  $-2$  dBi to 12 dBi and above—these being related to the effectivity of the directors. The received and detected voltages as coming from the diode output were typically above 1 mV in the main lobe and thus easily monitored with modest instrumentation. Some performance figures of the prototype system are summarized in Table 7.1. There was no special need for selectivity in the frequency domain due to the high power levels involved, but meaningful measurements were not possible in the vicinity of another active TX, which is using the same band. For example, the ILS localizer must be temporarily switched off for a conductivity measurement at that specific frequency.

Occasionally we already observed quite dramatic phenomena on a clay-like, grass-covered soil where the gain of the Yagi was virtually 0 dBi without the calibration hatch thus indicating a practical absence of any VHF conductivity. Also, an estimated decrease of the soil water level of 100 mm was detectable through a change in the main beam elevation angle during several weeks of dry weather. The spatial resolution along the surface depends on the selected transmitting Yagi's length or number of directing elements, which for the prototype meant about 4.2m. Theoretically also the ground, which is not just under the directors, had an effect on the pattern but our experiments with separate metal sheets (1.5m by 2.5m in size) positioned at arbitrary locations near the transmitting antenna indicated a relative field strength contribution below 10%. On the other hand it was not practical nor methodologically correct to shorten the test range distance by bringing the receiving monopole closer because we then lose the far-field characteristics and start to have a too strong mutual coupling between the two antennas.

As was tried during the very first evaluation runs, a further extension of the described method is to monitor the Yagi's lobe pattern above horizon. This, however will require considerably more mechanical installation work because then the receiving monopole should have a calibrated but adjustable vertical mount with a maximum height up to about 16m (equals  $40^\circ$  of beam tilt). It is also feasible to construct a completely automated, integral system with a PC control by using conventional interface cards and possibly a short-range fiber optic link whereby even adverse effects, such as heavy snowfall, are reliably handled. The suggested, and currently tested configuration is illustrated in block diagram form in Figure 7.8. If

**Table 7.1** Prototype Performance Figures

<i>Parameter of Interest</i>	<i>Value</i>
Output power	10 mW
Dynamic range	>30 dB
Frequency range	88–108 MHz
Spatial resolution (2D)	4m



**Figure 7.8** The trial version is easily upgraded to have a common PC interface and a fiber-optic link. Obviously indeed, cost and complexity increase.

needed, a continuous 24-hour or day-by-day monitoring is readily provided and thus seasonal variations can be observed.

### 7.5.2 High-Power HF VNA

The input impedance of most HF antennas operating in the frequency range from 100 kHz to 30 MHz and used for air or maritime navigation [10] or military communications has to be measured after installation. This is simply due to their size preventing any indoor tests in shielded rooms or to the site itself, which has unavoidable effects on the performance through such factors as ground conductivity [13]. A typical NDB installation site—with the antenna in the middle of snow and frozen vegetation—is illustrated in Figure 7.9. The NDB concept was treated in Section 3.3.1. Such an installed antenna freely receives a wide spectrum of external transmissions and interference that will all be analyzed by a conventional network analyzer or impedance meter as reflected power. Alternatively, if a high PA is added for measurement purposes, the risk of test equipment damage will be substantial because of the unknown return loss of the device under test [3]. Extensive tuning may be necessary to eliminate the misleading effects of the amplifier couplings. If the network analyzer could automatically adjust its test port power according to the



**Figure 7.9** Seasonal variations of vegetation and soil affect the impedance of many HF antennas. So does snow, which may be wet or dry, depending on temperature.



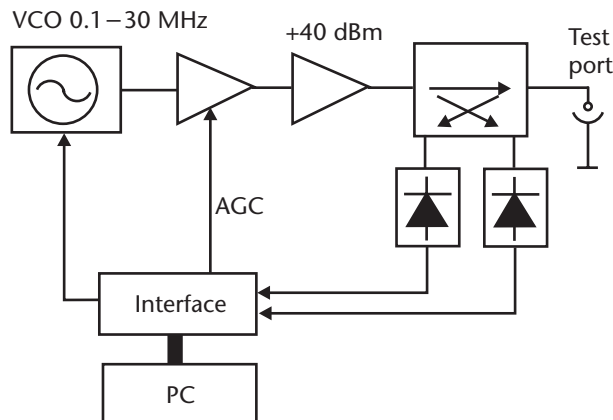
observed return loss, much better performance would be anticipated. A lowest possible tuning power would be beneficial in certain military applications, too.

The hardware setup described in this example has been assembled for evaluation purposes from commercial RF building blocks and low-cost data acquisition devices as illustrated in Figure 7.10. A key element is an AGC amplifier mounted just before the output stage and the two-way directional coupler. According to the particular case, the test port power now varies automatically between 0 dBm and +40 dBm, controlled by the PC through its D/A-card. The current prototype, shown in Figure 7.11, can produce a test signal frequency from 0.1 to 31 MHz as a continuous sweep by utilizing two VCOs and one mixer. If needed, the upper limit is easily extended to 1 GHz with only minor hardware modifications. The rejection of misleading external signals can be enhanced by using the built-in identifying modulation, which has been found suitable against widespread noise and spurious emissions.

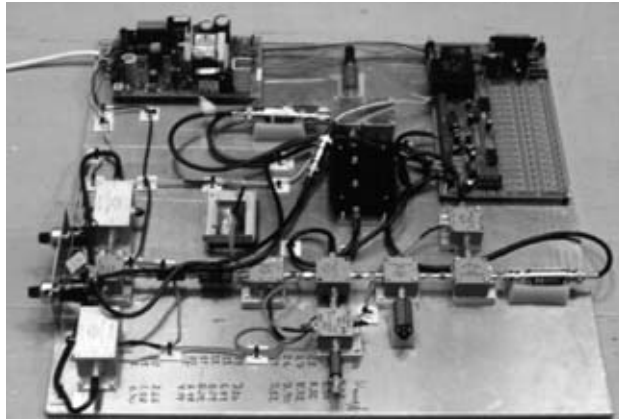
The PC interface utilizes the simple I<sup>2</sup>C bus and has three analog inputs for the forward, reflected, and phase signals and two outputs for the frequency adjustment and naturally for the power control. Both the forward and reflected powers are computed from the measured dc voltages through a curve-fitting algorithm of exponential nature, including also a temperature compensation feature. Observed return loss uncertainty is better than 1 dB up to 30 dB but naturally depends on the choice of the coupler. In the software loop the test port power is adjusted initially from its lowest value so that the reflected component stays below a preset limit—say +10 dBm, which is a convenient value for a reliable operation of the detectors. Currently the algorithm uses averaging over 100 individual samples which leads to a maximum control loop delay of 20 ms. Phase measurements can be done after this but require a slightly longer (40 ms) time constant for stable results. Calibration against known standards is easily adopted as a lookup table. The present setup needs short 6-, 12-, and 20-dB pads and two arbitrary delay lines for phase.

### 7.5.3 Pattern and Impedance Measurements of Compact Antennas

Small antennas cause considerable measurement problems because their performance is heavily affected by the immediate environment. For example, many cellular



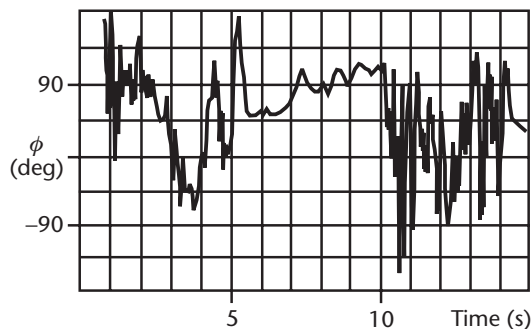
**Figure 7.10** The idea in the tested scheme is the use of feedback to adjust output power according to the observed return loss. Here this is the task of a 30-dB AGC amplifier.



**Figure 7.11** The assembled prototype uses commercial RF blocks and a data acquisition card.

phone pattern measurements are still partly based on relative values and the effects of the user's hand, head or other nearby obstacles is often neglected when the real output power is concerned. One case example is illustrated in Figure 7.12. Partly this is quite justified, because the real operational conditions differ a lot and a laboratory test can thus only very approximately mimic the user's behavior. One reason to omit realistic scenarios has been the cumbersome feed arrangements, typically relying on coaxial cables. In this example, two alternative simple arrangements for the pattern and SAR measurement of small mobile handset antennas and associated device structures are demonstrated.

Especially true radiated power measurements are known to be hampered if the antenna is very small compared to the wavelength [15]. The main ideas here are a simultaneous recording of impedance matching—which is known to be heavily affected by the proximity of the human body [16] and to some extent by the physical test arrangement itself [17]—and the substitution of the common coaxial feeder cable by plastic optical fiber as the reference channel. This practically excludes unwanted pattern interactions due to surface currents between the mobile device and the test setup. Additionally, some of the adverse effects of unintentional phase changes in coaxial cables due to bending—see [18] for an ultimate example—can be avoided. However, the proposed scheme cannot alone reduce the disturbing effects



**Figure 7.12** The fast and slow phase changes of the reflected signal coming from a mobile phone in an unsteady hand near a person's head.

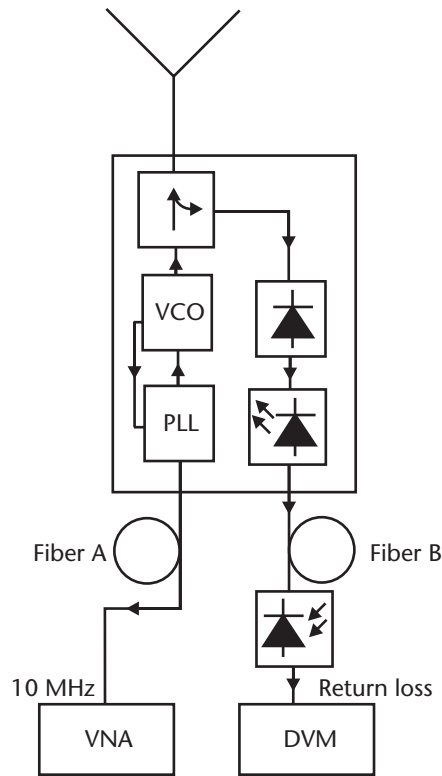
of the probing system, which typically consists of a considerably large robotic arm and some supporting hardware.

Two slightly different schemes have been outlined for this specific task. The fundamental requirement has been that the system must be able to provide meaningful results also in the case of near-field scanning. This implies a coherent instrumentation setup. Besides this, we have to be able to squeeze all the “onboard” electronics so small as to fit inside the tiny hand portables without any physical distortion or modifications. This is obvious as most mobile handsets currently utilize their enclosure as a vital part of the antenna system. Amplitude stability is a necessity, too, because otherwise the design margin must be respectively increased. Finally, interfacing to existing antenna measurement systems—both home made and commercial—should be as straight forward as possible and preferably be based on industrial off-the-shelf gear.

The TX is made of a commercial VCO chip working at 2 GHz (extendable up to 3.5 GHz) and having a nominal output power of +10 dBm. There is a frequency-related change of about 3 dB in the output level, but once a reference data set has been established, the real fluctuations as a function of time or loading conditions can be reduced below 0.2 dB. More information about VCOs is given in Section 4.3.5. For the lower VHF frequencies there is an option to use a 20- to 50-MHz circuit, which has an identical layout. From here on, there are two possibilities. We can phase-lock the VCO to a 10-MHz reference, which is brought into the *device under test* (DUT) from the network analyzer via a fiber. A directional bridge, connected between the DUT antenna and the VCO, feeds the reflected sample to a diode detector, the output of which directly dc-modulates an optical TX diode. All this needed hardware fits inside an 8-cm<sup>3</sup> box whereby even the smallest known handset geometries can be tested. Of course, a true measurement of the scalar part of  $S_{11}$  would require a sample of the forward signal as well, but the selected VCO seems to be able to sustain very severe mismatch without degradation in the output level. Finally, a second fiber carries this return loss information to the pattern measurement system. This arrangement is illustrated schematically in Figure 7.13. The apparent drawback of this scheme is the need for a twin fiber.

The second possibility is to totally omit the phase-locking electronics and to use a wideband laser diode TX [19], which is directly connected at the output of the directional bridge. This means that the RF signal is modulated by the 2-GHz carrier, which originates from a free-running oscillator or from a DRO; see Section 4.3.5 for further DRO details. The feasibility of this fiber optical process is based on [20], where, however, a direct interaction between a relatively strong microwave field and the optical signal has been used. The reflected VCO sample at 2 GHz contains both momentary frequency data and return loss information, which can be extracted in the pattern processing electronics, marked “vector RX” in Figure 7.14. In this second scheme, the phase noise and frequency stability of the VCO chip set the definitive accuracy limit.

The VNA, which is our standard device for measuring complex impedances, gains, phases and group delays, can be used as a vector front end only or, depending on its own features, we can push most of the processing to be handled by its internal circuits. In fact, there is no need for a VNA as such if the vector RX comprises two channels and thus is able to perform the phase and amplitude detection of the



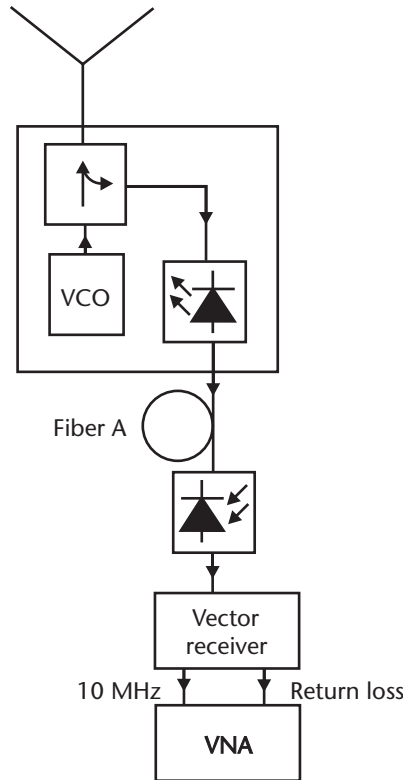
**Figure 7.13** The experimental TX includes a directional bridge and a phase-locked VCO, which is connected through optical fiber to avoid cable effects on measurement accuracy.

radiated signal as well. This approach, however, requires considerable calibration runs and may not be easy to maintain within standardized performance limits.

As shown, a coherent system for the recording of mobile phone radiation patterns without any coaxial RF feeder cables is realistic indeed. Existing off-the-shelf hardware can be extensively utilized. High-speed laser TXs from optical data networks look particularly interesting as building blocks. Realistic three-dimensional pattern evaluations seem feasible also in the near field because the same reference frequency is used in the receiving network analyzer system. Besides this, it is possible to define the real output power or return loss at the DUT antenna connector as a function of momentary alignment and phantom configuration, which have become practical test parameters with the most sophisticated positioning systems; see [21]. Without any further improvements in the measuring hardware, even the enhancements suggested here are unable to reduce the disturbances caused by the mechanical probe positioner (the robotic arm) or the phantom stand.

#### 7.5.4 Test Instrumentation for Air Navigation Facilities

This section describes—as a kind of conclusion for the whole book—the key elements of two computerized flight inspection systems designed and constructed by the author during a 5-year period for regular testing of basic RF air navigation ground facilities. Flight inspection is a process through which national aviation authorities

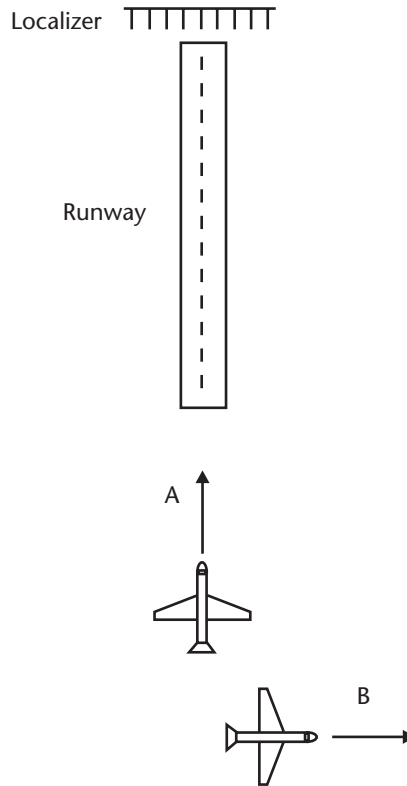


**Figure 7.14** The phone side does not need a phase lock if, for example, a fast diode laser TX can be installed as the return path.

periodically check the quality of navigation information. This is provided by ground radio installations and special visual indicator systems such as different approach light arrangements. Typically a dedicated aircraft and its skilled crew are used for these purposes. Measuring procedures include approaches and “slices” of the azimuth and elevation guidance signals as indicated in Figures 7.15 and 7.16. Measurements are made at predefined intervals, which can be, for example, 3 months for the ILS but just 1 year for NDB. Some nations have combined the adjustment (e.g., alignment of the glide path angle) and certain maintenance tasks to flight inspection work but many aviation authorities want to keep these two vital actions separated. One of the benefits is that in this way we can have an autonomous “control” function, which might reveal occasional human errors or test equipment biases.

Although the first requirement for successful flight inspection work is a suitably educated and motivated team, there is also an apparent need for special instrumentation. The coming paragraphs try to show some important issues, which appeared during one sample development project. The target was to create a simple computer-controlled flying system with IEEE-488-bus equipment and a theodolite tracker. It was to provide a basis for accurate and repeatable measurements according to ICAO Annex 10 requirements.

Choice of proper power supply, connection of different RXs, their characteristics, antenna installations, and data processing algorithms are discussed. Aircraft ground tracking techniques are considered together with achievable positioning

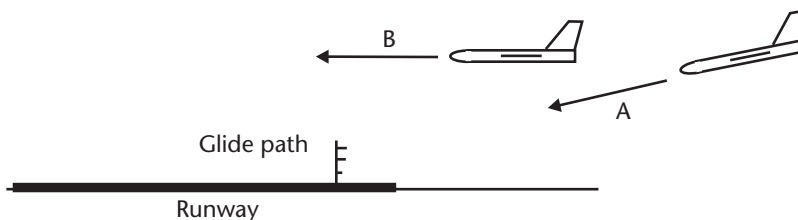


**Figure 7.15** A flight inspection task of the ILS localizer signal quality includes a center line approach (A) and a cross-flight (B) following predefined routes.

uncertainty. An autonomous method for measuring two frequency ILS-installations with a combination of two communication test RXs is illustrated as a byproduct of the development project. Two slightly different systems were compiled and thus selected features (and a couple of not so fortunate decisions) are compared as well.

#### 7.5.4.1 Countermeasures Against Shocks and Vibration

As was stated in Chapter 2, one of the first very general things in system design is to figure out the environment in which the designed entity is to be used. In this case both of the tested flight inspection systems were installed in a light twin turbo-prop aircraft so this was a major guideline for the mechanics. Two types of mechanical stress are encountered in an aircraft environment. During takeoff and landing we



**Figure 7.16** Glide slope inspection is similar to localizer except that now the flight tracks are referred to the local horizon. Approaches (A) and level runs (B) are used.

may observe shocks, which are mainly due to the runway roughness. When in the air, vibration from the engines—mainly because of propeller imbalance—and sometimes also from the aerodynamic surfaces is dominating. One of the first questions was if special precautions were needed.

All the airborne instruments were fitted into a single 19-inch rack because no more space was available. The height of both system racks was 1.2m, width 0.6m, depth 0.7m, and the weight of system I 150 kg and that of system II 200 kg, respectively. Initially the measuring team operated with a continuously equipped aircraft but later—mainly due to airliner's requirements—requests for an easily remountable system became frequent. The original system I had no special shock absorbers—it was just ordinary “rigid” laboratory hardware and mounting assemblies. Easy mounting, which was planned to take only half an hour or so, was only possible through dividing the first system into smaller physical pieces. This idea of using separable units in system II led, as expected, to an unfortunate weight increase of almost 50 kg. Besides, the individual units themselves, such as a VHF/UHF test RX, were, however, too heavy for transport by personnel, so each piece of instrumentation must be further installed one-by-one anyhow. No difference was finally found in the behavior or reliability of the two systems regarding the effect of shock absorbing. This seems to be rather obvious because in terms of time duration, vibration is dominating and absorbers having an airworthiness certificate that is based on shock survival cannot do much against limited-amplitude wobble.

Parallel field tests confirmed that the more lightweight and electromechanically stable (no frequent reconnections between units) system I was superior. The hardware of system II was thus dismantled. Commercial instruments such as VHF/UHF test RXs were “recycled” to be spares and upgrades of system I, but the rather expensive shock-absorbing mechanics and its mounting elements were completely discarded. Figure 7.17 shows the completed system I during its commissioning tests.

#### 7.5.4.2 Quality of Electrical Power

The second problem in the aircraft was providing sufficient amounts of electricity for the test instruments. The quality and availability of electrical power is an essential item, because many instruments prefer long stabilization times. A general idea was to provide for the radio measuring equipment an environment as near a



**Figure 7.17** After successive upgrades and modifications, flight inspection system I was put into commissioning tests.

laboratory as possible in this respect, too. This yielded to selecting a no-break system. The primary power supplied to the instruments was 220-V 50-Hz true sine wave ac produced by a 1-kVA inverter plant. The converter was fed by 28-V dc coming from the aircraft battery when airborne. On ground the system was connected directly to normal mains supply through a 16 A fuse, as electrical heating was provided through the same inlet—a mandatory feature in the arctic environment in winter. Inside the *flight inspection* (FI) console normal rack-mounted laboratory power supply units were used to convert 220-V ac to required dc voltages. This procedure also improved remarkably the isolation of the *VHF omnidirectional rangelinstrument landing system* (VOR/ILS) RX from the aircraft's electrical system. Three noticeable points were recorded during the system development phase:

- A square wave type PWM chopper power converter is not suitable for this kind of a system because of serious EMC problems and the fact that many laboratory instruments have an ac blower, which does not run with square wave;
- Forced air cooling of power supplies is to be avoided because the power unit normally sits at the bottom of the rack and gets all the dust and sand inside with increased flow of air;
- Any ac/dc converters should preferably be of linear type to avoid disturbances to ADF reception, details of which are also given in Section 3.3.1;
- A dc ground power supply (common in jet aviation) is not very favorable in flight inspection because this kind of service is normally not available at small community airports and it could not support cabin heating anyhow.

Feeding power to the ground theodolite, which defines the angular position of the aircraft and to the associated tracking unit was a big problem. The ultimate equipment size limit regardless of functional principle comes from the heavy snow conditions of arctic airports. All three possible types of ground power (ac, ac and batteries, battery only) were tested during the development phase. The combination of mains and battery was found the least satisfying solution, because the size and weight of the ac-section was quite remarkable if the recharge period was short enough. This together with the batteries made the system far too heavy. The direct ac-supply with a 100-m connecting cable, associated reel and a small dc-based feeding unit was in use for a long time. Mains supply provided an unlimited operational time and 400 VA of connectable power. The battery-only system was capable of 8 hours of transmission with limited power but the setup time was only a few minutes. The direct mains feed caused problems in the presence of high electric fields in the vicinity of the ILS localizer antenna system as the electrical shielding of nongrounded tracking equipment was not an easy task. Besides, nearby thunderstorms could have been a potential safety hazard to the tracking system operator due to the long mains cable.

#### 7.5.4.3 VOR/ILS-RX

Many aviation authorities consider precision approaches in bad weather to be one of the most risky phases of flying. Therefore also most of the flight inspection activity has been focused at ILS, which tends to be, even after GPS has finally reached a



mature state, the main supporting ground facility. It is foreseen, that *microwave landing systems* (MLSs) will be put into a similar role in the near future. Anyhow, a special ILS RX is needed. It contains actually two separate RXs—one around 110 MHz for demodulating the azimuth information (localizer) and another around 320 MHz for elevation (glide path). Flight inspection teams often use quite similar RXs to those found in larger airliners' aircraft, with some possible modifications. Due to historical reasons the ILS RX unavoidably contains a VHF omnidirectional range RX (RX/VOR) as well, but the intensity of flight inspection is generally lower.

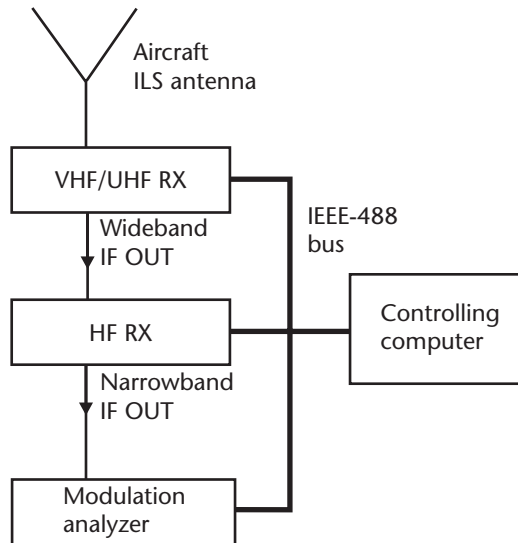
An ILS RX handles a signal, which is amplitude-modulated by two baseband frequencies: 90 and 150 Hz. This is again due to historical reasons and nobody would today select such a combination. The angular information is “hidden” as a difference in the modulation depths of these two, but to work appropriately, the modulation sum should be fairly constant and predefined.

The main factors considered at this stage were the following:

- Cross coupling between *difference of depth of modulation* (DDM) and *sum of depth of modulation* (SDM);
- Capability of measuring both TXs in a dual configuration (course and clearance TX);
- Time constant of DDM measurement;
- Total number of RXs.

Because the ILS concept comes from the late 1930s, almost any well-engineered RX will do the job. The flight inspection systems discussed here used initially Rockwell Collins 51RV2/4 units, but when the first IC RXs came available for flight inspection tasks, an immediate attempt was started. In terms of published data, the selected special new RX was supposed to be superior in accuracy, but the behavior soon turned out to be far from optimum—be it a sample fault or not—with high SDM/DDM values. The newcomer's measured DDM tended to decrease after a certain point with increasing SDM, where as the old Collins RX behavior was as expected by historical background. The lesson learned here by the project team was simple. New commercial constructions almost always have unknown defects—more or less serious. If you do not belong to a particularly wealthy team or organization, it is perhaps wise to let others do the field-testing for you.

Because such a feature was not required at that time by any major aviation authority, commercial RXs could not distinguish between the two close-by carriers of a two-frequency ILS. Actually, they should not do it, because the whole idea of two-frequency systems was in “fooling” the airborne RX. However, flight inspection should be able to find out the individual state of the two. The latest design of an ILS-RX was a combination of two high-grade communication test RXs and a modulation analyzer; see Figure 7.18. This system had the advantage of measuring separately the effect of both RF carriers and associated modulations in two-frequency systems (course and clearance TX), where the necessary selectivity is provided by the IF filters of the second (HF-) RX. This arrangement is capable of measuring both the course and clearance TX RF radiation patterns, too, and simultaneously—which saved a lot of airborne time (\$1,500 per hour at that time).



**Figure 7.18** This is an alternative scheme to measure two frequency ILS-installations. An HF-RX provides selectivity when fed with the 10-MHz IF from the front-end device.

The number of VOR/ILS-RXs in a typical airborne console has traditionally been high (3–5). Many national teams not only carry a bank of RXs but also calibration generators in their aircraft. This is of course possible, if the interior space is large enough. In this particular case the size of aircraft was the most limiting factor. On the other hand, a system based on a dual configuration is not of high statistical value because a decision-making process is not possible. What if two ILS RXs disagree for example about course alignment by  $0.02^\circ$ —which one is correct? Taking a generator on board looked very questionable. If the stability of RXs is so poor that they cannot tolerate a 1- to 2-hour transfer flight, how could we think about reliable measurements, which anyhow take 1 to 2 hours per runway? Besides, a generator on board will have to take all the shocks and vibration and as a laboratory unit it is hardly designed for that. Thus the whole development was based on a single RX configuration—and with good results.

The important system related parameter is the time constant or—to be more exact—the time constants of an RX DDM output. An example of this is a case where the input signal steps 35 mA, which is equal to the maximum allowable bend in glide path structure. The main target here was to adjust the sampling rate so that during an approach flight pattern we could get a realistic but not too large data set of the RF parameters. Respective analog filtering time constants were still toward the beginning of 1990s defined by the approach speed of a DC-3, which flew (under military designation) already during World War II. Now the computer sampling rate was adjusted to match the 63% rise time of the navigation RX DDM output.

#### 7.5.4.4 HF-RX

Due to the great number of NDBs installed (about 3 for every ILS or airport/runway) and the relatively long distances between airports it was found necessary to have a possibility for NDB inspection by air. For this purpose a special modified HF

RX was installed into the console. Typical commercial HF test RXs were in many respects very suitable for this kind of task. However it turned out, that their IF bandwidths were still by a factor of 1.5 wider than in the best available HF communication RXs. Also the weight of a good quality test RX is near or even exceeds that of a spectrum analyzer. Therefore we selected an HF-communication RX, which gave a selectivity of 150 Hz at 3-dB points with 70-dB attenuation measured at a distance of 150 Hz off the carrier. The factory model came with a dynamic amplitude range of 120 dB and a resolution of 5 dB. In the normal measuring situation this is not a satisfying combination. Therefore the internal A/D-converter was modified to give a resolution of 1 dB and a dynamic range of 20 dB. To this can be added the programmable attenuator of 20 dB to bring a total dynamic range of 40 dB. The hysteresis of the A/D-process could not be reduced respectively because of the heavy noise content in airborne measurements. This HF-RX had also an important role in the new ILS-method as described above.

#### 7.5.4.5 Field Strength Measurements

With an aviation VHF/UHF-RX the normal way of recording RF behavior of VOR/ILS-installations during a flight inspection mission has been through AGC-voltages. The clear disadvantages of such a system are the highly nonlinear characteristic, limited dynamic range and unknown frequency dependence (e.g., selectivity). To provide more defined conditions the special VHF/UHF-RX was chosen as a part of the console. One commercial RX type was suitable for this purpose but the RF level measurement circuits had to be filtered better than normal by averaging over time to avoid the influence of high VOR/ILS modulations (30–150 Hz) in the dBm output value. The AFC-function could not be used with ILS-localizer cross flights if the ground installation was a two-frequency system, because then the AFC would mistune the RX at points about  $\pm 10^\circ$  offset from the actual centerline. The small IF spectrum display of this RX was found very useful when tracking broadcasting-related interferences to ILS/VOR installations. However, the resolution was not good enough to resolve between the two carriers of an ILS.

When the measurement of NDB/LOC beacon field strengths became vital, we initially relied on airborne instrumentation only, but later the task was largely moved to the ground measuring van illustrated in Figure 7.19. It had best possible, commercially available, test gear at that time and utilized high performance loop antennas for HF measurements. Comparisons performed at random sites indicated a correlation within 0.5 dB between expensive and tedious flight tests and much more fluent ground observations.

#### 7.5.4.6 Antenna Installations

As indicated in Section 5.1, antennas are a good way to greatly enhance or totally spoil an RF system's performance. In an aircraft installation this is very critical, because aerodynamic restrictions tend to lead to curious designs. In this system project, all flight inspection antennas were totally separated from the aircraft's own navigation system. The navigation measurement antenna was a V-dipole mounted under the fuselage just behind the wing edge to provide adequate signal strength for



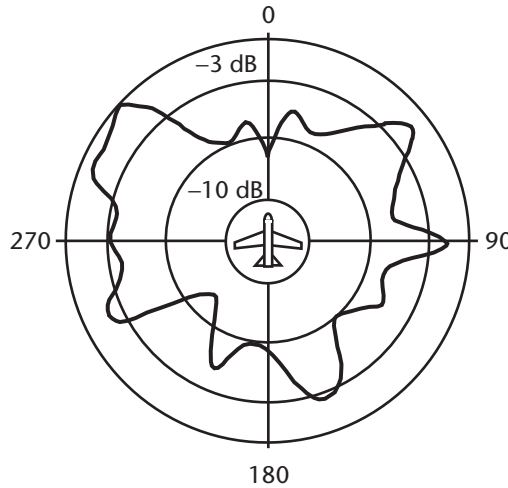
**Figure 7.19** Airborne HF field strength measurements were supplemented and later replaced to a large extent by ground measurements, which were performed with a dedicated van. Note the loop antenna on the roof.

orbital flights (e.g., around VOR stations). It was an “oversized” type normally used in large passenger aircraft but in this way we were able to get a better radiating efficiency. The associated vertical communication antenna was previously used for telemetry RX but was later reserved for measurements of VHF ground-station coverage patterns, when the microwave ISM telemetry was taken into use.

The *glide path* (GP) antenna was mounted far on the nose in front of the weather radar position but the weather radar dish was removed because of reflections. Unfortunately there was no weather radar available after that, although it would have been very beneficial during long transfer flights. The HF measuring antenna was a modified ADF-loop with its own power supply. The antenna was mounted at the same level with the NAV-dipole. Capacitive loading due to the feeder coax (90 pF/m; see Section 5.4.1) required the use of an additional booster amplifier, but this tended to cause unwanted spurious emissions. Additionally featured were microwave telemetry and limited-range *distance measuring equipment* (DME), a commercial design by Motorola; antennas were also beneath the fuselage but in front of the wing and slightly offset from the centerline to avoid shading by the aircraft nosegear. This was mandatory, because national authorities selected to use ILS measurements extended to the touchdown point. Various possible NAV-antenna positions were tested prior to building the fuselage as a 1:20 scale model. An example of those scale test results is Figure 7.20. They indicate severe lobing around the azimuth circle caused by local multipath and shadowing from the fuselage, tail and wings. Computer control allowed us to use this prerecorded pattern later as a lookup table during real measurements. It was combined with attitude information from the aircraft’s inertial system. In this way, the momentary amplitude error caused by a particular lobe of the aircraft antenna’s pattern could be subtracted from the measurement result of a VHF ground TX.

#### 7.5.4.7 Aircraft Position Reference System

At the time of the project’s initiation, many European aviation authorities were using partly manual aircraft tracking techniques for flight inspection purposes. One



**Figure 7.20** The azimuth pattern of an airborne VHF antenna (scale model measurement).

of the reasons to this might have been very conservative international standardization. The old verbal method of distributing theodolite data to flight inspection aircraft, suggested at that time by the ICAO, suffered from many problems. During a normal ILS-GP approach only very few samples could be obtained so that the aircraft was really at the nominal elevation angle, whereas the respective number of corrected samples during the same approach is very high. However, correction assumes linearity and well-known “gain” or displacement sensitivity of the ILS ground system under test, which of course is not a proper way of doing technical inspections or measurements. Also the apparent time delay caused by human factors is not to be overlooked. As an alternative brief tests were carried out with a visible-light television camera and associated TXs and RXs. The idea was to provide an approach view also to the pilot. Despite of using ultimate optical magnification we were not able to maintain adequate angular resolution. Also the analog television transmission turned out too vulnerable regarding onboard interference and aircraft attitude.

Theodolite tracking with a telemetry system reduces these delays considerably. It also increases the number of data, but suffers from visibility problems and requires always at least one human operator on ground. A typical demonstrated tracking distance was 21 km. The theodolite-based tracking system with telemetry was chosen as a standard for both flying installations. A microwave link operating in the S-band was used to carry the telemetry information. Data from the optical transducers was used directly to FM-modulate the carrier whereas the analog information was first converted to an audio frequency in a normal manner. A measured uncertainty for the analog system was  $0.01^\circ$  and resolution  $0.003^\circ$  for both axes and  $0.003^\circ/0.001^\circ$  for the digital theodolite.

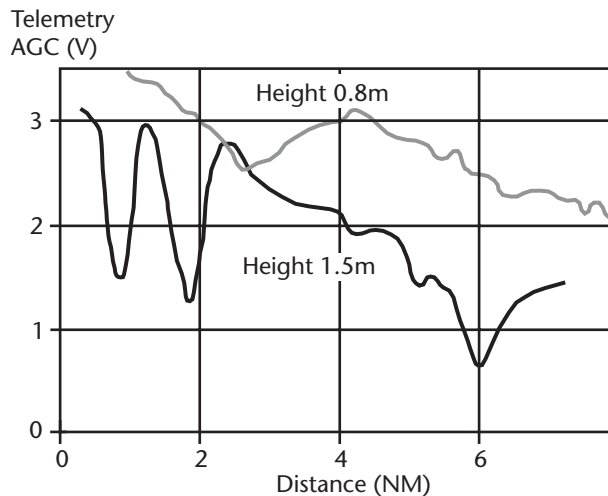
Separate distance measuring devices had already been used by some flight inspection units for a couple of years. A clear benefit is to get approach-mode results independent of aircraft speed variations. The achievable resolution of 3m (e.g., with Motorola Miniranger) was quite adequate for the purpose, but optimum microwave coverage is difficult to achieve, due to two facts. The ground station antenna gets

high lobing patterns, if raised above things like snow and grass. The aircraft antenna is often shaded by the wings and the fuselage. A recorded radiation pattern with two ground antenna elevations is shown in Figure 7.21 and is a solid proof for the theoretical presentation in Section 3.2.3.

#### 7.5.4.8 System Computer and Data Processing

Both systems were totally based on computer processing of information with no additional (manual) recording or display facilities. This was quite new at the time of the project because many national authorities still used multichannel ink recorders or tapes. During a particular measurement the computer on board was gathering data from the flight inspection sensors with maximum speed, which varied between 2 to 8 bytes/second depending on computer and software choice. After a flight procedure (e.g., approach) was completed, the results were shown on the display for operator judgment. If the result was acceptable as a measurement it was stored on digital tape. Some information was copied also on the thermal printer for future reference. No normal aircraft instruments were available to the operator. All data needed was displayed also during the procedure.

Very few computer designs could fulfill all the wanted parameters at the time of construction. A compromise was made by choosing two commercial computers from the same vendor but having different processors and mass storage arrangements. Connection to all peripheral devices was made through IEEE-488 interface except the telemetry channel, which utilized a 16-bit parallel bus provided by the computer plug-in module. Computer A had an extra 256-kb memory module and an advanced programming *read-only memory* (ROM). Computer B had a 4-Mb memory module and ROM BASIC. The measuring programs were originally developed in system A. Each separate measurement [e.g., *glide path* (GP) level run] had its own software module. After completing the development all the programs were transferred to system B by a converter program.



**Figure 7.21** Microwave telemetry antennas too high above ground (black) suffer from severe lobing but when lowered to 0.8m (gray) the RX AGC (V) behaves properly.

#### 7.5.4.9 Calibration of the System

Vital parameters of the flight inspection system were defined as coefficients in the respective program. Two different structures were in use. Because of the slow access of the data cartridge in system A the actual calibration constants were part of the program lines and each time these had to be modified the whole BASIC-program must be edited. On the other hand the disk drive of system B was fully utilized with a separate calibration file, which was updated by hand or directly under program control. Practice showed however, that the former method is more reliable, because during the calibration run one gets a ready document of the work performed. System stability was so good that changes were needed roughly twice a year. The periodic calibration interval used was 7 days, when the system was in operative use.

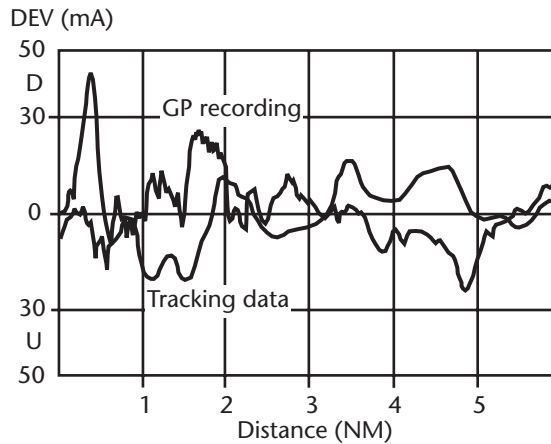
The key laboratory instruments, shown in action in Figure 7.22, used in periodic calibration were the following:

- A high-performance VOR/ILS generator (e.g., Rockwell Collins 479S-6A);
- A microwave synthesizer/sweeper (e.g., Rohde & Schwarz SMP22);
- A VOR/ILS modulation analyzer (e.g., Rohde & Schwarz FMAV);
- A synthesized high-end spectrum analyzer (e.g., HP 71200 C, now Agilent);
- A sampling oscilloscope (e.g., Tektronix TDS3000).

A real-life ILS-GP-approach measured with the described system is shown as an example in Figure 7.23. It describes both the corrected DDM-curve and the theodolite tracking registration. Also a mean value for SDM is calculated. The rough GP-angle and displacement sensitivity were calculated first by the common level-run-software and then corrected by the approach results. This naturally dictated the order of flight procedures. A typical correction was about  $0.04^\circ$ , but varied considerably with site and time. Two modes were available for NDB: a coverage orbit and a radial flight. No bearing information was presented, as NDB is a nonprecision system.



**Figure 7.22** Essential calibration test instruments needed to keep the flying installation within its specifications.



**Figure 7.23** An example of an ILS GP measurement with the developed system. The gray trace shows tracking information and the black curve represents the true GP recording DEV (mA).

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# List of Acronyms

<b>A/D</b>	analog-to-digital
<b>ADF</b>	automatic direction finder
<b>ADTS</b>	advanced detection technology sensor
<b>AGC</b>	automatic gain control
<b>AIS</b>	alarm indication signal
<b>AJ</b>	antijamming
<b>AM</b>	amplitude modulation
<b>APC</b>	adaptive power control
<b>ARM</b>	antiradiation missile
<b>BER</b>	bit error rate
<b>BITE</b>	built-in test equipment
<b>BS</b>	base station
<b>BSC</b>	base station controller
<b>C/I</b>	carrier-to-interference ratio
<b>C/N</b>	carrier-to-noise ratio (in communications)
<b>CDMA</b>	code division multiple access
<b>COHO</b>	coherent local oscillator
<b>COTS</b>	commercial-off-the-shelf
<b>CPU</b>	central processing unit
<b>CW</b>	continuous wave
<b>D/A</b>	digital-to-analog converter
<b>dBc</b>	decibels referred to the carrier amplitude
<b>dBi</b>	decibels referred to an isotropic antenna
<b>dBm</b>	decibels referred to 1 mW
<b>dBW</b>	decibels referred to 1 W
<b>DBPSK</b>	differential binary phase shift keying

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DCXO	digitally compensated crystal oscillator
DDM	difference in depth of modulation
DF	direction-finding
DME	distance measuring equipment
DOA	direction of arrival
DRO	dielectric resonator oscillator
DSP	digital signal processing
DUT	device under test
EA	electronic attack
ECM	electronic countermeasures
ELT	emergency locating transmitter
EMC	electromagnetic compatibility
EMI	electromagnetic interference
EIA	Electronic Industries Association
EIRP	equivalent isotropically radiated power
ERP	effective radiated power
EPIRB	emergency position indicating radio beacon
ES	errored second, electronic support
ESM	electronic support measures
ETSI	European Telecommunications Standards Institute
EW	electronic warfare
FAA	Federal Aviation Authority
FCC	Federal Communication Commission
FDA	Food and Drug Administration
FET	field effect transistor
FFT	fast Fourier transform
FFH	fast frequency hopping
FH	frequency hopping
FM	frequency modulation
FMCW	frequency-modulated CW
FSK	frequency-shifted keying
GPS	global positioning system

<b>GSM</b>	Groupe Spéciale Mobile
<b>GUI</b>	graphical user interface
<b>HF</b>	high frequency
<b>HV</b>	high voltage
<b>IC</b>	integrated circuit
<b>ICAO</b>	International Civil Aviation Organization
<b>IEC</b>	International Electrotechnical Commission
<b>IF</b>	intermediate frequency
<b>I2C</b>	intelligent industrial communication standard
<b>ILS</b>	instrument landing system
<b>ISM</b>	industrial, scientific, and medical
<b>ITU</b>	International Telecommunication Union
<b>JAN</b>	Joint Army Navy
<b>LO</b>	local oscillator
<b>LNA</b>	low-noise amplifier
<b>LOS</b>	line-of-sight
<b>LPD</b>	low probability of detection
<b>LPI</b>	low probability of intercept
<b>LSB</b>	lower sideband
<b>MLS</b>	microwave landing system
<b>MMIC</b>	monolithic microwave IC
<b>MTBF</b>	mean time between failures
<b>MTTR</b>	mean time to repair
<b>MUF</b>	maximum useful frequency
<b>MW</b>	medium wave
<b>NDB</b>	nondirectional beacon
<b>NF</b>	noise figure
<b>OCXO</b>	oven-controlled crystal oscillator
<b>OFDM</b>	orthogonal frequency division multiplexing
<b>PA</b>	power amplifier
<b>PCB</b>	printed circuit board
<b>PDH</b>	plesiochronous digital hierarchy

PEP	peak envelope power
PIM	passive intermodulation
PLL	phase-locked loop
ppb	parts per billion
ppm	parts per million
PRF	pulse repetition frequency
PSK	phase-shifted keying
PTFE	polytetrafluoroethylene
PWM	pulse width modulation
QAM	quadrature amplitude modulation
QMBOK	quadrature <i>m</i> -ary biorthogonal keying
QoS	quality of service
RAM	random access memory
RCS	radar cross-section
RF	radio frequency
RL	return loss
RLAN	radio local area network
rms	root mean square
RX	receiver
SAR	specific absorption rate, synthetic aperture radar
SAW	surface acoustic wave
SCPI	standard commands for programmable instruments
SDM	sum of depth of modulation
SES	severely errored second
SMD	surface mount device
S/N	signal-to-noise ratio
SPDT	single-pole dual-throw
SPST	single-pole single-throw
SSB	single sideband
SSR	secondary surveillance radar
STALO	stable local oscillator
STAP	space time adaptive processing

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SW	short wave
SWR	standing wave ratio
TCXO	temperature-compensated crystal oscillator
TDMA	time division multiple access
TEM	transverse electric and magnetic field
THAAD	theater high-altitude area defense
TTL	transistor-transistor logic
TWT	traveling-wave tube
TX	transmitter
UAT	unavailability time
UAV	unmanned aerial vehicle
UHF	ultrahigh frequency
ULA	uniform linear array
ULSA	ultralow sidelobe antenna
UMTS	Universal Mobile Telecommunications System
USB	upper sideband
UWB	ultrawide band
VCO	voltage-controlled oscillator
VEE	visual engineering environment
VHF	very high frequency
VNA	vector network analyzer
VOR	VHF omnidirectional range
WCDMA	wideband code division multiple access
YIG	yttrium iron garnet



# List of Symbols

$a$	target dimension
$A$	difference of element or subarray feeding amplitude
$A_T$	target's RCS
$A_{\text{eff}}$	effective area of an antenna
$B$	processing bandwidth
$c_0$	velocity of light in a vacuum
$d$	distance between array elements or subarrays, diameter of reflector
$D$	antenna or antenna array diameter
$e$	rms surface error of an antenna
$f$	frequency
$F$	array factor of an antenna array, NF
$G$	gain of a subarray, element, or front end
$G_a$	radiation pattern of an array
$G_D$	directive gain of an antenna
$G_R$	receiving antenna gain
$G_T$	transmitting antenna gain
$h_{\text{eff}}$	effective height of the electron layer in the ionosphere
$i$	summation index
$j$	index
$k$	wave number
$k$	Boltzmann's constant (1.38 E-23 J/K)
$K$	scaling factor
$L$	attenuation
$L_{\text{atm}}$	attenuation in the atmosphere (in addition to free-space loss)
$L_C$	cable attenuation
$L_{CR}$	RX cable attenuation



$L_{CT}$	TX cable attenuation
$L_{fs}$	free-space attenuation
$L_p$	propagation path loss
$L_r$	attenuation at a distance of $r$
$L_0$	attenuation at a reference distance of $r_0$
$N$	number of elements in an array
$P_{IN}$	input power
$P_L$	TX output power
$P_n$	noise power
$P_{OUT}$	output power
$P_T$	TX power
$R$	radius of a cylinder
$r$	radius, radial distance
$r_s$	inner radius of a conductor
$r_u$	outer radius of a conductor
$r_0$	reference distance for attenuation
$S$	power density
$S_r$	power density at a distance of $r$
$S_{ij}$	scattering parameter
$T$	temperature, noise temperature
$T_a$	antenna noise temperature
$T_{LOSS}$	temperature of lossy element
$T_{sys}$	system noise temperature
$T_0$	reference temperature (290K)
$t_i$	$i$ th instant of time
$U$	voltage, element feeding voltage
$x$	coordinate position (east-west)
$y$	coordinate position (north-south)
$z$	element positioning error from reference plane
$Z_{ij}$	mutual impedance
$\epsilon_r$	dielectric constant
$\phi$	phase angle

---

$\lambda$	wavelength
$\lambda_0$	reference wavelength
$\eta$	antenna aperture efficiency
$\phi_i$	incidence angle of a wave
$\theta$	physical angle
$\theta_{3\text{ dB}}$	3-dB beamwidth of an antenna



## About the Author

**Pekka Eskelinen** is a professor in the radio laboratory and the head of the Institute of Digital Communications at the Helsinki University of Technology, Finland. He left the position of the head of the Electronics and Information Technology Department at the Finnish Defense Forces' Technical Research Center in 2000. Before that, he was the head of the microwave laboratory and a professor of electronics at the Lappeenranta University of Technology in Lappeenranta, Finland. He received an M.Sc. in 1979 and a D.Sc. in electrical engineering in 1992 from the Helsinki University of Technology. He worked at the Technical Research Center of Finland in Espoo designing subsystems for satellite ground stations and military radar equipment. In 1984, Professor Eskelinen joined the National Board of Aviation in Vantaa, Finland, where he was chief of the avionics and flight inspection section. Since 1988, he has been involved in accident investigation, and he is a technical expert in the planning commission for investigation of major accidents. Professor Eskelinen has also been a member of the board of governors in the IEEE Aerospace and Electronic Systems Society and the chairman of the IEEE MTT/AP/ED chapter in Finland. He is an associate editor of the IEEE *AES Systems* magazine and is a member in the scientific committee of the European Frequency and Time Forum in Switzerland. Professor Eskelinen has been nominated to the International Advisory Committee of the JOM Institute in Denmark and has gained a docent's status of RF engineering both at Lappeenranta University of Technology and at Tampere University of Technology in Tampere, Finland. He is also the vice chairman of the communications section in the Scientific Advisory Council for Defense and a board member in ESA's external laboratory, Millilab, in Espoo, Finland. In addition, he has been a representative of Finland in the Western European Armament Group's Panel 6. Professor Eskelinen has coauthored *Microwave Component Mechanics* (Artech House, 2003) and *Digital Clocks for Synchronization and Communications* (Artech House, 2003) and published more than 200 papers in scientific journals and conference proceedings.



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