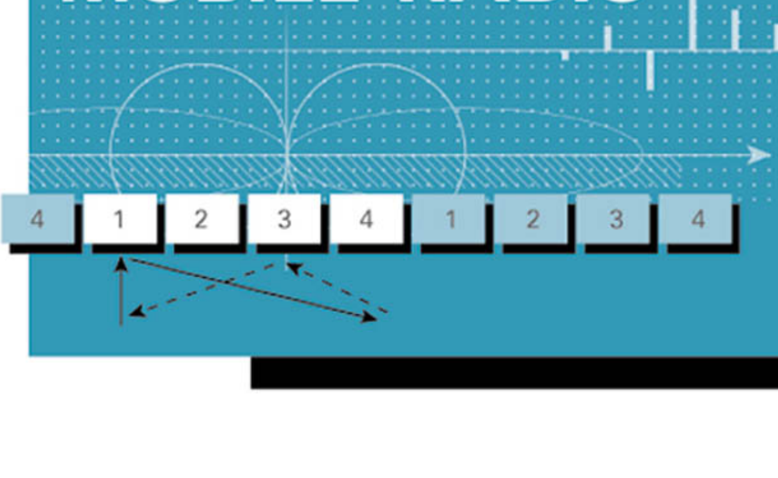
mobile communications series

Hans-Peter A. Ketterling

INTRODUCTION TO DIGITAL PROFESSIONAL MOBILE RADIO



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In memory of my parents, who granted me every necessary freedom to develop my personality and my skills without unnecessary limitations, and to my wife, whose never-ending patience and encouragement enabled me to write this book

Contents

Foreword

No praise is too high for Hans-Peter A. Ketterling's new book. It is a first-class piece of work by an acknowledged expert in the field and is essential reading for any user, operator, marketeer, engineer, or consultant who is actively interested in professional mobile radio (PMR), whether analog or digital.

The title of his work, *Introduction to Digital Professional Mobile Radio,* does not altogether do his book justice, as he has included excellent supplementary material on public mobile systems, including GSM and UMTS, in the chapter that deals with alternatives for PMR.

I first met Peter in the 1980s when, as the representatives of our respective national trade associations (ZVEI in Germany and the EEA in the United Kingdom), we participated in the work of the European Telecommunications and Professional Electronics Industry Association (ECTEL). Through ECTEL's PMR Group, Peter and I rapidly found common ground along with the representatives of other trade associations in our efforts to harmonize the European market for PMR, which at that time was a disorderly collection of different national technical standards calling for expensive and time-consuming approvals in each country, different, and in some countries, restrictive, licensing arrangements, and frequency bands that were only partially coordinated.

The group, with Peter as a leading member, played a major role in the ECTEL campaign to persuade the Conférence Européenne des Postes et des Télécommunications (CEPT) to harmonize the technical standards, accept a single approval process, and introduce sensible licensing arrangements. A liaison with the Directorate General XIII of the Commission of the European Community (CEC DG XIII) was also an essential part of the strategy, and with the formation of the European Telecommunications Standards Institute (ETSI), work on a set of European PMR standards began, again with Peter on the frontline.

Some 20 years later, we now have a realistic set of European PMR technical standards for which products only have to be evaluated once before being placed on the entire European market. With the appropriate CEC directives also in place, licensing arrangements have also been vastly improved and harmonized. Peter looked back over this long period over which he was always a major player and was thus able to write this authoritative book using his hands-on experience.

In his final chapter, "Outlook: PMR in the Decade Until 2010," Peter uses his crystal ball very effectively to suggest a number of market scenarios for digital professional mobile radio. The prospects look like an interesting possibility for the as-yet incomplete Digital Interchange of Information and Signaling (DIIS) ETSI standard to create a European nontrunked system well matched to the needs of the smaller PMR user, leaving TETRA for medium and large operators. The important question is whether it will receive the backing of at least two or three big-name manufacturers. Another important question is whether the voice-only, high-volume, analog, short-range hand portable-to-hand portable PMR 446-type market will ever change to digital, even though it would provide the valuable benefit of lower power consumption, which is just one of several interesting features.

We will have to wait and see.

Alan D. Hudson
Retired Director of Corporate Affairs, Motorola
Director, Federation of Communication Services
Hindhead, Surrey, United Kingdom
December 2003

Preface

Why has this book been written? There are several reasons, but the main reason is because I wanted to tell readers what *professional mobile radio* (PMR) really is. Moreover, I collected a number of interesting PMR issues that were helpful for me and may also be helpful for other people deeply involved in PMR.

In my long career as a professional engineer in the field of land mobile radio, I started with analog equipment design and I did not care for system issues. I would get an equipment specification and I then have to create it at the lowest possible cost with a maximum of quality. At that time, I learned the right balance between DM (deutsche mark) and dB (decibel).

Later, I was assigned tasks related to the design of new analog and digital systems. Suddenly I was faced with the problem of optimizing an entire system by playing around with a lot of technical, financial, and project management parameters. Once I needed a clever trade-off between different modulation and coding parameters, so I approached one of my engineers and asked him if he could provide me with a rough estimate regarding the trade-off by the next day or the day after. He replied that he had to run a number of simulations, but before he could even do that he had to set up the simulation system and he estimated that he would need at least a quarter of a year to give me a sound answer. However, I needed the answer much sooner than that for making a decision. On this occasion, I saw clearly that I had to find a solution on my own and I ended up with one of the quick and rough estimates I have used so often since then. It was not precise to the last decibel, but it enabled me to make the right decision.

From this and other experiences, I learned the lesson that one must always be able to help oneself. That is possible by understanding the basics of the theory related to one's own work and being able to apply those basics quickly but exactly enough whenever necessary. The precise calculations can very often be done much later. Another point is that design engineers are using more and more complex software packages for various tasks without always knowing exactly how they work. However, they should be able to detect if, for whatever reason, such a tool fails and produces unreasonable results. Therefore, a lot of experience and a sound understanding of the fundamentals of our work are necessary. This is another reason why I tried to create a technical toolbox for my own work.

Many people perceive analog and digital radio communication systems as quite different matters. However, in real communication channels all digital signals look at first glance more or less like analog ones because they are softly shaped to gain a considerable reduction in bandwidth. This is why even in digital channels the signals being transmitted have to be handled and processed at least partly like analog signals. In particular, this is true for radio communication in the RF channel. Hence, the digital world is not replacing the analog world, but is instead complementing it. Demonstrating some of the relationships between these two worlds and how to cross the bridge from the analog to the digital domain was one more reason to write this book.

In the past few years, I have heard many discussions about the pros and cons of PMR. I recognized that it is not an easy task to define PMR and to differentiate it clearly from other radio communication services. This is particularly true if the counterpart receiving the explanation is not an expert in radio communication. Thus, a description of the properties of PMR and how it is distinct from other services may provide arguments for decision making if somebody has to make a choice.

More than 10 years ago, I started to give lectures on PMR topics in parallel to my daily work and in the past few years this became a more comprehensive part of my work. Eventually somebody asked me if I would like to write a textbook on PMR, and so some years ago I wrote a small German booklet on digital PMR and TETRA. Since then, I have been asked frequently if there would be an English version. This was the stimulus to put together a more comprehensive collection of PMR issues and to write a completely new book for a much bigger audience.

I will be happy if readers of this book take away from it an improved understanding of many PMR-related problems, in particular, the basic relations between analog and digital radio transmission. Moreover, I hope readers will be able to use one or the other of the tools presented when they have to make quickly the right decision without being able to exhaust fully the problem in question. Even though readers might not need all of the theory presented, they should be able to benefit from the main thoughts and facts presented. Last, I would like to invite readers to contact the publisher if they find errors or want to suggest improvements because I would be grateful for all of the help I can get to promote this useful tool called professional mobile radio. May this book find interested readers who can benefit from it at least a little bit.

Acknowledgments

Writing this book would not have been possible without the patience of my family and all of the colleagues who let me share their experiences during my whole professional life. In particular, I appreciate very much the kind support of Alfred Bronner, Hans Burkhardt, Roland Göttel, Geert Herold, Hans-Joachim Langermann, Philippe Mège, Gottlieb Schwarz, Brian Seedle, Gerhard Tamm, Hans-Hartmut Thiem, Wolfgang Ließ, and Christopher Wright, who all very carefully read the whole manuscript or at least those chapters concerning specific issues with which they are familiar. They all gave various hints, pointed out many errors, and drew my attention to various facts unknown to me.

I am also very grateful to Adrian Mahlkow, who checked all the formulas and proposed many improvements. Eventually David Britland, a well-known PMR expert, not only polished the text and removed countless language mistakes but also pointed out many inaccuracies and gave innumerable valuable hints to improve and update the whole text. Moreover, I am very grateful that Alan Hudson, the grand old man of European PMR, who worked for several decades in the field of PMR and promoted it on the international scene in an exceptional way, taking advantage of his great amount of experience, was kind enough to write the foreword. I would also like to thank the publisher and, in particular, Dr. Julie Lancashire, who supported and encouraged me all the time, beyond even missed deadlines, and who helped me to overcome many hurdles. Without this strong support from all of these people, this book would never have been completed.

CHAPTER 1

PMR Roots and User Requirements

What is driving mobile radio communications besides market demand? It is the fast developing semiconductor integration technology in combination with digital signal transmission and processing. These advances have changed radio communications dramatically because they allow the realization of new technical features at low cost, thus meeting the demand of a mass market.

Professional users of nonpublic mobile radios have seen the public digital cellular networks evolving quickly and successfully, but for their own work they were reluctant to change to *digital professional mobile radio* (DPMR) even though it has been available since the late 1980s. This reluctance stems from the fact that they depend on their communication tools and they cannot afford for them to fail. This might be why many professionals waited until DPMR technology became commercially feasible and mature enough to perform well even under the worst operational and environmental conditions. However, now that they have been convinced by the public networks that the time for DPMR has come, they want DPMR to be fast, reliable, and cheap.

By the way, in the past PMR was the acronym for *private mobile radio*, but this does not stress clearly enough that professional applications were intended to be the primary application; therefore, *professional mobile radio* is the better name [1].

In the middle of the 1980s, proprietary DPMR systems had been developed and launched by several manufacturers. Since about 1990, attempts have been made to specify DPMR standards. In the meantime the *Association of Public Safety Communication Officials, Project 25* (APCO 25) and *Terrestrial Trunked Radio* (TETRA) have been launched and *Digital Interchange of Information and Signaling* (DIIS) is under preparation. These standards will change the PMR market considerably. Moreover, *Public Access Mobile Radio* (PAMR) and *Specialized Mobile Radio* (SMR) operators have emerged mainly in the last 10 years and they push digital PMR to supersede analog trunked PMR systems.

Now it is time to take a closer look at the roots and properties of PMR technology, to compare it afterward with public mobile telephone systems, and to investigate why DPMR systems are going to conquer the PMR market and replace existing analog PMR technology.

1.1 Brief Review of Historical PMR Development

Analog PMR technology surfaced in the United States as early as the 1920s. The first trials were performed by Detroit Police in 1921 and then by London

Metropolitan Police in 1923. The Detroit Police started regular service in April 1928, but in the beginning they used only one-way radios and *amplitude modulation* (AM). However, as early as 1934 two-way radios had become commonplace for police. *Frequency modulation* (FM) was invented by E. H. Armstrong in 1935 and its superior quality was so convincing that already by around 1940 nearly all U.S. police organizations had converted from AM to FM [2–4].

PMR use remained very restricted for a long time, primarily for public safety reasons. In the 1950s analog PMR for civil applications began to spread in Europe, and in the 1960s and 1970s it became by far the dominating mobile radio communication technology. PMR equipment was not very complex and it was reliable and available at acceptable costs. Most of the radio sets were able to work on only one or a few channels in simplex or half-duplex mode mostly in the 4-m and 2-m bands. Eventually the 0.7-m band was opened for PMR in the 1960s. For specific applications, such as police and fire brigades, in some countries duplex radios with up to 100 channels and more were used. Microelectronics was in its infancy at that time and the radios were therefore bulky and very power consuming.

The first frequency allocations for *land mobile radio* (LMR) in the 800/900-MHz band were made by the Federal Communications Commission (FCC) as early as 1970. The introduction of SMR in the United States in 1974 was another important move by which the FCC intended to stimulate the use of spectrally more efficient trunking technology and to encourage market competition. SMR offered commercial service by private carriers (wireline carriers were excluded) for various PMR applications. This situation attracted many new users and was very encouraging for the mobile radio industry [3].

Semiconductor technology advanced and the components for radio sets became more complex at the same time as they began to shrink in size. Hence additional features became economically and technically feasible such as selective calling and data transmission, which ran with subcarrier modulation at speeds from 300–1,200 bit/s and in exceptional cases at 2.4 kbit/s. In some cases, even prior to the 1980s, 4.8 kbit/s had been achieved with direct carrier modulation based on *frequency shift keying* (FSK).

The progress in components technology allowed manufacturers to create lightweight and small paging receivers, which in the 1980s turned into very small and elegant devices with long battery duty cycles. At the beginning of the 1980s the first digital paging systems appeared and the British *Post Office Standardization Advisory Group* (POSAG) code became the Radio-Paging Code No. 1 specified by the *Comité Consultatif Internationale des Radiocommunications* (CCIR), now the *International Telecommunication Union–Radio Sector* (ITU-R). In the 1990s ERMES (originally *European* but now *Enhanced Radio Messaging System*) was standardized by the European Telecommunications Standards Institute and paging services based thereon were introduced. Paging technology demonstrated very early capabilities that later were also achieved in the DPMR sector.

Cordless telephones (CTs) that used a very simple analog technology were first introduced at low frequencies. In the mid-1980s analog cordless telephones were launched in the 900-MHz band. Finally, in the 1990s complex *digital enhanced cordless telephones* (DECTs) appeared at 1,900 MHz. The DECT standard would not have been technically feasible without highly integrated chips and *radio-*

frequency (RF) components adequate to the 1,900-MHz range. High production numbers made devices like DECTs cheap enough for entry into consumer mass markets.

Public mobile telephony was introduced in the United States as late as 1946 at 35 MHz, and less than 10 years later the first European systems appeared in Sweden and Germany; of course, most of them were manually operated at that time. However, the Swedish *Mobile Telephone A* (MT A) system, the first fully automatic system in the world, started commercial service in 1956. The first ideas for cellular systems surfaced as early as 1947, but more than two decades passed before the first fully automatic commercial cellular systems began operation in the United States and Europe [2, 5].

For even the simplest analog solutions, public mobile telephone systems require many signaling and control functions. This meant that in the late 1960s and early 1970s the only option was to build very expensive, heavy, voluminous, and power-devouring radio sets with virtually no user comfort. This began to change when competing analog cellular systems such as the *Advanced Mobile Phone System* (AMPS) in the United States, *Total Access Communications System* (TACS) in the United Kingdom, *Nordic Mobile Telephone System* (NMT) in Scandinavia, and *Net C* in Germany were introduced, which became very successful in the 450-, 800-, and 900-MHz bands. The real breakthrough for cellular technology happened in 1981–1982 when the four Scandinavian countries opened a common analog but fully automatic mobile telephone network at 450 MHz with, at that time, advanced features like handover and international roaming [3]. Very soon, market acceptance worldwide far exceeded all expectations. This huge success was mainly the consequence of improved user comfort, miniaturized components, and acceptable cost of the mobile radio sets.

Semiconductor technology and miniaturization progressed steadily, and in 1987 those advancements encouraged the real start of standardization and development of the *Global System for Mobile Communications* (GSM). The keys to GSM's incredible success were (1) setting aside a European-wide coordinated frequency band in the 900-MHz region and later another at around 1,800 MHz and (2) creating a highly sophisticated digital system with European-wide operation, capable of voice and data transmission and a variety of mobile telephone services that considerably exceeded those of the analog systems. Decreasing market prices for the mobile phones due to exploding production numbers and subscriber figures formed a self-accelerating feedback loop that has driven the market up to the present time. GSM spread successfully throughout the world and regional derivatives even appeared, such as GSM-1900 in the United States. Today's mobile phones are much more developed than ever and compared to the 1970s they are available at unbelievably low costs.

Moreover, GSM technology has proven in the meantime to be much more flexible than originally expected. *High-Speed Circuit-Switched Data* (HSCSD) and *General Packet Radio Service* (GPRS) together with changes in modulation and coding schemes by *Enhanced Data Rates for GSM Evolution* (EDGE) provide additional flexibility and improved performance. On top of it a more powerful new system is rising on the horizon. The *Universal Mobile Telecommunications System* (UMTS), also called *International Mobile Telecommunications System 2000*

(IMT-2000; formerly the *Future Public Land Mobile Telecommunications System,* abbreviated as the tongue-breaker FPLMTS) on the global level, will offer a further improved air interface employing the latest *time division multiple access* (TDMA) and *code division multiple access* (CDMA) technologies in the 2-GHz band. UMTS will be more than an advanced mobile telephone system; it will provide fast mobile Internet access and act as a multimedia transport system.

On the other hand, GSM will not disappear from the market when UMTS is launched. On the contrary, it will be developed further to compete with UMTS. The *GSM/EDGE Radio Access Network* (GERAN) with all of its mobility management functions and improved data capabilities will have properties similar to those of the *UMTS Radio Access Network* (UTRAN). Hence UTRAN and GERAN can both be connected to the same terrestrial fixed backbone network [6]. Mobile and fixed networks are melting together anyway, forming a world wide *intelligent network* (IN) for communication in which advanced mobility management will be commonplace.

In the United States standardization is managed by the *American National Standards Institute* (ANSI), which is the national counterpart of *International Standardization Organization* (ISO) and the *International Electrotechnical Commission* (IEC). Furthermore, some of the ANSI-accredited organizations should be mentioned: the *Institute of Electrical and Electronics Engineers* (IEEE), the *Electronic Industries Association* (EIA), and the *Telecommunications Industry Association* (TIA). The latter pursues EIA's telecommunications interests. The *National Telecommunications and Information Administration* (NTIA) is responsible for frequency management for all U.S. federal agencies, and last but not least, the FCC sets general standards for mobile radio concerning the use of equipment and frequencies [3, 7, 8].

In Europe the situation has changed considerably in the last 20 years. Early on, for general standardization tasks, national standardization organizations were responsible while the national administrations, loosely coordinated by the Conférence Européenne des Postes et des Télécommunications (CEPT), regulated radio matters and performed national frequency management. In 1987 the *European Telecommunication Standards Institute* (ETSI) was founded and most radio standardization tasks were then transferred from the CEPT to ETSI. Initially the *Technical Committee* (TC) *Radio Equipment and Systems* (RES) and its *subcommittees* (STCs) were responsible for nearly all radio matters except GSM and some minor topics. Later some of their tasks were transferred to TC *Electromagnetic Compatibility and Radio Spectrum Matters* (ERM) and a number of newly formed *ETSI projects* (EP), such as EP TETRA. Many other organizations are involved in national and international standardization, but space allotments allowed only the most important ones to be mentioned here.

Of course, the fast development of public radio communication including all involved standardization and regulation issues has had a major impact on the development of PMR. However, the present situation for PMR is also determined by the very specific requirements of various user groups and their diverging needs. These have led to very differently structured systems from simple and small local applications to very complex regional and sometimes nationwide multisite networks with very different properties.

1.2 Typical PMR User Groups

The world of PMR is manifold and colored due to the very distinct operational requirements of the different user groups. The most important ones are shown in Figure 1.1. Note that the figures given are only rough estimates distilled from various sources and countries to give a certain feeling about the size of the different PMR market shares [5, 8–15].

Organizations devoted to public safety such as police, fire brigades, ambulances, catastrophe recovery, mountain rescue, and technical emergency services operate large hierarchically structured networks with regional or even nationwide coverage. These networks often employ multichannel radio sets and duplex operation. Non-emergency authorities such as customs, government departments, public health, environmental protection, *postal telephone and telegraph administrations* (PTTs) also have similar networks, sometimes also operating in duplex mode.

However, the vast majority of PMR applications is of a civil nature. Examples include large organizations such as utilities for electricity, oil, gas, and water supply or public transport, including airlines, railways, commuter trains, subways, buses, and trams. These types of organizations are all running big networks with at least regional coverage and a big number of mobile stations. For the most part, they use simplex or half-duplex equipment.

Networks of smaller size are in use by manufacturing plants, construction and mining companies, and various others such as airport operators, private transportation and taxi companies, maintenance and security services, and doctors in rural areas. These smaller organizations are for the most part running local simplex or half-duplex systems. There are countless other applications, some of them just at

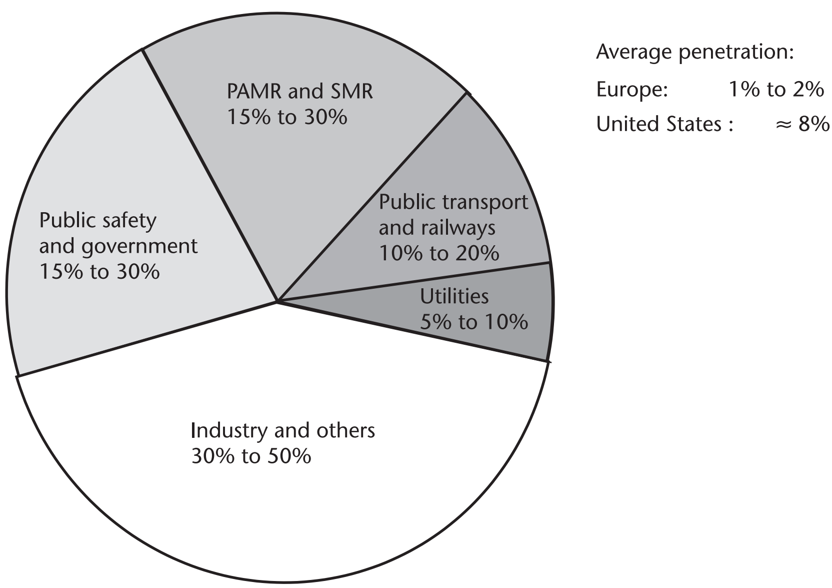

Figure 1.1 Average market shares of the most important PMR user groups. (*Source:* MRC Berlin, 2003, after various other studies from 1997–1999.)

the border of PMR, such as remote meter reading and remote machine control, to mention merely two examples.

And last but not least, many private applications exist in the leisure and sports areas that use one of the various short-range PMR services. Various applications such as hotel security systems, small airports, shopping centers, building sites, and sports and various other outdoor arenas are in this category.

PAMR or SMR is a peculiar animal right on the border between PMR and public cellular systems. It relies on PMR technology but is mostly also open to the public. The main distinction from conventional PMR is that PAMR improves the frequency efficiency by using bundles of RF channels to achieve trunking gain, but it therefore needs additional signaling and channel controls. Today many PAMR networks are based on the analog de facto standards MPT 1327 and 1343. However, in contradiction to cellular systems, these offer only half-duplex operation, there is no handover, and roaming is provided only in specific cases. The interconnection to the *Public Switched Telephone Network* (PSTN) or *Integrated Services Digital Network* (ISDN) has been available in many countries only under severe restrictions if at all, which is also true for most of the PMR applications. Another distinction from self-provided PMR is that trunked systems are mainly operated by third parties, which relieves the subscribers from all initial investments. However, monthly charges have to be paid for compensation, but these rates are usually much lower than for public cellular networks.

This is going to change with trunked digital PAMR networks based on TETRA, which have been introduced recently throughout Europe and have begun to spread over the entire world. They are capable of international roaming and can be used like public cellular networks but also provide comprehensive PMR features. Again, the subscriber need not bear heavy initial costs, but the monthly fees are not negligible.

Mobile data networks like *Advanced Radio Data Information Service* (ARDIS), *Mobile Text Transmission* (MOBITEX II), and *Motorola Data Communication* (MODACOM) are for data transmission and dedicated to specific applications only. The subscriber numbers are relatively small and the importance of these networks is fading.

Paging is seen as a specific area different from general PMR and therefore not treated here. Marine, amateur, and CB radio as well as short-range and low-power devices are also not regarded as parts of PMR, despite the fact that at least in some cases similar radio equipment is used.

1.3 Specific PMR Requirements

To find out why PMR is so versatile and unique, we must take a detailed look at the underlying operational and economical requirements and the resulting properties. Before we elaborate on these further, a short general definition of PMR should be given, even if it is nearly impossible to define such a complex matter with a single sentence [16–18]:

> PMR offers two-way radio communication carrying speech, data or a mix of both in non-public networks tailored to the specific operational needs of professional

mobile user groups for efficient and flexible communication within their area of daily operation.

The different categories of PMR users have widely varying requirements and operational needs to which their communication tool has to be matched in the best and most economical way. One of the most demanding issues is economical feasibility, which is related to a number of important points, such as these:

- Low terminal and infrastructure costs (good cost to coverage ratio);
- Easy and cheap installation of terminals and infrastructure;
- Low operating fees and costs, charges, and service and maintenance costs;
- High reliability, short nonavailability times, and long equipment lifetimes;
- Over-the-air updating and reprogramming of mobile software;
- Remote-controlled software updating of fixed network elements;
- System enlargement and upgrading;
- Analog-to-digital migration systems.

Because PMR is a professional communication tool, reliability, long-term availability for repair, parts replacement and system enlargement, upgrading to improved new networks (also from analog to digital transmission) and low operating costs are decisive.

To reduce the operational costs, low power consumption is necessary, in particular in the mobile radio terminals. Therefore, transmitting power control, energy-saving techniques, long battery duty cycles and lifetimes, and short battery charging times are important issues in terms of the operating costs.

Production numbers for radio terminals and infrastructure equipment cannot be increased arbitrarily due to the limited size of the PMR market, but in combination with the low operating costs, total PMR systems costs become remarkably low within a relative short time of operation. Moreover, there are no running call charges; therefore, good cost control can be achieved with PMR systems, whereas in public cellular and PAMR systems the charges are dependent on the traffic and, hence, are difficult to control by the user.

What is the benefit for an enterprise if it introduces PMR for internal mobile communication? According to several studies in different countries, the improvement in efficiency depends on the application and has been found to be in the range of 10–30%, and on average about 15%. For example, a small transportation company with 12 trucks will need only 10 after starting to use PMR for on-track route optimization. Moreover, depending on the penetration rate, the use of PMR increases a country's *gross domestic product* (GDP) on the order of 0.3–0.6%, which is often on the order of yearly GDP growth.

A very specific issue is that of migration from analog to digital systems. State-of-the-art analog PMR systems provide at least the possibility for further use of parts of the infrastructure when analog terminals are replaced by digital ones. However, some problems cannot be traded against costs, for example, when safety and human lives are involved or when valuable material is endangered [13, 15].

Frequency is a valuable and rare resource and therefore its utilization should be as efficient as possible, but the utilization depends on many variables:

- Demand-oriented and reliable service in coverage areas of different size;
- Service well suited to densely populated conurbations with high traffic volume as well as to large rural areas with low traffic volume;
- Simplex and half-duplex operation or, exceptionally, full-duplex operation;
- Good and reliable coverage even under bad propagation conditions;
- Coverage extension through repeaters or simulcast systems;
- Robustness against multipath propagation and the Doppler effect;
- Low or switchable transmitting power to reduce interference;
- Automatic transmitting time limitation and protocols for efficient channel use;
- Trunking, adequate modulation, and coding for good frequency economy;
- Interference-free cooperation of new (digital) and established conventional (analog) systems.

One single type of system can hardly be expected to meet the requirement to supply high traffic in conurbation areas as economically as low traffic density in large rural areas. This is a good reason to tailor different systems to these contradicting requirements. In the United States, APCO 25 and iDEN are two solutions well suited to these diverging requirements, whereas in European standardization the answer was TETRA for the first case and DIIS for the latter case.

However, all of these systems meet the requirement for good service quality under difficult propagation conditions if the coverage is properly customized to the needs. In buildings the lower floors and the cellars are particularly critical. Buildings made of reinforced concrete are difficult to penetrate, and tunnels and mines need specific means such as leaky cables. These requirements also have an impact on the economical feasibility of PMR systems.

In PMR the access to and the use of the radio channels is related to a variety of specific requirements:

- Easy and very fast channel access and short transmission delays;
- Short dialing, individual and selective calling;
- Channel busy indication or automatic collision prevention;
- Direct links between mobiles without base station or repeater involvement;
- Broadcast, fleet, group, and subgroup calls;
- Multilevel priority and emergency calls;
- Fleet control by dispatch, closed user groups, and late entry;
- Conference calls and open channels;
- Call repetition, call diverting, and automatic callback facilities;
- Transmitter interrupt to prevent faulty equipment from channel blocking;
- Automatic and manual acknowledgment;
- Handover and roaming.

Efficient mobile teamwork is based on a quick and simple connection among group members. They can all be kept informed by the use of, for the specific group, an open channel such that other users of the same carrier frequency cannot participate or interfere; hence, mutual annoyance is avoided.

The main items concerning efficient and immediate radio channel access and utilization are simple and fast or extremely fast channel access and short transmission delay. In particular, the latter is an issue of major importance with regard to digitized speech for specific applications, such as for police and fire brigades in dangerous situations. In many applications, fast channel access is provided by simply pressing the *push-to-talk* (PTT) button. Additionally, flexible and reconfigurable individual, group, and fleet call facilities, priority and emergency calls, identification, acknowledgments, open channel, conference and broadcast calls, and dispatch-controlled and direct mode operation are often demanded. New features that are becoming increasingly important are handover, roaming, flexible, and over-the-air reconfigurable numbering schemes and group organization and fleet management.

The preceding list comprises a mixture of typical PMR requirements and various telephone services. These are somewhat related to operational requirements, but there are additional ones:

- Ownership of network and full control over network operation;
- Fall-back facilities during malfunction or breakdown of network control;
- Access to terrestrial fixed or mobile telephone networks (PABX, PSTN, ISDN, IDN, and so on);
- Flexible and reconfigurable group building, preferably over the air;
- Status and short message transmission;
- Powerful error protection and correction;
- Good speech intelligibility and quality even under poor propagation conditions;
- Transparent and nontransparent data transmission of different kinds;
- Voice and data encryption at the air interface or end to end;
- User friendly and secure encryption key management.

System ownership and control are very important issues that are related to customized coverage, features and facilities, guaranteed performance, and cost control.

In many applications, it is very important that malfunctions or network control breakdowns do not block radio communication totally, in particular in catastrophic situations and when human lives are endangered. Direct mobile-to-mobile links are one of these possibilities to run at least a local basic service. Another is that base stations are able to operate autonomously and keep local services running in case network nodes or fixed links between them fail.

The need for a large range of voice and data services including interworking with other networks like PSTN, the Internet, and intranets via gateways is inevitable today. The most important data services include fax, status and other short or precoded messages, e-mail, database access, packet data transmission, file transfer, and e- and m-commerce transactions. Satellite and public mobile telephone networks may be accessed via the fixed network so that direct gateways are not necessarily needed.

Public safety and business transactions require encryption to ensure privacy of voice and data transmission. A weak point in many systems is key management, which should be secure but nevertheless user friendly. Moreover data compression, picture, and slow scan video transmission for the surveillance of endangered objects and so on are important for the usefulness of modern DPMR networks.

Finally PMR has specific application-related requirements that are concerned in particular with the properties of the mobile equipment:

- Hands-free operation of mobile radio terminals;
- Identity cards or modules for authentication;
- Remote disabling of lost or stolen mobiles;
- Data and control interfaces;
- *Man/machine interface* (MMI) suited to tasks and user requirements;
- Heavy-duty terminals for operation under severe environmental conditions;
- Operation in dangerous environments such as an explosive atmosphere;
- Long battery duty cycle (at least 1 working day).

PMR equipment in many cases has to resist severe environmental conditions such as shock and vibration, an extended temperature range, high humidity, dust, and chemicals. On tank ships, in oil refineries, and in chemical plants the equipment must be safely operated in explosive atmospheres. Thus, equipment for such specific applications must be carefully prepared by means of appropriate construction techniques, but it is needed only in limited quantities—factors which of course drive up costs.

These lists of features and requirements are by far not exhaustive. For each real application, another combination of properties is needed but never all of them simultaneously. In former times this requirement led inevitably to a big number of expensive distinct hardware variations. Hence, a flexible technology base at affordable cost is needed that can provide a number of standard building blocks to create all hardware and software modules needed for a specific application [17, 19, 20]:

> Today many of these desired features are not available from public radio telephone networks and it is questionable if even some of the most demanded of them will be provided by future public mobile telephone systems. If such functions are indispensable for users to fulfill their operational needs, then there is no alternative to PMR for them.

1.4 Basic PMR System Configurations

To understand better how the different needs of the various user groups and applications are being fulfilled, we should take a look at some basic PMR system structures. A number of elements exist from which all of the necessary system configurations can be composed according to the varying technical needs and services required in an acceptable cost framework [3, 7, 21, 22].

The simplest and cheapest case is the mobile radio link or direct mode between two *mobile stations* (MSs) or *hand portables* (HPs), which usually run in simplex mode on one single carrier frequency f_1. Such links have only limited coverage but they can be set up quickly and are well suited to tasks performed by a small group in a local area. *Simplex mode* means that one person is speaking while another one or parts of the group or the whole group is listening. The state of the mobiles is controlled by a transmit key (PTT) to be pressed by the person who starts to speak. Of course, the coverage area of all stations involved in a call must overlap.

If a larger coverage distance is needed than is achievable in the direct mode, then a *repeater* (RE) can be used. Assuming the allocation of one channel per carrier frequency, a simple FDMA repeater needs to operate on two frequencies, receiving one frequency f_1 and retransmitting each message simultaneously on another frequency f_2. However, for TDMA systems a single frequency repeater can be made, but this is an issue for Chapter 6. Consequently, all mobiles in such a *Frequency Division Multiple Access* (FDMA) system are working in simplex mode but on two different frequencies for reception and transmission as illustrated in Figure 1.2. This mode therefore is called *semiduplex* or *half-duplex*. Accordingly, four frequencies are needed for full-duplex mode.

In many cases there is a central *base station* (BS) with a dispatcher who controls a group of mobiles. This can be performed in small systems in simplex mode on one frequency f_1, but often it is done in semiduplex or full-duplex mode on two frequencies f_1 and f_2. As Figure 1.3 shows, the configuration can be much more complex if additional base stations or repeaters belong to the network, if gateways to the PSTN or to other fixed or mobile networks are present, if direct calls between mobiles is a requirement, or if a combination of all of these requirements is needed.

The communication from the base station to the mobiles takes place on the so-called downlink using a first frequency f_1. Usually this is the higher frequency separated from the second frequency f_2 used in the reverse direction on the uplink by the duplex separation Δf_D. In many cases the communication is performed only between the mobile and the dispatcher. Alternatively, all communication between the mobiles runs via the base station, which is used as a repeater only or in a situation where the dispatcher keeps listening and may step in at any time. In addition, if permitted by the system, direct calls between mobiles can be made on

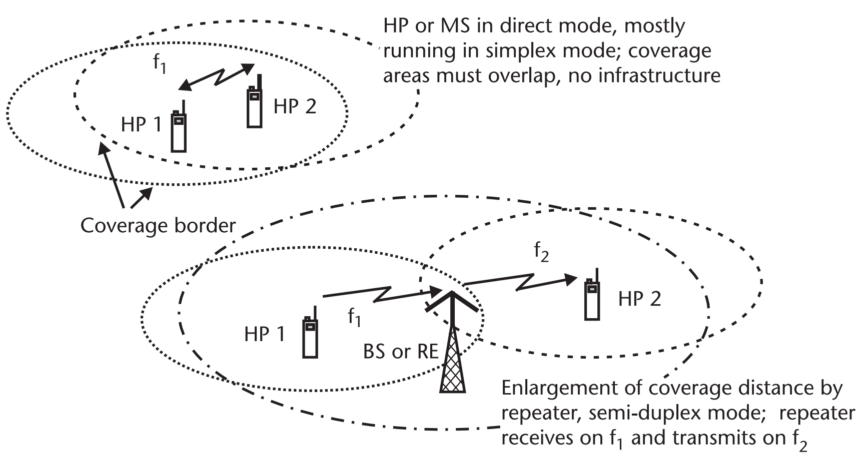

Figure 1.2 Direct mode and repeater operation.

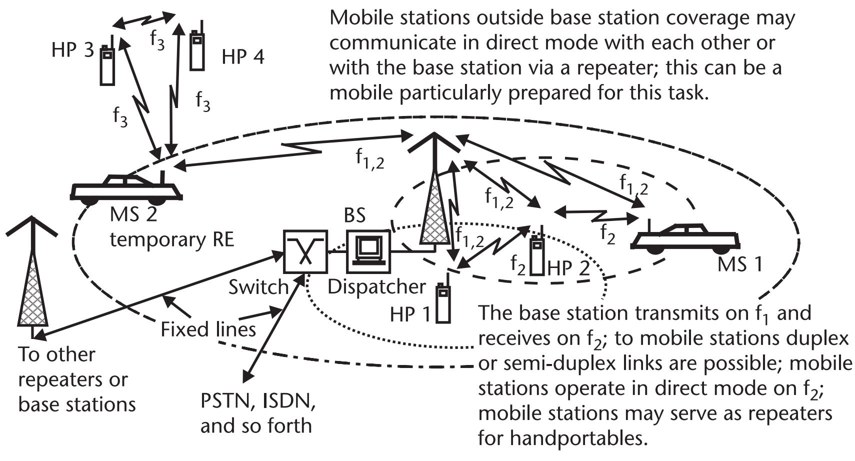

Figure 1.3 Single and multisite systems.

the uplink frequency f_2. Sometimes car-borne mobiles can also be used temporarily as repeaters. Hence, HPs out of base station range can be reached via such a mobile at the coverage border. However, the direct mode between out-of-range mobiles, HP3 and HP4 in Figure 1.3, for example, needs a third frequency f_3. Repeaters, which should appear to all mobiles like the base station itself, may be connected to the base station via a wireline or an additional frequency pair.

Finally, in large systems with many subscribers, only one carrier frequency pair for both uplink and downlink tasks does not suffice to carry all traffic and, thus, multichannel systems are built. Then the traffic must be carefully distributed to the different channels to obtain a balanced load on all of them, which is the necessary precondition for the most effective use of the system in the main duty hour. This requires a sophisticated automatic protocol to serve all users quickly and with a minimum of interference, waiting time, and lost calls or blockage.

People do not always operate their radio sets in duplex mode like a telephone. Why is that? The reason simply is that simplex or semiduplex radios can be built much cheaper for several technical reasons. We list only the most important ones: Instead of a simple antenna switch, a duplex filter and careful shielding and decoupling of the receiver and transmitter require additional effort to avoid blocking of the receiver during transmission because the transmitting power may exceed weak receiving signals by up to 160 dB, and two different frequencies for the receiver and transmitter have to be generated simultaneously, whereas for a simplex radio only one at a time is needed.

The frequency assignment in simple systems is no problem. However, in more complex cases it might be a very difficult task to achieve a maximum of communication capacity with a minimum of frequencies. This is of importance because the availability of mobile frequencies is very limited and the frequency managers and regulators for the most part handle the frequency assignments very restrictively. Therefore it becomes clear that major difficulties are introduced by direct mode and repeater use in centralized systems controlled by a base station. In particular, in full-duplex systems the number of frequencies is increasing dramatically: A half-duplex repeater operates on two frequencies, whereas a full-duplex repeater needs four carrier frequencies.

Figure 1.3 has already indicated that a simple system may be extended to a multisite system. To avoid interference between the different base station signals at the mobiles, the base stations must use different carrier frequencies. Thus, the frequency assignment in multisite systems with many channels per base station needs specific engineering skills to minimize the number of frequencies needed, to limit interference from adjacent channels and intermodulation, and to maximize the communication throughput [23].

Sometimes PMR systems have to be tailored to particular needs. This applies, for example, to public safety tasks with small numbers of users in large service areas that far exceed the coverage of one base station. Then a multicast system in which all base stations are transmitting on the same frequency might be a beneficial solution. However, all base station carrier frequencies must be well synchronized or they must have a small but precisely controlled frequency offset to minimize interference in the overlap regions of different base stations. Moreover, the modulation must not be differently delayed by the fixed lines connecting all base stations with the dispatch center; this requires well-adjusted line delay equalizers.

Other problems appear for mobile communication in tunnels and mines, in particular if a part of the communication happens on the surface and another part underground. Interference appears mainly at those places where trains or cars enter or leave a tunnel mouth. The propagation in tunnels is usually bad, particularly at low frequencies. Therefore, leaky cables are used most often, but unfortunately their loss allows the equipment to bridge only relatively short distances and thus specific means and skills are needed to build properly working large tunnel networks [24].

Our intent here was not to describe all possible cases, but merely to give the reader an idea about the many different system configurations PMR comprises and how it has to solve very different and sometimes also difficult communication tasks. Finally, the reader should note that digital systems with more than one traffic channel per carrier offer additional design and system layout flexibility. Also the frequency assignment can sometimes be done with fewer restrictions as will be explained in Chapters 5 through 8.

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CHAPTER 2

Alternatives to PMR

At present many different land mobile radio services exist that have differing but partly overlapping properties. Taken together, they cover all reasonable needs for professional mobile communication, but none of them meets all of the requirements and covers all possible applications simultaneously. In Chapter 1 we showed why PMR is inevitable for many tasks. When making the final decision to choose one or the other system, it is necessary to compare their main properties more in detail. In this chapter some guidance for conducting such an analysis is offered [1–3].

2.1 Traffic Scenarios and Types of Systems

Disregarding cordless telephones and paging, the different radio communication systems can be grouped according to their service area size and traffic volume as shown in Figure 2.1 [2]. PMR with its small service area and low number of users per system is one of the extremes. Public cellular systems with nationwide or continental service areas and very high subscriber figures represent the other one. Specific PMR networks also exist with large service areas but low traffic volume such as systems devoted to public safety or railway on the one hand and trunked PAMR systems with limited service areas but relatively high subscriber density on the other.

To fulfill the very divergent requirements of the various professional user groups, different types of PMR systems are needed, of which the most important categories are wide-area PMR and PAMR in contrast to short-range PMR and *short-range business radio* (SRBR) [4].

SRBR is a relatively new and fast growing analog service that utilizes simple, low-cost equipment based on analog PMR technology. In Europe this specific service was introduced in the second half of the 1990s in the United Kingdom and experienced overwhelming success there. Based on MPT 1383 and EN 300 296, as of the end of the 1990s, it had been harmonized throughout Europe as PMR 446. There are no base stations, but only groups of walky-talkies that operate on eight preassigned 12.5-kHz channels in the 446.0- to 446.1-MHz range with coverage restricted to 100m inside buildings and up to 400m outdoor due to the integral antenna and the limited transmit power of 500 mW (27 dBm). For operation, only a general license is necessary [5–8]. *Family radio service* (FRS) is a similar two-way radio service in the United States also with a transmitting power of 500 mW. It is operated on 14 frequencies with a channel separation of 25 kHz in two

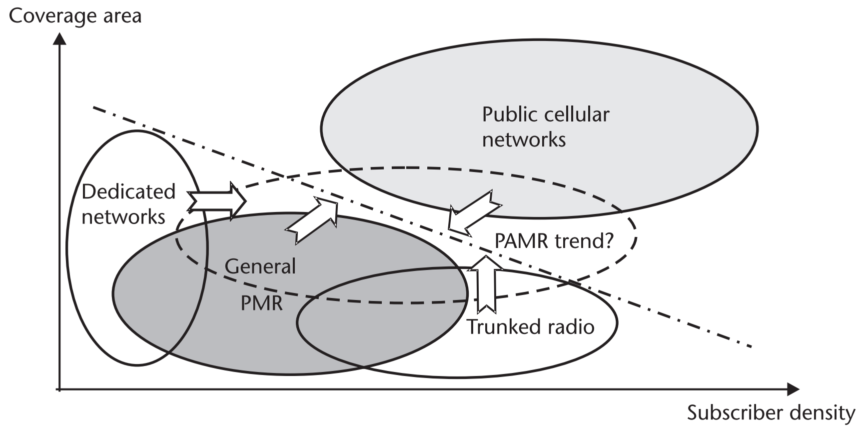

Figure 2.1 Traffic scenarios.

frequency blocks from 462.5625 to 462.7125 MHz and from 467.5625 to 467.7125 MHz.

Short-range PMR comprises very different applications. On-site PMR provides radio communication at an organization's premises for distances of typically less than 3 km, meaning that transmitting power and antenna height and gain are very often limited. The mode of operation is predominantly point-to-multipoint operation with simplex or half-duplex equipment. This is a domain of small systems with just a few mobiles. Sometimes it is also called *local communications,* and there is an overlap with on-site paging, which more and more also provides simple two-way speech and also frequently employs point-to-point operation and links to PABXs. An individual license is usually needed.

Wide-area PMR usually covers distances above 3 km and comprises single and multisite systems with one or more nontrunked or trunked radio channels. The number of mobiles per system can vary significantly, and up to 100 mobiles can be served per radio channel. Voice transmission is its predominant use, but short messages and data transmissions mixed with voice and automated dispatch are increasing. Many systems are operated in simplex or half-duplex mode, but a considerable amount of duplex operation is also seen. Very often point-to-multipoint communication is needed. With a transmitting power of 5W (37 dBm) and a base station antenna height of 30m in urban areas (depending on the frequency), distances up to 10 km per base station can be covered. In open areas and with high-gain base station antennas on multifloor buildings or hilltop sites, distances up to 50 km can be achieved in the lower PMR frequency bands. An individual license is usually required. PSTN access is not permitted or is at least very restricted, but this is going to be relaxed now in many countries.

PAMR and *SMR* offer professional users public access to trunked regional or nationwide systems. Usually there is no communication among the different user groups within one system, but a wide range of operational features needed, partly including PSTN access, is offered. An operator's license is necessary to run such a system.

Analog SMR and PAMR have been introduced in the United States and Europe to conserve spectrum and promote competition. Its development in the 1990s slowed down unexpectedly, particularly in Europe. This might have occurred for various reasons, among which the fast and successful rollout of GSM is surely not the least. Moreover public digital cellular networks became able to meet the requirements of a considerable number of the PMR users and, therefore, attracted many of them. At the same time a new strong demand for professional mobile communications emerged in many Asian countries that was driven by fast economic growth and a strong demand for mobile communications. This resulted in a renaissance of trunked radio systems, in particular in regions where public cellular systems were not yet available or were underdeveloped. Again the benefits of PMR played an important role: quick and easy channel access and comfortable group communication. The use of protocols such as MPT 1327, *logic trunked radio* (LTR, an open protocol introduced by E. F. Johnson in the United States), and SmarTrunk (a low-cost protocol from SmarTrunk Systems, United States) spread. In contrast to Europe, trunked systems in the United States, that is, SMR and since 1993 *enhanced SMR* (ESMR), a digital trunked system based on iDEN and operated by Nextel, still show considerable growth [9–11].

Recently, a new trend has evolved to install large digital PAMR networks based on the TETRA technology in many European countries and all over the world. Some people expect large digital PAMR networks that straddle the border between PMR and cellular services to emerge that satisfy many PMR requirements and also provide mobile phone services such as public cellular networks, as suggested by Figure 2.1.

2.2 Competition Between PMR and Public Cellular Systems

The world of radio communications would look very differently without GSM, which has triggered the transition from analog to digital technology in mobile radio communications. Now PMR—trunked or not—has also entered the transition phase from conventional analog to digital radio technology. It is the fastest and most extensive change to the nature of PMR since decades, and it seems to be more a revolution than an evolution.

Public mobile telephone networks are by definition nothing else but telephone networks with added mobility. They offer all of the services of terrestrial fixed telephone networks, but in contrast to them they are only capable of limited data speed. In Chapters 3 and 5 we explain why the radio path is a bottleneck that provides only a limited bandwidth in contrast to the much wider bandwidth of fixed networks.

Conventional analog and modern digital PMR networks have many properties in common although they differ considerably from cellular systems. Modern digital PMR and digital cellular such as GSM are systems with a very comprehensive and complex set of features, but they exhibit also many differences and they have different limitations. To gain better insight, public cellular and PMR systems will be compared based on the facts discussed in Chapter 1 but under slightly different headers now. It is necessary to regard the basic functions and all services and features point by point to make a sound decision about which of the different systems should be selected for a particular application. Unfortunately, in many cases there are only a few or even no killer criteria and thus there is no escape from a detailed analysis [1–3]. The comparison concerns mainly standardized DPMR and standardized digital cellular systems. However, we will show in Chapter 8 that many characteristics of proprietary digital PMR systems are very similar to those of the standardized systems that have been used as examples.

To provide a first rough overview, Table 2.1 lists some basic transmission modes and the main propagation properties. We assume for all comparisons that *analog PMR* means pure analog radios or systems that may include a set of sophisticated functions and services like those of MPT 1327. The main limitation of these

Table 2.1 Basic Transmission and Propagation Properties

System Characteristic	*Analog PMR*[1]	*DIIS*	*TETRA*	*GSM 1 and 2*	*GSM-R and 2+*
User data rate/ time slot[2]	≤1.2 kbit/s	≤9.6 kbit/s[2]	≤7.2 kbit/s[2]	≤9.6 kbit/s	≤59.2 kbit/s[3]
Packet data	Rare	Yes	Yes	No	Yes[4]
Digital speech	No	Yes	Yes	Yes	Yes
Efficient short message transmission	Yes	Yes	Yes	No	Conditional
Status transmission	Yes	Yes[5]	Yes; 32,768	No	Yes
Short messages	≤100 characters	Yes[5]	SDS[6]; ≤255 characters	SMS[7]; ≤160 characters	SMS[7]; ≤160 characters
Maximum usable mobile speed	Not critical	150 km/hr[8]	200 km/hr	250 km/hr[9]	250 km/hr[9,10]
Permissible delay spread	Not critical	50 μs	100 μs	16 μs	16 μs
Frequency bands [MHz]	30–520	30–520[11]	380–470; 900[12]	900; 1,800; 1,900[12]	450[12]; 900; 1,800; 1,900[13]
Coverage distance[14] at					
80 MHz	Extraordinary	Extraordinary	Not applicable	Not applicable	Not applicable
160 MHz	Very good	Very good	Not applicable	Not applicable	Not applicable
450 MHz	Fair	Fair	Fair	Not applicable	Fair[12]
900 MHz	Not applicable	Not applicable	Moderate	Moderate	Moderate
1,800 MHz	Not applicable	Not applicable	Not applicable	Poor	Poor

[1]Without or with five-tone sequential or MPT 1327.
[2]Transparent data.
[3]With EDGE.
[4]With GPRS and EGPRS.
[5]Not yet specified.
[6]Short data service.
[7]Short message service.
[8]Small degradation at 250 km/hr.
[9]125 km/hr at 1,800 MHz.
[10]500 km/hr for GSM-R.
[11]Very small degradation above 500 MHz.
[12]GSM-400 versions between 450 and 500 MHz were under consideration but have been skipped.
[13]Allocations and width of GSM bands are widely varying.
[14]Mainly frequency range dependent.

conventional systems is their very restricted data rate, which prohibits digital speech transmission of acceptable quality.

Of course, good voice quality is of interest, but for any PMR application intelligibility, in particular under bad propagation conditions, is of major importance and for many applications speaker recognition is inevitable. One of the most important examples is police organizations, where it must be ensured that only authorized persons can give commands. Additionally digital voice allows privacy by encryption. Short and status messages are other interesting features because these can accelerate the exchange of routine messages and save transmission capacity that becomes available for other purposes.

Two issues are of particular interest. First, the vulnerability of digital transmission to multipath distortion increases with the bit rate and mobile speed and requires better channel equalization and stronger coding for improved error correction. Hence, one of the major drawbacks of GSM is the limited delay spread compensation capability due to the high symbol rate, whereas PMR systems are much more robust against multipath impairment. Second, coverage is not very dependent on the system properties, but is heavily dependent on the carrier frequency, as will be explained in Chapter 3, provided that the transmitting power and the receiver sensitivity (in short, the link budgets) are comparable.

In Table 2.2 the elements of efficient frequency usage are compared. This is mainly a point of importance for administrators and frequency managers because frequencies are a very rare resource. However, users should also ask for frequency-economic systems because wasting of capacity may later lead to market blocking as a result of a lack of frequencies. Thus efficient modulation and coding schemes and efficient transmission protocols are prerequisites for modern digital PMR systems to achieve the necessary extraordinary good frequency economy, which is further improved by efficient group call schemes that allow many users to be addressed in parallel. Channel access and call establishment times are much shorter in PMR networks, which is necessary because the average call duration is also much shorter. By the way, comparing the frequency economy of different systems is not an easy task, but this is also explained more in detail in Chapter 3.

Although duplex operation is commonplace in public mobile radio telephone networks, it is the exception in PMR and is used only in specific cases because duplex radio sets are more expensive and PMR users are familiar with PTT-controlled radio operation.

Another important requirement is that a breakdown of network control or of links between network nodes should not paralyze system operation totally. Thus a fall-back scheme must be available in many cases to ensure that radio communication can be provided even during network malfunction, although this may be possible only with a restricted set of features. This can be achieved by direct mode and local BS service if the links from the BS to other parts of the network do not work.

Fast or even extremely fast channel access is of paramount importance for many PMR applications, for instance, to control a taxi fleet or for tactical police operations. However, fast PTT-controlled communication in open channels requires much discipline, but the channel busy indicator helps to avoid collisions. In trunked

Table 2.2 Efficient Use of Frequencies

System Characteristic	*Analog PMR*[1]	*DIIS*	*TETRA*	*GSM 1 and 2*	*GSM-R and 2+*
Channel separation [kHz]	12.5, 20, 25	12.5, 20,[2] 25[2]	25	200	200
Channel access	FDMA	FDMA	TDMA	TDMA	TDMA
Channels per carrier	1	1	4	8 or 16[3]	8 or 16[3]
Duplex operation[4]	Rare	Yes	Yes	Yes	Yes
Simplex, half-duplex	Majority	Yes	Yes	No	Yes
Repeater operation	Yes	Yes	Yes	Specific cases only	Conditional
Simulcast operation	Yes	Yes	Yes	No	No
Transmitting time limitation	Possible	Possible	Possible	No	Possible
Transmission protocol efficiency	Low to high	Medium to high	Medium to high	Medium	Medium
Modulation and coding efficiency	Very low	High	High	Moderate	High[5]
Regional trunking	Yes	No	Yes	No	Yes
Frequency economy[6]	Medium	High	High	Low	Low to medium[5]

[1]Without or with five-tone sequential or MPT 1327.
[2]Standardization on demand.
[3]Full-rate channel provides two half-rate channels.
[4]Usually not for group communication.
[5]With EDGE and GPRS.
[6]According to CEPT ERC Report 52, for GSM with normalized channel separation

and digital systems, the transmission protocol usually automatically resolves collisions in cases of unsolicited channel access. Channel access issues are compared in Table 2.3.

In contrast to analog speech transmission, all digital systems need a considerable amount of time to process and transmit digital speech. This processing time increases with decreasing transmission rates because efficient data rate reduction needs very sophisticated speech coding algorithms, which require an enormously increasing amount of computational power for a small reduction of the transmission rate.

For many applications privacy is urgently required. Privacy can be achieved by scrambling or much more securely by encryption of speech on the air interface or end to end. Except for some few and specific cases, analog systems usually offer only speech scrambling if privacy is an issue at all. However, strong encryption of speech and data is available in nearly all modern DPMR systems, whether they are standardized or not. Additionally there is also often a strong need for identification and authentication, which is provided by most DPMR systems.

Dispatcher control of a fleet of mobiles is used in many PMR applications but not in point-to-point mobile telephone links. This is also valid for broadcast transmission of voice and data messages.

Table 2.3 Channel Access

System Characteristic	*Analog PMR*[1]	*DIIS*	*TETRA*	*GSM 1 and 2*	*GSM-R and 2+*
Channel access[2]	PTT, short dialing	PTT, short dialing	PTT, short dialing	Call number dialing	PTT, call number dialing
Channel busy indication	Yes	Possible	Possible	No	Yes
Channel access time	30 ms–1 second	≤360 ms[3]	0.3–1 second[4]	4–10 seconds	300 ms[5], 1 second[6]
Speech processing delay	<5 ms	<300 ms[3]	≤210 ms	90 ms[7]	90 ms[7]
Privacy	Scrambling, rare	Scrambling[8]	Encryption[8]	Encryption[8]	Encryption[8]
User identification and authentication	Rare	To be specified	Yes	Yes	Yes
Dispatcher control	Yes	Yes	Yes	No	Yes
Broadcast speech/ data transmission	Conditional	Yes	Yes	No	Yes
Fall-back operation[9]	Yes	Yes	Yes	No	Possible

[1]Without or with five-tone sequential or MPT 1327.
[2]PTT = transmitting key.
[3]Estimation, not yet specified.
[4]Infrastructure dependent, 300 ms for intracell call setup.
[5]With preassigned channel.
[6]With ASCI.
[7]With full-rate speech coder.
[8]End-to-end encryption possible.
[9]Direct mode, automated repeaters, and BS.

In many PMR systems, it is of paramount importance to ensure that the system cannot be blocked by unimportant call attempts in cases of high traffic peaks or, in particular, during a catastrophe or the recovery from it. This is achieved in PMR by priority schemes that preferably have several levels. Of course, emergency calls have the highest priority and must always be served immediately. Several additional features concerning channel use are compared in Table 2.4.

Table 2.4 Channel Use

System Characteristic	*Analog PMR*[1]	*DIIS*	*TETRA*	*GSM 1 and 2*	*GSM-R and 2+*
Emergency calls	Yes	Yes	Yes	Yes	Yes; <1 second
Multilevel priority calls	Rare	Yes	Yes	No	Yes
Direct mode	Yes[2]	Yes	Yes	No	No
Direct mode via repeater or BS	Yes	Yes	Yes	No	Yes
Roaming[3]	Rare	No	Yes	Yes	Yes
Handover[4]	Very rare	No	Yes	Yes	Yes

[1]Without or with five-tone sequential or MPT 1327.
[2]Not for MPT 1327.
[3]In TETRA, this is called *migration*.
[4]In TETRA, this is called *cell reselection*.

Many PMR applications need direct links between mobiles either for point-to-point or group call applications. Hence direct mode is then inevitable and very often it must be combined with open channel operation. An example is use by a fire brigade when infrastructure coverage is not reliable or even not available; very often a remote central dispatcher is not appropriate due to fast changing conditions during the operation of the group, for example, in dangerous situations in mines or tunnels for rescue operations in which human lives are endangered.

Roaming and handover are very important for large public networks. In PMR they are not necessary in many cases due to the restricted area of operation of the particular service. Even if roaming is needed, in the case of large cells the necessity for handover will occur very infrequently because the calls usually are short and the probability of a cell change during a call is therefore low. However, this is going to change with the increasing mobility of the users, and modern DPMR systems can provide these features if needed. The additional means to provide them are complex and costly, and this is one of the reasons why most conventional PMR systems do not offer them.

Table 2.5 lists some of the services that are usually available and shows that modern DPMR systems are well suited to group communication. One user requirement might be specifically addressed—that of *closed user groups* (CUGs). These allow a radio communication system to serve several distinct groups without mutual interference or annoyance.

However, for all telephone-like services only small differences exist between cellular and modern DPMR systems. When we compare all available services, it turns out that Table 2.5 shows many more commonalties between the digital systems than between analog and digital PMR systems. The reason is that digital

Table 2.5 Available Services

System Characteristic	*Analog PMR*[1]	*DIIS*	*TETRA*	*GSM 1 and 2*	*GSM-R and 2+*
Individual calls	Yes	Yes	Yes	Yes	Yes
Group calls[2]	Yes	Yes	Yes	No	Yes[3,4]
CUGs	Yes	Yes	Yes	No	Yes
Conference calls	Possible	Yes	Yes	Limited	Yes
Late entry	Possible	Yes	Yes	No	Yes
Open channel	Yes	Yes	Yes	No	Yes
Call repetition	Possible	Yes	Yes	Yes	Yes
Call diverting	Possible	Yes	Yes	Yes	Yes
Automatic call-back	Yes	Yes	Yes	Yes	Yes
Transmitter identification	Yes	Yes	Yes	Yes	Yes
Manual acknowledgment	Yes	Possible	Possible	No	Possible
Automatic acknowledgment	Yes	Yes	Yes	Yes	Yes
PSTN/ISDN access	Rare	Yes	Yes	Yes	Yes

[1]Without or with five tone sequential or MPT 1327.
[2]Reconfigurable groups during service.
[3]Also manually.
[4]Maximum of 10 users.

transmission offers more flexibility, based on which a more comprehensive set of functions and services can be built. However, conference calls and open channels are functions available even in simple analog PMR systems but not in the basic GSM functionality. The latest GSM variations, however, have begun to catch up and GSM Phase 2+ and GSM for railways (GSM-R) are more powerful in this respect. Table 2.6 compares the operational and economical characteristics of different mobile systems and shows the considerable differences between PMR and public systems.

As far as the utilization of frequencies is concerned it is important to remember that reliable coverage in the service area is of major importance. If the user group in question owns the system, then the coverage can easily be tailored to its needs. In particular, difficult propagation conditions can be mastered and local coverage problems can be solved individually. If, however, the system is run by an operator, the situation is much different because the operator will only resolve such specific problems when additional profit is involved, if the operator is willing to offer a solution at all.

Table 2.6 Operational and Economic Characteristics

System Characteristic	*Analog PMR*[1]	*DIIS*	*TETRA*	*GSM 1 and 2*	*GSM-R and 2+*
Economical at low subscriber density	Yes	Yes	Conditional	No	No
Economical for large coverage area	Yes	Yes	Conditional[2]	Conditional[2]	Conditional[2]
Optimal traffic density	Low to medium	Low to medium	Medium to high	High	Medium[3] to high
Tailored coverage	Yes	Yes	No[4]	No	No[3]
System ownership	Yes	Yes	No[4]	No	No[3]
Operational system control	Yes	Yes	No[4]	No	No[3]
Initial system invest[5]	Yes	Conditional[6]	No	No	No
Mobile equipment costs	Medium	Medium	Medium	Low[6]	Low[6,7]
Ongoing operational costs	Low	Yes	Medium[8]	High	High
Total costs[9]	Medium	Medium	High	Very high	Very high
Heavy-duty equipment	Yes	Yes	Yes	No	Conditional[3,10]

[1]Without or with five-tone sequential or MPT 1327.
[2]At high traffic density.
[3]GSM-R only.
[4]Not in networks operated by a third party.
[5]For the total system.
[6]Subsidized.
[7]Very high for GSM-R.
[8]For PAMR.
[9]After, say, 5 years.
[10]Specific GSM-R design.

Besides operational considerations, commercial considerations are also of major interest when the decision is to be made as to which system should be introduced. Unfortunately, GSM has raised unrealistic expectations concerning PMR terminal prices. However, the PMR market is more than one magnitude smaller than that of GSM. Hence the first look into PMR terminal costs might be strongly misleading if these are compared with those of GSM terminals, which are produced in very large numbers. Moreover, their price of, say, $1, £1, or €1 is subsidized by the operators and has to be regarded together with the monthly charges and the particular tariffs, which depend on the type of contract chosen. In many countries the ongoing operating costs due to fees and charges have been and remain much smaller for PMR systems than for public mobile phones. On the other hand, the use of PMR requires a high initial invest and therefore the break-even in total costs can only be achieved after some years of operation.

Additional financial issues must be considered, such as costs for maintenance and repair, equipment replacement and upgrading, and the extension of features. Full system control and the reliability of the service are in specific cases of paramount importance, for example, for public safety and similar applications. Moreover reliable radio communication may save operators money due to shortened reaction times for maintenance or repair of machines, highly perishable goods, or even human lives [12].

PMR radios are usually designed for professional use and they must be operated very often under unfavorable environmental conditions, in contrast to mobile phones for public networks. Of particular importance is the MMI. The equipment must be easy to use and reliable to operate, for example, when the user has to operate the radio set while wearing gloves. Volume and weight must also be matched to practical use.

Today another question concerning the necessary investments for the transition from conventional analog to modern DPMR systems has arisen. In many cases complete system replacement within a short time interval is not feasible due to cost reasons. Therefore, it is beneficial if the manufacturer offers the option of keeping at least parts of the system rather than replacing the entire system. Another solution would be *multimode equipment* as it is currently called, although *multi-standard equipment* would be a better and more appropriate name. This issue is taken up again in Chapter 9.

2.3 Impact of Advanced GSM Versions and UMTS

There is no doubt that the competition of public mobile telephone networks will have a strong influence on the future evolution and success of PMR; in particular, increased data rates and improved data services together with low terminal costs will attract many PMR users. The situation is complicated further by the fact that GSM Phase 2+ and GSM-R, the last releases of the GSM standard, offer the *Advanced Speech Call Items* (ASCI), which are additional PMR-like features [13]. Finally, GSM now also provides *virtual private networks* (VPNs), which means that the user of such a VPN is carefully separated from other user groups.

The ASCI features permit improved group communication. Queuing and multilevel priority features are available to avoid and resolve call collisions. The *enhanced Multilevel Precedence and Preemption* (eMLPP) service provides accelerated call establishment in three classes with 1–2.5-, 5-, or 10-second duration and it distinguishes seven priority levels for individual and group calls, five from ISDN, and two additional ones for mobile purposes. *Voice Group Call Service* (VGCS) allows group communication, but without queuing, while *Voice Broadcast Service* (VBS) is suited to voice and data transmission in group communication, also permitting late entry and acknowledgments. In VGCS and VBS, a virtually unlimited number of mobiles can listen on the downlink to a speaker active on the uplink, as in PMR group communication. This results in improved frequency utilization because it is no longer necessary to assign to every group member a full-duplex channel. Of course, the mode of operation is then half-duplex.

With some delay GSM radios for heavy-duty purposes equipped with a robust housing such as those needed for many PMR applications have become available in the past few years. GSM products for professional use under rough environmental conditions that also offer the latest GSM features are available; one in particular is GSM Pro by Ericsson [14].

While GSM-900 has been designed for a maximum mobile speed of 250 km/hr, meaning 125 km/hr for GSM-1800 due to the increased impact of the Doppler effect as explained in Chapter 3, GSM-R works up to 500 km/hr as needed for fast trains. It has been promoted by the *European Radio Enhanced Network Project* (EIRENE) run by the *Union International des Chémins des Fer* (UIC). However, GSM-R is only available for railways in the European bands of 876–880 MHz paired with 921–925 MHz.

With *Circuit-Switched Data* (CSD) the initial GSM data rate was restricted to 9.6 kbit/s but later it was extended to 14.4 kbit/s by the introduction of a new coding scheme. Moreover it has become possible to assign more than one time slot simultaneously to a single user by means of HSCSD. The next step was the introduction of packet transmission by GPRS, and after that the new modulation scheme EDGE will allow the data rates to further increase. With *Enhanced Circuit Switched Data* (ECSD) and *Enhanced General Packet Radio Service* (EGPRS), GSM now approaches theoretical bit rates up to 384 kbit/s or even 473.6 kbit/s [15].

In short, GSM tries to catch a considerable part of the PMR market. Admittedly a fraction of the PMR users could also be technically well served by these enhanced GSM systems, but this is not true for all of them. Some major functions are still missing, for example, real fast call establishment and direct mode. Specific GSM equipment well adapted to rough environmental conditions as well as to specific frequency ranges, as is necessary in the case of GSM-R, is only needed for a limited market and is therefore not available at the low price of ordinary GSM sets. Thus the economical feasibility of those GSM extensions has been questioned [16].

The introduction of UMTS or IMT-2000 was scheduled for 2002–2003 and therefore the comparison of PMR with public mobile systems also has to cover UMTS. However, UMTS will be offered in the beginning years only in conurbation areas and, second, it is again a mobile telephone system well suited to point-to-point communications but less adequate for the communication and control of large groups of mobile subscribers with their specific needs.

Admittedly, UMTS will offer interesting *Global Multimedia* (GMM) capabilities with bit rates up to 2,048 kbit/s exceeding the data transmission rates of current DPMR systems by far. However, the maximum data rates of advanced GSM versions and UMTS can only be provided in the vicinity of the base stations for nonmoving mobiles or those moving at low speed. Otherwise the maximum usable data rates will be severely limited. At medium distance and medium speed, the maximum usable bit rate of UMTS will be reduced to 384 kbit/s and under worst case conditions at the coverage border and at high mobile speed only 144 kbit/s remains available. For GSM similar restrictions are valid. In contrast to that, DPMR systems are able to deliver their maximum data rates even under the most unfavorable conditions at maximum mobile speed and during severe multipath propagation distortion and at the coverage border. Moreover, future versions of TETRA and other DPMR systems with higher bit rates have already been envisaged.

Furthermore, the demand for steadily increasing user bit rates does not mean that the gross bit rates of the systems have to increase at the same order. Presently powerful compression techniques with steadily increasing compression rates are being developed for data, audio, picture, and video transmission; for example, with M-JPEG and MPEG-4 audio and video compression ratios up to 300 have been achieved. Another well-known example is the MP3 format, which sacrifices only a very small amount of quality but permits CD music files to be compressed to about 10% of their original size.

2.4 Other Alternatives to PMR

Possibilities other than public mobile networks exist to replace PMR. However, they are real competitors to PMR in only a very few cases.

PMR applications for low mobility requirements sometimes can be served with telephone-like features. For example, inventory control in a warehouse benefited from data transmission but can be done with low mobility and limited range. For such applications digital cordless telephones such as DECT, which provides up to about 500 kbit/s, are an interesting and competitive alternative to PMR if a maximum distance of 100–300m restricted to the low transmitting power of 10 mW (10 dBm) or 1W (30 dBm) suffices. In the last few years, DECT equipment for heavy-duty purposes and an industrial environment has been developed and already launched [17].

Some PMR applications may also be well served by Bluetooth, a new type of *radio local-area network* (RLAN) for temporary mobile computer networks at 2.5 GHz, and the *High Performance Radio Local Network* (HIPERLAN) at 5.2 GHz. Both systems provide only limited mobility and coverage due to the high frequency and the limited transmitting power of 100 mW (20 dBm) or 1W (30 dBm), respectively, but they offer bit rates of some hundreds of kilobits per second [18].

HIPERLAN/2 (the second generation) is a mobile short-range access network standardized by ETSI EP *Broadband Radio Access Networks* (BRAN) that provides bit rates up to 54 Mbit/s and can interwork with several core networks including *asynchronous transfer mode* (ATM), Ethernet, and IEEE-1394. It is well suited to high-speed Internet and multimedia access, real-time video services, and the like. In EP BRAN two additional projects are under way. The first is *High Performance Radio Access* (HIPERACESS), a fixed wireless broadband access network optimized for point-to-multipoint operation in the 40.5- to 43.5-GHz band that employs a single TDMA broadband carrier. The second is *High Performance Radio Metropolitan Area Networks* (HIPERMAN), a fixed wireless access network that operates below 11 GHz. Both are usable only at very limited distances [19].

Despite the fact that all of these new standards offer considerably higher bit rates than DPMR systems, they are primarily intended for other purposes and can therefore compete with PMR only to a very limited extent. However, it is really remarkable that in recent years several standardization initiatives from different parts of the world encouraged mutual consultation and, in some cases, real cooperation. Such an example is the cooperation on broadband radio access networks between the ATM Forum, HIPERLAN/2 Global Forum, IEEE Wireless LAN Committees P 802.11a and 802.16, the Internet Engineering Task Force, ITU-R, MMAC-PC High Speed Wireless Access Systems Group, ETSI EP BRAN, and other ETSI technical bodies [19].

Mobile Satellite Services (MSSs) with their extremely large service areas are useful only for very specific cases like big trucks on continental or intercontinental routes not fully covered by terrestrial mobile radio communication systems. In this case neither PMR nor cellular systems will suffice, and MSS has been created exactly for such applications. However, the transmission rate is currently limited to 2.4 kbit/s, most of the specific PMR features are not supported, and the service is expensive. MSS systems that provide higher bit rates are under development but they will not be introduced very soon and they will be even more expensive. This means that MSS is a real competitor to PMR only for a limited number of applications.

Finally, some initiatives extend the coverage range and mobility of DECT to compete in some fields with GSM and PMR. There is also a trend to install home base stations for GSM that are going to compete with DECT. Additionally, PAMR tries to win away GSM subscribers and vice versa. Hence, in the mobile communications market all of the different systems are trying to extend their shares and to invade new fields to compete with the incumbent technology base.

2.5 Basic PMR Market Structure

Disregarding GSM-R, PMR is a collection of very different applications based on a more or less common basic radio technology. Nevertheless, the PMR market is very complex. Figure 2.2 aims not to reflect market figures or the size of market shares, but to instead show the general PMR market structure, which has more than one dimension [20].

Three different market segments can be distinguished. The first is a big segment of conventional analog PMR and PAMR including MPT 1327 and a variety of

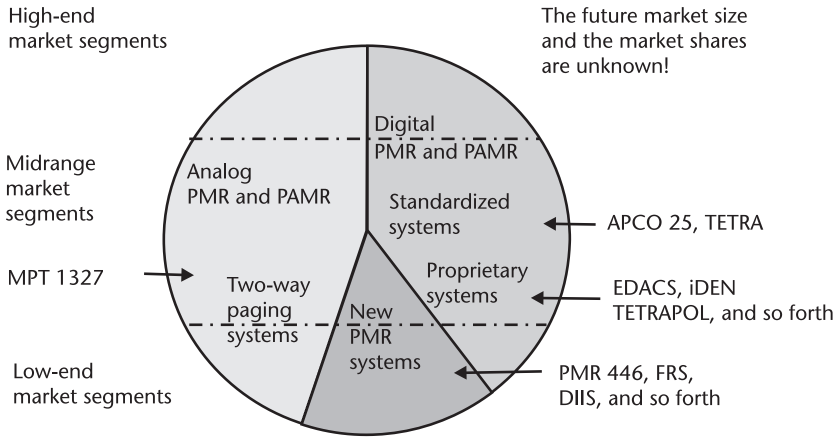

Figure 2.2 Basic PMR market structure.

two-way paging systems. It is expected that this segment will fade away slowly but the last analog equipment will not disappear before 10–15 years from now.

The second segment, comprising DPMR systems, is currently small but is growing very fast. Here various proprietary systems such as ASTRO, EDACS, iDEN, TETRAPOL, and others are to be found. On the other hand, there are also standardized DPMR systems such as APCO 25 and TETRA. In the long run the major part of this segment will presumably be occupied mainly by iDEN and TETRA [21].

Finally, there is the segment of new systems where even new analog services like PMR 446 and FRS appear. In 2002 in the United Kingdom, where it was introduced in 1999, 1 million subscribers have chosen PMR 446, so it is well on its way. Therefore expectations have arisen that this system will have around 10 million users in Europe 10 years after its introduction. Another new system is DIIS, which will be the right choice for those desiring cheap and small systems. This segment is the most unknown and its future evolution is hard to predict. Its growth will depend on many regulatory and market-related issues. Moreover, it is likely that PMR 446 and DIIS features will be combined, and this may attract totally new PMR user groups, which may enter the market by initially buying such equipment and turning later to more sophisticated PMR equipment [20, 22].

The whole PMR market may also be subdivided in another way. At first there is the high-end equipment with a broad bunch of features for very demanding applications. The technical and operational user requirements are high, and the equipment costs may not play a primary role in some of these cases. Secondly, there is the medium range where equipment for ordinary applications is to be found. In this segment competition from other services is high and the users are for the most part sensitive to costs. Finally, there is the low-end market with less demanding applications that require only a limited set of PMR features. Usually the users of such equipment are expecting very low equipment prices.

Provided the general conditions in the PMR markets remain stable or at least do not get worse in Europe, small but steady overall growth can be expected there. In particular, the PAMR segment may exhibit considerable growth. However, all of the predictions were made before the Dolphin Telecom commitment to digital PAMR based on TETRA. On the other hand, this initiative has not succeeded as expected and the impact of this failure has not yet been fully recognized. Another fact that has not been fully taken into account is the change in the regulatory environment in Europe, mainly the long-term consequences of changes in the approval and licensing regime. In particular the R&TTE directive of the *European Union* (EU) and the envisaged refarming of the European PMR frequency bands will influence the development of the European PMR market in a way that is not easy to assess. In contrast to the very cautious estimates for general PMR, expectations for TETRA and PMR 446 in Europe are much more optimistic. iDEN and FRS may play a similar role in the United States. If all this becomes true, then overall PMR growth may exceed recent predictions by a considerable amount.

Even if the author hesitates to cite market figures (because the different sources do not always deliver consistent market data and these data are changing quickly and are hardly predictable), some rough figures about the magnitude of the PMR market can be given. The reader should have at least a slight idea about the market size, which is by far not comparable to the many hundreds of millions of mobile phones in use worldwide in public cellular networks. In the United States in 1998, there were more than 16.3 million PMR users and 4.6 million of them belonged to SMR. A market study stated 21 million PMR users worldwide in 1999. It is surprising that the PMR penetration in the United States is about four times higher than in Europe where it is in the range between 1% and 2% in the different countries. The reasons for the limited growth in Europe were for a long time a restrictive regulatory and licensing regime, relatively high fees and long waiting times for licenses in some countries, and a well-organized and nearly perfectly working European network of public digital cellular systems with handover and international roaming. The latter has also hampered the development of European analog PAMR networks. In contrast, in the United States the conditions for PMR growth have all along been much more favorable than in Europe. This is due to the different geographical situation and the lack of full coverage and easy access to a public cellular system with fully developed roaming capabilities [12, 21, 23–25].

Finally, considerable growth rates are being experienced outside the United States and Europe in countries where DPMR systems are currently being introduced. The fast PMR growth there is not in contradiction to the growth of cellular systems but complements it. This growth will be driven mostly by iDEN and TETRA; all other DPMR systems including TETRAPOL are expected to be less successful in the long run [21].

Due to the digitalization of PMR and severe competition from other mobile systems, the PMR world market is currently changing very fast and it is very difficult to predict the growth and even more difficult to estimate the development of each particular market segment. Regulatory issues and—at least in Europe—harmonization and refarming of the PMR frequency bands will have a nearly unpredictable impact on the future PMR evolution.

2.6 Consequences

Many PMR users want or need additional functions and services usually not provided by PMR systems. In this case they should use different systems in parallel.

The main conclusion of the considerations made above is that PMR and cellular complement each other more than they compete with each other.

We now see multistandard and multiband equipment evolving, which means that different services may be integrated in a single radio set. Handsets for different GSM versions (dual- or triple-band handsets) and combinations of cellular radios with cordless telephones or palm-top computers are already available. As it is explained in Chapter 10, it can be predicted that it is only a question of time until the first combinations of cellular and PMR appear, for example, combinations of GSM and TETRA sets [1].

The result of all of our investigations and comparisons is that PMR can be replaced by other radio communication systems only in very specific cases. Consequently, PMR will be needed in the future—at least as long as it provides operational and economical features that are not available from other systems [1, 3]:

> As long as new mobile services cannot fully meet the requirements of DPMR with its ability to be tailored to the user's needs, PMR will not become obsolete. However, DPMR and other mobile services will complement each other. The user does not necessarily have to make a choice; he can use different services in parallel and in the future he may do this with multistandard radios.

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CHAPTER 3

Propagation and Traffic-Related Fundamentals of Digital PMR Technology

The properties of PMR systems are determined by a complex regulatory framework that reflects legislative, standardization, and licensing issues on one hand and radio propagation issues on the other. Planning for radio systems, therefore, requires comprehensive knowledge of all of the issues involved, of which one of the most important is radio propagation. Unfortunately, due to the constraints imposed by propagation effects, not all frequencies available for PMR are useful in the same way for the various PMR applications.

The basic parameters of current PMR equipment are more or less independent of the frequency. The static receiver sensitivity is usually in the range between –115 and –120 dBm, whereas the mobile transmitter peak power dependent on the application varies roughly between 100 mW (20 dBm) and 25W (44 dBm). The difference between transmitting power and receiver sensitivity results in a maximum permissible propagation loss on the order of about 160 dB, which means a power ratio of 10^{16} or 10,000,000,000,000,000:1, respectively [1]. At first glance this huge ratio seems to promise nearly unlimited coverage, but the laws of radio propagation introduce large losses because of the heavy interdependence of frequency, distance covered, terrain properties, and required transmitting power.

3.1 Basic Propagation Models, Static Path Loss, and Link Budget

In free space the propagation of radio waves can be modeled without difficulty because free space is lossless. This is also true for air and the frequencies used for PMR. Well above 1 GHz, molecule resonance might cause additional loss at certain frequencies but for the frequencies regarded here this effect is negligible [2, 3]. Therefore the power density S_P can be written as a function of the distance d, the transmitting power P_T, and the antenna gain G_T [4–13]:

$$S_P = \frac{P_T \cdot G_T}{4\pi d^2} \tag{3.1}$$

The transmitting antenna gain in a given direction is referenced to that of an isotropic radiator. The effective aperture A_R of the receiving antenna is determined by its gain G_R (which for lossless antennas is equal to the directivity D_R) and the wavelength $\lambda = v_C/f_C$ where v_C is the speed of light and f_C the carrier frequency:

$$A_R = \frac{\lambda^2 \cdot G_R}{4\pi} \tag{3.2}$$

This effective aperture collects the received power density and thus the total power can be calculated:

$$P_R = S \cdot A_R = \left(\frac{\lambda}{4\pi d}\right)^2 \cdot G_R G_T \cdot P_T \tag{3.3}$$

Due to the reciprocity law of antenna systems, it is usually not necessary to distinguish between transmitting and receiving antennas [7, 9, 12–16]. For further considerations it is useful to introduce the logarithmic antenna gain $g_{T,R}$ and to rewrite the total loss a_{TR} between the transmitter and receiver depending on the distance d:

$$g_{T,R} = 10 \cdot \log_{10} G_{T,R} \tag{3.4}$$

$$a_{TR} = 10 \cdot \log_{10}(P_T/P_R) = 20 \cdot \log_{10}(4\pi d/\lambda) - (g_R + g_T) \tag{3.5}$$

Such calculations can be done easily with a scientific pocket calculator. Besides the antenna gains, additional losses a_{CR} and a_{CT} caused by cables and couplers between the transmitter and receiver and their antennas have to be taken into account, whereas the antennas themselves are nearly lossless:

$$a_{TR} = 20 \cdot \log_{10}(4\pi d/\lambda) - (g_T - a_{CT} + g_R - a_{CR}) \tag{3.6}$$

However, all of these antenna gains and cable losses can be added and represented by the resulting logarithmic gain g_Σ or linear gain G_Σ:

$$g_\Sigma = (g_T - a_{CT}) + (g_R - a_{CR}) \quad \text{and} \quad G_\Sigma = 10^{g_\Sigma/10} \tag{3.7}$$

The static path loss a_{LS} for unobstructed free-space propagation depends only on the distance, and the link budget a_B can now be defined as the maximum permissible loss between transmitter and receiver [9]:

$$a_{LS} = 20 \cdot \log_{10}(4\pi d/\lambda) \tag{3.8}$$

$$a_B = a_{TR} + g_\Sigma \geq a_{LS} \tag{3.9}$$

All of these basic considerations remain valid for mobile radio frequencies if we turn from free-space propagation to propagation in the mobile environment.

3.2 Propagation in the Mobile Environment

In mobile services vertical polarization and in many cases vertical thin-wire antennas with an omnidirectional propagation characteristic in the horizontal plane are used. A vertical whip antenna of length l over the ideal reflecting horizontal ground can be treated mathematically like a dipole of length $2l$ with $l \neq 0$ in free space as Figure 3.1 shows.

For very thin whips and dipoles, the following formula for the vertical radiation pattern in terms of relative field strength has been derived where the angle ϑ is measured against the vertical axis z [7, 12–17]:

$$F_{2l}(l/\lambda,\ \vartheta) = \frac{1}{\sin\ \vartheta} \cdot \left[\cos\left(\frac{2\pi l}{\lambda} \cdot \cos\ \vartheta\right) - \cos\left(\frac{2\pi l}{\lambda}\right)\right] \tag{3.10}$$

Evaluating this equation reveals that with increasing length l up to $\lambda/2$ the width of the lobe in the main radiation direction r, which means $\vartheta = 90°$, becomes smaller and the antenna gain therefore becomes higher. If only the relative field strength in the main radiation direction is of interest, then it is $\vartheta = 90°$ and the following equation results:

$$F_{2l}(l/\lambda,\ \vartheta = 90°) = 1 - \cos(2\pi l/\lambda) \sim \sqrt{G_{2l}} \tag{3.11}$$

The antenna gain depends on the effective antenna aperture, which increases with the antenna length. The important point, however, is that for $l > 5\lambda/8$ additional lobes exist and the gain in the main direction might be reduced. Table 3.1 gives the properties of some common mobile antennas. The antenna gain G_i is given there relative to the isotropic radiator dependent on the length l and the aperture $A_{T,R}$.

In practice, no isotropic radiator can be built and its best approximation is the Hertzian dipole with top capacity. The general problems with very short antennas are that their gain is low and their input impedance is very different from the cable impedance that in PMR is usually 50Ω, and the required matching introduces additional losses.

A whip antenna of length l mounted on the metallic roof of a car behaves similar to a whip on ideal ground. Compared with a dipole of length $2l$ it has only

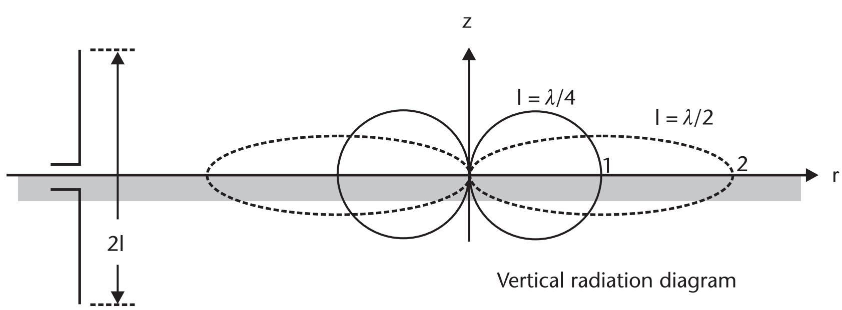

Figure 3.1 Whip antenna and equivalent dipole.

Table 3.1 Characteristics of Common Mobile Antennas

Antenna	l	$A_{T,R}$	G_i	$g_i = 10 \log_{10} G_i$ *[dBi]*
Isotropic	Fictitious	$\lambda^2/4\pi = 0.08 \cdot \lambda^2$	1	0
Hertzian dipole[1]	$2l << \lambda$	$1.5 \cdot \lambda^2/4\pi$	1.5	1.76
Short dipole	$2l << \lambda$	$1.5 \cdot \lambda^2/4\pi$	1.5	1.76
$\lambda/2$ dipole	$2l = \lambda/2$	$1.64 \cdot \lambda^2/4\pi$	1.64	2.15
λ dipole	$2l = \lambda$	$2.41 \cdot \lambda^2/4\pi$	2.41	3.82
Short whip	$l << \lambda$	$0.75 \cdot \lambda^2/4\pi$	3	4.77
$\lambda/4$ whip	$l = \lambda/4$	$0.82 \cdot \lambda^2/4\pi$	3.28	5.16
$\lambda/2$ whip	$l = \lambda/2$	$1.21 \cdot \lambda^2/4\pi$	4.82	6.83

[1]Loaded with top capacity, the best simple practical approximation to the theoretical isotropic radiator.

half the effective aperture, but the radiated power is concentrated in the upper hemisphere and therefore doubled, thus an additional gain of 3 dB should occur [9, 14]. In practice, such a configuration never is ideal and an additional gain less than 3 dB will be experienced. The antenna characteristic has to be matched to the geometry of the car and the precise location of the antenna, and it has been found that in many cases instead of $l = \lambda/2$ a better match is $l = \lambda 5/8$. The results given in Table 3.1 refer to the *ideal* configuration without distortion of the field by an imperfect geometry, for example, by a ground plane that is limited or not totally flat such as a car rooftop. Therefore the exact antenna properties have to be taken from the manufacturer's leaflet and the mounting instructions have to be observed carefully.

Longer whips deliver higher gain, but that gain is dependent on the angle ϑ against the vertical axis. Thus tilting the antenna somewhat, which cannot be avoided in mobile applications, may result in even less gain in the wanted direction. These antennas also might have input impedances very different from 50Ω, and the necessary matching network may devour at least a part of the gain. Hence high-gain antennas are only feasible when they are not tilted and when proper matching without too much loss is feasible. Usually that is only the case at the base station. For base stations, antennas with a gain of 10 dB or even much more are often installed. Sometimes the horizontal radiation characteristic is not omnidirectional. This is the case for sector antennas or coverage areas with a shape very different from a circle.

Keep in mind that a receiving antenna should have a length l proportional to the wavelength. This means that the received voltage is $U \sim l \cdot E \sim \lambda \cdot E$ and therefore the received power is $P_R \sim 1/f^2$.

To evaluate the link budget, all antenna gains and cable losses have to be added. For car-mounted mobiles the cable loss is small and so is the antenna gain, and a resulting gain $g_M \approx 0$ dB is a reasonable assumption. In practice, a good antenna installation will provide somewhat better results.

For hand portables shortened whip antennas or sometimes integrated antennas are normally used, and the antenna gain is on the order of 1 to 1.5 dBi. However, very often the antenna is tilted and that reduces the gain. Consider this example: A 45° tilt angle theoretically results in a 2-dB loss for a $\lambda/4$ whip and a 5.5-dB loss for a $\lambda/2$ whip because the beam is much smaller. However, the presence of the human body will change the gain in a nearly unpredictable way. Dependent on the carrier frequency and how the equipment is worn—in the hand, on the body, or close to the ear—a body loss in the range of 2–12 dB will be experienced, with it averaging on the order of 2.5 dB [7, 18]. All of these effects taken together result in the rough average of $g_{HP} \approx -3$ dB for handhelds. For more accurate system and coverage evaluation, the reader is encouraged to ask the equipment manufacturer for more reliable data.

At the base station, considerably higher antenna gain is usually possible but longer cables and in many cases additional couplers and filters are installed. In the end all of these gains and losses might nearly cancel out. Hence $g_B \approx 0$–6 dB can be expected. However, to determine the correct link budget, the real values of all gains and losses have to be evaluated carefully.

Service is often needed in tunnels and mines, which are rife with poor propagation conditions. Therefore, the best coverage is achieved primarily by installing leaky cables along the tunnel. The maximum distance covered depends mainly on the cable loss, which is roughly on the order of 30 dB/km, and the coupling loss from the cable to the mobile, which is about 50–70 dB [19].

To understand what really happens in land mobile services, it is helpful to look at a specific case. Let us consider propagation from a thin vertical transmitting antenna placed high above ground, where the ground is represented by an ideal plane with infinite conductivity, to a vertical mobile antenna at a low height above ground. As shown in Figure 3.2, there is always a direct ray and another ray reflected by the ground plane [7, 11, 13, 20].

Hence the total received power is the sum of the power P_{Rd} from the direct path and that of the reflected path P_{Rr} where r_r is the reflection coefficient and $\Delta\Phi$ the phase difference between the two rays.

$$P_R = P_{Rd} + P_{Rr} = P_T \cdot \left(\frac{\lambda}{4\pi d}\right)^2 \cdot \left[1 + r_r \cdot e^{j\Delta\Phi}\right]^2 \tag{3.12}$$

If the distance between the base and mobile stations is d while their antenna heights above ground are h_B and h_M, then the length d_d of the direct path and the length d_r of the reflected path can be calculated using the law of Pythagoras:

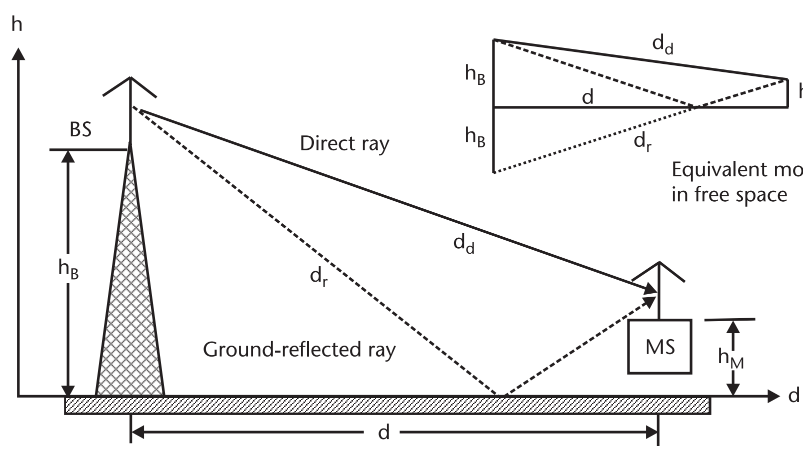

Figure 3.2 A simple two-ray propagation situation.

$$d_{d,r} = \sqrt{d^2 + (h_B \mp h_M)^2} = d \cdot \left[1 + \frac{1}{2} \cdot \left(\frac{h_B \mp h_M}{d}\right)^2\right] \tag{3.13}$$

With $(1 + x)^{1/2} \approx 1 + x/2$ because $d >> h_{B,M}$ the distance difference Δd and the phase difference $\Delta\Phi$ of the two paths can be determined [21]:

$$\Delta d = d_r - d_d = \frac{2h_B h_M}{d} \tag{3.14}$$

$$\Delta\Phi = \frac{2\pi \cdot \Delta d}{\lambda} = \frac{4\pi \cdot h_B h_M}{\lambda \cdot d} \tag{3.15}$$

For vertical polarization as is usually found in PMR and flat angles the reflection is nearly ideal and therefore the reflection coefficient is $r_r = -1$.

$$P_R = P_T \cdot \left(\frac{\lambda}{4\pi d}\right)^2 \cdot \left[1 - e^{j\Delta\Phi}\right]^2 \tag{3.16}$$

Using Euler's relation $e^{jx} = \cos(x) + j \cdot \sin(x)$, the trigonometric theorem $[1 - \cos(2x)] = 2 \cdot \sin^2(x)$ and $\sin(x) \approx x$ for $x << 1$, we get the resulting dependency of the received power on the distance [21]:

$$\left[1 - e^{j\Delta\Phi}\right]^2 = 4 \cdot \sin^2\left(\frac{\Delta\Phi}{2}\right) \tag{3.17}$$

$$P_R \approx P_T \cdot \left(\frac{h_B^2 \cdot h_M^2}{d^4}\right) \tag{3.18}$$

Hence the received power falls with the fourth power of the distance because the energy of both rays at the receiving antenna for large distances is more or less equal while there is an obstructive phase difference of about π caused by the ground reflection. In a real mobile environment, the antenna gains and the dependency on the carrier frequency f_C also have to be taken into account, which leads to the propagation model proposed by Egli [2]:

$$P_R \approx P_T \cdot G_\Sigma \cdot \left(\frac{h_B^2 \cdot h_M^2}{d^4}\right) \cdot \left(\frac{f_P}{f_C}\right)^2 \tag{3.19}$$

The propagation constant f_P has been estimated by Egli to be $f_P = 40$ MHz. This idealized propagation model is only valid where there is a direct *line of sight* (LOS) between transmitter and receiver and it provides only limited accuracy for absolute path loss calculations.

The received signal power drops with the fourth power of the distance while the influence of frequency and antenna height is governed by the second power of these quantities. Hence some basic conclusions can be drawn that are also true in conjunction with more refined propagation models. Three simple examples will make these relations more transparent [1]. First, the covered distance is proportional to the fourth root of the transmitter power. This means that the transmit power has to be increased by a factor of $2^4 = 16$ if the distance needs to be doubled:

$$d \sim \sqrt[4]{P_T} \tag{3.20}$$

Second, receiver resonant antennas ($l \sim \lambda$) have to be used, causing severe frequency dependence. Doubling the carrier frequency leads to a reduction of the covered distance to $1/\sqrt{2}$ or 70% or to a required increase in the transmitting power by a factor of $2^2 = 4$ to maintain the original coverage:

$$d \sim \frac{1}{\sqrt{f}} \tag{3.21}$$

Finally, the antenna height and the maximum covered distance share an interdependence, which simply means that the covered distance is proportional to the root of the base station antenna height because the mobile station height is usually more or less constant and not much above ground level. The base station antenna might be found somewhere between 10m and 200m in the case of very high buildings or hilltop sites, and doubling the base station antenna height gives a gain of $\sqrt{2}$ or 40% for the covered distance:

$$d \sim \sqrt{h_B \cdot h_M} \tag{3.22}$$

Why have these relationships been highlighted? Transmitter power, receiver sensitivity, antenna heights, and carrier frequency are the main factors determining the useful coverage area and not every frequency is equally useful for the same service or application. Hence the carrier frequency and all the other technical parameters have to be tailored to the service under consideration.

Additionally it becomes obvious that coverage is limited to about 20 km under good propagation conditions, but might be reduced under certain circumstances to distances between 100m and a few kilometers. But even where good coverage can be achieved, it might be limited artificially in order to allow frequency reuse at a certain distance by other users. Therefore frequency managers have introduced rules that unfortunately are different in the various countries and make the harmonization of PMR—particularly in Europe—a difficult but nevertheless necessary task.

3.3 Refined Propagation Model for Large Distances

For practical use a refined model with better accuracy than the Egli model is needed. Therefore, the propagation or path loss exponent α is introduced that characterizes the propagation conditions and is somewhat different from 4. Equations (3.6) and

(3.9) can now be rewritten for real propagation conditions and deliver the necessary condition for the average static path loss a_{LS}:

$$a_{LS} = 10\alpha \cdot \log_{10}(4\pi d/\lambda) + a_P \leq a_{TR} + g_\Sigma = a_B \tag{3.23}$$

Of course, the path loss a_{LS} has to be smaller than the link budget a_B to ensure reception with sufficient quality. In free space where $\alpha = 2$ and $a_P = 0$ the well-known $1/d$ law results for the field strength. For any other value of α the attenuation can be determined if the constant a_P is also known. However, a_P depends on many parameters, in particular, the heights of base and mobile station antennas, the frequency, the environment, and other parameters like those already discussed for the Egli model. Note that the result is true for a probability of 50% in location and time only. Moreover, the overall result for the average path loss is independent of the direction, which means it is equal for uplink and downlink.

Figure 3.3 shows the field strength over the distance on a linear scale. Near the base station it falls very steeply and becomes flatter at larger distances and in particular at the coverage border. For comparison the inset shows the same signal level on a double logarithmic scale.

For relative comparisons, (3.23) can be simplified because a_P and g_Σ are canceled. Provided the field strength E_0 or the power P_0 is known at a certain distance d_0, then the unknown field strength E or the power P at another distance d can be determined easily on a linear scale or in decibels [10]:

$$P = P_0 \cdot (d_0/d)^\alpha \quad \text{or} \quad E = (d_0/d)^{\alpha/2} \tag{3.24}$$

$$20 \cdot \log_{10}(E/E_0) = 10 \log_{10}(P/P_0) = 10\alpha \cdot \log_{10}(d_0/d) \tag{3.25}$$

What does all of this mean in practice? Assuming there is poor coverage between the cell border and a distance where the receiving power is 10 dB greater and

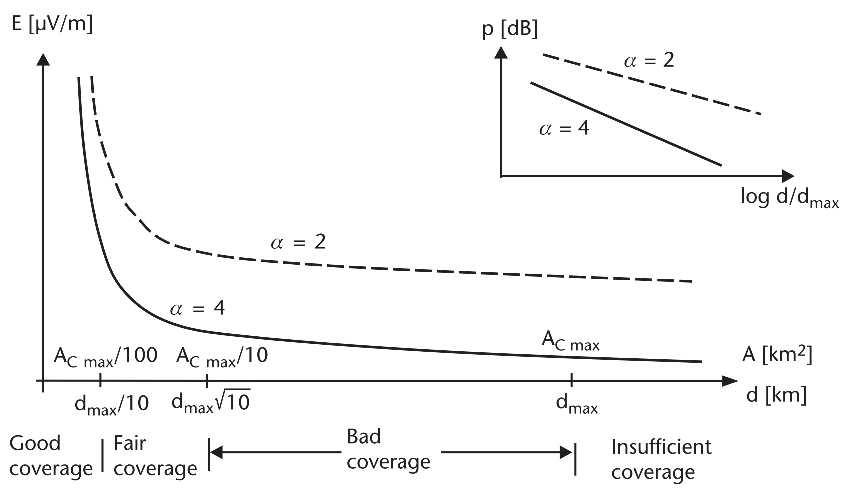

Figure 3.3 Illustration of the field strength dependence on distance.

$\alpha = 3.5$, the percentage of the cell area with good conditions can be calculated by $(d/d_0)^2 = (1/10)^{2/3.5} = 0.27$. The surprising result is that about a quarter of the cell surface (27%) has good coverage, which means only slightly more than half the maximum possible distance (52%).

For typical PMR propagation scenarios, the following values of the propagation exponent α have been experienced:

Free-space propagation:	$\alpha = 2$
Typical land mobile radio environment:	$\alpha = 3$ to 4
Indoor propagation:	$\alpha \approx 5$ to 6

Figure 3.4 shows how α changes with the distance. For short distances $\alpha = 2$. For the usual outdoor propagation conditions beyond a certain distance d_0, α is in the range of 3–4. In many cases the approximation $\alpha = 3.5$ is a good choice. If the radio signal penetrates a building at the distance d_i, then the propagation conditions indoors would deteriorate and α would become 5–6 or even worse beyond d_i [11, 18, 22]. This is caused by multiple reflections and absorption by walls and furniture. Indoor propagation is very poor and therefore weak radio signals entering a building may drop below the noise level very soon.

Earlier we showed that there is a constant a_P missing to determine the real path loss, and for decades facts about radio propagation were not as accurately known as today. Therefore Okumura et al. performed an extensive measurement campaign in the 1960s to evaluate propagation losses under various propagation conditions. More than a decade later, Hata analyzed Okumura's results and put them in a very easy-to-use format [4, 6, 9, 11, 13, 18, 22–27]:

$$a_{LS} = 10\alpha \cdot \log_{10} d + a_{1\,\mathrm{km}} + a_T \tag{3.26}$$

The path loss $a_{1\,\mathrm{km}}$ is that at a distance of 1 km. Compared with (3.23), it is $a_P = a_{1\,\mathrm{km}} + a_T$ where a_T is a correction depending on the type of terrain. The

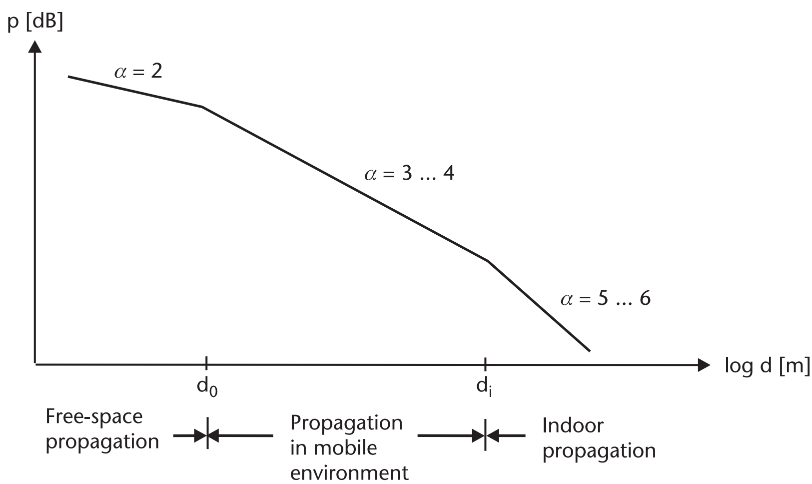

Figure 3.4 Illustration of the propagation exponent dependence on distance.

constants α and $a_{1\,\mathrm{km}}$ can be retrieved from Hata's formulas (3.27) to (3.32) or from Table 3.2, which compiles the results for the main PMR frequency bands.

Table 3.2 is valid for built-up city centers and for a mobile antenna height h_M of 1.5m.

The distance d is given in kilometers, the carrier frequency f_C in megahertz, the base station antenna height h_B in meters, and all losses in decibels. The relation is valid for frequencies from 150 to 1,500 MHz, distances from 1 to 20 km, mobile and base antenna heights from 1–10m and 30–200m, respectively, and a probability of 50% in location and time due to fading and shadowing. Because it is mainly dependent on the geometry of the radio path, α does not vary much with the frequency and in Hata's formula this has been fully neglected. The loss $a_{1\,\mathrm{km}}$ for a distance of 1 km varies considerably with the antenna height due to the changing geometry, and with the frequency due to the wavelength-related receiver antenna length and some other small effects. Note that the $a_{1\,\mathrm{km}}$ values for 400 MHz are about 1 dB smaller than those for 450 MHz.

For readers who wish to make their own calculations, Hata's formulas are given below. For α and $a_{1\,\mathrm{km}}$ Hata derived the following empirical relations that approximate the reality very well and allow the prediction of radio coverage with reasonable accuracy:

$$\alpha = 4.49 - 0.655 \cdot \log_{10} h_B \tag{3.27}$$

$$a_{1\,\mathrm{km}} = a_{f_C} - 13.82 \cdot \log_{10} h_B - a(h_M) \tag{3.28}$$

where

$$a_{f_C} = 69.55 + 26.16 \cdot \log_{10} f_C \tag{3.29}$$

Hata's complete propagation formula therefore looks somewhat complicated:

$$\begin{aligned} a_{LS} = 69.55 &+ 26.16 \cdot \log_{10} f_C - 13.82 \cdot \log_{10} h_B \\ &- a(h_M) + (44.9 - 6.55 \cdot \log_{10} h_B) \cdot \log_{10} d \end{aligned} \tag{3.30}$$

The correction factor $a(h_M)$ is zero for $h_M = 1.5$m, which is an appropriate assumption for vehicle-mounted antennas as well as for handheld and body-worn

Table 3.2 Path Loss According to Okumura and Hata

h_B [m]	α All Bands	$a_{1\,\mathrm{km}}$ [dB][1] 75 [MHz][2]	150 [MHz]	450 [MHz]	900 [MHz]	1,500 [MHz]
30	3.52	100.1	106.1	118.5	126.4	132.2
50	3.38	97.0	103.0	115.5	123.4	129.2
70	3.28	95.1	101.0	113.5	121.3	127.1
100	3.18	92.8	98.8	111.3	119.2	125.0
150	3.06	90.4	96.4	108.9	116.8	122.6
200	2.98	88.7	94.7	107.2	115.0	120.8

[1] $h_M = 1.5$m.
[2] With (3.32) derived from 150 MHz by −6 dB.

equipment. For the range of 1–5m, it can be approximated for large cities by the following relation:

$$a(h_M) \approx 3.2 \cdot (\log_{10} 11.75 h_M)^2 - 4.97 \quad \text{with} \quad a(h_M)_{1.5\text{m}} = 0 \text{ dB} \quad (3.31)$$

The influence of the human body was discussed earlier. For PMR frequencies below 150 MHz are very often of interest. Hata's formula (3.30) then has to be modified by the following substitution [27]:

$$26.16 \cdot \log_{10} f_C \rightarrow 26.16 \cdot \log_{10}(150) - 20 \cdot \log_{10}(150/f_C) \quad (3.32)$$

In the case where the terrain is different from a built-up city center, the Hata formula has to be extended by the correction term a_T, which is shown in Table 3.3 for the most important cases. Because the terrain characteristics are based on details much bigger than the wavelength, the carrier frequency has very little influence and in some cases none at all.

In suburban areas without high and densely arranged buildings, the coverage range is superior, and in flat rural areas without buildings it is still even wider. It is less important that the coverage in town centers is much inferior to that of open areas because usually the service area is limited by the number of subscribers and their traffic, which must not surpass the system capacity. In hilly terrain the propagation is worse because the service area is to a large extent shadowed. In areas with heavy foliage, severe additional loss will be experienced. It depends strongly on distance and frequency, the height and type of trees, the season of the year, and other factors, but no simple relation can be given. However, it is recommended that the base station antenna height exceed that of the trees [7, 8].

Reordering Hata's formula allows calculation of the maximum coverage distance for a given link budget a_B with $d_0 = 1$ km:

$$d_{\max} = d_0 \cdot 10^{(a_B - a_{1\,\text{km}} - a_T)/10\alpha} \quad (3.33)$$

Note that there are differences between various cities due to the different types and heights of buildings. Therefore other authors have given further propagation models, but this means only that the values given for the constants α and $a_{1\,\text{km}}$ differ somewhat. However, for first rough calculations Hata's formula yields fairly useful results.

Table 3.3 Correction a_T in Decibels for Various Types of Terrain

Terrain and Development	*150 MHz*	*450 MHz*	*900 MHz*
Built-up city center	0	0	0
Suburban area	−5.5	−8.5	−10.0
Quasi open terrain	−19.0	−21.0	−23.5
Flat open terrain	−24.0	−26.0	−28.5
Hilly terrain Δh = 30m	7.0	7.0	7.0
Hilly terrain Δh = 100m	12.0	12.0	12.0

3.4 Static Path Loss for Small, Medium, and Large Distances

The propagation conditions in the vicinity of a transmitter designed to transmit less than 1 km have to be calculated using a different method. To evaluate the average path loss, first the loss at 1m is calculated based on free-space propagation conditions. Then an estimate has to be made as to where free-space conditions remain valid. This is related to street width, obstacle distances, antenna height, and so on. Say it is valid as far as d_0 [7, 8, 11, 13, 17, 20, 22, 24, 25]. Then the additional path loss between 1m and d_0 can be calculated without difficulty. Finally the whole static path loss a_{LS} and thus the maximum covered distance $d_{\max}$ can be determined:

$$a_{LS} = a_B \approx a_{1\text{m}} + a_0 + 10\alpha \cdot \log_{10}(d_{\max}/d_0) \tag{3.34}$$

$$d_{\max} \approx d_0 \cdot 10^{(a_B - a_{1\text{m}} - a_0)/10\alpha} \tag{3.35}$$

This model is called the *dual-slope propagation model*. It is useful for the calculation of the path loss at small distances and for small antenna heights, for instance, in the case of path loss between two mobiles. The reference value $a_{1\text{m}}$ for the loss at 1m of distance from the antenna can be calculated according to (3.5):

$$a_{1\text{m}} = 20 \cdot \log_{10}(4\pi \cdot 1\text{m}/\lambda) \tag{3.36}$$

For $f_C = 75$ MHz the result is $a_{1\text{m}} = 10.0$ dB, and each doubling of the frequency results in an increase of 6 dB. In reality, and in particular for low frequencies, $a_{1\text{m}}$ may not be fully correct due to the near-field conditions that govern the field strength distribution in the vicinity of the antenna. For the Hertzian dipole the error compared to the $1/d$ law can be determined from the general field strength formula as given in the literature. It is less that 0.1 dB at λ [15, 16, 28]. Hence it can be assumed that for relatively short thin whip or dipole antennas, such as those used in PMR, the error of the amplitude of the field strength is negligible for all distances exceeding λ.

Based on (3.25) the path loss for the path from 1m to d_0 can also be calculated without difficulty:

$$a_0 = 20 \cdot \log_{10}(d_0/1\text{m}) \tag{3.37}$$

For d_0 = 3m we find a_0 = 9.5 dB, and a factor of 10 for d_0 means an increase of 20 dB for a_0. To get reasonable results, it is necessary to find a good approximation for d_0. The limit of the free-space propagation distance can be found according to the sketch of a typical situation, as shown in Figure 3.5.

The determining issue is the so-called first Fresnel zone, which is given by the location of all possible reflectors on a propagation path that is $\lambda/2$ longer than the direct propagation path d_{01}. Inside this ellipsoid there is free-space propagation with $\alpha = 2$, whereas outside the value is $\alpha \approx 4$. The maximum distance $d_{d\max}$ for

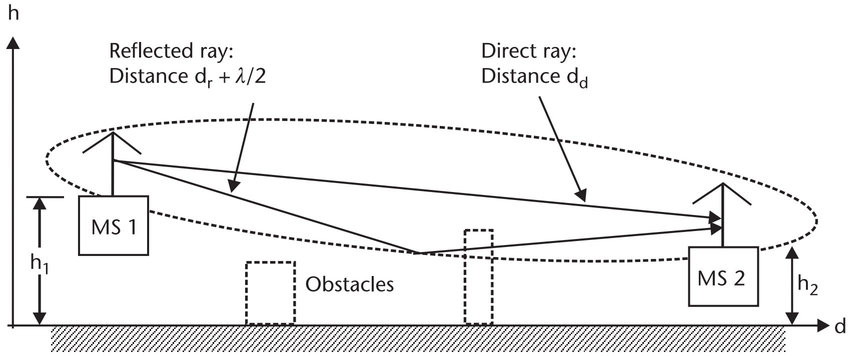

Figure 3.5 Limit of free-space propagation.

free-space propagation can be determined from the geometry. For nearly equal antenna heights $h_1 \approx h_2 \approx h$, which are small compared to the direct distance d_d, the length of the reflected ray $d_r = d_d + \lambda/2$ and flat ground the following relation is valid:

$$h^2 + \left(\frac{d_d}{2}\right)^2 = \left(\frac{d_r}{2}\right)^2 \tag{3.38}$$

Therefrom the maximum free-space distance d_{01} can be obtained after a short calculation:

$$d_{01} = d_{d\,\max} = \frac{4h_1h_2}{\lambda} - \frac{\lambda}{4} \quad \text{for} \quad h_1 \approx h_2 \quad \text{and} \quad h_{1,2} << d_d \tag{3.39}$$

The wall of an adjacent big building, a large truck, or another vertical obstacle might act as a reflector and therefore another critical path d_{02} may exist. Assuming mainly the situation between buildings, the street width d_S determines the maximum free-space propagation path:

$$d_{02} = d_S - \frac{\lambda}{4} \tag{3.40}$$

The smaller of the two distances dominates the propagation, and for the link between two mobiles ($h_1 = h_2 = 1.5$m) the results are presented in Table 3.4:

$$d_0 = \min(d_{01}, d_{02}) \tag{3.41}$$

Hence for low frequencies the antenna heights are dominant due to the ground reflections, whereas for high frequencies it is the horizontal distance to buildings or other reflecting objects. For practical use it is reasonable to take the rounded minimum of d_{01} and d_{02} because too much accuracy does not make sense since it all is an approximation. Table 3.5 gives the final results.

Table 3.4 Estimation of the Maximum Free-Space Propagation Distances d_{01} and d_{02}

Frequency Range [MHz]	λ *[m]*	d_{01} *[m]*[1]	d_{02} *[m]*[1] $d_S = 3m$	$d_S = 10m$	$d_S = 30m$	$d_S = 100m$
75	4.00	1.25	2.0	9.0	29.0	99.0
150	2.00	4.0	2.5	9.5	29.5	99.5
450	0.67	13.3	2.8	9.8	29.8	99.8
900	0.33	26.9	2.9	9.9	29.9	99.9

[1]All values of d_{01} and d_{02} are based on $h_1 = h_2 = 1.5$m and d_S represents the street width.

Table 3.5 Estimation of the Resulting Maximum Free-Space Propagation Distance d_0

Frequency Range [MHz]	d_0 *[m]*[1] $d_S = 3m$	$d_S = 10m$	$d_S = 30m$	$d_S = 100m$
75	1.0	1.0	1.0	1.0
150	2.0	4.0	4.0	4.0
450	3.0	10.0	10.0	10.0
900	3.0	10.0	30.0	30.0

[1]All values of d_{01} and d_{02} are based on $h_1 = h_2 = 1.5$m and d_S represents the street width.

To get an idea of what this means, the resulting coverage distance $d_{\max}$ for a link between two mobiles and 50% probability is given in Table 3.6, assuming $\alpha = 4$ beyond d_0, a transmitting power of 35 dBm (3W), and a static receiver sensitivity of −115 dBm (0.4 μV_{cc} at 50Ω), which means a link budget of 150 dB.

The dual-slope model is valid for the vicinity of the transmitter not only for the wanted signal but also for unwanted signals from spurious emissions, intermodulation, and the like. It is also applicable for low antenna heights, particularly for distances below 1 km where the Hata model is not applicable.

The evaluation of the propagation conditions within city centers with the dual-slope model for small distances and with the Hata model for distances beyond 1 km gives the results shown in Figure 3.6, which provides an overview of the total propagation loss for urban areas for short distances, in the transition range and for larger distances, for the three most important PMR frequency bands: 150, 450, and 900 MHz.

The critical area is that between the end of free-space propagation and the start of the Hata model. In reality, for every path there is a jump of roughly 10 dB somewhere between d_0 and 1 km [22]. It appears for instance when moving around

Table 3.6 Maximum Coverage Distance Between Two Mobiles

Frequency Range [MHz]	$d_{\max}$ *[km]*[1] $d_S = 3m$	$d_S = 10m$	$d_S = 30m$	$d_S = 100m$
75	3.2	3.2	3.2	3.2
150	3.2	4.5	4.5	4.5
450	2.2	4.1	4.1	4.1
900	1.6	2.9	5.0	5.0

[1]All values of d_{01} and d_{02} are based on $h_1 = h_2 = 1.5$m, no fading or shadowing is assumed, and d_S represents the street width.

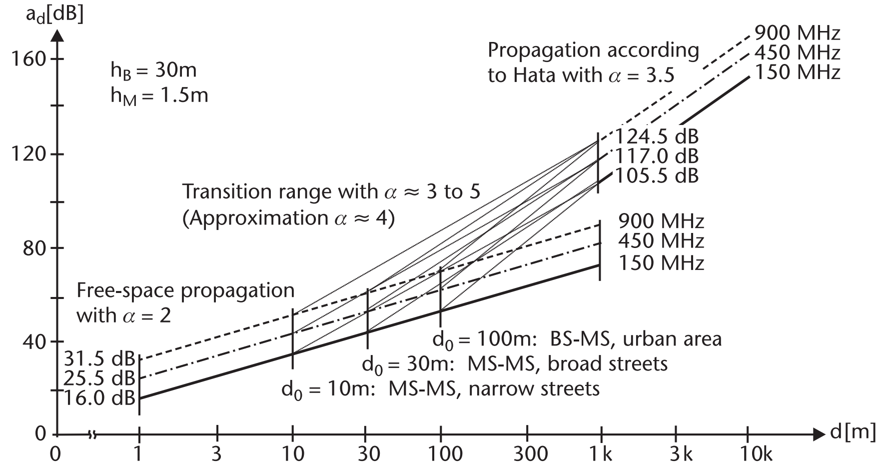

Figure 3.6 Static path loss in cities.

a corner. On average, there is nothing better than inserting a straight line between the maximum free-space propagation distance and 1 km, where the Hata formula becomes valid, and the crude approximation $\alpha = 4$ [10].

3.5 Fading and Shadowing: Dynamic Path Loss

Coverage predictions based solely on the static path loss will provide misleading results if fading and shadowing are not taken into account. These phenomena are the result of multipath propagation. Besides a direct path, which only exists if it is not shadowed, a large number of indirect paths of different length are present with reflections, scattering, and diffraction between the transmitter and receiver. All of these different signals arrive at the receiver input with varying amplitude and phase, and their addition leads to any result between a certain maximum and zero [3, 8, 11, 12, 28].

Figure 3.7 shows a simple example. One of the reflected rays is coming from the ground and the other is equivalent to a ray coming from a virtual base station. If only two rays with equal amplitude interfere, one direct and one totally reflected signal, then compared to the direct ray the resulting signal exhibits fluctuations between nearly doubling (+6 dB) and nearly zero (−∞ dB; in practice, −20 to −40 dB) with a spatial repetition of maximum values and zero proportional to $\lambda/2$. The angle between the ray and the direction of movement of the mobile will change the precise distance of the fading nulls. If a higher number of rays are involved, then the resulting pattern no longer appears to be regular. Nevertheless, the distance between the nulls is still on the order of $\lambda/2$.

Another important effect is shadowing. If an obstacle like a hill or a big building sits between the transmitter and receiver, an additional amount of attenuation will be experienced. These obstacles are usually much larger than the wavelength and therefore their radio shadow is also much larger and is on the order of 10 to 100 wavelengths or more. As Figure 3.8 shows schematically, both effects have to be combined to model general propagation conditions.

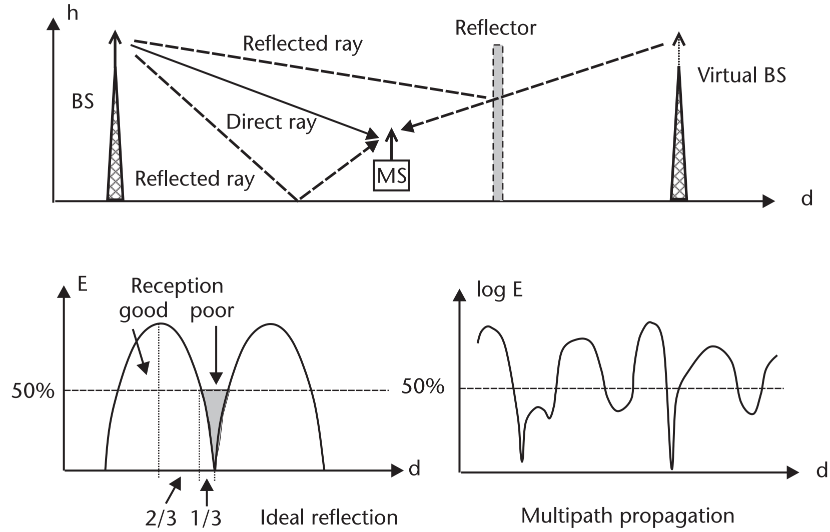

Figure 3.7 Multipath propagation situations.

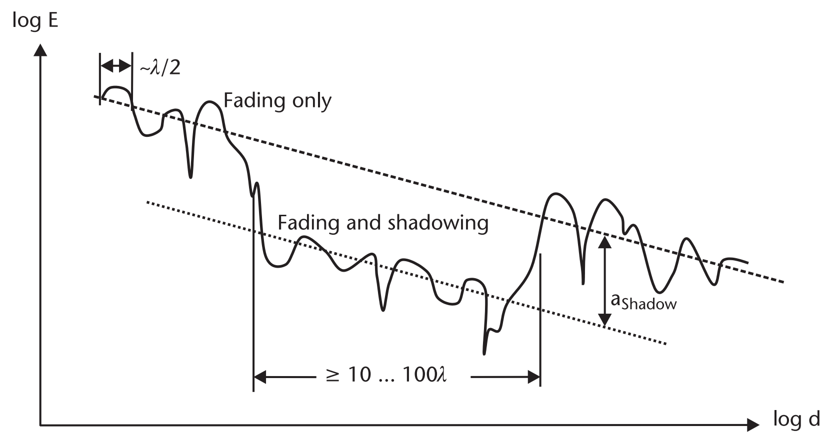

Figure 3.8 Fading and shadowing.

Fading and shadowing can be characterized by specific statistical distributions. For shadowing this is usually a log-normal distribution, whereas multipath Rayleigh fading is modeled by a Gaussian normal distribution. For scenarios with an additional direct path, a Rician distribution should be taken instead of the Rayleigh distribution. Of course, the Rician distribution turns into a Rayleigh distribution if the direct signal becomes zero. Fading and shadowing altogether can be modeled by a Suzuki distribution, which is achieved by multiplying both [5, 8, 11, 17, 24, 29–31].

The distributions for fading and shadowing can both be characterized by their logarithmic standard deviations σ expressed in decibels, which have to be squared before addition due to their statistical independence. This is an approximation because the underlying distributions are different. However, for many practical considerations it is accurate enough and allows easy handling [18, 29].

$$\sigma_P^2 = \sigma_F^2 + \sigma_S^2 \rightarrow \sigma_P = \sqrt{\sigma_F^2 + \sigma_S^2} \tag{3.42}$$

Dependent on the type of fading (i.e., Rayleigh or Rician fading), $\sigma_F \approx$ 3–6 dB can be estimated as an average value. For shadowing there is a much wider range of $\sigma_S \approx$ 6–12 dB, but values mostly around 8 dB are found. Hence, the standard deviation for both effects combined will be in the range of $\sigma_P \approx$ 7–13 dB. If a rough estimation about the coverage is intended and if there is no specific information about the terrain characteristics, a sufficiently accurate approximation might be $\sigma_P \approx 9$ dB. If, however, more comprehensive and detailed information about the terrain is available then, in particular, σ_S can be determined more precisely and the accuracy of the results will be improved. In the end the losses a_F and a_S due to fading and shadowing have to be added to the static path loss resulting in the dynamic path loss a_{LD}:

$$a_{LD} = a_{LS} + a_F + a_S \tag{3.43}$$

The *average* losses a_F and a_S caused by fading and shadowing can be set to about zero. If the most frequent statistical distributions are regarded more accurately, $a_F \approx 0$ to -2 dB (due to reflections) and $a_S \approx 0$ have to be assumed. The probability for sufficiently good reception is dependent on the difference Δp_R of the received signal level from that power level $p_{R\,50\%}$ that results in 50% probability for good reception. Assuming a log-normal distribution as an approximation, the following equation describes this probability p_G:

$$p_G(k\sigma_P) = \frac{1}{\sigma_P\sqrt{2\pi}} \cdot \int_{-k\sigma_P}^{\infty} e^{-\frac{(\Delta p_R)^2}{2\sigma_P^2}} \, d(\Delta p_R) \tag{3.44}$$

Here the level in decibels is $\Delta p_R = 10 \cdot \log_{10}(P_R / P_{R\,50\%})$, where P_R is the actual power and $P_{R\,50\%}$ is its average. Figure 3.9 shows the normal or Gaussian probability density function. Its integral is the normal distribution; see also Chapter 5 [21, 32]. Look-up tables are available for the normal distribution, but Table 3.7 gives some typical values to provide the reader with better insight into its properties.

If the availability should be, for example, 85% roughly σ_P has to be added, meaning $k = 1$. In practice, mobile radio systems are designed for an availability of 95–98%, which means that k becomes 1.6–2. Many PMR systems provide about 95% availability, but for police, fire brigades, and the like a much higher availability is required, for example, the U.K. Public Safety Radio Communication Service has been designed to provide 99% for mobiles and 96% for HPs.

Table 3.7 suggests that 100% availability can never be achieved, and extremely high availability requires an unrealistically high receiving power level, which in turn means that the average coverage distance becomes larger and consequently frequency reuse will become more critical. If a very high availability really is needed, then other means are preferable, for example, additional fixed transmitters and receivers to fill coverage holes.

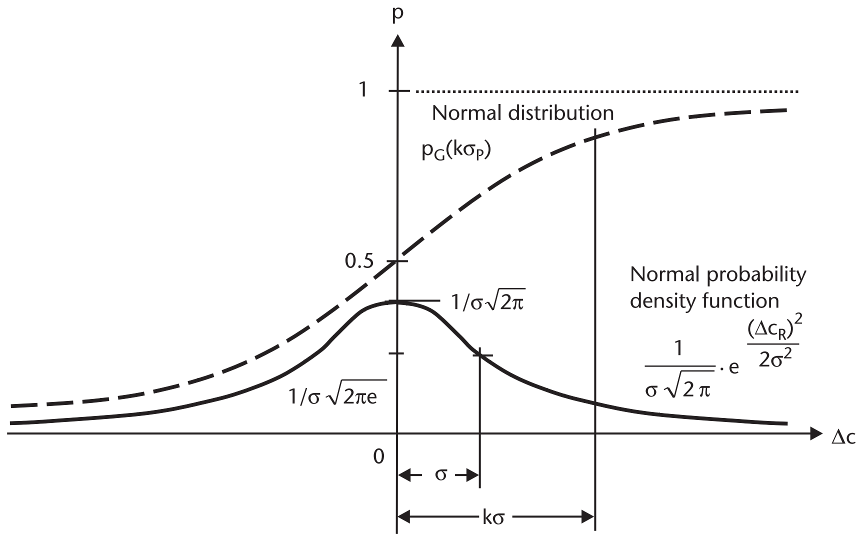

Figure 3.9 Normal probability density function and normal distribution.

Table 3.7 Probability of Good Reception

p_R [%]	50	75	85	90	95	98	99	99.5	99.9	99.99	99.999
k	0	0.67	1.04	1.28	1.64	2.05	2.33	2.58	3.09	3.75	4.27

Now the dynamic path loss a_{LD} can be determined by simply adding $k\sigma_P$ to the static path loss. For good reception the link budget must exceed the static path loss, but for the maximum coverage distance both become equal:

$$a_{LD} = a_{LS} + k\sigma_P \tag{3.45}$$

$$a_B = a_{TR} + g_\Sigma \geq a_{LS} + k\sigma_P = a_{LD} \tag{3.46}$$

Now a more realistic calculation of the maximum coverage distance including fading and shadowing can be made. For Hata's relation we find with $d_0 = 1$ km:

$$d_{\max} = d_0 \cdot 10^{a_{LD}/10\alpha} = d_0 \cdot 10^{(a_B - a_{1\,\text{km}} - a_T - k\sigma_P)/10\alpha} \tag{3.47}$$

For the dual-slope model the result is:

$$d_{\max} = d_0 \cdot 10^{a_{LD}/10\alpha} = d_0 \cdot 10^{(a_B - a_{1\text{m}} - a_0 - k\sigma_P)/10\alpha} \tag{3.48}$$

To get an impression about what this means in practice, the dynamic link budgets and the coverage distances have been calculated for some examples and the results are given in Table 3.8.

Compared to the 450-MHz band, the maximum distance covered at 150 MHz is considerably higher. The constant $a_{1\,\text{km}}$ in Hata's table is reduced by 11.5 dB and $a_{1\text{m}}$ in the dual-slope model is 9.5 dB smaller. The increases in distance are therefore factors around 2.1 for the Hata model and 1.7 for the dual slope model.

Table 3.8 Coverage Distances for Some Examples

Type of Link[1]	*Static Link Budget* $a_{TR} - g_{\Sigma}$ *[dB]*[2]	*Maximum Covered Distance* d_{max} *[km]*[3] *Okumura-Hata Model*	*Dual-Slope Model*
BS → MS[4]	161	6.3	—
BS → HP[4]	158	5.2	—
MS → MS[5]	157	—	4.6
MS → HP[5]	154	—	3.9
HP → HP[5]	144	—	2.2

[1]BS, base station; MS, mobile station; HP, hand portable; p_{TBS} = 44 dBm; p_{TMS} = 40 dBm; p_{THP} = 33 dBm; static receiver sensitivity p_R = –117 dBm.
[2]$g_{\Sigma BS,MS}$ = 0 dB and $g_{\Sigma HP}$ = –3 dB per station, to be combined appropriately.
[3]With availability = 95%, $k = 1.6$, and σ_P = 9 dB, we get $k\sigma_P$ = 14.4 dB.
[4]According to Hata with f_C = 450 MHz, h_B = 30m, and h_M = 1.5m; hence, $\alpha = 3.52$ and $a_{1\,km}$ = 118.5 dB.
[5]Dual-slope model with $d_S \approx d_0$ = 10m, $\alpha = 4$ for $d \geq$ 30m, a_{1m} = 25.5 dB, and a_0 = 29.5 dB.

3.6 Specific Problems Related to Multipath Propagation

The situation is similar for analog and digital transmission as long as they are both narrowband systems. However, it may change considerably if high data rates and broadband transmission are involved. Figure 3.10 shows the different effects that can occur and can impair radio signals. Besides the direct path there are numerous indirect paths involving reflection, diffraction, and scattering.

Due to multipath propagation, the mobile radio channel is dispersive. Therefore, the received signal $s_R(t)$ can be represented by a virtually infinite number of replicas of the transmitted signal $s_T(t)$, which arrive with differing delays τ_i and have experienced different (linear) path losses A_i:

$$s_R(t) = \sum_{i=0}^{\infty} A_i \cdot s_T(t - \tau_i) \tag{3.49}$$

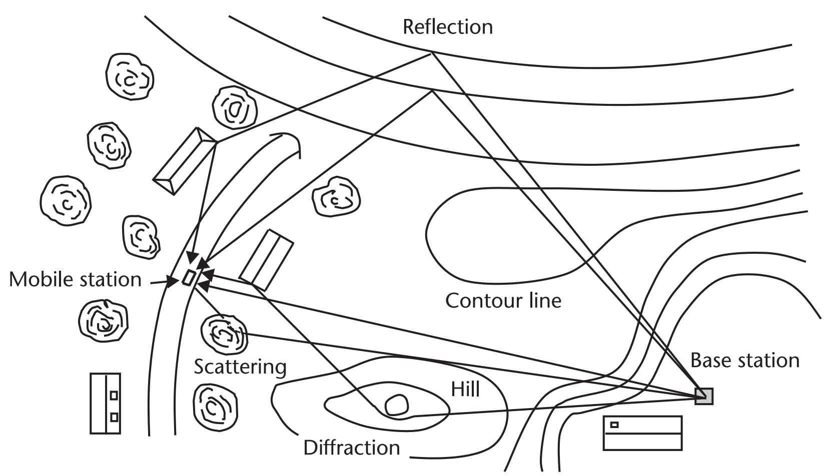

Figure 3.10 Multipath propagation.

Because very small portions of the whole received power do not contribute very much to the overall result in practice, only a limited number $i << \infty$ of signals have to be regarded at the receiver input. If the transmitter sends a sharp pulse, then at the receiver a number of echo pulses will be obtained. This sharp transmitted pulse can be represented by the Dirac function $\delta(t)$ and the signal pattern appearing at the receiver input is called the channel response $h(t)$:

$$h(t) = \sum_{i=0}^{\infty} A_i \cdot \delta(t - \tau_i) \tag{3.50}$$

A more generalized approach, which provides the same result as (3.49), is to regard the received signal as the convolution of the transmitted signal with the channel pulse response:

$$s_R(t) = \int_{-\infty}^{\infty} s_T(t - \tau) \cdot h(t)\, d\tau \equiv s_T(t) * h(t) \tag{3.51}$$

This is one of the most important relations used in system theory for the analysis of linear time-invariant systems for which the following rules are valid:

$$a * b = b * a \text{ and } a * (b * c) = (a * b) * c \text{ and } a * (b + c) = a * b + a * c \tag{3.52}$$

A moving mobile experiences ever-changing propagation conditions and the channel properties are varying quickly, but only slowly compared to the desired variations of the transmitted signals. Hence for a short time during which a mobile does not change its location too much, the propagation conditions are virtually constant and thus the tools of system theory are applicable.

Figure 3.11 shows two cases of how the channel response may look. Urban areas usually have many echoes with a small delay, but only rarely will echoes delayed by more than 10 μs exceed a receiving level p_R, which is 10 dB below the strongest received signal. If there is a direct LOS very often this will produce the first and strongest signal. All other signals travel via longer paths and will arrive later. If there is *no direct line of sight* (NLOS) or if there is a strong destructive two-ray interference, then the strongest component may stem from a short indirect path traveling via a good reflector [3, 11, 17, 24, 25, 29, 33–36].

In hilly or mountainous terrain, strong echoes with long delays may appear that are not much more attenuated than the main signal. This may cause severe intersymbol interference that can only be cured by a channel equalizer. This becomes more critical as the bit rate increases. Hence, as Figure 3.11 suggests, GSM signals for instance are much more vulnerable to multipath propagation distortion than are TETRA signals.

The scatter function describes what happens to a radio signal traveling along its transmission path. The received total power density $S(t, \tau, f_D)$ is collected over a time interval and the frequency band over which the power of the transmitter

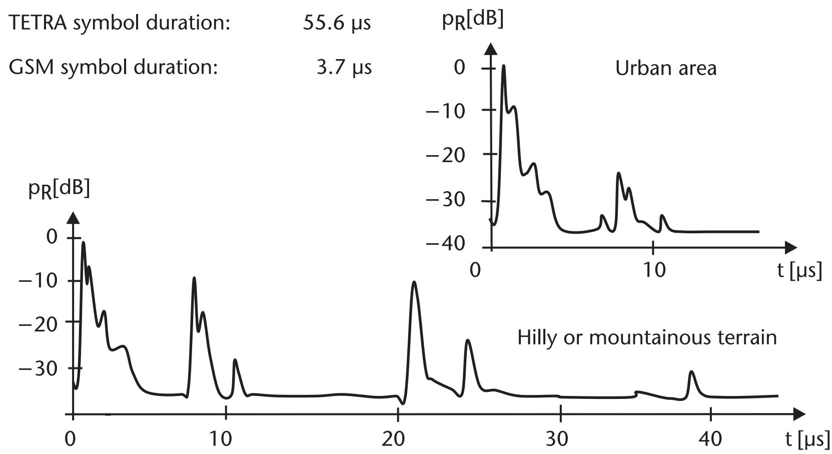

Figure 3.11 Delay profile examples.

signal has been widened. This is equivalent to an integration over time and frequency and delivers the total received power $P_R(t)$ [37]:

$$P_R(t) = \int_{f_D=-\infty}^{\infty} \int_{t=-\infty}^{\infty} S(t, \tau, f_D)\, d\tau\, df_D \tag{3.53}$$

If the integration is carried out only in the time domain, then the Doppler power density spectrum is the result:

$$S_D(t, f_D) = \int_{t=-\infty}^{\infty} S(t, \tau, f_D)\, d\tau \tag{3.54}$$

The Doppler power density spectrum is limited by the Doppler frequencies $\pm f_D$ (see below) and can often be approximated by the following equation with a minimum at the carrier frequency f_C:

$$S_D(t, f_D) = S_0 \Big/ \sqrt{1 - (\Delta f / f_D)^2} \tag{3.55}$$

Integration of the power density over the frequency range delivers the delay power density spectrum, which is nothing but the square of the channel pulse response:

$$S_\tau(t, \tau) = \int_{f_D=-\infty}^{\infty} S(t, \tau, f_D)\, df_D \sim h^2(t, \tau) \tag{3.56}$$

If propagation conditions are to be investigated, these quantities are measured and displayed by highly specialized equipment. Nevertheless, in modern digital radio receivers they have to be computed to estimate the channel behavior and to compensate for the impairments caused by multipath propagation. Figure 3.12 shows how the energy of a transmitted pulse is smeared over time and frequency and how Doppler power density spectrum and delay power density spectrum (delay profile) together produce the scattered signal.

If a signal of constant power density in a certain bandwidth during a certain time interval is transmitted it will not be constant either in time or in frequency at the receiver input as Figure 3.13 shows.

Again the situation for GSM is very different from that for TETRA. The GSM broadband burst signal is very short and exhibits strong impairments in the frequency domain. The narrowband TETRA signal, in contrast, shows negligible frequency response impairments but the amplitude varies considerably over burst time. Hence, channel equalizing requires mainly the removal of frequency impairments in one case and time variations in the other.

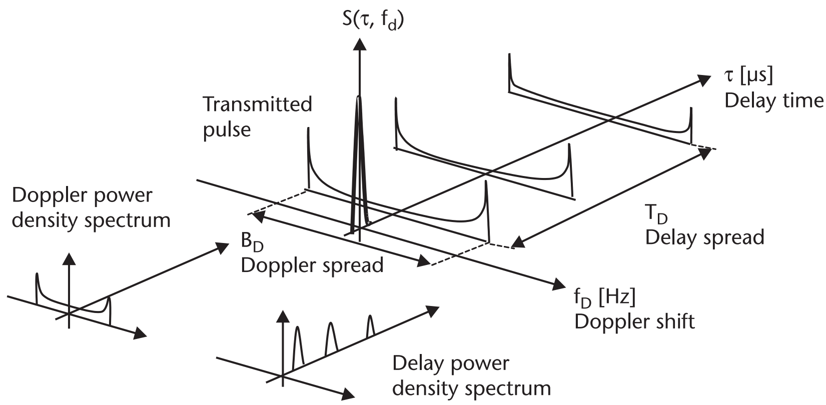

Figure 3.12 Energy scattering by multipath propagation.

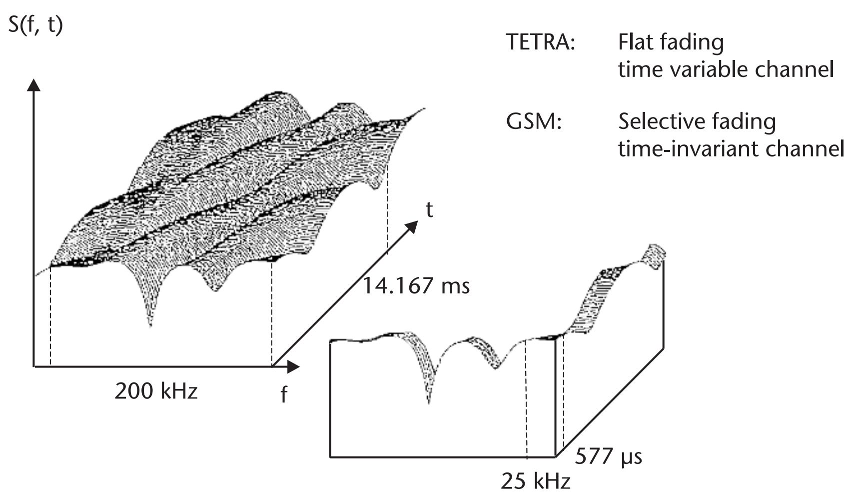

Figure 3.13 Fading with GSM and TETRA.

Multipath propagation effects can be characterized by several typical quantities. The first is the Doppler frequency or Doppler shift f_D caused by the movement of one or both stations constituting a radio link. It is at its maximum if the motion is on a direct line between the two stations, which means that the angle $\delta = 0$, and it is zero if both lines are perpendicular to each other meaning $\delta = 90°$. The Doppler shift depends on the carrier frequency f_C, the speed of light v_C, and the vehicle speed v_M. The Doppler spread B_D is the bandwidth over which each single frequency is smeared. Due to multipath propagation it generates a band from $-f_D$ to $+f_D$ [11].

$$f_D = \frac{v_M}{v_C} \cdot f_C \cdot \cos\,\delta = \frac{v_M}{\lambda} \cdot \cos\,\delta = f_D \cdot \cos\,\delta \tag{3.57}$$

$$B_D = 2 \cdot f_D \tag{3.58}$$

The coherence bandwidth is that bandwidth within which there is good correlation between the different components of the modulated signal. It depends on the propagation conditions and the properties of the modulated signal [5, 11, 24, 35, 38, 39]. The simplest case is represented by the two-ray model where the reflected ray is delayed by τ. The reflection coefficient may vary between 0 and 1, while its phase jump may in principle be somewhere between $\pm\pi$ depending on the polarization and the properties of the reflector. The correlation of two signals is defined by the correlation coefficient ρ, which can be written for sinusoidal signals as follows [37, 40]:

$$\rho = \frac{2}{S_{C1} \cdot S_{C2} \cdot T} \cdot \lim_{T\to\infty} \int_{-T/2}^{T/2} s_1(\omega_1 t) \cdot s_2(\omega_2 t + \Delta\Phi)\, dt \tag{3.59}$$

where S_{C1} and S_{C2} are the signal amplitudes and T is the integration time. If $s_1(\omega_1 t)$ and $s_2(\omega_2 t)$ are different signals, then this is a cross-correlation. If, however, both signals disregarding a time shift τ are identical, then it is an autocorrelation. This general definition becomes much simpler for signals with equal amplitude but a frequency difference $\omega_1 - \omega_2 = \Delta\omega = 2\pi \cdot \Delta f$ or a phase difference $\Delta\Phi$ [21]:

$$\rho = \begin{cases} \cos(\Delta\Phi) & \text{for } \Delta f \equiv 0 \\ si(2\pi\Delta f T) & \text{for } \Delta\Phi \equiv 0 \quad \text{with} \quad si(x) = \dfrac{\sin(x)}{x} \end{cases} \tag{3.60}$$

due to

$$\int \cos\,ax \cdot \cos(ax + b)\, dx = \frac{1}{4a} \cdot \sin(2ax + b) + \frac{x}{2} \cdot \cos\,b$$

and

$$\int \cos\,ax \cdot \cos\,bx\, dx = \frac{\sin[(a-b)x]}{2(a-b)} + \frac{\sin[(a+b)x]}{2(a+b)}$$

Because radio waves travel with the speed of light v_C, that is, 300m in 1 μs, the phase difference between two signals from one transmitter spaced by the coherence bandwidth B_C and differently delayed by τ is:

$$\Delta\Phi = 2\pi \cdot B_C \cdot \tau \quad \text{with} \quad \tau = \Delta d / v_C \tag{3.61}$$

In the general case there are a large number of differently delayed and attenuated signals and τ has to be replaced by the delay spread T_D. For phase modulation schemes, a better correlation than for amplitude modulation schemes is usually necessary and therefore the precise definition of the coherence bandwidth depends on the grade of correlation ρ between the different parts of the modulated signal:

$$B_C = \frac{\Delta\Phi}{2\pi T_D} \rightarrow \begin{cases} \infty & \text{for } \rho = 1 & \text{and} & \Delta\Phi = 0 \\ 1/12T_D & \text{for } \rho = \sqrt{3}/2 & \text{and} & \Delta\Phi = \pi/6 \\ 1/8T_D & \text{for } \rho = 1/\sqrt{2} & \text{and} & \Delta\Phi = \pi/4 \\ 1/6T_D & \text{for } \rho = 1/2 & \text{and} & \Delta\Phi = \pi/3 \\ 1/4T_D & \text{for } \rho = 0 & \text{and} & \Delta\Phi = \pi/2 \\ 1/2T_D & \text{for } \rho = -1 & \text{and} & \Delta\Phi = \pi \end{cases} \tag{3.62}$$

Hence the coherence bandwidth is inversely proportional to the delay spread T_D, which is nothing but the duration of the channel response $h(t)$. For an amplitude correlation of $\rho = 0.5$ or $\Delta\Phi = \pi/3$, Lee gives [11, 24, 35]

$$B_{CA} = \frac{1}{2\pi \cdot T_D} \tag{3.63}$$

From (3.62) we obtain a factor of 6 instead of 2π, which is a pretty good estimate. For phase modulation a phase correlation of $\Delta\Phi = 0.5 \approx \pi/6$ is assumed and therefore a smaller bandwidth results:

$$B_{CP} = \frac{1}{4\pi \cdot T_D} \tag{3.64}$$

From our approximation we get $\rho = 0.88$ and a factor of about 12 instead of 4π, which also is very close. To estimate the coherence bandwidth without regard to the type of modulation, Lee recommends the following assumption:

$$B_C = \frac{1}{8 \cdot T_D} \tag{3.65}$$

A typical path length in an urban area may be about 3 km. Depending on the radio environment, we can estimate that $T_D \approx \Delta d \approx d/30$ to $d/3$ and therefore usually $T_D = 0.3$–3 μs and only rarely is it larger. Thus a coherence bandwidth B_C of 40–400 kHz results, which for typical PMR systems is on average considerably larger than the modulation bandwidth. Hence, PMR systems suffer usually from flat fading and only very seldom from selective fading.

We now look at some parameters that characterize the channel stability in the time domain. The coherence time T_{CC} is the time during which the channel propagation properties remain nearly constant:

$$T_{CC} = \frac{1}{f_D} \tag{3.66}$$

If only the signal amplitude is taken into account, then a much shorter stability time T_{CA} results [10]. The distance between a fading maximum and a null in the worst case is $\lambda/4$. If an incremental distance of $\leq \lambda/8$ as a maximum is assumed then a mobile traveling with velocity v_M needs $T_{CA} = \lambda/8v_M$ for this distance and we get, using (3.57):

$$T_{CA} \leq \frac{1}{8 \cdot f_D} \tag{3.67}$$

Another point is freedom from intersymbol interference, which restricts the maximum usable symbol rate R_S. A symbol is usually sampled at its center and the symbol duration T_S should therefore exceed twice the constantly changing maximum delay spread T_D:

$$R_S = \frac{1}{T_S} < \frac{1}{2 \cdot T_D} \tag{3.68}$$

Depending on the modulation scheme in practice, the aperture of the eye pattern is somewhat shorter than the symbol duration and, therefore, $R_S \leq 1/4T_D$ is more realistic. This is true in particular for multilevel modulation schemes. If the symbol rate exceeds this limit, a channel equalizer becomes necessary. Finally, one often-used rule of thumb for the relation between symbol rate and Doppler shift is as follows:

$$100 \cdot f_D < R_S \tag{3.69}$$

This means that a bit rate of 10 kbit/s in a binary modulation scheme would suggest that the Doppler shift should not exceed 100 Hz to avoid problems.

3.7 Some Specific Propagation Problems with Digital Radio Transmission

For linear modulation schemes, the dynamic range must be known to properly design the receiver. The dynamic range is the difference between the maximum possible power level at the receiver input and the sensitivity level. The maximum level can be calculated from the transmitter power level and the minimum path loss to the mobile assuming free-space propagation (refer back to Figure 3.2). As an example the minimum static path loss between base and mobile station can be calculated:

$$d_{d\,\min} = \sqrt{(h_B - h_M)^2 + d^2} \tag{3.70}$$

If $h_B = 30\text{m}$, $h_M = 1.5\text{m}$ and the horizontal distance $d_h = 30\text{m}$ are assumed, then the path length becomes $d_{d\,\min} = 41.4\text{m}$. Because the radiation from both antennas is not in the main lobe direction but tilted by 45° from the horizontal plane, an additional loss of 3 dB on average has to be added for each antenna to approximate the minimum static path loss $a_{LS\,\min}$:

$$a_{LS\,\min} \approx 20 \cdot \log_{10}(4\pi d/\lambda) + 6 \text{ dB} \tag{3.71}$$

The result is 48 dB for 75 MHz and 70 dB for 900 MHz. For $d < h_B$ the antenna gain reduction increases and tends to compensate the decreasing loss due to the shorter path length. In the case of wide-area coverage, the transmitter power might be greater but usually the base station antenna height will then also be greater. Both tend to compensate for each other with regard to the minimum path loss.

Assuming a transmitter power of 40 dBm (10W) the necessary maximum dynamic range of the receiver can now be calculated, but due to the additive effects of fading the maximum receiver input level may temporarily be up to about 10 dB higher. Hence, depending on the carrier frequency, transmitter power, receiver sensitivity, and so on, the necessary dynamic input level range of the receiver can vary roughly between 80 and 120 dB.

In the preceding discussions, we have always assumed that the resulting antenna gain and the transmit power at the base station and the mobile station are roughly equal and that the link budget is therefore balanced. However, mobile stations mostly transmit lower power than base stations and therefore the link budget is often different for downlinks versus uplinks. In the case of hand portables, the imbalance can be on the order of 10 dB or more due to their much smaller transmit power as determined by limited battery capacity and health considerations.

Several possibilities exist to avoid different maximum coverage distances for downlink and uplink. If the receiving and the transmitting antennas are different at the base station for whatever reason, then the receiving antenna may be one with a higher gain. Spatial diversity is another means often employed. For this purpose a second base station antenna is installed for the receiver at a distance that ensures sufficient decorrelation between the two antennas. Due to fading the signals at both antennas are varying in time and it is most likely that one is always better than the other. The trick is to switch the receiver to that antenna which receives the better signal. Very sophisticated algorithms and hardware devices have been developed to achieve this, but the easiest method is to select one antenna arbitrarily and to keep it until the quality decreases below that minimum level where the radio link would collapse. Most likely the other antenna then gets a better signal and should be kept until the quality gets too bad and so on. The improvement of spatial two-branch diversity is roughly on the order of 4–10 dB [3, 25]. If a perfect algorithm is assumed, that is, one that always switches the receiver to that of an infinite number of antennas with the best signal, the influence of fading would not only be totally removed, but a certain additional gain could be found due to always selecting the situations of constructive interference where the signal level is above the average. In reality, the biggest step is made by introducing a second antenna.

If the link imbalance exceeds the maximum possible gain by high-gain antennas or spatial diversity, then so-called *satellite receivers* (SRs) can be used as shown in Figure 3.14. Therefore, the main base station receiver is complemented by some remote receivers in different directions and at a distance that compensates the link budget imbalance. In the example shown in the figure, for all MSs a balanced link budget is assumed. Their smaller transmitting power, however, results in a smaller coverage distance for the HPs, but the remote BS satellite receivers are placed in such a way that at least one of them can always receive a signal with sufficient quality. At the base station the signals from all satellite receivers are evaluated and only the best is processed further. Due to multipath propagation impairment, sometimes the strongest signal is not the one that provides the best quality—a much weaker signal at another receiver might be significantly less distorted.

No specific problems are usually related to the Doppler shift for conventional analog systems. However, some digital transmission schemes are vulnerable especially to large Doppler shifts. Specifically, a fast change of the Doppler shift may cause synchronization difficulties and this problem increases with the frequency.

At a maximum vehicle speed of v_{max} = 360 km/hr or 100 m/s as can be found, for example, with very fast trains, at f_C = 450 MHz a Doppler shift of f_D = 150 Hz results. This is small compared to the maximum permissible center frequency difference between transmitter and receiver in the unsynchronized state, that is, 2.5 kHz for a 12.5-kHz channel separation in the 450-MHz band (see also Chapter 4). It is also small compared to the maximum permissible peak frequency deviation of 2.5 kHz. Hence, the height of the eye pattern will not be significantly affected by the Doppler spread. The Doppler shift also should not, as long as it does

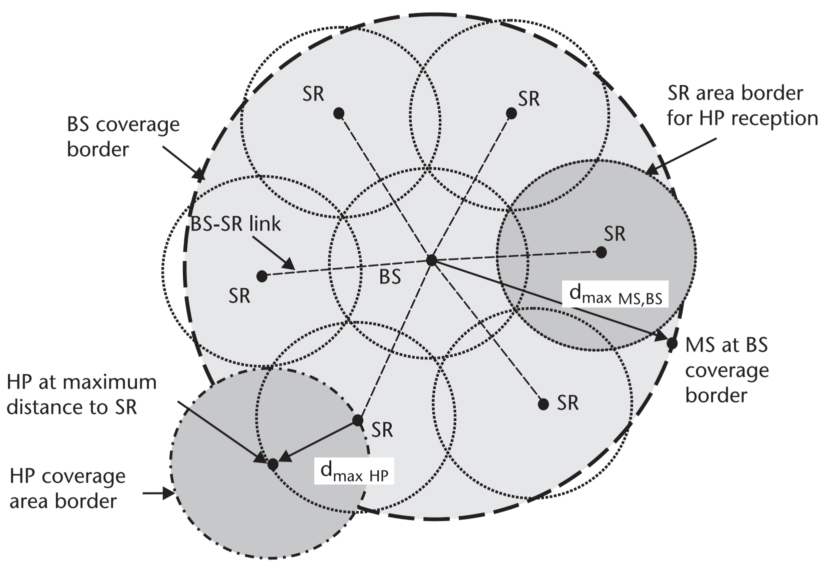

Figure 3.14 HP coverage improvement by BS satellite receivers.

not change too fast, affect the symbol synchronization. Noncoherent slow FSK modulation schemes are robust against such impairments. However, for high bit rates and coherent demodulation, which are usually employed if maximum possible sensitivity is targeted, the situation is somewhat different. The most critical case is a fast change in Doppler shift that cannot be compensated for quickly enough by the synchronization circuitry and that therefore may cause a temporary error burst and subsequent loss of information when the coherent demodulation fails.

An example is a car heading directly toward a BS antenna at a frequency of 450 MHz at maximum speed trying to stop as quickly as possible or to start and accelerate to maximum speed as fast as possible. Assuming that the vehicle gets from 0–30 m/s or 108 km/hr within 3 seconds, or that it is able to slow down from 30 m/s to 0 within the same time, then it would need an acceleration of $b_M = \pm 10$ m/s$^2 \approx$ 1g. Hence the Doppler shift would change within 3 seconds from 45 Hz to 0 or from 0 to –45 Hz. The maximum rate of the Doppler shift change will then be

$$\frac{df_D(t)}{dt} = \frac{f_{D\max} \cdot b_M}{v_{\max}} \tag{3.72}$$

For our example the result is $df_D(t)/dt = \pm 15$ Hz/sec. In reality a car will rarely be able to perform this maneuver and the speed of the Doppler shift change will be significantly smaller. This case will not present any real problem.

A much more critical situation is a train passing a BS antenna with a constant but very high speed at a very small distance as shown in Figure 3.15. Provided that the angle between the direction of the railway and the line between the vehicle and the BS antenna location is δ and that at the time $t = 0$ at minimum distance the Doppler shift dependency of the Doppler $f_D(0) = 0$, we get

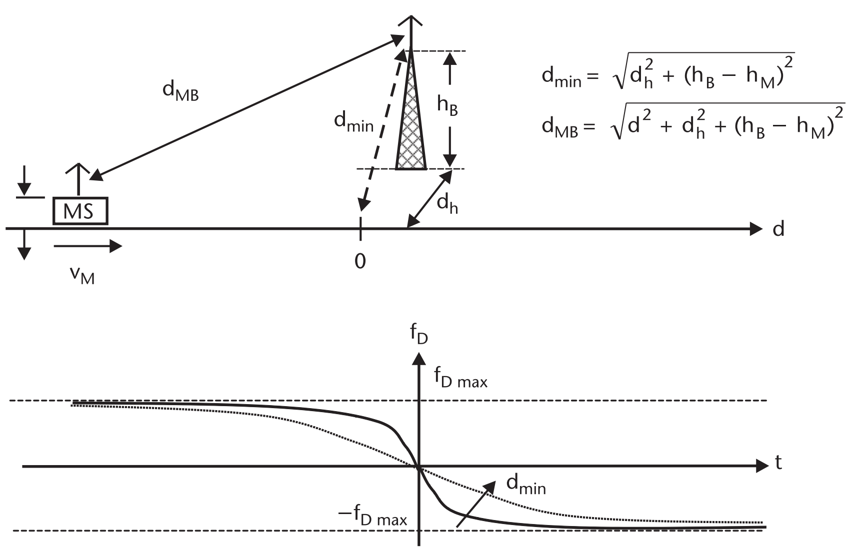

Figure 3.15 Doppler shift change for a mobile passing the base station.

$$f_D(t) = -f_{D\max} \cdot \cos\,\delta = \frac{-f_{D\max} \cdot v_M \cdot t}{\sqrt{d_{\min}^2 + (v_M \cdot t)^2}} = \frac{-f_{D\max}}{\sqrt{1 + \left(\frac{d_{\min}}{v_M \cdot t}\right)^2}} \tag{3.73}$$

By differentiation we get the speed of the change of the Doppler shift, and the maximum rate of change of the Doppler shift k_D will be found at $t = 0$:

$$\frac{df_D(t)}{dt} = \frac{-f_{D\max} \cdot v_M}{d_{\min} \cdot \left[\sqrt{1 + \left(\frac{v_M \cdot t}{d_{\min}}\right)^2}\right]^3} \tag{3.74}$$

$$\frac{df_D(0)}{dt} = \frac{-f_{D\max} \cdot v_M}{d_{\min}} = k_D \tag{3.75}$$

Assuming a horizontal minimum distance $d_h = 5$m and an antenna height of $h_B = 20$m, a minimum distance of $d_{\min} = 21$m results. For $f_C = 450$ MHz and $v_M = 360$ km/hr = 100 m/s the train approaching from a great distance will exhibit the maximum positive Doppler shift of $f_{D\max} = 150$ Hz, which reduces to zero at the smallest distance and then will change its sign instantly to $-f_{D\max}$ with decreasing rate of change. The maximum Doppler shift rate of change is $k_f = -714$ Hz/sec and this is much more by far than in the first case. Hence, a fast moving vehicle passing close by a base station temporarily causes a very high rate of change of the Doppler shift, which has the potential to cause a severe synchronization problem.

For coherent demodulation, a fast frequency and phase control (channel tracking) are needed to compensate around the zero crossing of the Doppler shift during the time T a phase error of $\Delta\Phi_D$. Here we assume that around zero the Doppler shift is nearly linear in time:

$$\Delta\Phi_D = 2\pi \cdot \int_{-T/2}^{T/2} f_D(t)\,dt \approx 2\pi k_f \cdot \int_{-T/2}^{T/2} t\,dt = \pi k_f \cdot T^2 \tag{3.76}$$

Different digital PMR systems typically exhibit a time slot duration between 15 and 30 ms. Assuming a 20-ms slot, a phase error of $\Delta\Phi_D \approx \pi/4$ may result in our last example during one time slot if the channel tracking does not work well. For *Gaussian minimum shift keying* (GMSK) and *quaternary phase shift keying* (QPSK), the minimum phase difference between two symbols is only $\pi/2$. The maximum permissible tolerance depends on the implementation and is $\pi/8$ to $\pi/4$. Hence a fast change as found in our last example may introduce severe synchronization errors that may last in the worst case for several time slots until resynchronization is achieved.

3.8 Frequency Economy and Frequency Management

In radio communications the efficient use of frequencies is an issue of paramount importance. However, the task of evaluating frequency economy is more complicated than it appears at first glance because the economic use of frequencies can be defined in different ways. A simple one-dimensional approach assesses the bit rate per bandwidth in bits per second per hertz. This is a good measure if no spatial frequency reuse is intended because this is the case for *noise-limited* or *coverage-limited* systems. Examples are fixed radio links, satellite systems, and many single-site PMR systems [10, 11, 41, 42].

The spectral efficiency of selected mobile radio communication systems in terms of bits per second per hertz is compared in Table 3.9. For all systems only the original system properties have been listed and none of their more advanced versions. D-AMPS exhibits the best spectral efficiency of the standardized current mobile radio systems. However, due to the high user bit rate but weak error correction, it is more vulnerable to multipath impairments than is, for instance, GSM, which also looks very good at the first glance, however, it does not meet the rigid requirements for the adjacent channel power that PMR systems have to fulfill. In contrast, TETRA, APCO 25, and DIIS really make very efficient use of the assigned bandwidth. The old analog systems including the AMPS, TACS, German Net C, and the NMT system have certain data transmission capabilities but at much lower spectral efficiency [10, 43–46].

The situation is different for cellular systems and PMR in densely populated regions where spatial frequency reuse is a must. Such systems are called *interference limited* and frequency economy has to be regarded there as a two-dimensional quantity and appropriate measures are bits per second per hertz per cell or bits per second per hertz per square kilometer. In the end radio capacity expressed in channels per megahertz per cell or channels per megahertz per square kilometer is the measure in which radio engineers are interested. Hence, a closer look at frequency economy is desirable [2, 11, 17, 20, 24, 35, 39, 41, 42].

The spectral efficiency η_N of noise- or coverage-limited systems expressed in bits per second per hertz depends mainly on the channel separation ΔF_{Ch} and the user bit rate R_{BU}. System designers may use different types of trade-offs between the modulation rate, the type of modulation, and the error correction by appropriate channel coding, but for the user only that bit rate that he or she really can use is of interest and not the gross bit rate. Additionally, the factors N_A and N_M have to be taken into account. Hence,

$$\eta_N = \frac{N_A \cdot N_M \cdot R_{BU}}{\Delta F_{\mathrm{Ch}}} \tag{3.77}$$

The access factor N_A depends on the number of traffic channels that are modulated onto one carrier frequency. $N_A = 1$ for FDMA systems such as analog PMR or APCO 25 while it is equal to the number of time slots in a TDMA frame for TDMA systems not employing *time division duplex* (TDD). It is 4 for TETRA, 8 for GSM full-rate, and 16 for GSM half-rate channels. In a CDMA system it is the number of usable spreading codes and this can be very large:

Table 3.9 Spectral Efficiency Examples for Mobile Radio Systems

Radio Communication System	*Channel Access*	*Mode of Operation*[1]	*Type of Modulation*	*Channel Separation* ΔF *[kHz]*	*Gross Bit Rate* R_B *[kbit/s]*	*Spectral Efficiency*[2] η_R *[bit/s/Hz]*	*User Bit Rate* R_{BN} *[kbit/s]*
D-AMP, IS-54/136	3TDMA	D	π/4-DQPSK $r = 0.35$	30	48.6	1.62	3 × 9.6[3]
TETRA V+D, PDO	4TDMA, Packet	S, D, SD	π/4-DQPSK $r = 0.35$	25	36	1.44	4 × 7.2 ≤ 28.8[3]
GSM 900, DCS 1800[4]	8/16TDMA (FR/HR)[5]	D	GMSK BT = 0.3	200	270.83	1.35	8 × 9.6 (FR)[6]
DIIS[7]	FDMA	S, D, SD	4GFSK BT = 0.28	12.5	12	0.96	≤9.6[3]
APCO 25	FDMA	S, D, SD	4FSK	12.5	9.6	0.77	7.2[3]
DECT[4]	24TDMA	TDD	GMSK BT = 0.5	1728	1152	0.67	24 × 32[6]
DSRR[8]	FDMA	SD, S	GMSK BT = 0.3/0.5	25	16/4	0.64/0.16	≤13[3]
AMPS	FDMA	D	FSK	30	10	0.33	—
TACS	FDMA	D	FSK	25	8	0.32	—
C 450[9]	FDMA	D	FSK	20	5.28	0.26	—
PMR FM, PM	FDMA	S, D, SD	SC-FSK[10]	12.5, 20, 25	2.4/1.2	0.048–0.192	≈0.6/0.3[6]
NMT-450, NMT-900	FDMA	D	SC-FSK[10]	12.5, 20, 25	1.2	≈0.048–0.096	—

[1] D, duplex; TDD, time division duplex; S, simplex; SD, semiduplex.
[2] Based on gross bit rate.
[3] Transparent.
[4] Higher bit rates are available in advanced versions.
[5] Full-rate/half-rate channels.
[6] Protected.
[7] Under preparation.
[8] Withdrawn December 1995.
[9] Net C.
[10] Subcarrier FSK.

$$N_A = \begin{cases} 1 & \text{for FDMA} \\ >1 & \text{for TDMA and CDMA} \end{cases} \tag{3.78}$$

The mode factor N_M is a measure of the number of frequencies needed for one radio link. In the case of a simplex channel, $N_A = 1$. The capacity of one simplex channel continuously used for 1 hour is called a *radio traffic channel* (RTC) while for comparison 1 Erlang (Erl) originally meant the capacity of a wireline duplex channel used for 1 hour [41]. Hence one duplex radio channel without a repeater means $N_M = 1/2$. Duplex is rare in conventional PMR but commonplace in cellular systems where even the calls within one cell are running via its base station:

$$N_M = \begin{cases} 1 & \text{for simplex} \\ 1/2 & \text{for semi-duplex or full duplex} \\ 1/4 & \text{for full duplex with repeater} \end{cases} \tag{3.79}$$

In a noise-limited system with the system bandwidth B_{Syst} besides the spectral efficiency η_N the number N_N of channels per carrier in RTC can be calculated in a manner similar to that for η_N:

$$N_N = \frac{N_A \cdot N_M \cdot B_{\text{Syst}}}{\Delta F_{\text{Ch}}} \tag{3.80}$$

FDMA- or TDMA-based cellular systems and PMR systems, which use frequencies in a regular spatial reuse pattern, are interference limited and a frequency usually cannot be reused in adjacent cells but only at a larger distance. Therefore, in comparison to noise-limited systems additionally the cell cluster size N_C has to be taken into account, which reduces considerably the efficient use of frequencies expressed as η_I in bits per second per hertz per cell or as the number of radio channels N_I in RTC per cell:

$$\eta_I = \frac{N_A \cdot N_M \cdot R_{BU}}{N_C \cdot \Delta F_{\text{Ch}}} \tag{3.81}$$

$$N_I = \frac{N_A \cdot N_M \cdot B_{\text{Syst}}}{N_C \cdot \Delta F_{\text{Ch}}} \tag{3.82}$$

For clear differentiation from the *spectral efficiency* η_N the two-dimensional measure η_I is better called *spatial efficiency* or simply *frequency economy* to avoid confusion with spectral efficiency. The critical point of interference-limited systems that determines the cluster size is *cochannel rejection*. The interfering cell must be far enough away to ensure that the interfering cochannel signal at the cell border is at least smaller by the cochannel rejection than the wanted signal, as Figure 3.16 shows [10, 24, 42].

The radii of all cells are assumed to be equally R while the distance of the centers of cells with equal frequencies should always be D. Then the P_C/P_I ratio

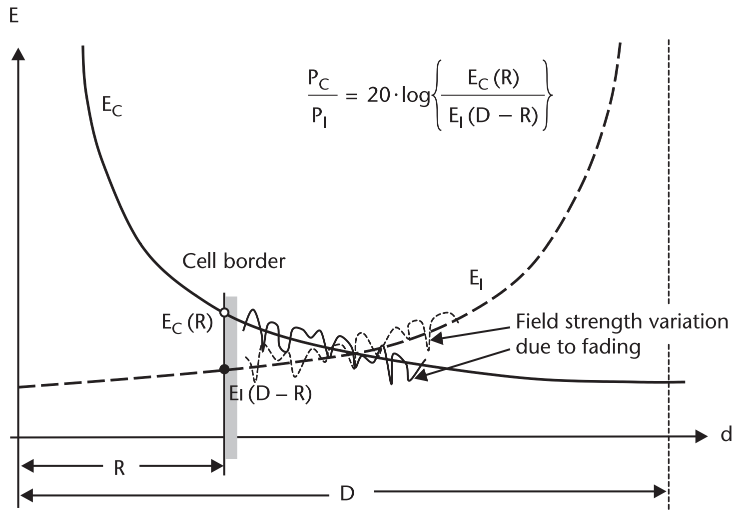

Figure 3.16 Field strength at the cell border.

at the cell border can be calculated directly or be determined from the field strengths E_C of the wanted carrier and E_I of the interferer.

$$(p_C/p_I) = 10\ \log_{10}(P_C/P_I) = 20 \cdot \log_{10}\left\{\frac{E_C(R)}{E_I(D-R)}\right\} \tag{3.83}$$

On flat and homogeneous terrain, all cells would ideally be circles with equal diameters if all base stations transmit equal power from equal omnidirectional antennas with identical height. Unfortunately, it is impossible to cover a surface with circles without any gaps. Therefore, hexagons are chosen instead because their shape is very close to a circle. Figure 3.17 shows as an example clusters with seven cells [17, 24, 39, 42].

In all clusters cells with identical frequencies are at a maximum distance from each other in all directions and form a pattern of triangles with identical side lengths. Based on simple geometric relations, cluster size N_C and D/R ratio are linked:

$$(D/R) = \sqrt{3N_C} \quad \text{or} \quad N_C = \frac{1}{3} \cdot (D/R)^2 \tag{3.84}$$

It is usually $P_C/P_I > 1$, which means $D > 2R$ and that the same frequency cannot be reused in adjacent cells. Paving the plane without gaps is only possible for certain integer numbers of N_C cells per cluster, which can be found with the equation below where i and k are integers and at least one of them is greater than zero:

$$N_C = i^2 + ik + k^2 \tag{3.85}$$

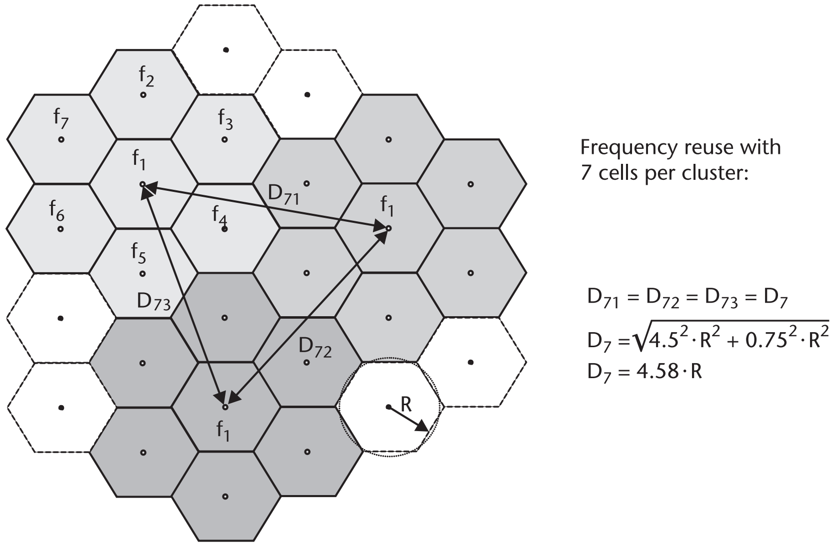

Figure 3.17 Frequency reuse.

Table 3.10 lists permissible values of N_C. However, for a certain number N_C more than one configuration may exist for the cluster shape. For example, for $N_C = 9$ two different shapes have commonly been used. Moreover, identical N_C may also be found by different combinations of i and k as $N_C = 49$ proves.

Cluster sizes of 7 are applicable for some mobile radio systems; for others a cluster size of, say, 12 is more appropriate. In addition to the terrain conditions additional features of the radio system have to be taken into account. Systems with power control and voice activity detection exhibit reduced interference and therefore the cluster size can be smaller than calculated without these features.

For CDMA $N_C \approx 1$ is claimed and Lee reports $N_C = 1.33$ for one particular example, which means that the number of usable radio channels per cell is somewhat reduced due to the influence of the surrounding CDMA cells, generally resulting in $1.0 < N_C < 2.0$ [35]. However, CDMA is not well suited for PMR for various reasons, mainly because it only performs well for a bandwidth exceeding 1 MHz, whereas PMR systems often require only a much smaller radio capacity and bandwidth.

In most PMR systems based on FDMA or TDMA, $7 \le N_C \le 19$, where N_C depends on the application, the propagation conditions, the transmission quality, and the properties of the radio channel. Hence, for PMR systems the capacity difference between noise- and interference-limited systems is 1:7 or even considerably worse.

The calculation of the necessary D/R ratio according to (3.83) must be based on the dynamic carrier-to-interferer power ratio $(P_C/P_I)_D$ including fading and shadowing:

$$(P_C/P_I)_D = (D/R - 1)^\alpha \tag{3.86}$$

For current analog PMR systems, the static carrier-to-interferer level difference $(p_C - p_I)_S = 10 \cdot \log_{10}(P_C/P_I)_S$ is 8–12 dB, depending on the different modulation

Table 3.10 Possible Cluster Sizes

i	1	1	2	2	3	2	3	4	3	4	5	3	4	5	6	4	5	6	4	5	7
k	0	1	0	1	0	2	1	0	2	1	0	3	2	1	0	3	2	1	4	3	0
N_C	1	3	4	7	9	12	13	16	19	21	25	27	28	31	36	37	39	43	48	49	49

index values for the channel separations of 25, 20, and 12.5 kHz. However, for further considerations the dynamic value including fading and shadowing is needed, and a standard deviation of $\sigma_P = 9$ dB can be assumed as already discussed if the precise propagation conditions dependent on the properties of the environment are not accurately known. However, for P_C/P_I considerations both signals, carrier and interferer, are subject to fading and shadowing with a standard deviation σ_I:

$$\sigma_I^2 \approx \sigma_P^2 + 10 \cdot \log_{10}\left(1 + \frac{P_I}{P_C}\right) \tag{3.87}$$

For the most part, $P_I << P_C$ and therefore $\sigma_I \approx \sigma_P$. The necessary dynamic cochannel protection $(p_C - p_I)_{AD}$ for analog transmission can now be calculated from the static value $(p_C - p_I)_{AS}$ as has been done for the dynamic link budget with (3.46):

$$(p_C - p_I)_{AD} = (p_C - p_I)_{AS} + k\sigma_I \tag{3.88}$$

For good analog reception at least $k \geq 1.5$ with $\sigma_I \approx 9$ dB and 90–95% probability is necessary:

$$(p_C - p_I)_{AD} \geq (p_C - p_I)_{AS} + 1.5 \cdot \sigma_I \approx (p_C - p_I)_{AS} + 13 \text{ dB} \tag{3.89}$$

Therefore, 21–25 dB, dependent on the channel separation of 25, 20, or 12.5 kHz, is needed for fairly good reception. For 30-kHz analog systems with $(p_C - p_I)_{AS} \approx 6$ dB this becomes approximately the well-known 18-dB rule of thumb, which for 15-kHz systems results in 24 dB because the deviation is halved [38, 39].

For the unprotected or transparent digital channel, very often a maximum BER of 1% can be tolerated for good transmission, which means that 98% of the time $(p_C - p_I) >> (p_C - p_I)_{DS}$ must be fulfilled, where the latter is the necessary static protection ratio. For the remaining time, we have $(p_C - p_I) < (p_C - p_I)_{DS}$, resulting in a BER of ≤50%. Hence for 2% of the time nearly all the errors occur giving an overall BER of ≈1%. Therefore, $k \cdot \sigma_I \approx 2.0 \cdot \sigma_I$ can be estimated:

$$(p_C - p_I)_{DD} \geq (p_C - p_I)_{DS} + 2.0 \cdot \sigma_I \approx (p_C - p_I)_{DS} + 18 \text{ dB} \tag{3.90}$$

In practice, $(p_C - p_I)_{DD}$ depends heavily on the required BER, the type of modulation, and the coding scheme used and may range from 10 to 30 dB. Hence 20 dB is a good first estimation for lightly error-protected channels.

For GSM and TETRA $(p_C - p_I)_{DD}$ is 9 or 19 dB, respectively, but under fading conditions only, which means $\sigma_I \approx 6$ dB. If shadowing is also taken into account, about 3 dB has to be added and the result is roughly 12 and 22 dB. Because very often $(p_C - p_I)$ is only known under fading conditions, this addition of 3 dB might generally serve for the first rough system comparisons.

Digital symbols are more robust than analog ones and the necessary bit energy-to-noise power density ratio E_B/N_0 is typically much smaller for digital radio channels than the necessary carrier-to-noise power ratio P_C/P_N for sufficiently good analog transmission. For the unprotected digital radio channel, the resulting figures may therefore not differ much from the analog case. However, the introduction of *forward error correction* (FEC), interleaving, and proper coding can improve the robustness of the digital radio channel dramatically. In heavily protected digital systems, a resulting BER of up to 20% on the gross transmission rate might be tolerable and it may therefore then be

$$(p_C - p_I)_{DD} \approx (p_C - p_I)_{DS} \tag{3.91}$$

For $D >> R$ the interfering field strength for larger cluster sizes does not vary very much within a cell, whereas the more remote cochannel cells add only negligible contributions to the total interfering power at the cell border [11, 35, 38, 42]. In the case of omnidirectional antennas, the nearest six interfering cells contribute nearly equally to the resulting $(P_C/P_I)_D$ ratio. If, however, the different distances are really taken into account, the total interfering power P_I for a mobile at the cell boundary can be calculated according to (3.24):

$$P_I = \sum_{k=1}^{6} (D/d_k)^{\alpha} \cdot P_{Ik} \tag{3.92}$$

The precise evaluation is difficult without providing much more accuracy than the following good approximation for the total interfering power, which assumes equal P_{Ik}:

$$P_I = \sum_{k=1}^{6} (D/d_k)^{\alpha} \cdot P_{Ik} \approx \left\{ \left(\frac{D}{D \pm R} \right)^{\alpha} + 2 \cdot \left(\frac{D}{D \pm R/2} \right)^{\alpha} \right\} \cdot P_{Ik} \tag{3.93}$$

With the assumption $d_k \approx D >> R$ we get $P_I \approx 6 \cdot P_{Ik}$ as it is mostly used in the literature. However, $\alpha = 4$ and a cluster size of $N_c = 9$ with $D/R = (3N_C)^{1/2} = (27)^{1/2} = 5.2$ [see (3.84)] results in $P_I = 7.2 \cdot P_k$. For large cluster sizes, the difference becomes negligible but for smaller ones it might become even larger:

$$P_I = \sum_{k=1}^{6} P_{Ik}(d_k) \xrightarrow[d_k \approx D >> R]{} 6 \cdot P_{Ik}(D) \tag{3.94}$$

From (3.86), it immediately follows that

$$(P_C/P_I)_D \approx \frac{1}{6} \cdot (D/R)^{\alpha} \tag{3.95}$$

Because the interfering six cells are normally only partly loaded an additional load factor N_L has to be introduced [10]:

$$0 \le N_L \le 1 \tag{3.96}$$

In congested areas $N_L \approx 0.3$ may be used for nontrunked systems, whereas an estimate of $N_L \approx 0.7$ might be more appropriate for very heavily loaded trunked systems with a large number of available traffic channels and short transmission times. For first estimates for a well-loaded trunked system $N_L = 0.5$ may be assumed. N_{LI} is the load factor of the interfering cells. If these belong to the same system, then $N_{LI} = N_L$ can be assumed and from (3.95) we get

$$(P_C/P_I)_D \approx \frac{1}{6N_{LI}} \cdot (D/R)^{\alpha} \tag{3.97}$$

With (3.84) the dependence of the necessary cluster size on the permissible $(P_C/P_I)_D$ ratio can now be derived:

$$N_C = \frac{1}{3} \cdot [6N_{LI} \cdot (P_C/P_I)_D]^{2/\alpha} \tag{3.98}$$

Equations (3.82) and (3.82) for the frequency economy and the available channels for interference-limited systems, respectively, can be rewritten with this relation:

$$\eta_I = \frac{3N_A \cdot N_M \cdot R_{BU}}{\Delta F_{\text{Ch}} \cdot [6N_{LI} \cdot (P_C/P_I)_D]^{2/\alpha}} \tag{3.99}$$

$$N_I = \frac{3N_A \cdot N_M \cdot B_{\text{Syst}}}{\Delta F_{\text{Ch}} \cdot [6N_{LI} \cdot (P_C/P_I)_D]^{2/\alpha}} \tag{3.100}$$

The latter is the *radio capacity formula* as defined by Lee giving the system capacity as the number of maximum available channels in RTC/cell [11, 35, 38]. However, here we have also taken into account the fact that the interfering cells are not always fully loaded and that often $\alpha \neq 4$. For system comparisons B_{Syst} should be set to 1 MHz to get the radio capacity in RTC per megahertz or RTC per cell per megahertz.

Table 3.11 gives the cluster size and the radio capacity for a number of standardized and proprietary mobile radio communication systems. For all calculations we have assumed that all systems are interference limited with $\alpha = 3.5$ and $B_{\text{Syst}} =$ 1 MHz under fading *and* shadowing conditions. The values for cochannel rejection are taken from Chapters 6 to 9. In the table the calculated cluster size N_C has been replaced by the nearest possible integer value. However, the radio capacity has been calculated without rounding to get a more precise measure.

From the results in Table 3.11 some interesting conclusions can be drawn, for example, note that the channel splitting from 25 to 12.5 kHz by no means resulted in doubling the capacity because the cochannel rejection got significantly worse. Moreover, the vulnerability against ignition noise and other interference has become higher, and the AF bandwidth for analog transmission had to be reduced from 3.0 to 2.55 kHz, all of which resulted in a considerably worse speech quality.

The small radio capacity of DSRR does not matter very much because this system was intended to be used mainly for mobiles and without infrastructure.

Table 3.11 Frequency Economy of Selected Mobile Radio Systems

System	N_A	ΔF_{Ch} *[kHz]*	$(p_C - p_I)_F$[1] + 4 *[dB]*	$(P_C/P_I)_F$[2]	N_C	N_I *[RTC/MHz/cell]*
DSRR	1	25	31	1258.9	$36.89 \rightarrow 37$	1.08
GSM FR	8	200	13	20.0	$3.46 \rightarrow 4$	1.93[3]
GSM HR	16	200	13	20.0	$3.46 \rightarrow 4$	3.85[3]
PM 25	1	25	20	100.0	$8.68 \rightarrow 9$	4.61
PM 12, PMR 446	1	12.5	25	316.2	$16.75 \rightarrow 16$	4.78
PM 20	1	20	21	125.9	$9.90 \rightarrow 9$	5.05
EDACS	1	25	18	63.1	$6.67 \rightarrow 7$	6.00
SR 440	1	25	17	50.1	$5.85 \rightarrow 7$	6.84
DIIS	1	12.5	22[4]	158.5	$11.29 \rightarrow 12$	7.09
EN 301 166	1	5	28[5]	631.0	$24.86 \rightarrow 25$	8.05
APCO 25	1	12.5	20.5	112.2	$9.27 \rightarrow 9$	8.63
EDACS	1	12.5	20	100.0	$8.68 \rightarrow 9$	9.22
SR 440	1	12.5	19	79.4	$7.61 \rightarrow 7$	10.51
TETRAPOL	1	12.5	19	79.4	$7.61 \rightarrow 7$	10.51
TETRA V+D	4	25	23[6]	199.5	$12.88 \rightarrow 13$	12.42
iDEN	6	25	23	199.5	$12.88 \rightarrow 13$	18.63
RVE (TTIB)	1	5	21[7]	125.9	$9.90 \rightarrow 9$	20.20

[1] Inverse cochannel rejection with fading plus 4 dB to account for shadowing.
[2] Linear power ratio.
[3] GSM results corrected by 1/6 to account for duplex and 600 instead of a 200-kHz carrier separation for comparable adjacent channel power.
[4] Typical.
[5] For 7.2 kbit/s.
[6] Typically 1 dB better.
[7] For speech and transparent analog channel.

Due to the resulting small coverage area we did not have to worry that a congestion of frequencies would occur. For the same reason it does not matter that PMR 446 also has a weak radio capacity similar to conventional PM12 systems.

For GSM three times the carrier separation should be taken to meet the stringent adjacent channel power limits of PMR systems. Moreover, GSM is a full-duplex system, which means an additional factor of 2, disregarding the fact that all radio links between two mobiles are running via base stations, which would mean another factor of 2. Hence, to compare the radio capacity of GSM with PMR systems operating in simplex mode, at least a factor of 1/6 has to be taken additionally into account to get the results shown in Table 3.11.

Finally the most interesting modern digital PMR systems EDACS, iDEN, TETRA, and TETRAPOL provide roughly two to three times the radio capacity of conventional 12.5-kHz PMR systems. Narrowband SSB systems exhibit a weak cochannel rejection but due to their extremely small carrier separation a very good radio capacity results nevertheless. Despite this benefit, narrowband systems have not been able to conquer big market shares yet because they require a very complex and expensive radio technology.

From the radio capacity formula a simple condition for constant frequency efficiency or radio capacity respectively can be derived:

$$\eta_I \sim N_I \sim \frac{1}{\Delta F_{\mathrm{Ch}} \cdot (P_C/P_I)_D^{2/\alpha}} \xrightarrow[\alpha=4]{} \frac{1}{\Delta F_{\mathrm{Ch}} \cdot \sqrt{(P_C/P_I)_D}} \tag{3.101}$$

A reduction of the modulation bandwidth B_M and hence the nearly equal channel separation $\Delta F_{\mathrm{Ch}} \approx B_M$ normally leads to an increased minimum permissible P_C/P_I ratio and therefore requires an increased cluster size N_C. Hence, an improved frequency economy always requires a reduction of the product $B_M \cdot (P_C/P_I)_d^{2/\alpha}$ and not only of the modulation bandwidth.

If only a rough estimation for the radio capacity of a system is needed, then the approximations $\alpha = 4$ and $N_{LI} = 1$ leading to Lee's original formula and giving the best case result in RTC per cell can be used [11, 35, 38].

$$N_I = \frac{B_{\mathrm{Syst}}}{\Delta F_K \cdot \sqrt{\frac{2}{3} \cdot \left(\frac{P_C}{P_I}\right)_D}} \tag{3.102}$$

If the frequency economy for two systems is equal, then additional measures should be taken into account. For example, digital radio transmission systems with longer symbol duration or lower P_C/P_I protection ratios are less vulnerable to multipath distortion and need less or sometimes even no equalization. Data compression and improved voice coding techniques reduce the necessary net transmission rate and will therefore allow significant improvements to the frequency economy of digital radio transmission systems. Additionally *voice activity detection* (VAD), *discontinuous transmission* (DTX), and *discontinuous reception* (DRX) might be used. Because power control decreases the overall interference, the cluster size can be reduced and trunking allows an increase in the channel load. Besides all of these means, the mode of operation, including group call features and so on, will also have considerable influence on the frequency economy. Speech coder properties and bit rates also have a great amount of influence on the frequency economy. If systems with speech coders of similar speech quality but different bit rates are compared, the stated improvement may depend solely on the better speech coder.

3.9 PMR-Related Frequency Economy Considerations

In PMR a considerable percentage of the traffic employs point-to-multipoint links, that is, group calls and broadcast messages. Addressing more than one party simultaneously means that the frequency economy is multiplied by the number of called parties. This might be described by a group call factor N_G that depends on the group size n_G and the ratio k_G of the total group call traffic A_G to the total traffic A_T [42]:

$$k_G = A_G/A_T < 1 \tag{3.103}$$

The group call factor therefore becomes

$$N_G = n_G \cdot \frac{A_G}{A_T} + \frac{A_T - A_G}{A_T} = 1 + k_G \cdot (n_G - 1) > 1 \tag{3.104}$$

Hence, for big groups even a limited number of group calls considerably improves the efficient use of frequencies. Spectral efficiency, frequency economy, and radio capacity for noise- and interference-limited systems with group calls can now be recalculated by multiplication with N_G.

Besides the already discussed general system limitations, we need to consider others. In contrast to the limitations mentioned above, which are based on physical constraints, the following limitations exhibit difficulties that can be overcome by investing in some increased technical effort. Some of these limiting factors have a greater effect on, in particular, simulcast systems and therefore greater care must be taken in such cases.

Delay-limited systems exhibit a bad ratio of burst to guard time, which is a problem associated with TDMA and CDMA but not with FDMA. For large coverage areas and their associated propagation delays, the duration of guard and burst ramping times must be shortened if the burst time cannot be made longer. As with GSM the guard time can be considerably reduced if timing advance methods are introduced in which the mobile transmits its bursts with a variable timing advance compared to the received base station frame to compensate for varying signal propagation times. However, the guard and ramping times together cannot be made shorter than the delay spread as determined by the propagation conditions.

Dispersion limitations occur where intersymbol interference is introduced by multipath propagation. This appears when the delay spread exceeds a considerable percentage of the symbol duration [see (3.68)]. However, as GSM demonstrates, this limitation can be overcome by equalizing methods in which each burst contains a known training sequence from which the channel propagation conditions can be calculated and then used to restore the distorted message symbols. The necessary effort is not negligible and the use of training sequences reduces the spectral efficiency and frequency economy. However, at the same time the permissible P_C/P_I ratio will be improved by equalization, which in turn increases the spectral efficiency and frequency economy. Hence, the net result must be carefully examined.

Depending on the type of modulation and bandwidth the *Doppler spread* may also limit system performance if it is not negligible compared to the modulation bandwidth or if it causes synchronization problems. Again suitable equalizing causing additional effort is needed.

3.10 System Design and a Short Introduction to Traffic Theory

In many cases multicell systems with high traffic have to be planned and additional traffic considerations become necessary. Therefore, the number of users and their behavior in terms of generated traffic and its statistical properties must be known and the required grade of service has to be defined. To evaluate the number of channels needed, traffic theory has to be applied. Different traffic models are used for terrestrial or mobile telephone networks and in PMR. If typical PMR operation as well as point-to-point telephone traffic is offered, then the mixture has to be known or estimated. In PMR systems the ratio of typical PMR traffic to telephony traffic often is about 90%:10%. In addition data traffic also has to be taken into account.

The traffic generated during the busy hour represents the peak load of a system and determines the necessary capacity. Hence, only the traffic in the busy hour needs to be discussed further. For PMR traffic, call duration on the order of 10–30 seconds is typical, and up to several calls can be made generating a total voice call traffic of up to 25 mErl per user. For the most part, the required short message capacity is negligible in comparison with speech because there are only a few messages/user and usually their length is restricted to some dozens of bytes. However, this does not apply to general data transmission, and the needed data transmission capacity depends strongly on the application so that no general sound estimate can be given. The telephone traffic is often to be found on the order of one call per user but mostly with considerable longer duration on the order of up to 180 seconds, thus creating up to 50 mErl per user.

A recent investigation in Germany revealed a typical group call duration of 90 seconds for police and fire brigades, while police dealing with criminal activities mainly used mobile phones in public networks with a typical call duration of 150 seconds. In CHEKKER networks the average call duration was 40–45 seconds and mainly individual calls were made, whereas in the fixed networks the typical call duration is about 180 seconds. In this example the percentage of voice transmission was exceeding 98% and data transmission was less than 2%, but this ratio is sure to change in the future because more useful data applications will evolve.

The group call behavior also depends heavily on the application and therefore only rough figures can be given. For example, as many as two-thirds of the calls might be group calls and only one-third individual calls. In large multicell systems, for about 50% of the calls the user has to be called in one or two cells, for 30% in up to five cells, and for 20% in up to 10 cells; actual system-wide calls seldom occur.

To arrive at the proper system layout, application-related traffic data are needed and these can only be achieved by means of thorough investigations of user behavior. Moreover, it is difficult to make sound assumptions if user behavior changes due to the introduction of new system features or if the application environment varies. Thus the system planner must always be aware that user behavior may change and he must be prepared to modify or enlarge the system later. If the number of users in a system is known and if reasonable assumptions about their behavior have been made, the total traffic in the peak traffic hour and the number of necessary channels can be determined [5, 20, 24, 25, 47–50].

The users are the traffic sources. In a radio system this is identical to the number of mobiles M. These generate a certain call arrival rate λ_C with an average time h_A between call arrivals and a certain average call duration T_A. These form the offered traffic A to be served by a given number N of channels:

$$A = \lambda_C \cdot T_A \tag{3.105}$$

The channels could be radio channels or telephone channels or both in a complex mobile communication system. The different channels are occupied by the traffic where the channel occupancy obeys a statistical distribution resulting in an average occupancy ρ_C.

$$\rho_C = A/N \tag{3.106}$$

The two basic types of systems to consider are *loss systems* and *queuing systems*. Figure 3.18 shows an example of the occupancy of the individual channels and also the sum of the offered traffic of all channels. If the number of users making a call attempt is smaller than the number of available channels, there is no problem serving them. If, however, the number of users temporarily exceeds that of the available channels, then the call attempts can be put in a waiting queue of length Q. In a loss system the calls are simply not served ($Q = 0$) and are discarded while in an ideal queuing systems ($Q = \infty$) they are held in the queue until they can be served. However, in practical systems the queue cannot be of unlimited length and, therefore, all real queuing systems are a mixture of queuing and loss systems [50].

To calculate the number of necessary channels based on a certain intended *grade of service* (GoS), specific statistical tools have been developed. For a loss system the offered traffic A exceeds the carried traffic Y by the loss characterized by B, which depends on the RF outage probability p_S due to shadowing and so on and the channel blocking probability p_B [24]:

$$Y = A \cdot (1 - B) = A \cdot (1 - p_B) \cdot (1 - p_S) \tag{3.107}$$

The RF outage probability is a design parameter and it may be 1%, 2%, or 5% for example. If the protocol on the RF path repeats call attempts that are lost because of an RF outage, then $p_S = 0$ can be set. The channel blocking probability can be calculated by the Erlang B formula (also known as Erlang's loss formula) from the offered traffic and the total number of channels in all the radio cells that are accessible to a certain mobile subscriber:

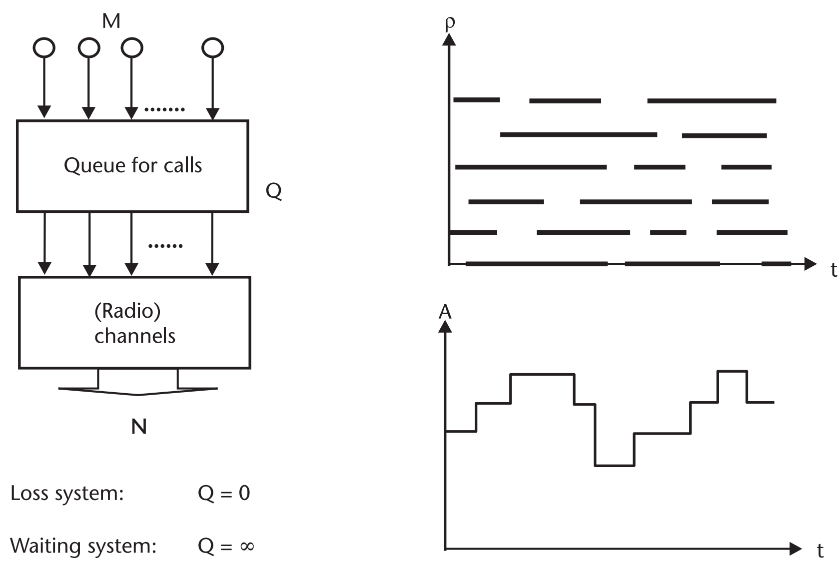

Figure 3.18 System configurations and traffic.

$$p_B = \frac{A^N / N!}{\sum_{k=0}^{N} \frac{A^k}{k!}} \tag{3.108}$$

Unfortunately, most often the GoS in terms of blocking probability is given and the number of necessary channels has to be determined from the GoS. However, it is not possible to reorder the Erlang B formula to calculate the number of channels directly and therefore look-up tables have to be used. For every combination of a number of channels N and any acceptable blocking probability p_B, the carried traffic Y can be determined in Erlangs (1 Erl = 1,000 mErl) as Table 3.12 shows. For example, 12 Erl can be carried on a 20-channel system with 1% blocking probability, but if 18 Erl were attempted, then the probability would exceed 10%. Of course, the convergence criterion $A < N$ has to be fulfilled in that the offered traffic must be smaller than the number of channels. Otherwise the blocking probability p_B approaches 1 or 100% and the system becomes totally blocked.

The normalized traffic or trunking gain Y_L shows that the carried traffic increases faster than the increase in the number of channels. This means that with more channels a higher average load per channel can be achieved:

$$Y_L = Y/N \tag{3.109}$$

The graph on the left side of Figure 3.19 shows that the carried traffic Y increases at a faster rate than the increase in the number of channels. If a very low blocking probability is required, then the carried traffic is much smaller than would be the case for a higher acceptable blocking probability. The second diagram illustrates the trunking gain Y_L, which shows how the carried traffic per channel increases with the number of channels. If a higher blocking probability is acceptable, then a higher trunking gain can be achieved.

Table 3.12 Offered Traffic and Blocking Probability Using the Erlang B Formula

Number of Channels	*Blocking Probability p_B*				
N	*1%*	*2%*	*5%*	*10%*	*20%*
1	0.01	0.02	0.05	0.11	0.25
2	0.15	0.22	0.38	0.60	1.00
3	0.46	0.60	0.90	1.27	1.93
4	0.87	1.09	1.52	2.05	2.95
5	1.36	1.66	2.22	2.88	4.01
6	1.91	2.28	2.96	3.76	5.11
8	3.13	3.63	4.54	5.60	7.37
10	4.46	5.08	6.22	7.51	9.68
15	8.11	9.01	10.63	12.48	15.61
20	12.03	13.18	15.25	17.62	21.63
25	16.12	17.51	19.99	22.84	27.72
30	20.34	21.93	24.80	28.11	33.85
35	24.64	26.44	29.68	33.43	39.98
40	29.01	31.00	34.60	38.78	46.14
45	33.43	35.61	39.55	44.17	52.32
50	37.90	40.26	44.53	49.56	58.50

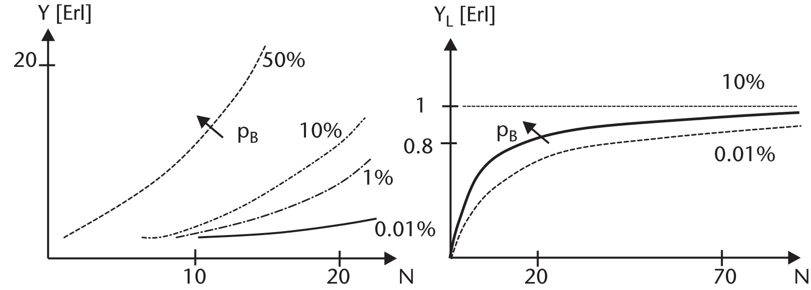

Figure 3.19 Carried traffic Y and trunking gain Y_L in a loss system.

The grade of service is defined as the quality of the system in terms of lost calls [24]:

$$\mathrm{GoS}_B = [1 - (1 - p_B) \cdot (1 - p_S)] \tag{3.110}$$

This formula can be simplified if the RF outage and channel blocking probability are much smaller than 1:

$$\mathrm{GoS}_B \xrightarrow[p_S,\, p_B \ll 1]{} p_B + p_S \tag{3.111}$$

In an ideal queuing system the carried traffic Y is equal to the offered traffic A, which must be smaller than the number of channels N; otherwise, the waiting lime T_W will become unlimited and the system will be blocked. The waiting probability p_W in this case can be calculated by the Erlang C formula, which can be derived from Erlang's loss formula [47]:

$$p_W = \frac{p_B}{1 - \frac{A}{N} \cdot (1 - p_B)} = \frac{A^N}{A^N + N! \cdot \left(1 - \frac{A}{N}\right) \cdot \sum_{k=0}^{N-1} \frac{A^k}{k!}} \tag{3.112}$$

The Erlang C formula allows the calculation of the carried traffic for any waiting probability p_W and any number of channels N. Table 3.13 gives the carried traffic $Y = A$ in Erlangs as a function of the number of channels N.

If the desired value is not to be found in the Erlang tables, it can be determined by interpolation or the Erlang formulas can be evaluated with a mathematics program on a computer. Comprehensive Erlang B and Erlang C tables are also available in the literature, but sometimes a source contains only one of the two [17, 20, 24, 25, 47]. However, reordering of (3.112) allows us to determine p_B from p_W:

$$p_B = \frac{N \cdot p_W - A}{N - A \cdot p_W} \tag{3.113}$$

Table 3.13 Offered Traffic and Waiting Probability Using the Erlang C Formula

Number of Channels N	*Waiting Probability p_W* 1%	2%	5%	10%	20%
1	0.01	0.02	0.05	0.10	0.20
2	0.15	0.21	0.34	0.50	0.74
3	0.43	0.55	0.79	1.04	1.39
4	0.81	0.99	1.32	1.65	2.10
5	1.07	1.50	1.91	2.31	2.85
6	1.76	2.05	2.53	3.01	3.62
8	2.87	3.25	3.87	4.46	5.21
10	4.08	4.54	5.29	5.99	6.85
15	7.39	8.04	9.04	9.97	11.09
20	10.97	11.77	13.00	14.11	15.46
25	14.72	15.64	17.08	18.36	19.89
30	18.59	19.64	21.25	22.69	24.39
35	22.55	23.71	25.48	27.06	28.92
40	26.58	27.84	29.77	31.48	33.48
45	30.67	32.03	34.10	35.93	38.07
50	34.80	36.26	38.47	40.41	42.69

In an ideal queuing system the grade of service GoS_W is equal to the waiting probability p_W:

$$GoS_W = p_W \quad \text{with} \quad Y = A \tag{3.114}$$

The waiting time is of a statistical nature but an average time T_{WA} can be determined [49]:

$$T_{WA} = \frac{T_A}{N - A} \cdot p_W \tag{3.115}$$

What is of interest is to know the probability that the waiting time will exceed a certain given time t [20]:

$$p_W(t) = e^{-t(N-A)/T_A} \tag{3.116}$$

In a queuing system the carried traffic also increases faster than the increase in the number of channels. If a higher waiting probability can be accepted then a higher traffic load can be carried. Conversely, fewer channels are needed to carry a given traffic load if a higher waiting probability can be tolerated. Figure 3.20 shows that the normalized waiting time rises steeply if the traffic approaches the available number of channels. This effect is smaller for systems employing a greater number of channels.

What is an acceptable grade of service? For PMR waiting systems often, a GoS or *quality of service* (QoS) of $p_W = 2\%$ and an average waiting time of $T_{WA} = 5$ seconds are judged to be acceptable.

Modern digital PMR systems like TETRA attempt to avoid the loss of calls and therefore the Erlang C formula is applicable. Planning considerations depend

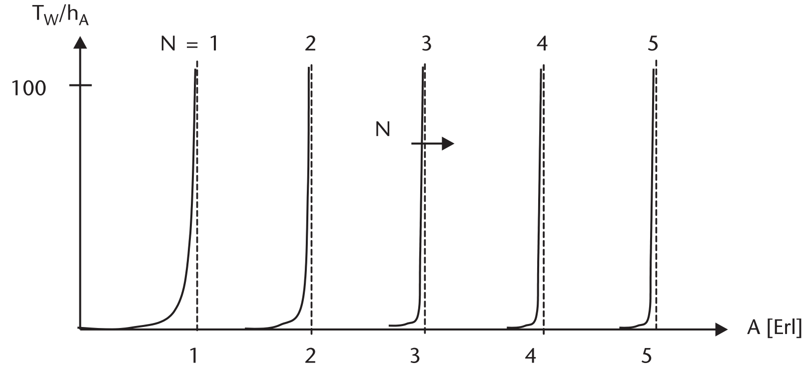

Figure 3.20 Average waiting time in a queuing system.

on the traffic and not on the mode of transmission. Hence, traffic optimization for digital systems can be made in the same way as for analog systems.

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CHAPTER 4

Review of the Basics of Analog PMR Technology

Most people believe that analog and digital radio technology are completely different. However, this is not the case. In this and the next chapter, we will show that analog and digital transmission have many commonalities. Moreover, very often the same signal processing function can be performed by analog or digital means, and in many cases, when choosing one solution or the other, it comes down to a question of finding the best trade-off between physical and economic constraints and benefits. An understanding of the fundamentals of conventional professional radio communication systems will make it easier to understand the modern digital PMR equipment and systems. Hence, a short review of classical radio technology will help the reader to become more familiar with the functioning of modern digital PMR systems.

4.1 Classical PMR Frequency Bands and Their Use and Properties

There is no single PMR frequency band but a number of them due to the historical development of radio communication technology and frequency management. Below 1 GHz only 20% of all frequencies are allocated to land mobile radio services, whereas 30% are devoted to military applications, 40% to broadcast and television, and 10% to other services. The PMR frequency allocations are the smaller part of all land mobile services and they are not only very fragmented but also different in the various countries. PMR penetration varies widely. It is 1–2% in most of the European countries, but about 8% in the United States [1–4].

Remarkably PMR, paging, and cellular services not only compete but they also complement each other. That is, usually, they either do well together or they starve simultaneously. The reasons are, to a great extent, of a regulatory nature. It is no surprise that PMR penetration is low in countries with high fees and long license application times, but there GSM penetration is also relatively low [4, 5].

Besides the already-mentioned regulatory hurdles, there is another reason for the differing PMR penetration in the various countries and that is simply a lack of frequencies. In the different European countries, from less than 20 up to nearly 100 MHz are allocated to PMR, and about 150 MHz is allocated in the United States. In Europe, Germany is the country with the highest PMR mobile station density. Around 1995–1996, approximately 1.2 million mobiles were in use, of which 925,000 were in the 2-m band, which represents about one-third of all

German PMR allocations. The average density of 3.35 mobiles/km^2 is up to eight times higher than anywhere else in Europe. The United States is another example. In the United States PMR frequency allocations are split into a number of very different bands, but nearly half of the PMR users are in the UHF range [1, 2, 4].

A lack of capacity exists for civil radio services below 1 GHz. A large amount of that spectrum is allocated to military services so in times of peace it is only used to a small extent. Of course, during times of war military services will need much more capacity, and the use of civil applications will necessarily be significantly reduced. Hence, the shared use of frequency bands for civil and military use could provide additional capacity for civil radio services including PMR. The rules to be respected for band sharing between military and civil radio services are listed in CEPT Report 51: time sharing, use of different parts of a common band, geographical separation, and simultaneous use at the same location and time on a noninterference basis [6]. One example for the implementation of these principles is the sharing of parts of the NATO band between 380 and 400 MHz in Europe, where TETRA can be used for public safety services. Additionally the envisaged release of TV frequencies between 470 and 520 MHz would provide further spectrum for PMR.

Before the introduction of a (new) PMR service, we need to take the differing properties of the envisaged PMR bands into account to avoid a mismatch between the requirements of the service and the properties of the frequency band. It is simply not true that every frequency is equally useful for PMR purposes, as has been shown in Chapter 3. To stress this further, the existing European PMR bands and their properties are summarized as follows:

- *27.4–41 MHz:* The 7-m band offers large coverage areas but also frequent overshoot interference. Today the PMR use of this band is rapidly declining.
- *47–68 MHz:* Television will be or has already been removed from this band. However, the situation is different from country to country and removal from TV usage will require much time in some cases. The enhancement of PMR services in this band has been discussed but most manufacturers are not very interested.
- *68–87.5 MHz:* Until now the 4-m band has been heavily used for different PMR services, including public safety and railways. This band exhibits large coverage distances but also overshoot interference, and both effects are considerably smaller than in the lower bands.
- *146–174 MHz:* In many countries the 2-m band is the classical PMR band for civil applications and public safety. Only a few other services representing a small total bandwidth are also allocated there, for example, some maritime and satellite services. For a fast and orderly transition to digital PMR, a consistent refarming strategy for the whole band is urgently needed but is not yet available in Europe. Therefore, analog systems may stay in this band for some time, serving users that mainly need speech communication.
- *174–216 MHz:* In the United Kingdom, the lower part of TV Band III was opened for PMR a long time ago. Currently, however, this is not the case for many other countries. In total, increasing PMR utilization of this band can be expected in the long run. In the adjacent range from 216–230 MHz, *Digital Audio Broadcasting* (DAB) has been established in several countries.

- *380–400 MHz:* In many European countries this so-called "NATO band" has been opened partly for public safety according to the Schengen Treaty. However, CEPT and NATO have ruled that the precondition is the introduction of one common European standard for this purpose. In many European countries, therefore, the implementation of TETRA is planned, but the problem is that other countries have opted for TETRAPOL [7].
- *410–430 MHz:* In many countries analog PAMR networks based on MPT 1327 in 12.5-kHz channels have been established in this band. Because these networks are now on the decline, the introduction of TETRA is envisaged in parts of this band. In some countries the whole band can be used for PMR down to 406 MHz [8].
- *440–470 MHz:* Besides PMR, analog cellular networks are also in operation, for example, NMT-450 in many European countries and Net C in Germany and Portugal. In the meantime, some of them have been shut down; for example, Net C was shut down at the end of 2000. At 445 MHz, some frequencies for direct mode have recently been assigned Europe-wide for DPMR by CEPT ERC Decision (01) 21, and the same has happened at 446 MHz with eight frequencies for PMR 446 by CEPT ERC Decision (98) 25 [9, 10].
- *870–876 MHz:* This band is available for PMR in many European countries but the introduction of TETRA there currently is supported only by very few administrations and manufacturers. Thus, the predicted market figures are small. However, in many countries worldwide the 900-MHz band is the only available one for new DPMR services and this fact will push TETRA development for 900 MHz. Nevertheless, the limited coverage distance is a major drawback for many applications.
- *876–880 MHz:* Throughout Europe this band has been allocated to GSM-R, which is a PMR service for railways based on the GSM technology.
- *915–921 MHz:* Via the duplex separation of 45 MHz, this band is paired with the 870- to 876-MHz band.
- *921–925 MHz:* Via the duplex separation of 45 MHz, this band is paired with the 876- to 880-MHz band.

The widths of the PMR bands and their precise allocations and borders are different in the various countries, but this does not matter very much because the properties are only changing slowly with the frequency. In other parts of the world, PMR allocations have been made in similar bands.

4.2 Analog Modulation Techniques

Because digital modulation uses the same basic mechanisms as analog modulation, it is worthwhile to first review analog modulation techniques [11–14]. There is another reason for looking back to the fundamentals: In the radio channel digital transmission no longer exhibits discrete levels but looks very similar to analog signals.

Because every baseband signal can be represented as a sum of signals with different frequencies, amplitudes, and phases, it is sufficient to replace a more complicated baseband signal by a single frequency signal $s_B(t)$ where f_B is the baseband frequency, S_B is the baseband signal amplitude (in volts), and $\omega = 2\pi f_B$ is the baseband frequency expressed in radians:

$$s_B(t) = S_B \cdot \cos(2\pi f_B t) = S_B \cdot \cos(\omega t) \tag{4.1}$$

This baseband signal should be modulated onto a carrier frequency f_C well suited to the transmission in one of the radio bands, where S_C is the unmodulated RF signal amplitude and $\Omega = 2\pi f_C$ is the carrier frequency in radians:

$$s_C(t) = S_C \cdot \cos(2\pi f_C t) = S_C \cdot \cos(\Omega t) \tag{4.2}$$

All of the basic relations of interest can be derived from this simple model case of one sinusoidal high-carrier frequency modulated by one low-signal frequency. There are different possibilities: The message can modulate either the amplitude or the frequency or phase of the carrier signal. In the case of *amplitude modulation* (AM), the baseband frequency modulates the carrier amplitude without affecting the frequency or the phase of the carrier:

$$s(t)_{\mathrm{AM}} = S_C \cdot [1 + m \cdot \cos(\omega t)] \cdot \cos(\Omega t) \quad \text{with} \quad 0 \leq m = \frac{S_B}{S_C} \leq 1 \tag{4.3}$$

The time-dependent grade of modulation $m(t) = (S_B/S_C) \cdot \cos(\omega t)$ carries the entire message content. The result of the modulation process, which ideally can be obtained by a multiplication of signal and carrier frequency, is the carrier frequency plus two sidebands:

$$s(t)_{\mathrm{AM}} = S_C \cdot \sum_{p=-1}^{+1} \left(\frac{m}{2}\right)^{p^2} \cdot \cos[(\Omega + p\omega)t] \tag{4.4}$$

This somewhat odd-looking format has been chosen to better show the similarities and differences between amplitude and angle modulation. The two sideband frequencies are $\pm f_B$ apart from the unmodulated carrier frequency f_C, and their amplitude depends on the grade of modulation m. Hence, the AM bandwidth is twice the maximum baseband frequency:

$$B_{\mathrm{AM}} = 2f_B = \frac{\omega}{\pi} \tag{4.5}$$

This is only true if the transmitter power amplifier is highly linear; otherwise, sideband frequencies of higher order multiples of the signal frequency will occur that would spoil the spectrum outside the intended modulation bandwidth. If the baseband comprises a mixture of different frequencies, then two complete sidebands will be created by the modulation process: One sideband is in the normal position, whereas the other one is inverted due to the minus sign. Each of the two sidebands contains the complete message and this is why *single sideband* (SSB) modulation using only one of the two sidebands was invented. Thus, the bandwidth of SSB is only half of that of ordinary AM:

$$B_{\text{SSB}} = f_B = \frac{\omega}{2\pi} \tag{4.6}$$

In AM and SSB it is possible to suppress the carrier frequency after the modulation process to avoid transmitting energy that does not contribute to the message content. This is mainly done in highly specialized long-range systems for specific purposes. During at least the last two decades, SSB has also been discussed for PMR applications without achieving a real breakthrough.

In the case of frequency and phase modulation (FM and PM) the angle of the carrier is modulated by the baseband signal instead of the amplitude. There are two different possibilities: either the frequency or the phase can be modulated. Because both cases are very similar the mathematical representation is also similar:

$$s(t)_{\text{FM}} = S_C \cdot \cos\{[\Omega + \Delta\Omega \cdot \sin(\omega t)]t\} \tag{4.7}$$

$$s(t)_{\text{PM}} = S_C \cdot \cos[\Omega t + \Delta\varphi_C \cdot \cos(\omega t)] \tag{4.8}$$

with

$$\Delta\Omega(t) = \frac{d}{dt} \cdot [\Delta\varphi_C \cdot \sin(\omega t)] = \omega \cdot \Delta\varphi_C \cdot \cos(\omega t) \tag{4.9}$$

For FM the frequency deviation $\Delta\Omega \cdot \sin(\omega t)$ is time dependent, whereas in the case of PM it is the phase deviation $\Delta\varphi_C \cdot \cos(\omega t)$. The carrier peak frequency deviation is $\Delta F_C = \Delta\Omega/2\pi$ and the carrier peak phase deviation is $\Delta\varphi_C$. The latter is also called the modulation index. A simple general relation exists between phase and frequency that for sinusoidal signals links the peak frequency and the peak phase deviation or modulation index:

$$\frac{d\varphi}{dt} = 2\pi f \tag{4.10}$$

$$\Delta\Omega = \omega \cdot \Delta\varphi_C \quad \text{or} \quad \Delta F_C = f_B \cdot \Delta\varphi_C \tag{4.11}$$

Note that in daily life people do not always distinguish clearly between FM and PM. In PMR, PM is usually meant even if FM is mentioned. However, there are also PMR systems employing true FM.

The mathematical representation of AM, FM, and PM looks a little bit difficult at first glance. The illustrations in Figure 4.1 may improve understanding. At the left, the baseband and the carrier signals are shown. The middle panel shows sketches of AM at the top and of FM and PM on the bottom. On the right side,

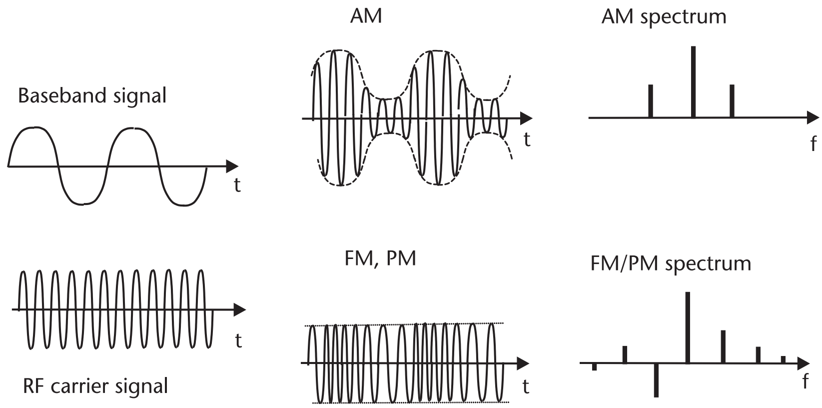

Figure 4.1 AM, FM, and PM in the time and frequency domains.

the modulation spectra are shown. AM has only two sideband frequencies, while the spectra of FM and PM are more complex due to the multiples of the sideband frequencies that occur. The picture would be less clear if the baseband contained several distinct frequencies and, therefore, only the simplest cases have been chosen for the illustration.

Calculating the modulation spectrum for FM or PM leads to a much more complex relation than in the case of AM. On both sides of the carrier frequency, all integer multiples of the baseband signal frequency will appear. However, their amplitudes decrease very quickly the higher the order of multiples, as will be shown later.

$$s(t)_{\mathrm{FM}} = S_C \cdot \sum_{p=-\infty}^{+\infty} J_p(\Delta\varphi_C) \cdot \cos[(\Omega + p\omega)t] \tag{4.12}$$

The amplitudes of the modulation products are determined by Bessel functions of the first kind and the pth order with integer p [15]:

$$J_p(\Delta\varphi_C) = \sum_{k=0}^{\infty} \frac{(-1)^k}{k! \cdot (p+k)!} \cdot \left(\frac{\Delta\varphi_C}{2}\right)^{2k+p} \quad \text{with} \quad -1 \le J_p(x) \le 1 \tag{4.13}$$

The calculation of the Bessel functions can be done with a scientific pocket calculator or more comfortably with a PC and a mathematics software package; however, the latter usually has the Bessel functions implemented so that they can be looked up directly. Because only the angle (phase or frequency) of the carrier is modulated in the ideal case, the amplitude is not affected. Thus, FM and PM modulation exhibits a constant envelope and it is obvious that the sum of all squared modulation products must be equal to the power of the unmodulated carrier (Parseval's theorem):

$$\sum_{p=-\infty}^{+\infty} J_p^2(\Delta\varphi_C) = 1 \tag{4.14}$$

For practical work it is sometimes useful to modify this formula slightly:

$$J_0^2(\Delta\varphi_C) + 2 \cdot \sum_{p=1}^{+\infty} J_p^2(\Delta\varphi_C) = 1 \quad \text{because} \quad J_p^2(x) = J_{-p}^2(x) \tag{4.15}$$

The modulation bandwidth can be calculated according to these formulas by determining how many of the sidebands are needed to obtain 99% of the transmitting power, that is, to determine that p_M for which 99% of the total power is reached or slightly exceeded:

$$J_0^2(\Delta\varphi_C) + 2 \cdot \sum_{p=1}^{p_M} J_p^2(\Delta\varphi_C) \geq 0.99 \tag{4.16}$$

Having evaluated this relation and having determined p_M, the modulation bandwidth can easily be calculated:

$$B_M = 2p_M \cdot f_B \tag{4.17}$$

Dependent on the phase deviation or modulation index, respectively, p_M will vary in practical cases from less than 1 to much more than 10, whereas in AM without nonlinear distortions the grade of modulation does not exceed 1. Hence, admittedly FM and PM exhibit a wider modulation bandwidth compared to AM, but both are much more robust against nonlinear distortions in the transmitter power amplifier, which therefore does not need to be very linear. Thus, FM and PM power amplifiers can be built with greater efficiency than AM or SSB power amplifiers. The reason for this robustness of FM and PM is that the modulation spectrum of both contains higher multiples of the modulating signal frequency, which hides possible small additional AM distortions, provided that the modulator does not cause too many other nonlinearities. It is interesting to determine the impact of the nonlinearity of the transmitter power amplifier. Let's assume that we cut away all negative half-waves from the carrier signal according to (4.2), then we get the following distorted carrier signal [16]:

$$s(t) = \frac{S_C}{\pi} + \frac{S_C}{2} \cdot \cos(\Omega t) + \frac{2S_C}{\pi} \cdot \sum_{q=1}^{\infty} \frac{(-1)^{(q+1)}}{4q^2 - 1} \cdot \cos(2q\Omega t) \tag{4.18}$$

If this carrier signal carries a PM, then we get, using (4.8) and (4.12),

$$\begin{aligned} s(t) = {} & \frac{S_C}{\pi} + \frac{S_C}{2} \cdot \sum_{p=-\infty}^{\infty} J_p(\Delta\varphi_C) \cdot \cos[(\Omega + p\omega)t] \\ & + \frac{2S_C}{\pi} \cdot \sum_{p=-\infty}^{\infty} \sum_{q=1}^{\infty} \frac{(-1)^{(q+1)}}{4q^2 - 1} \cdot J_p(\Delta\varphi_C) \cdot \cos[2q(\Omega + p\omega)t] \end{aligned} \tag{4.19}$$

The first term is a DC component that will be removed automatically because the RF circuitry is usually AC coupled. The second term represents ordinary PM on the carrier frequency, and all other terms are PM modulation products grouped around the harmonics of the carrier. All harmonics are removed by the harmonics filter because they have to be suppressed anyway. Hence, the linearity of the transmitter power amplifier really does not matter for FM and PM.

The Bessel functions determining the FM and PM spectrum are shown in Figure 4.2. If one is not familiar with such functions, they look somewhat strange. However, using sine or cosine functions for the first time results in a similar experience. At first glance, the Bessel functions look slightly similar to damped cosine and sine functions, but if we look at them more closely the differences become visible. It is of importance that all Bessel functions show zero crossings, which means that they disappear for certain arguments. For example, $J_0(\Delta\varphi_C) = 0$ for $\Delta\varphi_C \approx 2.4$. This fact in the past has often been used by laboratory engineers to calibrate their deviation meters.

What does all of this mean for the modulation bandwidth? In analog PMR systems, low frequencies might appear with a large modulation index and, therefore, many sideband frequencies have to be taken into account, but their frequency difference is small and thus the modulation bandwidth is limited. The highest frequencies have larger differences but smaller modulation indices and, therefore, fewer sideband frequencies are of interest. Again the total modulation bandwidth is limited. In practice, speech contains a mixture of different frequencies but, as has been shown, the modulation bandwidth will be limited to an acceptable size.

For measurements of analog PMR sets, a modulation frequency of 1 kHz at 60% peak deviation is usually taken. Assuming 5-kHz peak deviation and 25-kHz channel separation, the phase deviation is $\Delta\varphi_C = 3$. This also represents roughly a speech signal that has its peak intensity around 1 kHz, and its higher frequencies have weaker energy and thus a smaller modulation index. The highest modulation indices would result at the lowest modulation frequency, but there speech signals also exhibit a smaller energy and, therefore, the possible maximum of $\Delta\varphi_C = 5/0.3 = 16.67$ will never be reached. The situation might be different for tone signaling and data transmission, but even there the peak deviation and the maximum possible modulation index will never occur. Note that for broadcast and HiFi transmission,

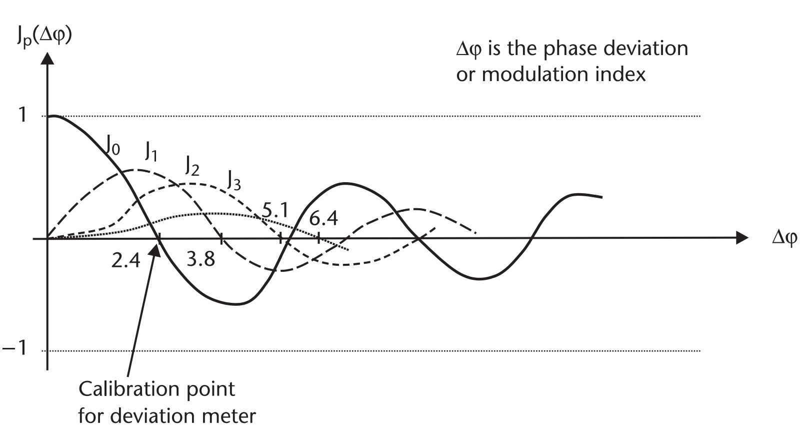

Figure 4.2 Bessel functions of the first kind and *p*th order.

a considerably higher bandwidth is required than for analog PMR because both the upper baseband cutoff frequency and the modulation index are much larger.

Because the modulation bandwidth is determined by 99% of the transmitting power, the total out-of-band power is 0.5% per side due to the symmetry of the spectra of symmetric functions. In practice, only the first sideband frequency outside the modulation bandwidth is usually of interest because the next one and all others are for the most part much smaller. Thus, all modulation products below –23 dBc can be neglected:

$$10 \cdot \log_{10}\left[\sum_{p=p_M+2}^{\infty} J_p^2(\Delta\varphi_C)\right] << 10 \cdot \log_{10}[J_{p_M+1}^2(\Delta\varphi_C)] \leq -23 \text{ dBc} \quad (4.20)$$

$$\text{because} \quad J_{p+1}^2(x) << J_p^2(x)$$

In Table 4.1 the logarithms of the Bessel functions are given in –dBc to find more quickly the sidebands relevant for modulation as well as for adjacent power bandwidth. Because all power levels below –80 dBc are not of interest, these have been suppressed, and the signs of the levels in dBc have also been suppressed to save space. Moreover, the limits for modulation products below –23 dBc and –60 dBc have been clearly marked. Finally, we should mention that the power levels are given with an accuracy not exceeding 0.1 dB because, in practice, the tolerances of nearly all measurements are usually much larger.

For digital modulation in particular, the Bessel functions of small arguments are of specific interest and these will, therefore, be discussed further in Chapter 5.

If the 99% modulation bandwidth will be normalized to the baseband frequency and the modulation index normalized in a manner similar to the way it will be normalized in Chapter 5 to the bit rate, then from the evaluation of the highest order p_M of the last sideband contributing to the modulation bandwidth, it becomes obvious that there is a close linear relation between p_M and the phase deviation $\Delta\varphi_C$. After the introduction of the two constants a and b, we get from (4.17):

$$\frac{B_M}{2f_B} = p_M \approx (a \cdot \Delta\varphi_C + b) \quad (4.21)$$

Evaluating (4.16) these two constants can be determined and the result is simply $a = b = 1$. Hence, we have found a very simple possible way of calculating the 99% modulation bandwidth without the need to evaluate the Bessel functions:

$$B_M \approx 2f_B \cdot (\Delta\varphi_C + 1) = 2 \cdot (\Delta F_C + f_B) \quad \text{for} \quad \Delta\varphi_C \geq 1 \quad (4.22)$$

This formula works well for $\Delta\varphi_C \geq 1$ and is very accurate above 2. It is nothing more than the Carson formula, which has been known of for a long time [12–14, 17]. The reason why this has been elaborated so thoroughly is that we need a modified relationship for those regions where the Carson formula is not applicable.

A certain bandwidth is needed to modulate the carrier with the message to be transmitted. For angle modulation, we use the general rule that the robustness

Table 4.1 Squared Bessel Functions of the First Kind and pth Order (in –dBc)

		$\lvert 10 \cdot \log_{10} \cdot \{J_p^2(\Delta\varphi_C)\}\rvert$ *[–dBc]*															
$\Delta\varphi_C$	$p = 0$	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
0.0	≡0	∞	∞	∞	∞	∞	∞	∞	∞	∞	∞	∞	∞	∞	∞	∞	∞
0.1	<0.1	26.0	58.1	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
0.2	<0.1	20.0	46.0	75.6	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
0.3	0.2	16.6	39.0	65.0	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
0.4	0.4	14.1	34.1	57.6	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
0.5	0.6	12.3	30.3	51.8	75.9	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
0.6	0.8	10.9	27.2	47.1	69.6	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
0.7	1.1	9.7	24.6	43.2	64.3	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
0.8	1.4	8.7	22.4	39.8	59.7	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
0.9	1.9	7.8	20.5	36.8	55.7	76.6	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
1.0	2.3	7.1	18.8	34.2	52.1	72.0	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
1.2	3.5	6.1	16.0	29.7	46.0	64.3	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
1.4	4.9	5.3	13.7	25.9	40.9	57.8	76.3	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
1.6	6.8	4.9	11.8	22.8	36.5	52.2	69.6	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
1.8	9.4	4.7	10.3	20.1	32.7	47.3	63.7	>80	>80	>80	>80	>80	>80	>80	>80	>80	>80
2.0	13.0	4.8	9.0	17.8	29.4	43.0	58.4	75.1	>80	>80	>80	>80	>80	>80	>80	>80	>80
2.5	26.3	6.1	7.0	13.3	22.6	34.2	47.5	62.2	78.1	>80	>80	>80	>80	>80	>80	>80	>80
3.0	11.7	9.4	6.3	10.2	17.6	27.3	38.9	51.9	66.1	>80	>80	>80	>80	>80	>80	>80	>80
3.5	8.4	17.2	6.8	8.3	13.8	21.9	31.9	43.4	56.2	70.1	>80	>80	>80	>80	>80	>80	>80
4.0	8.0	23.6	8.8	7.3	11.0	17.6	26.2	36.4	47.9	60.6	74.2	>80	>80	>80	>80	>80	>80
5.0	15.0	9.7	26.6	8.8	8.2	11.7	17.7	25.5	34.7	45.2	56.7	69.1	>80	>80	>80	>80	>80
6.0	16.4	11.2	12.3	18.8	8.9	8.8	12.2	17.7	25.0	33.5	43.1	53.8	65.3	77.5	>80	>80	>80
7.0	10.5	46.6	10.4	15.5	16.0	9.2	9.4	12.6	17.9	24.6	32.6	41.6	51.5	62.3	73.6	>80	>80
8.0	15.3	12.6	18.9	10.7	19.5	14.6	9.4	9.9	13.0	18.0	24.3	31.8	40.3	49.7	59.8	70.7	>80
9.0	20.9	12.2	16.8	14.8	11.5	25.2	13.8	9.7	10.3	13.4	18.1	24.1	31.2	39.3	48.2	57.8	68.1
10.0	12.2	27.2	11.9	24.7	13.2	12.6	36.8	13.3	10.0	10.7	13.7	18.2	24.0	30.8	38.4	46.9	56.1
12.0	26.6	13.0	21.4	14.2	14.8	22.7	12.3	15.4	26.9	12.8	10.4	11.4	14.2	18.4	23.7	30.0	37.1
14.0	15.3	17.5	16.4	15.0	22.4	13.1	21.8	16.4	12.7	18.8	21.4	12.6	10.9	11.9	14.6	18.6	23.6
16.0	15.1	20.9	14.6	27.2	13.9	24.8	15.6	14.8	43.1	14.4	13.7	23.3	19.0	12.5	11.3	12.4	15.0

against interference and multipath distortion increases with the modulation index or the modulation bandwidth, respectively. On the other hand it is desirable to put as many radio channels into a given bandwidth as possible because frequency is a rare resource. How do we cope with these contradicting requirements?

For AM the carrier spacing should only be a little bit wider than the modulation spectrum. Additionally, finite filter steepness and transmitter and receiver frequency tolerances have to be taken into account. Because PMR uses angle modulation almost without exception, the following considerations are restricted to FM and PM and their wider modulation bandwidth.

The modulation bandwidth B_M nicely matches the 20-dB bandwidth that the receiver needs at least to avoid severe modulation distortion due to insufficient IF pass bandwidth. To avoid adjacent channel impairment by modulation products, the adjacent channel power B_{ACP} should not exceed –60 or –70 dBc dependent on the channel separation and the type of system in question [18–20]. Unfortunately, again the related Bessel functions must be calculated by setting (4.16) to 10^{-6} or 10^{-7} to determine the adjacent power bandwidth. Due to the rapid decrease of the Bessel functions with increasing order, again only the first sideband modulation product falling into the adjacent channel needs to be evaluated:

$$10 \cdot \log_{10} \left| J^2_{p_{\mathrm{ACP}}+1}(\Delta\varphi_C) \right| \le \begin{cases} -60 \text{ dBc} \\ -70 \text{ dBc} \end{cases} \tag{4.23}$$

From Table 4.1 p_{ACP} can be determined and the adjacent channel power bandwidth, therefore, is

$$B_{\mathrm{ACP}} = 2p_{\mathrm{ACP}} \cdot f_B \tag{4.24}$$

Note that the ETSI standards suggest for the measurement of the adjacent channel power in analog PMR systems a modulation frequency of 1.25 kHz with an input level override of 20 dB more than is necessary for 60% peak deviation [18, 21]. Hence, the modulation signal is limited to maximum deviation and the phase deviation is $\Delta\varphi_C = 2$, 3.2, or 4 for a channel separation of 12.5, 20, or 25 kHz, respectively.

From modulation bandwidth, receiver channel filter properties, and the tolerances of transmitter and receiver center frequency, the necessary channel separation can now be determined as shown in Figure 4.3, which visualizes the relation between modulation bandwidth and receiver channel filter characteristics [11, 22]. Unwanted interference from the transmitter to a receiver in an adjacent channel can occur if either the adjacent channel power exceeds the limits due to insufficient premodulation filtering, or if strong spurious sidebands due to noise or distortion appear, or if the receiver channel filter stop-band attenuation is insufficient. The formula for the channel separation can directly be derived from this figure:

$$\Delta F_{\mathrm{Ch}} \ge \frac{1}{2} \cdot (B_M + B_{\mathrm{ACP}}) + \delta f_{\mathrm{Rx}} + \delta f_{\mathrm{Tx}} > B_M \tag{4.25}$$

Equation (4.21) has turned out to be very convenient for quick modulation and adjacent channel power bandwidth calculations. From Table 4.1 we can derive

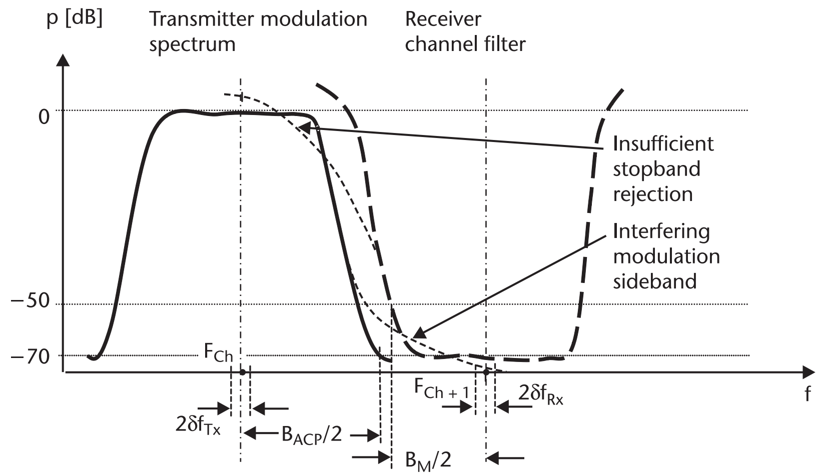

Figure 4.3 Necessary channel separation for PMR.

that it can be used with constants $a = 1.3$ and $b = 3.5$ to determine the adjacent channel power bandwidth. Hence, we get for the −60 dBc limit:

$$B_{\text{ACP}} \leq 2 \cdot (1.3 \cdot \Delta F_C + 3.5 \cdot f_B) = 2f_B \cdot (1.3 \cdot \Delta\varphi_C + 3.5) \quad \text{for} \quad \Delta\varphi_C \geq 1 \tag{4.26}$$

This formula works well above $\Delta\varphi_C = 1$ and very precisely above 2. For smaller modulation indices, as will be interest for digital modulation in Chapter 5, a modified approach is discussed. Because the higher multiples of the baseband frequency decrease very quickly with increasing order for the cases of interest in PMR (characterized by small modulation indices), usually no significant difference is seen between B_{ACP} for −60 and −70 dBc that are the different limits for the various PMR channel separations and systems.

4.3 Properties of Conventional Analog PMR Equipment

Decades of different analog PMR development in the European countries has led to a number of different PMR bands, several different channel separations, and a variety of other parameters being in use, as Table 4.2 illustrates.

Moreover, different licensing conditions are in use and frequency planning and allocation methods are everything but uniform. Hence, frequency tolerances, transmitting power levels, and maximum antenna heights and gains also exhibit considerable differences. Last but not least, two different types of modulation are currently in use, namely, PM and FM, and a number of different selective calling and data transmission systems have been introduced in the past few decades. All of this results in a nearly uncountable number of combinations and, therefore, a large number of different types of PMR radios. This limits the production numbers of each type and drives up costs [1, 5].

In the United States, the history of PMR has been quite different from that of Europe and has led in part to different technical parameters such as channel

Table 4.2 Main Properties of European PMR Systems

Frequency bands	40, 80, **160**, 390, 420, 440, **460**, and 900 MHz
Channel separation	5.0, 6.25, **12.5**, 20, 25 kHz
Duplex separation	Mainly 4.6, 5.0, 9.8, 10, 45 MHz
Modulation	**PM**, FM, and SSB
Other parameters	Transmit power, frequency tolerance, and antenna parameters vary widely
Selective calling	Tone squelch and various simultaneous and sequential tone systems
Data transmission	Various systems from 300 via **1,200** up to 2,400 bit/s; rarely 4,800 bit/s

Note: The most important characteristics are in bold.

separations of 40 and 30 kHz, but later also 25, 20, 15, and 12.5 kHz. For narrowband systems based on SSB, 7.5-, 6.25-, and 5-kHz channels were allocated. Different bands were also successively opened for PMR. In the beginning the 7m and later the 2-m and 70-cm regions were used. The 800- and 900-MHz bands were used for PMR much earlier in the United States than in Europe, and currently even frequencies around 1.5 GHz are in use in the United States for PMR. Because in the United States a proportion of the PMR systems is needed for long-range coverage, BS transmitting powers of up to several hundreds watts are in use—much higher powers than are permitted in many European countries. The limits for mutual interference, and here in particular the unwanted emissions of receivers and transmitters, have been fixed by FCC rules, whereas other parameters determining the transmission quality directly or indirectly have been set by TIA/EIA standards [23–25]. Moreover the methods of measurements developed in Europe by the CEPT and later ETSI, in the United States by the FCC and the EIA, and elsewhere are not identical.

Therefore, about three decades ago the IEC began a work item project to standardize the methods of measurement for PMR on an international level. Due to the different methods in use in Europe, the United States, and elsewhere, however, unification on the methods of measurement could not be reached. However, in the annexes of IEC Publication 60489 (formerly IEC Publication 489) all important methods have been collated [26]. If no consensus could be found on a specific method of measurement, then at least the number of different methods has been reduced to very few alternatives. Unfortunately, there was no official cooperation between IEC and CEPT and later ETSI on this topic and, therefore, many of the methods have never been brought in line. Despite this, the differences are much smaller than had been feared in the beginning due to many informal contacts between the experts in the different international bodies. However, in reality, the situation is much more complicated. Even if identical methods of measurement have been chosen by the administrations of different countries for type approval and licensing, in many cases different limits have been set.

Although the methods of measurements recommended by the IEC are to be found in Publication 60489, the ETSI methods of measurement and the evaluation of their uncertainties have been fixed in the ETSI PMR coexistence standards and specifically in technical reports ETR 027 and ETR 028, which are continuously updated, as are other ETSI documents [21, 27].

Finally, we should mention that the ITU-R (formerly CCIR) tried very early to harmonize the main characteristics and limits of PMR systems to avoid unnecessary interference and to ease frequency coordination at country borders by issuing several reports and recommendations on this topic. All of these cover analog and simple digital PMR systems and, most important, they have set up reasonable rules for the coexistence of different types of systems [28–31].

4.4 European Coexistence Standards for PMR

The long history of analog PMR has been very different in most of the European countries, and not before about three decades ago were the first attempts made to unify PMR services in Europe. At first, this was tried with the introduction of CEPT recommendations for PMR-type approval (mainly, CEPT Recommendation T/R 24-01 [32]), which since 1987 have been transferred to ETSI where they have been transformed into several ETSI PMR standards. Nevertheless, it appeared impossible to remove the results of decades of different PMR development in the different European countries. Hence, currently a number of different PMR bands and three different channel separations are in use. In fact, further channel separations are already in use or under discussion for the narrowband system.

The primary source of the European analog PMR standards is EN 300 086. This is not a system standard, but a coexistence standard that sets the rules for equipment design insofar as coexistence on adjacent channels and frequency reuse is concerned. This specification applies to equipment with antenna connectors. If selective calling devices are added to a PMR radio, EN 300 219 also has to be met. Finally, data transmission has been introduced and EN 300 113 ensures full coexistence with systems governed by EN 300 086 and 219 [18, 19, 33].

For equipment with integral antennas, similar coexistence conditions have been fixed by EN 300 296, 341, and 390. Each of them corresponds to one of the three first mentioned above. The methods of measurement to be applied to equipment with integral antennas are more complicated and more time consuming than those without because many measurements have to be carried out on radio test sites as radiation measurements [20, 34, 35].

Three more standards should be mentioned. The first is EN 301 489-5, which deals with EMC requirements imposed by the EMC Directive of the European Commission. The second one is EN 300 471, which is needed for an orderly access to the frequencies in case of systems with shared channels. Moreover, recently EN 301 166 has been issued for narrowband systems with a channel separation below 10 kHz. It determines coexistence conditions for a channel separation of, for example, 5.0 or 6.25 kHz and it merely specifies a transparent analog radio channel for analog or digital transmission but a radio modem may be added for data transmission [36–38].

Table 4.3 shows the main properties of conventional analog and simple digital PMR systems for the frequency range between 30 and 1,000 MHz. The limits, for example, for adjacent channel power and adjacent channel rejection have been set under the assumption that it must be possible for one radio set to transmit with maximum permissible power while another is just receiving a very weak signal in the adjacent channel within a small spatial distance of a few meters.

Table 4.3 Common Limits of European PMR Standards

Parameter	*Limit/Typical Value*[1]
Frequency range	30–1,000 MHz
Spurious emissions[2]	–36/–57 dBm for transmitter/receiver
Transmitter output power	Usually 100 mW to 25W, depending on frequency and application
Frequency tolerance	±0.6 to ±2.5 kHz, depending on frequency band, channel separation, and type of station
Adjacent channel power	–60/–70 dBc, depending on the channel separation (or ≤–37 dBm)
Transmitter interchannel modulation	10-dB conversion loss for mobile stations, 40-dB conversion loss for base stations
Receiver sensitivity[1,3]	–107 dBm (0 dB $\mu V_{CC}/50\Omega$) for analog equipment and selective calling devices, –110 dBm for data, typically –100 to –110 dBm with Rayleigh fading
Cochannel rejection[3,4]	–12 and –8 dB P_C/P_I for 12.5 and 20/25 kHz, approximately –21 and –17 dB with Rayleigh fading
Adjacent channel rejection	60 and 70 dB for 12.5 and 20/25 kHz for 14-dB SINAD
Spurious response	70-dB rejection
Receiver intermodulation	65-dB rejection for mobile stations, 70-dB rejection for base stations
Blocking	–23 dBm (84 dB $\mu V_{CC}/50\Omega$)
AF quality	See ZVEI Recommendation AKR 95-1D, no ETSI requirement
Ambient temperature	–20 to +55°C
Power supply tolerance	±10% for mains, maximum –15% to +30% for batteries, depending on the type

[1]Receiver limits for equipment with integral antenna, depending on frequency band.
[2]Below 1 GHz.
[3]Approximately 9-dB difference between static and dynamic values; see Chapter 9.
[4]Approximately ≈–6 dB for 30-kHz channels.

For 20- and 25-kHz channels on one hand and 12.5-kHz channels on the other, in some cases two limits are given. For a channel separation of 12.5 kHz, the limits of *adjacent channel power* (ACP), *adjacent channel rejection* (ACR), and *cochannel rejection* (CCR) that are applicable to 20 and 25 kHz are very difficult (ACP and ACR) or even nearly impossible (CCR) to meet. The limits of the main parameters comply with international specifications, such as ITU-R Recommendation 748-5 [31].

Unlike earlier approval requirements, those related to AF quality, such as distortion and audio frequency response, have been deleted. It has been assumed that the market force of competition will regulate these equipment characteristics. Nevertheless, in Germany the *Zentralverband der Elektrotechnik- und Elektronikindustrie* (ZVEI) has issued a recommendation concerning the AF parameters related to the quality of transmission, and the EIA standards contain similar parameters [39].

Note that the characteristics of narrowband equipment in accordance with EN 301 166 differ on some small points. However, equipment for 5.0- and 6.25-kHz channel separation is currently used only for specific applications.

4.5 Modulation-Related Aspects of Analog PMR Equipment Design

The overwhelming number of analog PMR sets employ angle modulation. AM is only used in nonprofessional radio equipment like CBs or in services others than land mobile. The only exception is SSB for narrowband systems. Angle modulation offers two possibilities: frequency or phase modulation. Only in a few cases is true FM used in PMR, with PM being predominant. There is a simple reason for this: PM offers higher receiver sensitivity.

The sensitivity $p_{\mathrm{Rx\,min}}$ of an analog receiver is the minimum input power level necessary to deliver a specified signal-to-noise (S/N) ratio at the output, for example, 12 or 20 dB defined as the ratio (S + D + N)/(N + D) or SINAD. Disregarding the distortion D, this can be replaced by (S + N)/N, or simply S/N if S >> N:

$$p_{\mathrm{Rx\,analog\,min}} = 10 \cdot \log_{10} N_0 + 10 \cdot \log_{10} B_{\mathrm{AF}} + k \cdot \sigma_F + 20 \cdot \log_{10} \frac{S}{N} + 10 \cdot \log_{10} F - g_M \tag{4.27}$$

where $10 \cdot \log_{10} N_0 = -174$ dBm/Hz is the thermal noise power density; B_{AF} is the baseband width (nearly equal to the upper baseband cutoff frequency f_B); $10 \cdot \log_{10} F$ is the noise figure (8 dB for good PMR sets), which usually contains all implementation losses; and $k \cdot \sigma_F$ determines the impairment by fading. Shadowing is usually not included because it is very much dependent on the propagation characteristics of the radio environment. If the lower cutoff frequency is not zero but 50 or 300 Hz as is usual, this does not alter the result very much. For receiver sensitivity measurements, the output signal in some cases (e.g., in the ETSI PMR standards) is psophometrically weighted to match the measurement result better to the characteristic of the human ear. Compared to calculations according to (4.27), this gives an improvement of about 2.5 dB compared to a flat audio response in the AF passband [40].

For the calculation of a realistic link budget and the coverage distance, $k \cdot \sigma_F$ has to be replaced by $k \cdot \sigma_P$ where σ_P accounts for fading and shadowing. If the receiver sensitivity with fading is known and only shadowing has to be included, then the following correction has to be made:

$$p_{\mathrm{Rx\,analog\,min}\,P} = p_{\mathrm{Rx\,analog\,min}\,F} + k \cdot \sqrt{\sigma_P^2 - \sigma_F^2} \tag{4.28}$$

Another important term that depends on the type of modulation and the modulation parameters is the modulation gain g_M, which is also called the receiver processing gain. Values of g_M for different modulation methods are listed in Table 4.4 [13].

Table 4.4 Modulation Gain in Decibels for Different Modulation Methods

Modulation Method	*Modulation Gain*
SSB	$g_{\mathrm{SSB}} = 0$
AM	$g_{\mathrm{AM}} = 10 \log_{10}(m^2) - 3$ dB
FM[1]	$g_{\mathrm{FM}} = -10 \log_{10}(3\Delta\varphi_C^2/2)$
PM[2]	$g_{\mathrm{PM}} = -10 \log_{10}(9\Delta\varphi_C^2/2)$

[1]With $\Delta\varphi_C = \Delta F / f_B$.

[2]For PMR with equal deviation ΔF at 1 kHz and $f_B = 3$ kHz.

Current basic PMR SSB systems suffer from bad noise and CCR. In practice, this nearly always is significantly improved by adding suitable companding techniques to reduce the dynamic range of the transmitted signal and expand it again after reception. The improvement from companding is not included in g_{SSB} in Table 4.4. The formulas given for FM and PM are valid only above the so-called FM threshold, which for PMR sensitivity definition is a correct assumption. Above the FM threshold, the following relation between FM and PM modulation gain is valid due to the differing noise weighting by having or not having de-emphasis in the receiver:

$$\frac{g_{PM}}{g_{FM}} = 20 \cdot \log_{10} \frac{\Delta\varphi_{CFM} \cdot \omega_B}{\Delta\Omega_{FM} \cdot \sqrt{3}} \tag{4.29}$$

In PMR systems for both FM and PM, an equal frequency deviation is usually chosen for the modulation frequency at 1 kHz. Assuming a channel separation of 20 kHz, a peak frequency deviation ΔF = 2.4 kHz, and a phase deviation $\Delta\varphi_C$ = 2.4, the result is an improvement of 20 $\log_{10} \sqrt{3}$ = 4.8 dB for PM. Calculating the receiver sensitivity for a channel separation of 20 kHz, a deviation of 2.4 kHz, a maximum AF cutoff frequency of f_B = 3 kHz, a noise figure of 8 dB, and a SINAD of 20 dB is assumed. The static sensitivity without fading ($k\sigma_F = 0$) of a FM receiver is –111.7 dBm, whereas a PM receiver offers –116.5 dBm. From that, the dynamic receiver sensitivity value can be derived by adding roughly 9 dB, as explained in Chapter 9, meaning that the sensitivity gets 9 dB worse. The channel separation of 20 kHz has been chosen for this example because both FM and PM systems are available.

Why should FM be used at all if PM delivers better receiver sensitivity? If the range below the sensitivity level is regarded, then it turns out that for very low signal-to-noise ratios below roughly 10 dB, FM performs better. Recalling that the outer parts of the coverage area always exhibit bad speech quality, it becomes clear why the use of FM might be beneficial in such situations. (By the way, one of the very rare FM user groups is the German organizations devoted to public safety). Figure 4.4 shows the AF signal amplitude together with the noise curve. Hence, for good coverage areas PM is superior, but at the cell border and beyond FM might be preferable.

In practice, many analog modulators and demodulators are of the FM type but PM is used primarily. What should the analog equipment designer do in this case? Fortunately, there is an easy method to convert analog FM to PM or vice versa as (4.11) suggests. If the baseband signal is passed through a differentiator, in the simplest case a passive RC highpass filter, its frequency response is raised by +6 dB/octave (pre-emphasis). Applying this to a linear FM modulator in the transmitter produces PM with a deviation proportional to the input voltage. An FM receiver converts frequency deviation linearly to output voltage. If PM is applied, it is only necessary to alter the frequency response of the baseband by –6 dB/octave (de-emphasis) with an integrator or a passive RC lowpass filter after demodulation to process PM properly with a FM receiver and to obtain at its output a flat frequency response. Likewise, FM can be generated from PM modulators and

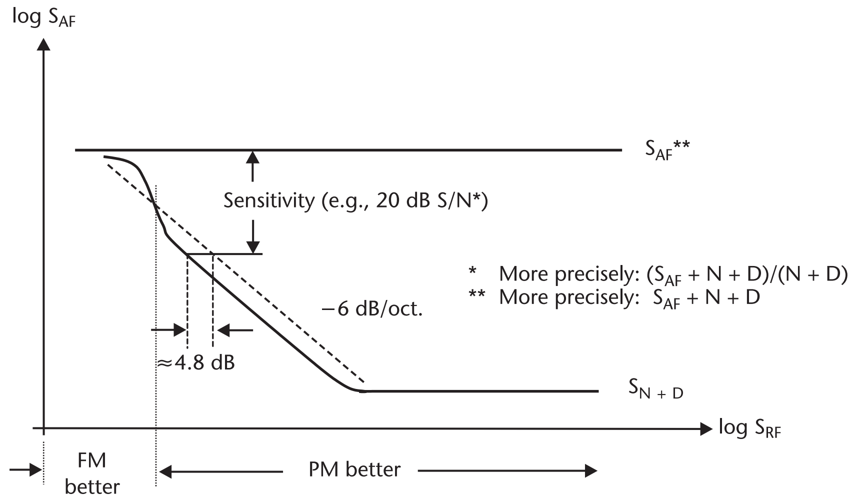

Figure 4.4 FM versus PM in analog PMR sets.

demodulators by weighting the baseband with –6 dB/octave in the transmitter and +6 dB/octave in the receiver.

PMR sets have a frequency deviation limiter, and in the case of FM-to-PM conversion the average baseband signal level has to be reduced to avoid frequency response distortion by unwanted deviation limitations for the higher AF range as Figure 4.5 suggests. However, the speech spectrum shows a natural decrease at higher frequencies. Thus, for FM and for PM an identical frequency deviation can be set up at 1 kHz for speech transmission.

Another question is why has AM been more or less completely abolished in PMR? There must be reasons to use the more complex modulation schemes of FM and PM instead of AM. FM and PM signals occupy a wider bandwidth, but show a better sensitivity compared to AM. An important additional reason becomes clear if an interference situation is analyzed.

In the case of AM, the ratio between the wanted and unwanted modulation introduced by an interferer within the modulation bandwidth is directly dependent

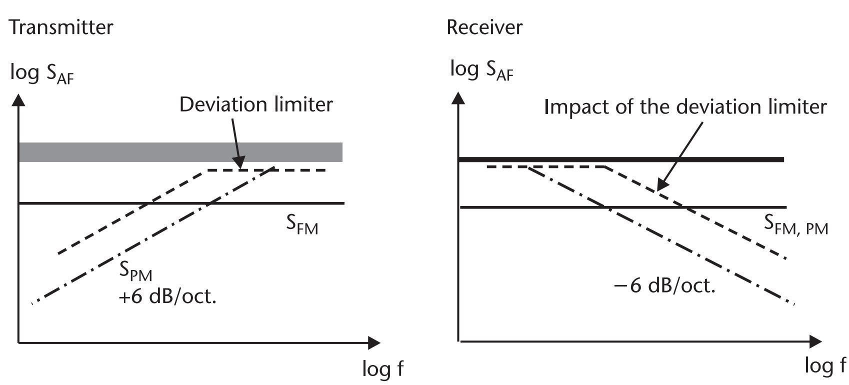

Figure 4.5 FM-to-PM conversion.

on the amplitude ratio of the carrier and the interferer. In the case of FM and PM, the signal envelope is constant. That means amplitude interference can be easily removed by hard limiting of the received signal inside the IF amplifier of the receiver. Unfortunately, the interferer produces not only an unwanted AM but also an unwanted PM component as Figure 4.6 indicates.

If the amplitudes of the wanted carrier and the interferer are S_C and S_I, respectively, then their ratio gives directly the unwanted grade of modulation m_I for AM:

$$m_I = \frac{S_I}{S_C} \tag{4.30}$$

If the grade of the wanted modulation is m, then the AF signal-to-interferer voltage ratio $(\mathrm{S/I})_{\mathrm{AF}}$ will be

$$(\mathrm{S/I})_{\mathrm{AM}} = \frac{m}{m_I} \tag{4.31}$$

For PM the unwanted phase deviation $\Delta\varphi_I$ due to the presence of the interferer can be derived from Figure 4.6:

$$\sin \Delta\varphi_I = \frac{S_I}{S_C} \tag{4.32}$$

Hence, the AF signal-to-interferer voltage ratio $(\mathrm{S/I})_{\mathrm{PM}}$ for PM can be determined from the ratio of the unwanted phase deviation to that of the wanted modulation $\Delta\varphi_C$ where for small arguments the arcsin function can be replaced by the argument itself:

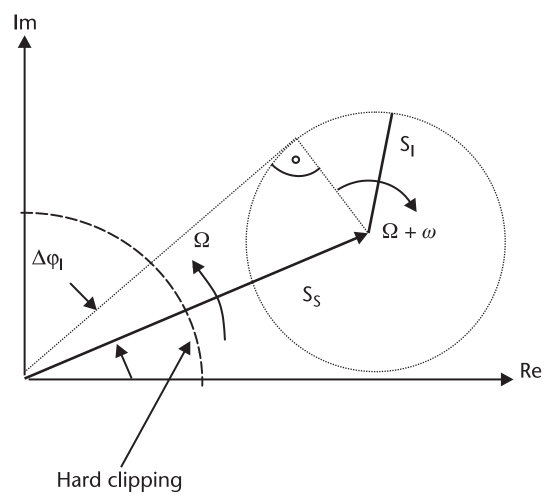

Figure 4.6 Interference reduction by hard clipping.

$$(S/I)_{\mathrm{PM}} = \frac{\Delta\varphi_C}{\arcsin\left(S_I/S_C\right)_I} \xrightarrow{S_I << S_C} \Delta\varphi_C \cdot \frac{S_C}{S_I} \tag{4.33}$$

This means that the unwanted interfering PM modulation is relatively small compared to the wanted modulation for the case of reasonable phase deviation of the wanted modulation. Hence, the benefit of FM and PM is total suppression of all amplitude variations including ignition noise and so on and the creation of only a small unwanted phase component. The larger the modulation index of the wanted phase or frequency signal, the better the suppression of the interferer will be. Let's look at some specific cases.

At first for AM the best case with $m = 1$ is regarded. If a $(S/I)_{AM}$ ratio of 20 dB is desired, the unwanted grade of modulation must not exceed $m_I = 0.1$. If for PM a wanted phase deviation of $\Delta\varphi_C = \pi/2$ is assumed for the same ratio of carrier to interferer the $(S/I)_{PM}$ ratio becomes $20 \cdot \log_{10}[(\pi/2)/0.1] = 24$ dB. For PM equal amplitude of carrier and interferer result in an unwanted phase deviation of $\Delta\varphi_I = \pi/2$ and 0 dB will be found for the $(S/I)_{PM}$ ratio as is found for equal amplitudes in AM. In both cases the channel will be completely disturbed. However, as long as the interferer amplitude does not exceed that of the carrier, the interfering phase deviation never can exceed $\pi/2$, but in principle the wanted phase deviation can be increased much further, resulting in a significant $(S/I)_{PM}$ improvement.

FM broadcasting employs a much higher modulation index than PMR and that is the reason for its excellent interference suppression. The general conclusion is that the ratio of signal phase deviation to interferer phase deviation improves with the modulation index at the expense of a wider occupied bandwidth. In PMR the compromise between interference resistance and occupied bandwidth has led to the parameters chosen for contemporary PMR systems.

4.6 Selective Calling and Slow Data Transmission Systems

Until around 1970 conventional analog mobile radio with one or few sites and one or few manually assigned channels was the only available PMR technique, and it was restricted more or less to only speech transmission. However, additional simple features were realized very early with single tones of different frequencies, but these could provide only a very limited set of additional functions.

With the transmission of several tones in parallel or different short tones in series, a wider range of functions could be implemented—even simple signaling became possible. Based on that, the next step was the introduction of selective calling, status and short data transmission, repeater control by radio commands, and the like at low data rates, mainly at 1.2 kbit/s [41]. For this purpose subcarrier modulation schemes (often called indirect) have been introduced that are not very bandwidth efficient, but they can be applied easily to existing analog equipment. Strictly speaking, all of these additional features have to be regarded as the first steps to digitalization, which started in PMR a long time ago even if nobody called it that.

One of the simplest but nevertheless widely used and reliable systems is the *Continuous Tone Controlled Squelch System* (CTCSS) according to the U.S. EIA standard RS-222-A from 1979. It uses tone frequencies below the lower AF cutoff frequency of 300 Hz for squelch control (tone squelch) and selective calling. There are three groups, A, B, and C, totaling 37 tones ranging from 67 to 250.3 Hz, but the tones of group C are less frequently used [11, 41–44].

Another squelch control method widely used in the United States is the *Digital Controlled Squelch* (DCS), which uses a continuous repetition of a Golay 23/11 code modulated with 134 bit/s onto a subaudio carrier of 67 Hz. Twelve of the 23 bits of each code word are data bits from which 3 bits are needed to initiate the code. Hence, 9 bits are left for 2^9 or 512 different addresses [17]. DCS has only a limited address capacity but it is a real digital system.

The German five-tone sequential selective calling system according to ZVEI AK SRDS 87-1E was invented in the 1960s and used in other countries as well, such as France. The specification was updated and reissued in 1987 [45]. It uses five consecutive tones from a pool of 10 frequencies f_1 to f_{10}, each representing one of the digits 1 to 0, with a duration of 70 ms each and offers, therefore, 100,000 different addresses. Moreover, group calls are provided. These are indicated by a specific group call frequency f_G, which replaces one of the last three digits, thus addressing a group of 10, 100, or 1,000 users simultaneously. In the case of addresses with two consecutive identical characters, the second is replaced by a specific frequency f_R known as a repeat tone. This was deemed necessary so that old, simple decoders would not confuse addresses with two or more consecutive identical tones due to their wide timing tolerances. The two highest tones of sequence *I* are only feasible for a channel separation of 20 kHz and above. The reduced upper AF cutoff frequency in 12.5-kHz channels required to replace the highest two by two other frequencies as tone sequence II in Table 4.5 shows. This came in use, for example, in France where it has been needed since the middle of the 1970s after the introduction of 12.5-kHz channels.

The emergency call scheme according to ZVEI AK SRDS 87-2E complements the ZVEI five-tone sequential system [46]. It provides cyclical switching between reception and transmission with 10-second duration each to get information from the mobile in an emergency and also to have the possibility of transferring information back. The transmission from such a mobile is always mixed with an alerting tone to ensure that everybody receiving such a signal recognizes at once that this is an emergency call. The emergency call cannot be switched off from inside the car; it is necessary to remove the power supply cable from the calling device. This guarantees that, for example, during a robbery in a taxi it can never be switched off. The emergency call usually is combined with a five-tone sequential signal to ensure proper automatic identification of the party in an emergency.

The five-tone sequential system is currently in widespread use in many countries. Similar systems have been specified, for example, for maritime applications by CCIR Recommendation 257-2 dating back as early as 1957 [47]. Five tones of 100-ms duration each are transmitted twice with a break of 900 ms, and 100 groups can be built, defined by all combinations of only two of the tone frequencies. A similar system with different tones has been specified by the CCITT. Moreover, a five-tone sequential selective calling system employing the CCIR tones, but shortened to 40 ms, has been recommended in the United Kingdom by the EEA in MPT 1316 where a comparison of various systems is also provided [48]. There a ZVEI

Table 4.5 Properties of Various Five-Tone Sequential Selective Calling Systems

		Tone Frequencies [Hz]											
System	T_{Tone} *[ms]*	f_1	f_2	f_3	f_4	f_5	f_6	f_7	f_8	f_9	f_0/f_{10}[1]	f_R	f_G
CCIR	100	1,124	1,197	1,275	1,358	1,446	1,560	1,640	1,747	1,860	1,981	2,110	2,400
CCITT	100	697	770	852	941	1,209	1,336	1,477	1,633	1,800	***400***[1]	2,300	—
ZVEI I	70	1,060	1,160	1,270	1,400	1,530	1,670	1,830	2,000	2,200	2,400	2,600	2,800
ZVEI II	70	1,060	1,160	1,270	1,400	1,530	1,670	1,830	2,000	2,200	2,400	***970***[2]	***885***[2]
DZVEI	70	970	1,060	1,160	1,270	1,400	1,530	1,670	1,830	2,000	2,200	***810***[2]	2,400
EEA	40	1,124	1,197	1,275	1,358	1,446	1,540	1,640	1,747	1,860	1,981	2,110	***1,055***[2]
EIA	33	741	882	1,023	1,164	1,305	1,446	1,587	1,728	1,869	***600***[1]	***459***[2]	2,151

[1] Always representing zero.
[2] Shifted to the lowerband end to avoid exceeding 2,556 Hz in 12.5-MHz systems.

version is also given (DZVEI), which never has been recommended by the ZVEI. Obviously it is one of the various unauthorized versions that have been developed and used since the 1970s mainly for 12.5-kHz applications or with shortened tones or both in various countries. Finally an even faster version with 33-ms tone duration but again different tone frequencies has been specified by the EIA [41, 43]. The duration of the complete telegram is of course five times the duration of one digit plus some ramping time for the transmitter keying on and off. The interested reader may want to take a look at Table 4.5, which provides a comparison of the main data for the five-tone sequential calling systems just discussed.

Binary coded digital selective calling systems provide superior address capacity, provide additional transmission of operational parameters and status information, and permit shorter transmission times. Some similar systems are based on *fast frequency shift keying* (FFSK) subcarrier modulation at 1.2 kbit/s where “1” is represented by 1,200 Hz and “0” by 1,800 Hz. [This a *minimum shift keying* (MSK) of a 1,500-Hz subcarrier frequency with $B_B T_B = \infty$ where B_B is the baseband width and T_B the bit duration. Due to the presence of the AF lowpass filter, $1 < B_B T_B < \infty$; further details are explained in Chapter 5.] The British MPT 1317 system specified initially by the EEA and Home Office Radio Regulatory Department in the United Kingdom in 1981, the French system according to ST/PAA/TPA/1382, and the ZVEI digital selective calling scheme defined in AK SRDS 87-3E are in widespread use. For all three systems the transmission speed and the type of modulation is identical but the message formats including the address space are different. Moreover, various similar systems exist even *within* one country. For example, the German electricity suppliers (VDEW) and the Funkmeldesystem (FMS) of public safety forces use modifications and extensions of the ZVEI digital calling system [41, 49–55].

Disregarding the first early trials, trunking was widely introduced at the end of the 1980s and the beginning of the 1990s for PAMR and SMR. In Europe the first trunked PAMR system based on MPT 1327 was opened in Band III (175–225 MHz) in the United Kingdom in 1987 with a license awarded by the Department of Trade and Industry. At that time additional spectrum for *common base station service* (CBS) was released. CBSs usually operate in single-site configurations with three voice channels on the average, covering distances of up to 50 km (30 miles). The different users are separately addressed by CTCSS or DCS. In France first PAMR licenses were awarded in 1989, and in Germany in 1990 Deutsche Telekom opened the CHEKKER network, which was taken over later by Dolphin Telecom. Other European countries were less progressive concerning PAMR introduction [56].

MPT 1327 is based on the same transmission speed and the same type of modulation as MPT 1317 and provides selective calling, status and slow data transmission, and signaling mechanisms for trunking. It is extended by the MPT 1343 specification (air interface), MPT 1347 (BS), and MPT 1352 (terminal approval). MPT 1327, which by the way exhibits many similarities to APCO Project 16, has become a quasi-standard that is in widespread use in Europe and elsewhere but not in the United States. In Germany the original specification was modified, reissued, and put into service in 1990 as RegioNet 43 [57–62].

MPT 1327 offers a considerable number of PMR features and its limitations today are mainly a consequence of the low data rate restriction of 1.2 kbit/s. Data transmission is, therefore, very limited and speech transmission is analog. Large cells are used for the most part, so handover is not necessary and, therefore, not implemented. The most advanced MPT 1327 systems show an interesting portfolio of features: up to 100 radios per channel, call queuing, roaming, *automatic vehicle location* (AVL) with evaluation of GPS data, e-mail, and many more. Intercell call setup needs about 2 seconds, whereas group calls in a single cell can be established in less than 220 ms. There are very small systems but also networks with up to two dozen channels per site. The service area can be local, but large hierarchical networks also exist that provide regional, interregional, or even country-wide coverage comprising up to many hundred base stations in total. Table 4.6 shows the most important features of modern MPT 1327 systems.

The interface between mobile radio terminals and data equipment has been specified by the *User Access Definition Group* (UADG), an independent interest group comprising manufacturers and users. This *Mobile Access Protocol for MPT 1327* equipment (MAP 27) has become a de facto standard for interfacing mobile radios and data terminals as well as other accessories, which, therefore, can also be used in many other systems [63]. MAP 27 describes a point-to-point connection based on subsets of the V.24, V.28, and RS-232 specifications for serial data interfaces.

Table 4.6 Main Properties of Modern MPT 1327 Systems

Type of System	*Trunking System with Organization Channel*
Frequency band	Mainly 403–470 MHz
Channel separation	12.5 kHz
Type of modulation	Indirect 1,200/1,800 Hz FFSK similar to MPT 1317
Speech transmission	Analog
Signaling and data	1.2-kbit/s gross transmission rate
Transmitting power	HP: 1–4W (30–36 dBm), MS: 5–24W (37–44 dBm) BS: according to national licensing conditions
Static sensitivity[1]	–117 dBm (0.3 $\mu V_{CC}/50\Omega$) at 12-dB SINAD
Dynamic sensitivity	–108 dBm
Dynamic CCR	–19 dB
Features and services	Individual calls with 1–10 digits PABX and PSTN access, DTMF dialing Group, conference, and broadcast calls Dynamic group reorganization Priority and emergency calls Call queuing Status and short message (46-bit) transmission AVL messages based on GPS Up to four messages to a single address at a time Transparent data transmission Ambiance listening Direct mode Callback facility Vote now advice Roaming but no handover

[1]Typically the values for speech and data are much better than demanded by the limits of EN 300 086 and EN 300 113.

Binary Interchange of Information and Signaling at 1,200 bit/s (BIIS 1200) was the attempt to bring the above-mentioned British, French, and German specifications for digital selective calling together. The aim was to combine them into a more comprehensive system that in the end would replace all three. Hence, the type of modulation and the transmission speed remained unmodified to provide the opportunity for a common hardware approach. The idea was developed in 1986 within the ECTEL PMR Working Group, where the basic work also has been carried out. Later the specification was transferred to ETSI where it was completed and issued in 1993 as ETS 300 230; an updated version followed 1996 [64]. The basic message format of BIIS 1200 was no new invention but a mixture of already well-proven formats as the comparison of bit allocations in Table 4.7 shows.

Of particular interest is the error protection procedure of BIIS 1200. The basic method is a 16-bit *cyclic redundancy check* (CRC). However, if improved error protection is needed, all of the information, including the CRC, is additionally protected with a tailbiting convolutional code with a code rate of 0.5. For data compression a specific packing method called Radix 40 is provided. Transparent data transmission according to the *Hierarchical Data Link Control* (HDLC) protocol is also available if needed. Like other systems, BIIS also has the possibility of creating concatenated messages for transmission of more comprehensive data messages. BIIS 1200 can be adapted to existing analog radio sets without problems and, therefore, it can be used in all PMR bands for all channel separations. The main technical properties of BIIS 1200 are similar to those of MPT 1327 as Table 4.8 indicates.

BIIS 1200 provides many important PMR features as listed in Table 4.9. Most of the elements of BIIS have been taken from other sources, which improves the

Table 4.7 Comparison of BIIS 1200 Code Word Formats with Its Predecessors

System	*LET [ms]*	*Bit Synchronization 01-seq.*	*Frame Synchronization*	*Info*	*CRC*	*FEC*	*Hang Bits*	*Total Number of Bits*
ZVEI Dig.[1]	≥25[2]	8[3]	16[4,5]	32	8[6]	—	—	72
PAA/1382	Undefined	16	16[7]	48	16[8]	—	—	96
MPT 1317	Undefined	≥16	16[9]	48	16[8]	—	—	≥96
MPT 1327	≥5	≥16	16[10]	48	16[8]	—	1	≥97
BIIS 1200	System specific	16	16[11]	48	16[8]	64[12]	1	97 or 161[12]

[1]Also FMS and VDEW 1987.
[2]Additional synchronization bits during carrier ramping-up.
[3]16 bits in FMS and VDEW 1991.
[4]1010 0010 1111 0111.
[5]FMS 8 bits: 0001 1010.
[6]$g(x) = x^8 + x^7 + x^4 + 1$.
[7]1011 0100 0011 0011.
[8]$g(x) = x^{15} + x^{14} + x^{13} + x^{11} + x^4 + x^2 + 1$.
[9]1100 0100 1101 0111.
[10]As MPT 1317 for CCH, inverted for TCH.
[11]As PAA/1382, but inverted if additional FEC is applied.
[12]Optional convolutional code.

Table 4.8 Technical Characteristics of BIIS 1200

Type of System	*Selective Calling and Data Transmission on Single Frequencies*
Frequency band	All PMR bands
Channel separation	12.5, 20, and 25 kHz
Type of modulation	Indirect FFSK with 1,200 bit/s similar to MPT 1317
Speech transmission	Analog
Transmitting power	According to national licensing conditions
Static sensitivity[1]	–117 dBm (0.3 μV/50Ω) for 12-dB SINAD and 12.5 kHz
Dynamic sensitivity	–108 dBm for 12-dB SINAD and 12.5 kHz
Dynamic CCR	–19 dB for 12.5 kHz
Telegram settling	System-specific *link establishment time* (LET), termination with hangover bit
Bit synchronization	1–0 sequence of 16-bit length
Block synchronization	Specific 16-bit pattern
Useful information	48 bit
Error protection	16-bit CRC, optional additional 64-bit tailbiting convolutional code with $r_C = 0.5$
Total message duration	97/161 bit or LET + 80.8/134.2 ms, concatenated bursts for longer messages

[1]Typically the values for speech and data are much better than demanded by the limits of ETS 300 230.

Table 4.9 Most Important Features and Services of BIIS 1200

Services and Features	Individual, group, fleet, priority, and emergency calls PABX and PSTN access Acknowledgments, status and short message transmission, transmitter identification, country codes Repeater and channel access control, channel change commands, remote on and off switching, system control, and comprehensive signaling functions
Data Transmission	Transparent data transmission (HDLC) and data compression (Radix 40) Transmission protocol well suited to the interface protocol MAP 27
Further Expansion	Extensions of the standard have been prepared and application-specific modifications are possible

compatibility with those. MAP 27 has been chosen for the data terminal interface of the mobile stations and, therefore, it is possible to use standardized MPT 1327 accessories. Note that the BIIS 1200 standard is open not only to user and application-specific modifications but also to future expansion. Hence, in principle, user-defined commands can be defined to build a trunked BIIS system despite the original intention to use BIIS only in nontrunked systems.

History repeats itself very often. When *complementary metal-oxide semiconductor* (CMOS) technology was introduced at the end of the 1960s it did not succeed well because 5 years before *transistor-transistor logic* (TTL) had conquered the market. In a similar way five-tone sequential systems have slowed down the introduction of early digital calling systems like MPT 1317 and these, another repetition, have hampered the broad introduction of BIIS 1200, which will now be superseded by fast standardized DPMR systems.

It is of interest to note that the *Nordic Mobile Telephone* (NMT) system also employs a signaling technique at 1.2 kbit/s based on FFSK with 1,200/1,800-Hz keying. Moreover, already by the 1970s subcarrier MSK systems using 2.4 kbit/s have been developed using shift keying between 1,200 and 2,400 Hz where the keying had been done at the maximum and not at the zero crossing, as in MPT 1317 and 1327 and all the other 1.2-kbit/s systems, to get a spectrum with fewer harmonics. For the German Net C, which has been in use since the mid-1980s, a direct *binary frequency shift keying* (BFSK) modulation of the carrier at a transmission rate of 5.28 kbit/s was designed around 1980, which has proven to be well suited to this application. If a low-speed digital voice coding algorithm of sufficient quality had been available, Net C would have become the first fully digital cellular public mobile telephone system in the world. However, the voice coder development failed and time-compressed analog speech was used instead. The U.S. AMPS and the U.K. TACS systems use even higher signaling rates of 10 and 8 kbit/s with FSK and Manchester coding; however, this has a coding efficiency of only 50% and, therefore, is not appropriate for digitized speech [65–67].

In conclusion, we can say that current PMR technology provides a sophisticated and powerful basic technology that can easily be tailored to very different PMR applications. Its limitations can be overcome by the introduction of digital PMR, which maintains all of the existing PMR features and adds some new ones based on higher data transmission speeds. These are the reasons why modern (digital) PMR is so versatile and unique.

4.7 The Path to True Digital PMR Systems

The very basic structure of a digital radio transmission system can be shown in a simple diagram like that of Figure 4.7 [68]. It is obvious that an existing system such as BIIS 1200 could be improved by simply changing the modulation scheme and increasing the bit rate. Hence, the transmission time for a given message could be considerably shortened while maintaining the complete functionality of the system.

One early idea was to improve BIIS 1200 and to create a BIIS 9600 system by doing exactly this, but later on this idea was dropped. It would not have been very reasonable to introduce a more powerful transmission scheme and to stay with the small information capacity and transmission speed limitations of the original system. Therefore, DIIS has been designed more or less totally afresh instead of merely

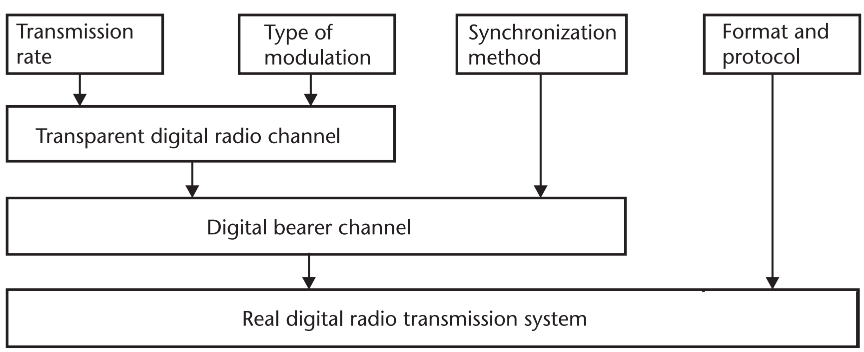

Figure 4.7 Basic structure of digital radio transmission systems. (*Source:* ECTEL PMR WG 1986.)

extending BIIS 1200, and it has become a real digital PMR system. DIIS and other true digital PMR systems will be described in detail in Chapters 6 to 8.

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CHAPTER 5

Fundamentals of Digital Modulation and Coding

Bandwidth is a limited resource and, therefore, frequencies for radio communication are a rare and precious commodity, but the need for bandwidth and new frequencies is growing progressively. Thus, new radio transmission systems need to have better frequency efficiency than older ones. Important methods for bandwidth reduction of digital radio signals are proper formatting of the baseband signal and appropriate premodulation filtering. Additionally, the combination of channel access method and modulation and coding scheme must be optimized. In combination with data compression, suitable coding and error correction digital modulation permits better quality than analog radio transmission.

The Fourier transform allows us to derive the spectrum from the shape of a function in the time domain or to generate the shape of a signal in the time domain from its components in the frequency domain. After reviewing the most interesting types of baseband signals, the results are applied to different modulation schemes to assess their bandwidths, and modulation schemes in widespread use are explained more thoroughly than others. Channel access, speech coding, channel coding, and data compression together with error protection and error correction are other important elements of the physical radio channel. However, these topics are only treated here in a very general way to explain the underlying principles.

5.1 A Tool for Spectrum Evaluation

To evaluate the behavior of a transmission channel in the time domain, the output signal $s_O(t)$ must be calculated from its input $s_I(t)$ and the convolution with the channel pulse response $h(t)$:

$$s_O(t) = \int_{-\infty}^{\infty} s_I(\tau) \cdot h(t-\tau)\, d\tau = s_I(\tau) * h(t) \tag{5.1}$$

The evaluation of this convolution integral is often difficult, and in many cases only the signal *spectra* are of interest. The main tool for the evaluation of spectra of nonperiodic signals is the *Fourier transform*, which permits the calculation of spectra from impulse shapes [1–10]. The results are immediately applicable to digital modulation because any digital message can be represented by a chain of single pulses. The operation to derive the spectrum $S(f)$ from a function in time $s(t)$ looks simple:

$$S(f) = \int_{-\infty}^{\infty} s(t) \cdot e^{-j2\pi ft}\, dt \tag{5.2}$$

Admittedly, some basic mathematical skills are necessary to perform this calculation for pulses that have a complicated shape. So why is this transform performed? It is performed because in the frequency domain only a simple multiplication of input signal spectrum $S_I(f)$ and channel response $H(f)$ suffices to get the output signal spectrum $S_O(f)$. In the end the behavior of the resultant signal in the time domain and not in the frequency domain is frequently of interest, and this can be achieved with the inverse Fourier transform:

$$s(t) = \int_{-\infty}^{\infty} S(f) \cdot e^{j2\pi ft}\, df \tag{5.3}$$

Calculating this inverse transform is often a difficult task. However, the calculation of the Fourier transform and its inverse can be eased by the use of look-up tables that are available for most of the cases of practical importance. To perform the transform only makes sense if it really exists. Therefore, some existence conditions have to be fulfilled and the most important one is that the following integral does not exceed all limits:

$$\int_{-\infty}^{\infty} |s(t)|\, dt < \infty \tag{5.4}$$

If this condition is not fulfilled, then the spectrum can be evaluated with the help of the Fourier transform in an indirect way or by using the Laplace transform [2, 11]. Additional conditions need to be fulfilled before a Fourier transform can be performed but these are of minor or even no significance in most practical cases. Very often the calculations become easier if angular frequency $\omega = 2\pi f$ is used. Then the Fourier transform and its inverse looks slightly different but this is the format being used here:

$$S(\omega) = \int_{-\infty}^{\infty} s(t) \cdot e^{-j\omega t}\, dt \quad \text{and} \quad s(t) = \frac{1}{2\pi} \cdot \int_{-\infty}^{\infty} S(\omega) \cdot e^{j\omega t}\, d\omega \tag{5.5}$$

In the time domain the integrals can be limited to the interval where the function really exists, for example, between T_1 and T_2 if the function is otherwise zero. The same applies for the frequency domain and the inverse transform if the spectrum is limited by a bandpass transmission system with the lower and the upper cutoff frequencies ω_L and ω_U. According to Parseval's theorem for any signal, its total energy must be the same regardless whether it is regarded in the time domain or in the frequency domain [1, 2]:

$$\int_{-\infty}^{\infty} s^2(t)\ dt = \int_{-\infty}^{\infty} |S(f)|^2\ df = \frac{1}{2\pi} \cdot \int_{-\infty}^{\infty} |S(\omega)|^2\ d\omega \tag{5.6}$$

It may also be useful to replace the complex exponential function by sine and cosine functions using Euler's relation [2]:

$$e^{\pm j\omega t} = \exp(\pm j\omega t) = \cos\ \omega t \pm j \cdot \sin\ \omega t \tag{5.7}$$

This allows the replacement of the complex exponential function $\exp(j\omega t)$ by $\cos\ \omega t$ for even and $\pm j \cdot \sin\ \omega t$ for odd functions. Accordingly, for even and odd functions the integral from $-\infty$ to ∞ can be simplified by taking it only from 0 to ∞ but doing it twice:

$$\int_{-\infty}^{\infty} s_{\text{even}}(t) \cdot e^{-j\omega t}\ dt = 2 \cdot \int_{0}^{\infty} s_{\text{even}}(t) \cdot \cos(\omega t)\ dt = S(\omega) \tag{5.8}$$

$$\int_{-\infty}^{\infty} s_{\text{odd}}(t) \cdot e^{-j\omega t}\ dt = -2j \cdot \int_{0}^{\infty} s_{\text{odd}}(t) \cdot \sin(\omega t)\ dt = S(\omega) \tag{5.9}$$

Figure 5.1 illustrates the cases of even and odd functions in the time domain.

Events like the rectangular pulse are discontinuous at those points in time t_D where steps occur. For the Fourier transform the value of a function is defined at this point by its average. Often it is necessary to calculate the values of $s(t)$ or

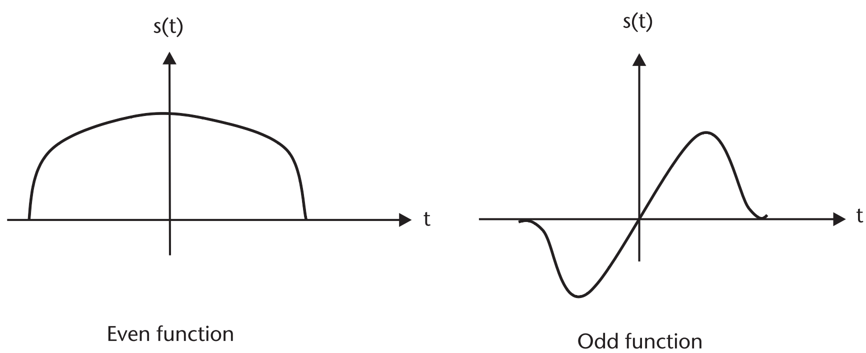


Figure 5.1 Even and odd functions in the time domain.

$S(\omega)$ at the origin. Applying $t = 0$ and $\omega = 0$ to the basic transform formulas we get due to $e^0 = 1$:

$$S(0) = \int_{-\infty}^{\infty} s(t)\, dt \quad \text{and} \quad s(0) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S(\omega)\, d\omega \tag{5.10}$$

Table 5.1 lists a number of operations associated with the Fourier transform that can be used to find the solution for a specific transform. The main benefit of these rules is that they allow the derivation of the solution to a new problem from a known one very often with only a little effort. It is particularly useful in such simple cases as multiplying the duration of a pulse or when the spectrum of the sum of two functions in the time domain is required.

The left column of Table 5.2 shows some important basic functions in the time domain (in volts), and the right column shows the spectral voltage density of these functions (in volts per hertz).

Next we take a look at some more complicated cases. Finally, if a solution cannot be found in one of the tables we have to do the calculations ourselves.

Table 5.1 Applications of Some Basic Operations to Fourier Transforms

Operation	*Time Domain*	*Frequency Domain*
Superposition	$a \cdot s_1(t) \pm b \cdot s_2(t)$	$a \cdot S_1(\omega) \pm b \cdot S_2(\omega)$
Scaling	$s\left(\frac{t}{a}\right)$	$\lvert a \rvert \cdot S(a\omega)$
Symmetry	$S(t)$	$2\pi \cdot s(-\omega)$
Complex conjugate	$s^*(t)$	$S^*(-\omega)$
Multiplication in time	$s_1(t) \cdot s_2(t)$	$S_1(\omega) * S_2(\omega)$
Convolution in time	$s_1(t) * s_2(t)$	$S_1(\omega) \cdot S_2(\omega)$
Time inversion	$s(-t)$	$S(-\omega)$
Time shift by T	$s(t - T)$	$S(\omega) \cdot e^{-j\omega T}$
Frequency shift by F[1]	$s(t) \cdot e^{j\Omega t}$ with $\Omega = 2\pi F$	$S(\omega - \Omega)$
Amplitude modulation at F	$\left.\begin{matrix} s(t) \cdot \cos(\Omega t) \\ s(t) \cdot \sin(\Omega t) \end{matrix}\right\}$ with $\Omega = 2\pi F$	$\frac{1}{2} \cdot [S(\omega - \Omega) + S(\omega + \Omega)]$ $\frac{1}{2j} \cdot [S(\omega - \Omega) - S(\omega + \Omega)]$
Differentiation in the time domain	$\frac{d^n}{dt^n} s(t)$	$(j\omega)^n \cdot S(\omega)$
Differentiation in the frequency domain	$(-j2\pi)^n \cdot s(t)$	$\frac{d^n}{d\omega^n} S(\omega)$
Integration in the time domain	$\int_{-\infty}^{t} s(\tau) \cdot d\tau$	$\pi \cdot S(0) \cdot \delta(\omega) + \frac{S(\omega)}{j\omega}$
Integration in the frequency domain	$\pi \cdot s(0) \cdot \delta(t) - \frac{s(t)}{jt}$	$\int_{-\infty}^{\omega} S(\varphi)\, d\varphi$

[1]Equal to an AM of a complex carrier; see also next case.

Table 5.2 Fourier Transforms of Some Fundamental Functions

Function	*Time Domain* $s(t)$	*Frequency Domain* $S(\omega)$
DC	$s_{DC}(t) = 1$	$S_{DC}(\omega) = 2\pi \cdot \delta(\omega)$
AC	$s_{AC}(t) = \begin{cases} e^{j\Omega t} \\ \cos(\Omega t) \quad \text{with } \Omega = 2\pi F \\ \sin(\Omega t) \end{cases}$	$S_{AC}(\omega) = \begin{cases} 2\pi \cdot \delta(\omega - \Omega) \\ \pi \cdot [\delta(\omega + \Omega) + \delta(\omega - \Omega)] \\ j\pi \cdot [\delta(\omega + \Omega) - \delta(\omega - \Omega)] \end{cases}$
Unit or Dirac pulse	$\delta(t) = \frac{d}{dt}\epsilon(t) = \begin{cases} \frac{1}{\delta} & \text{for } \lvert t\rvert < \frac{\delta}{2} \\ 0 & \text{else} \end{cases}$	$S_\delta(\omega) = 1$
Dirac pulse shifted by T	$s_\delta(t) = \delta(t - T)$	$S_\delta(\omega) = e^{-j\omega T}$
Dirac pulse train	$s_{\Sigma\delta}(t) = \sum_{n=-\infty}^{\infty} \delta(t - nT)$	$S_{\Sigma\delta}(\omega) = \Omega \cdot \sum_{n=-\infty}^{\infty} \delta(\omega - n\Omega)$ with $\Omega = 2\pi/T$
Sign in the time domain	$s_{sgn}(t) = \frac{t}{\lvert t\rvert} = \begin{cases} 1 & \text{for } t > 0 \\ -1 & \text{for } t < 0 \end{cases}$	$S_{sgn}(\omega) = +\frac{2}{j\omega}$
Sign in the frequency domain[1]	$s_{sgn\,\omega}(t) = \frac{1}{-j\pi t}$	$S_{sgn\,\omega}(\omega) = \frac{\omega}{\lvert\omega\rvert} = \begin{cases} 1 & \text{for } \omega > 0 \\ -1 & \text{for } \omega < 0 \end{cases}$
Unit step in the time domain[1]	$\epsilon(t) = \int_{-\infty}^{t} \delta(\tau)\, d\tau = \begin{cases} 1 & \text{for } t > 0 \\ 0 & \text{for } t < 0 \end{cases}$	$S_\epsilon(\omega) = \pi \cdot \delta(\omega) + \frac{1}{j\omega}$
Unit step in the frequency domain	$s_\epsilon(t) = \frac{1}{2} \cdot \delta(t) - \frac{1}{j2\pi t}$	$\epsilon(\omega) = \int_{-\infty}^{\omega} \delta(\varphi)\, d\varphi = \begin{cases} 1 & \text{for } \omega > 0 \\ 0 & \text{for } \omega < 0 \end{cases}$

[1]Note that $\text{sgn}(x) = 2\epsilon(x) - 1$

5.2 Baseband Pulse Spectrum Evaluation

To show how to use the Fourier transform, the calculation of the spectrum of a rectangular pulse with the duration T and amplitude 1 is shown. Its time domain representation is

$$s_{rect}(t) = \text{rect}\left(\frac{t}{T}\right) = \begin{cases} 1 & \text{for } |t| < T/2 \\ 0 & \text{for } |t| > T/2 \end{cases} \tag{5.11}$$

Carrying out the transform of this even function delivers the following, with $\sin(x)/x = \text{si}(x)$ [2], which is sometimes written $\text{sinc}(x/\pi)$:

$$S_{rect}(\omega) = \int_{-\infty}^{\infty} s(t) \cdot e^{-j\omega t}\, dt = 2\int_{0}^{T/2} \cos\, \omega t\, dt = 2 \cdot \left[\frac{\sin\, \omega t}{\omega}\right]_0^{T/2} = T \cdot \text{si}\left(\frac{\omega T}{2}\right)$$

$$\text{because} \quad \int \cos(ax)\, dx = \frac{1}{a} \cdot \sin(ax) \tag{5.12}$$

This si-shaped spectrum is decreasing with increasing frequency and it shows periodic nulls depending on the pulse duration at the frequencies $\pm n/T$ for $n \geq 1$ and an integer (see Figure 5.2). Recall that $\mathrm{si}(0) = 1$. This result can be used to derive the spectrum of the Dirac pulse $\delta(t)$, which is a very short but very high impulse with unity product of duration and amplitude. Replacing the amplitude 1 of the rectangular pulse by $1/\delta$ and its duration by δ the pulse area becomes unity:

$$\delta(t) = \begin{cases} \dfrac{1}{\delta} & \text{for } |t| < \delta/2 \\ 0 & \text{for } |t| > \delta/2 \end{cases} \tag{5.13}$$

From the result for the rectangular pulse, we can determine that the spectrum of $\delta(t)$ will become constant for all frequencies if the pulse duration δ tends to zero. Hence, the Dirac pulse has a constant voltage density all over the spectrum:

$$S_\delta(\omega) = si\left(\frac{\omega\delta}{2}\right) \xrightarrow{\delta \to 0} 1 \tag{5.14}$$

Trying the same procedure with the step function $\epsilon(t)$, we find that an infinite integral occurs and therefore the transform does not exist. However, replacing the step function for $t > 0$ by $s(t) = e^{-t/T}$ (needle pulse) leads to an integral that can be solved:

$$S_{\mathrm{ndl}}(\omega) = \frac{T}{1 + j\omega T} \tag{5.15}$$

With $T \to \infty$ the DC component has to be regarded separately. Because $s(t) = 1$ for $t > 0$ for the DC part $\pi\delta(\omega)$ has to be added and the spectrum as listed

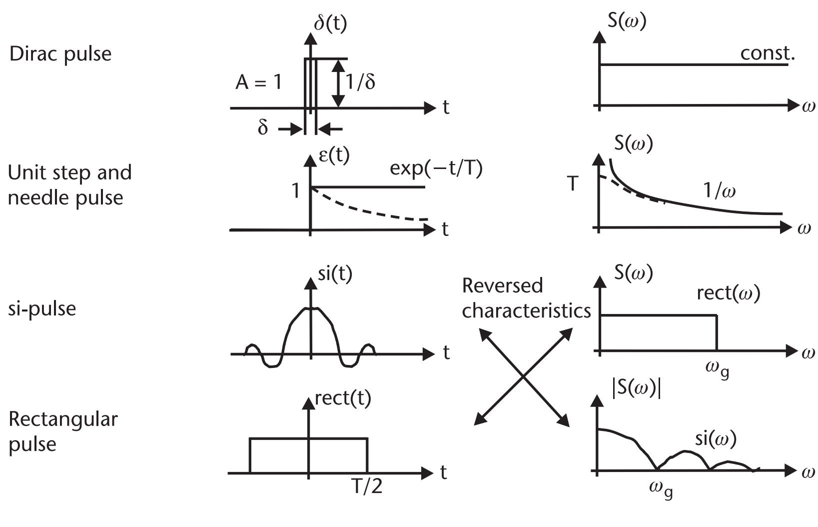


Figure 5.2 Pulse shapes and spectra for some basic cases.

in Table 5.2 will be obtained. An easier method is to use the following relationship between the Dirac pulse and the step function:

$$\delta(\omega) = \frac{d}{dt}\epsilon(t) \quad \text{and} \quad \epsilon(t) = \int_{-\infty}^{t} \delta(\tau)\, d\tau \tag{5.16}$$

Using the rule for integration in Table 5.1, we can achieve the result given in Table 5.2 for the unit step. Similarly, the ramping function can be derived from the rectangular pulse by integration. Likewise some functions are the derivatives of others as the examples of the *cos pulse* and the *cos*^{2} *ramp,* also called the *raised cosine* or *RC ramp,* shows. The duration of the RC ramp has been chosen to be $2T$ for closer comparison with the spectrum of the linear ramp. However, the integral over $2T$ would result in T for the linear ramp and $4T/\pi$ for the RC ramp [2]:

$$\int_{-T}^{T} \cos\left(\frac{\pi t}{2T}\right) dt = 2 \cdot \left[\frac{2T}{\pi} \cdot \sin\left(\frac{\pi t}{2T}\right)\right]_0^T = \frac{4T}{\pi} \tag{5.17}$$

Therefore, the necessary correction for the RC ramp ascending smoothly from 0 to 1 is $\pi/4$ and thus we get:

$$s_{\epsilon \mathrm{RC}}(t) = \frac{\pi}{4} \cdot \int_{-\infty}^{\tau} s_{\cos}(\tau)\, d\tau \tag{5.18}$$

All this confirms that there are many possibilities to derive more complicated functions from simpler ones by applying the rules shown above to simplify such calculations.

By comparing the rectangular pulse with the Dirac pulse, a general principle can be deduced: The duration of a pulse and the width of its spectrum are inversely proportional [6, 8]:

$$B_1 \cdot T_1 = 2\pi \tag{5.19}$$

Herewith the bandwidth B_1 and the pulse duration T_1 are defined for even functions as the bandwidth and the duration of rectangles with identical area:

$$T_1 = \frac{1}{s(0)} \cdot \int_{-\infty}^{\infty} s(t)\, dt \quad \text{and} \quad B_1 = \frac{1}{S(0)} \cdot \int_{-\infty}^{\infty} S(\omega)\, d\omega \tag{5.20}$$

The same insight can be gained from the scaling rule in Table 5.1 with a most important consequence: *Pulses limited in the time domain exhibit unlimited spectra in the frequency domain, whereas pulses unlimited in the time domain show limited spectra.*

The inverse transform can be calculated in a manner similar to that for the original Fourier transform. How it works will be shown with an example of the spectrum of the rectangular pulse and the known solution of the integral that appears [2]:

$$s_{\mathrm{rect}}(t) = \frac{1}{2\pi} \cdot \int_{-\infty}^{\infty} T \cdot \mathrm{si}\left(\frac{\omega T}{2}\right) \cdot e^{j\omega t}\, d\omega \tag{5.21}$$

$$s_{\mathrm{rect}}(t) = \frac{2}{\pi} \cdot \int_{0}^{\infty} \mathrm{si}\left(\frac{\omega T}{2}\right) \cdot \cos \omega t\, d\left(\frac{\omega T}{2}\right) = \begin{cases} 1 & \text{for } |t| < \dfrac{T}{2} \\ 0 & \text{for } |t| > \dfrac{T}{2} \end{cases} \tag{5.22}$$

$$\text{because} \quad \int_{0}^{\infty} \frac{\sin x \cdot \cos ax}{x}\, dx = \begin{cases} \pi/2 & \text{for } |a| < 1 \\ \pi/4 & \text{for } |a| = 1 \\ 0 & \text{for } |a| > 1 \end{cases}$$

This example also illustrates that the inverse transform may be difficult for more complicated spectrum shapes. The Fourier transform of the si-pulse (see also Figure 5.2) is similar to the inverse transform of the rectangular pulse and it delivers a strictly limited rectangular spectrum. Of course, this result could also have been depicted from Table 5.3.

Sometimes the spectrum of a function is given in a format that directly permits the calculation of its time domain representation. The unit step has been chosen as an example to demonstrate this:

$$\epsilon(t) = \int_{-\infty}^{t} \delta(\tau)\, d\tau = \begin{cases} 1 & \text{for } t > 0 \\ 0 & \text{for } t < 0 \end{cases} \tag{5.23}$$

The spectral voltage density and the reverse transform of this function are as follows:

$$S_{\epsilon}(\omega) = \pi \cdot \delta(\omega) + \frac{1}{j\omega} \qquad \epsilon(t) = \frac{1}{2\pi} \cdot \int_{-\infty}^{\infty} \left[\pi\delta(\omega) + \frac{1}{j\omega}\right] \cdot e^{j\omega t}\, d\omega \tag{5.24}$$

Reordering and evaluating the even and odd functions separately delivers the representation in the time domain based on the spectral voltage density:

Table 5.3 Fourier Transforms of Some Specific Pulse and Step Shapes

Function	*Time Domain* $s(t)$	*Frequency Domain* $S(\omega)$
Rectangular pulse	$s_{\text{rect}} = \text{rect}\left(\frac{t}{T}\right) = \begin{cases} 1 & \text{for } \lvert t\rvert < T/2 \\ 0 & \text{for } \lvert t\rvert > T/2 \end{cases}$	$S_{\text{rect}}(\omega) = T \cdot \text{si}\left(\frac{\omega T}{2}\right)$
Si-pulse	$s_{\text{si}}(t) = \text{si}\left(\frac{\pi t}{T}\right) = \frac{\sin\left(\frac{\pi t}{T}\right)}{\left(\frac{\pi t}{T}\right)}$	$S_{\text{si}}(\omega) = T \cdot \text{rect}\left(\frac{\omega T}{2\pi}\right) = \begin{cases} T & \text{for } \lvert\omega\rvert < \frac{\pi}{T} \\ 0 & \text{else} \end{cases}$
Triangular pulse (tri)	$s_{\Delta}(t) = \begin{cases} 1 + t/T & \text{for } -T < t < 0 \\ 1 - t/T & \text{for } 0 < t < T \\ 0 & \text{for } \lvert t\rvert > T \end{cases}$	$S_{\Delta}(\omega) = T \cdot \text{si}^2\left(\frac{\omega T}{2}\right)$
Gaussian pulse	$s_G(t) = e^{-2(t/T)^2}$ with $T/2 = \sigma$	$S_G(\omega) = T\sqrt{\frac{\pi}{2}} \cdot e^{-\frac{1}{2}(\omega T/2)^2}$
Needle pulse	$s_{\text{ndl}}(t) = \begin{cases} e^{-t/T} & \text{for } t > 0 \\ 0 & \text{for } t < 0 \end{cases}$	$S_{\text{ndl}}(\omega) = \frac{T}{1 + j\omega T}$
Peak pulse[1]	$s_{\text{peak}}(t) = \begin{cases} e^{(t/T)} & \text{for } t \le 0 \\ e^{-(t/T)} & \text{for } t > 0 \end{cases}$	$S_{\text{peak}}(\omega) = \frac{2T}{1 + (\omega T)^2}$
Cosine pulse	$s_{\cos}(t) = \begin{cases} \cos\left(\frac{\pi t}{2T}\right) & \text{for } \lvert t\rvert \le T \\ 0 & \text{else} \end{cases}$	$S_{\cos}(\omega) = \frac{4\pi T \cdot \cos(\omega T)}{\pi^2 - (2\omega T)^2}$
$\cos^2$-pulse[2]	$s_{\text{RC}}(t) = \begin{cases} \cos^2\left(\frac{\pi t}{2T}\right) & \text{for } \lvert t\rvert \le T \\ 0 & \text{else} \end{cases}$	$S_{\text{RC}}(\omega) = \frac{\pi^2 T \cdot \text{si}(\omega T)}{\pi^2 - (\omega T)^2}$
$\cos^2$-spectrum[3]	$S_{\text{RC}}(\omega) = \frac{\text{si}\left(\frac{\pi t}{T}\right)}{1 - \left(\frac{t}{T}\right)^2}$	$S_{\text{si}}(\omega) = \begin{cases} T \cdot \cos^2\left(\frac{\omega T}{2}\right) & \text{for } \lvert\omega\rvert \le \frac{\pi}{T} \\ 0 & \text{else} \end{cases}$
Linear ramp	$s_{\text{Ramp}}(t) = \begin{cases} 1 & \text{for } t > T \\ 1/2 + t/T & \text{for } \lvert t\rvert \le T \\ 0 & \text{else} \end{cases}$	$S_{\text{Ramp}}(\omega) = \pi \cdot \delta(\omega) + \frac{2T}{j\omega} \cdot \text{si}(\omega T)$
RC ramp	$s_{\epsilon\text{RC}}(t) = \begin{cases} 1 & \text{for } t > T \\ \frac{1}{2}\left[1 + \sin\left(\frac{\pi t}{2T}\right)\right] & \text{for } \lvert t\rvert \le T \\ 0 & \text{else} \end{cases}$	$S_{\epsilon\text{RC}}(\omega) = \pi\delta(\omega) + \frac{1}{j\omega}\,\frac{\pi^2 T \cdot \cos(\omega T)}{\pi^2 - (2\omega T)^2}$
Carrier keying-on	$s(t) = \cos(\Omega t) \cdot \epsilon(t)$	$S(\omega) = \frac{\pi}{2} \cdot [\delta(\omega - \Omega) + \delta(\omega + \Omega)] + \frac{1}{j\omega} \cdot \frac{2}{1 - (\Omega/\omega)^2}$

[1]Symmetrical needle.
[2]Also known as RC pulse.
[3]Also called RC spectrum.

$$\epsilon(t) = \frac{1}{2\pi}\left\{2\pi\int_0^\infty \delta(\omega)\cos(\omega t)\,d\omega + 2j\int_0^\infty \frac{1}{j\omega}\cdot\sin(\omega t)\,d\omega\right\} \tag{5.25}$$

$$\text{with} \quad \cos\left(\frac{\omega\delta}{2}\right) \xrightarrow{\delta\to 0} 1 \quad \text{and} \quad \int_0^\infty \delta(\omega)\,d\omega = \frac{1}{2}$$

$$\epsilon(t) = \frac{1}{2} + \frac{1}{\pi}\int_0^\infty \frac{\sin(\omega t)}{\omega}\,d\omega \tag{5.26}$$

This is another format of the unit step from which the time domain representation can be calculated, and the spectrum being $1/\pi\omega$ can also be easily recognized.

In practice, there are only a limited number of time functions and pulse shapes that can reasonably be used for transmission purposes. Therefore, the most important of them that are widely used in various transmission systems are listed in Table 5.3. The results for the spectrum evaluation of the most interesting cases will be discussed shortly to enable better understanding. If the time domain contains a DC component, then there is also a constant representing frequency zero (DC) in the frequency domain. This is nicely illustrated by the step and ramping functions.

The amplitude is always 1 while the amplitude-time area is always T for most of the pulse shapes to ensure good comparability of their spectra. Therefore, the duration of the rectangular pulse, the linear ramp, and the RC ramp has been set to T. For triangular, cos, and $\cos^2$ pulse of the total baseline duration has been set to $2T$. Hence, their average duration is also T, except for the cos pulse. The amplitude-time area of the latter is unequal to T, but its duration of $2T$ has been chosen so that it exhibits the nulls in the frequency domain corresponding to the other pulse shapes for easier comparison. For the same reason the first nulls of the si-pulse have been set to $\pm T$. It is more difficult to find the right duration for best comparability in case of exponential pulses, which show nulls neither in the time domain nor in the frequency domain. For needle and peak pulses the amplitude-time areas are T and $2T$, whereas the duration $2T = 4\sigma$, where σ is the standard deviation, has been chosen for the Gaussian pulse corresponding to about 95% of the total area. Note that for all step and ramping functions the amplitude-time area is unlimited. The Dirac pulse exhibits a constant spectrum; however, for the step function the amplitude density falls with the reciprocal value of the frequency and the spectrum of the linear ramp falls even faster with increasing frequency. That pulses with limited duration always exhibit infinite spectra, whereas pulses of unlimited duration possess finite spectra is proven by the rectangular and the si-pulse as shown in Figure 5.2.

Some pulses have reversed characteristics in the time and the frequency domain as shown by the rectangular pulse and the si-pulse. This is also true for DC and the Dirac pulse. Pulses also exist with shapes that are invariant against the Fourier transform. One is the Gaussian pulse, which also has a Gaussian-shaped spectrum. In principle, it is infinite in both the time and frequency domains. Due to the fact that this pulse drops rapidly at both sides in both domains, it has been selected for many transmission systems. Another example is a train of Dirac pulses in the time domain that produces a similar train in the frequency domain.

It is interesting to compare the spectra of pulses with similar shapes but different duration or equal average duration but different shapes. Figure 5.3 presents some important examples.

If a Gaussian-shaped pulse is made narrower with unchanged amplitude-time area, then the spectrum becomes wider. The comparison of differently shaped pulses with identical *average* length and equal amplitude-time area shows that with a smoother pulse shape the spectrum decreases faster with increasing frequency. Hence, bandwidth can be reduced by shaping pulses more smoothly or by increasing their duration or both.

To save bandwidth, pulses with a narrow bandwidth and short duration are of paramount interest, but these are contradicting requirements. Gaussian- and $\cos^2$-shaped pulses show narrow spectra and so have been selected for many radio transmission systems. The strange looking si-pulse has also been selected for many systems because it offers another good trade-off: It has a strongly limited rectangular spectrum. It has an additional very interesting property. If a chain of si-pulses is sent separated in time in such a way that the pulse center of one pulse falls onto the first null of the preceding pulse than there is no *intersymbol interference* (ISI) between the pulses if these are sampled in the pulse center. In the baseband, according to Nyquist this requires a fixed relation between the 6-dB channel bandwidth B_{Ch} and the pulse or symbol duration $2T_S$, which is the difference between the first nulls on both sides of the pulse center as Figure 5.4 shows. The sign and amplitude of the pulses do not matter and therefore this condition is also valid for multilevel signals [1, 9, 10, 12–14]:

$$B_{\text{Ch}} = \frac{1}{2T_S} = \frac{R_S}{2} \tag{5.27}$$

Unfortunately, si-pulses show severe ringing that may result in system impairments if the transmission channel is not designed carefully or if distortion occurs as in the case of multipath propagation. Therefore, instead of a very steep filter

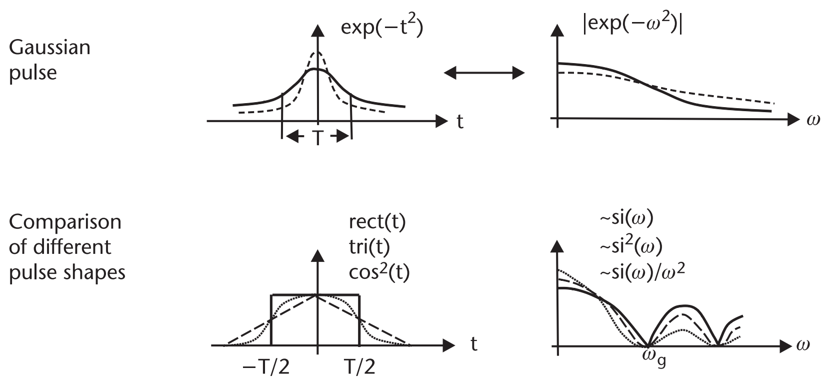


Figure 5.3 Comparison of some pulses of similar shape or average duration.

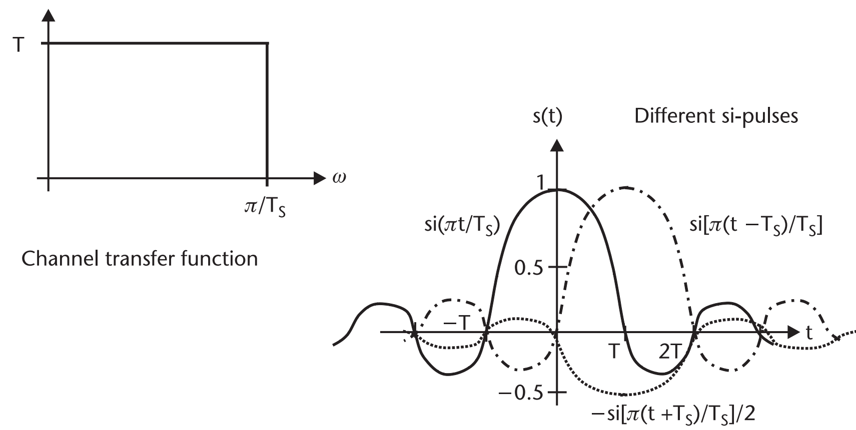


Figure 5.4 ISI free multilevel pulses in a Nyquist channel.

slope at the cutoff frequency, filters with a smoothly shaped characteristic, for example, those with a cosine roll-off characteristic and $0 < r_O < 1$, are used to reduce ringing at the expense of an increase in bandwidth. This method considerably reduces the sensitivity to channel impairments.

To conclude, it is necessary to apply appropriate pulse shaping to reduce the bandwidth of digital signals. After shaping these look very similar to analog signals and can be treated as such during transmission. The differences are of little importance up to the detection and further processing in a radio receiver.

5.3 Basic Binary Modulation Schemes

To find an appropriate modulation scheme for mobile radio communication, the first step is to select an appropriate baseband pulse format with two or more levels that has as narrow a spectrum as possible. This signal is then used to modulate the amplitude, frequency, or phase of a carrier frequency. Several modulation types may be applied simultaneously, for example, by modulating amplitude and phase simultaneously. An example would be *quadrature amplitude modulation* (QAM).

The modulation bandwidth should be as narrow as possible. The appropriate combination of baseband pulse shape, number of modulation levels, and type of modulation is a tricky question. Additionally, different types of coding are possible and a beneficial trade-off between code rate and modulation rate has to be found. In the end, the best receiver sensitivity and CCR together with robustness against multipath propagation impairments and general interference as well as low adjacent channel power have to be combined with easy synchronization, detection, error correction, and channel equalization.

The bandwidth of the baseband signal can be determined by finding the highest baseband signal frequency f_B, which has a fixed relation to the symbol rate. In correspondence with (5.27) and disregarding the harmonics in case of binary modulation, it is simply half the bit rate R_B [5, 6, 9, 10, 12, 14–23]:

$$f_B = \frac{R_B}{2} = \frac{1}{2T_B} = \frac{\omega}{2\pi} \tag{5.28}$$

Appropriate premodulation filtering removes those parts of the pulse spectra not needed for transmission and restricts the bandwidth to its minimum. If the bandwidth of the baseband signal is known, then its precise modulation bandwidth can be determined.

The simplest case of digital modulation is *amplitude shift keying* (ASK). This is binary AM with $m = 1$, where before premodulation filtering, only two levels, 0 and S_C, exist as Figure 5.5 shows. If the baseband signal is adequately filtered, then its smooth shape becomes the shape of the envelope of the modulated carrier, which looks very similar to an analog AM signal. Besides premodulation filtering, other methods of pulse shaping, for example, direct synthesis of an appropriately shaped pulse in the baseband and straightforward amplitude modulation afterwards, will lead to similar results. This is also true for other types of modulation.

In principle, the modulation bandwidth of ASK is double the maximum base bandwidth signal, but the properties of the channel filter have also to be taken into account. For instance, such a filter may have a $\cos^2$ or *raised cosine* (RC) characteristic:

$$H_{LP}(\omega) = \begin{cases} 1 & \text{for } \eta = \omega/\omega_0 < 1 - r_O \\ \cos^n\left[(\eta - 1 + r_O) \cdot \dfrac{\pi}{4r_O}\right] & \text{for } 1 - r_O \le \eta \le 1 + r_O \\ 0 & \text{else} \end{cases}$$

$$\text{with} \quad n = \begin{cases} 1 & \text{for RRC filter} \\ 2 & \text{for RC filter} \end{cases} \tag{5.29}$$

This filter is usually divided into two parts. Half of it serves as the transmitter premodulation filter and the second part is used at the receiver to reduce noise

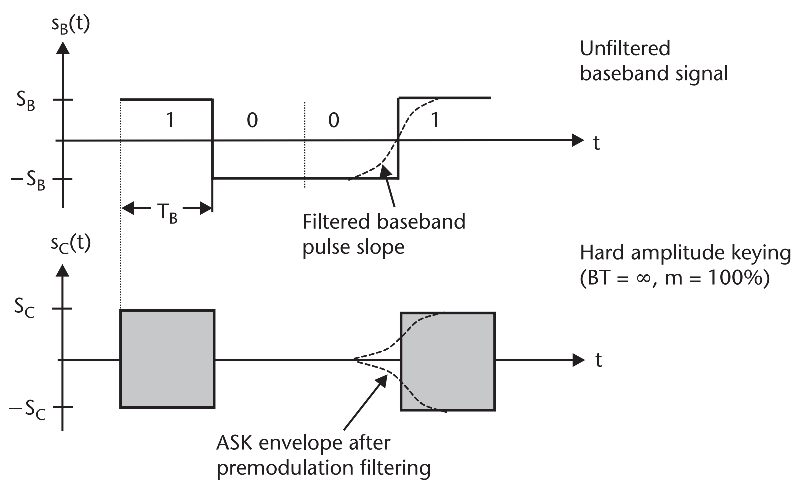


Figure 5.5 Baseband signal and ASK.

and to provide channel selection. (This strategy gives the best noise rejection, but there might be reasons for a different approach, for instance, to achieve a narrower modulation spectrum.) The $\cos^2$ filter would be replaced by two *square root raised cosine* (SRRC) filters, one in the transmitter and one in the receiver with $n = 1$ instead of 2. With SRRC premodulation filtering, the following ASK modulation bandwidth will result:

$$B_{\text{ASK}} \approx 2f_B \cdot (1 + r_O) = R_B \cdot (1 + r_O) \quad \text{with } 0 \le r_O \le 1 \tag{5.30}$$

Using a si-pulse, the narrowest bandwidth will be obtained for $r_O = 0$, but then maximum ringing occurs, while $r_O = 1$ leads to twice the bandwidth and a minimum of ringing. Hence, the roll-off factor r_O should be chosen in such a way that a good compromise between bandwidth and pulse shaping can be achieved as Figure 5.6 demonstrates [9, 10].

The inverse Fourier transform can be used to determine a single ASK signal pulse according to (4.4) and (5.5) and Table 5.1 (frequency shift) from its filtered baseband spectrum:

$$s(t)_{\text{ASK}} = \frac{1}{4\pi} \cdot \int_{-\infty}^{\infty} S(\omega) \cdot e^{j(\Omega \pm \omega)t} \, d\omega \tag{5.31}$$

Simply said the baseband pulse spectrum $S(\omega)$ occurs on both sides of the carrier with half the baseband amplitude for $m = 1$ but twice. Both parts are shifted from ω to $\Omega \pm \omega$, one in the normal and the other in the inverse position. This is the most general representation of the spectrum of an arbitrarily shaped single ASK pulse. A complete ASK transmission can, therefore, be represented by adding a number of such pulses to be transmitted one after the other with the pulse centers always separated by T_B at the instants nT_B where T_B is the bit duration:

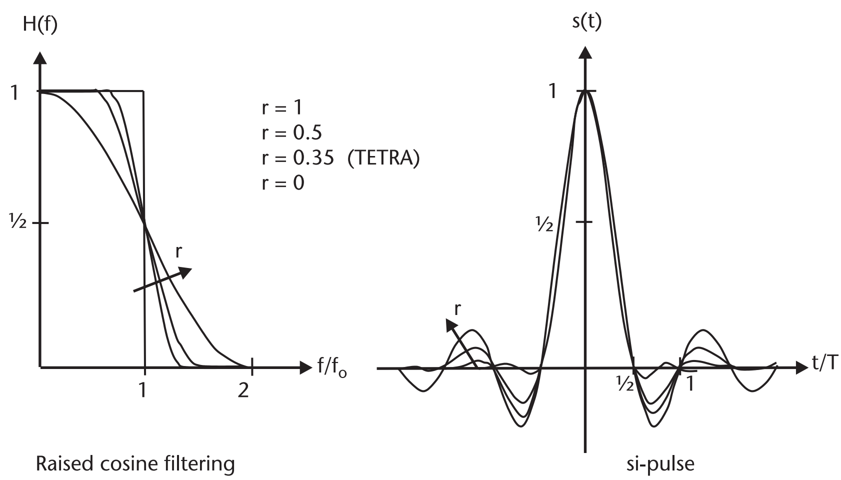


Figure 5.6 RC filtering and pulse ringing.

$$s(t)_{\text{ASK message}} = \frac{1}{4\pi} \cdot \sum_{n=1}^{N} \left[\int_{-\infty}^{\infty} S(\omega) \cdot e^{j(\Omega \pm \omega)(t - nT_B)}\, d\omega \right] \tag{5.32}$$

Keeping in mind that $1/2\pi$ comes from the Fourier transform and 1/2 from the amplitude, only the latter has to be squared to calculate the signal power according to (5.6) and we get for the adjacent channel power:

$$P_{\text{ACh}} = \frac{1}{8\pi} \sum_{n=1}^{N} \left[\int_{\omega_L}^{\omega_U} \left| S(\omega) \cdot e^{j(\Omega \pm \omega)(t - nT_B)} \right|^2 d\omega \right] \tag{5.33}$$

This is valid for the upper adjacent channel, while the lower one is situated between $-\omega_L$ and $-\omega_U$. The power in both adjacent channels is usually more or less equal and so only one of them needs to be calculated. Otherwise, the greater power has to be determined. The evaluation of such an equation by hand is unreasonable. This is the why we work with estimates wherever possible and if they are adequate. However, if a correct evaluation is really needed, then this can be done using a PC and a mathematics software package.

Modulating the frequency of the carrier digitally delivers *frequency shift keying* (FSK), which in the simplest case is a binary FM called *binary FSK* (BFSK). Basically it is hard switching from one frequency to another where one represents logic "0" and the other logic "1." The separation of these two frequencies is double the peak frequency deviation:

$$f_{0,1} = f_C \pm \Delta F_C \tag{5.34}$$

A beneficial FSK approach avoids phase jumps and relies on continuous phase progression; otherwise, the spectrum becomes more spread. Usually a frequency modulator is employed instead of a switch to ensure *continuous phase* (CP) during the transition from one frequency to the other as Figure 5.7 shows. Modulation without phase jumps usually is denoted *continuous phase modulation* (CPM) [10, 24, 25]. The modulation index h of FSK is defined similarly as for FM by the ratio of peak frequency deviation ΔF_C to maximum baseband frequency f_B or bit duration T_B or bit rate R_B, respectively:

$$h = 2\Delta F_C \cdot T_B = \frac{2\Delta F_C}{R_B} = \frac{\Delta F_C}{f_B} \tag{5.35}$$

Of course, there is also a fixed relation between the modulation index h and the phase difference between adjacent symbols of different logic contents, and in a periodic 0-1 sequence the peak phase deviation $\Delta\varphi_C$ is half the phase difference $\Delta\Phi$:

$$|\varphi_n - \varphi_{n-1}| = \Delta\Phi = \pi h \tag{5.36}$$

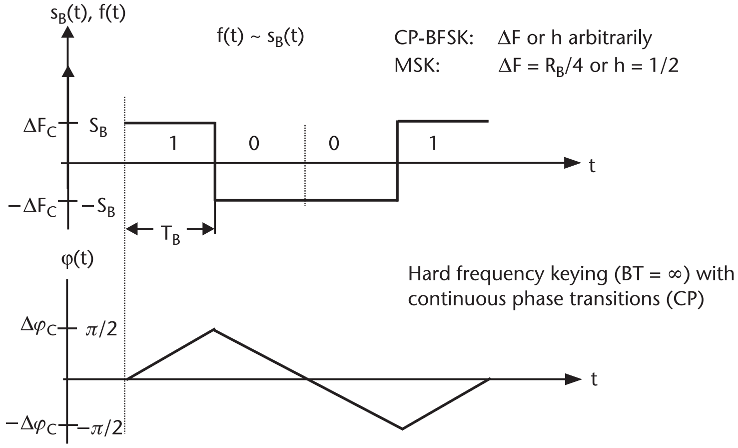


Figure 5.7 Baseband signal, CP-BFSK and MSK.

$$\Delta\varphi_C = \pi \cdot \Delta F_C \cdot T_B = \frac{1}{2} \cdot |\varphi_n - \varphi_{n-1}| = \frac{\pi h}{2} \tag{5.37}$$

MSK is a particular case of CP-BFSK. It is defined by $h = 1/2$ and, according to (5.35), is therefore $\Delta F_C = R_B/4$. MSK shows specific correlation properties that are of importance for proper demodulation and good sensitivity. In Figure 5.7 the relationships between CP-BFSK and MSK are shown.

BFSK can be regarded as two mixed ASK schemes on two frequencies f_0 and f_1 as shown in Figure 5.8. Rectangular baseband pulses exhibit a wide spectrum, and modulating them directly onto a carrier frequency results in a wide modulation spectrum irrespective of whether ASK or FSK is used. In the case of MSK there are no phase jumps but sharp corners in the phase trajectories, resulting in frequency jumps and therefore occupying a relatively wide bandwidth. Proper premodulation filtering removes these hard frequency jumps and reduces the modulation bandwidth. Moreover, the modulated carrier then exhibits automatically continuous

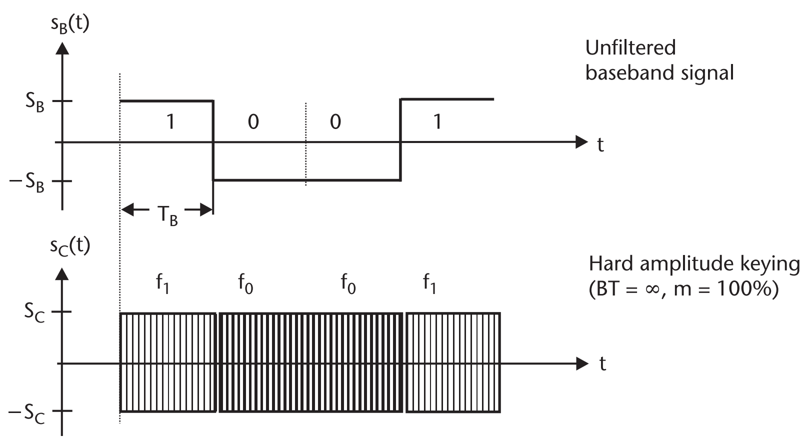


Figure 5.8 The relation between ASK and BFSK.

progression of phase *and* frequency. Hence, such schemes often are denoted *continuous frequency* (CF) modulation.

In addition to amplitude and frequency the phase of a carrier frequency can also carry modulation. In binary PSK or BPSK the two phase states are π (180°) apart from each other and the phase deviation in the case of a periodic 0-1 sequence is $\pi/2$:

$$\varphi_n - \varphi_{n-1} = \begin{cases} 0 \\ \pi \end{cases} \quad \text{and} \quad \Delta\varphi_C = \frac{\pi h}{2} \quad \text{with} \quad h = 1 \tag{5.38}$$

The easiest way to achieve BPSK is phase inversion with a simple switch. However, appropriate premodulation filtering should be applied, and because it is necessary to minimize the modulation bandwidth a phase modulator providing continuous phase transitions should be used.

Due to the nonlinearity of the constant envelope modulation process, the evaluation of the modulation bandwidth of FSK and PSK sometimes can be a difficult exercise. If, however, $B_B T_B \leq 0.5$ has been chosen, which is mostly the case and where B_B is the baseband width, then the harmonics of the baseband pulses decrease quickly as their order increases. The maximum baseband frequency of interest then is $f_B = R_B/2$. The worst case modulation bandwidth occurs if a periodic 0-1 sequence that looks similar to sinusoidal modulation is sent. For $B_B T_B \leq 0.5$ the filtered baseband signal exhibits a nearly perfect sinusoidal shape and thus an approximation using the Carson formula is convenient to calculate the modulation bandwidth of FSK and PSK.

The filtering reduces the amplitude of the maximum baseband frequency f_B and also the peak frequency deviation ΔF_C and the peak phase deviation $\Delta\varphi_C$, causing bandwidth reduction. If the cutoff frequency of the premodulation filter is equal to f_B, then this reduction is 3 dB or $1/\sqrt{2}$. Hence, the effective phase deviation $\Delta\varphi_C/\sqrt{2}$ has to be taken. Periodic 0-1 sequences are used, at least temporarily, for synchronization, for example, and therefore in many cases this worst case calculation is applicable. In practice, arbitrary 0-1 combinations usually have to be transmitted and the baseband signal can be regarded as a mix of high and low frequencies, meaning that the average modulation bandwidth is somewhat reduced. Its reduction has been found to be on the order of up to 10% for practical cases (similarly, for ASK the bandwidth of arbitrarily distributed logical states is also smaller than the case of periodic 0-1 sequences):

$$\Delta\varphi_{C_{eff}} \approx \frac{\pi h}{2} \cdot \frac{0.9}{\sqrt{2}} \approx h \tag{5.39}$$

If $B_B T_B$ is decreased further, then the effective peak frequency deviation will become even smaller. Nevertheless, we will use for quick bandwidth estimations the approximation $h \approx \Delta\varphi_C$ and arrive at the Carson formula (4.22):

$$B_{\text{FSK}} \approx 2(\Delta F_C + f_B) \approx 2f_B(h+1) = R_B \cdot (h+1) \tag{5.40}$$

PSK can be regarded as ASK or FSK to determine the modulation bandwidth depending on the characteristics of the phase transitions: CP or hard keying. A benefit of FSK and some PSK schemes is the constant envelope of the modulated carrier signal allowing the use of class C amplifiers.

Premodulation filtering does not reduce by much the modulation bandwidth in terms of 99% of the carrier power; instead it reduces much more the modulation products, which are more than 30–40 dB down from the carrier. As has already been discussed, the carrier separation in a mobile radio communication system requires a sideband power below −60 or −70 dBc in the adjacent channel. A smaller bandwidth for the region of high-order modulation products is possible with ASK, SSB, and some linear PSK schemes at the expense of a highly linearized transmitter power amplifier. However, in recent years appropriate amplifier techniques have been developed and therefore linear modulation schemes for land mobile radio systems have become feasible.

In some cases it may be interesting or necessary to evaluate the modulation bandwidth a little more accurately. In the case of FSK and unfiltered rectangular pulses, a modified Carson formula can be used to estimate the 95% or 99% power bandwidth:

$$B_M \approx 2 \cdot (\Delta F_C + n \cdot f_B) \quad \text{with} \quad n = \begin{cases} 3 & \text{for } 95\% \\ 9 & \text{for } 99\% \end{cases} \text{ of the total power} \tag{5.41}$$

The precise calculation of the modulation bandwidth in the case of a periodic but nonsinusoidal signal is difficult. If the baseband signal can be represented by a Fourier sequence a PM signal would look like

$$s(t) = S_C \cdot \cos\left(\Omega t + \sum_{i=1}^{I} \Delta\varphi_i \cdot \sin \omega_i t\right) \tag{5.42}$$

The related frequency domain representation is much more complicated [26]:

$$s(t) = S_C \cdot \sum_{k_i=-\infty}^{\infty} \left[\prod_{i=1}^{I} J_{k_i}(\Delta\varphi_i)\right] \cdot \cos\left\{\left(\Omega + \sum_{i=1}^{I} i\omega_i\right) \cdot t\right\} \tag{5.43}$$

This equation is not easy to evaluate, however, for small arguments, as used in mobile radio communication systems, the Bessel functions decrease rapidly with increasing order. Consequently, their products will decrease even faster. As the simplest case only two signal frequencies are taken into account:

$$s(t) = S_C \cdot \cos(\Omega t + \Delta\varphi_1 \cdot \sin \omega_1 t + \Delta\varphi_2 \cdot \sin \omega_2 t) \tag{5.44}$$

$$= S_C \cdot \sum_{k=-\infty}^{\infty} \sum_{i=-\infty}^{\infty} J_k(\Delta\varphi_1) \cdot J_i(\Delta\varphi_2) \cdot \cos[(\Omega + k\omega_1 + i\omega_2) \cdot t] \tag{5.45}$$

Investigating this more in detail, we find for the carrier signal at Ω and for all sidebands related to ω_1 and to ω_2:

$$s_C(t) = S_C \cdot J_0(\Delta\varphi_1) \cdot J_0(\Delta\varphi_2) \cdot \cos(\Omega t) \tag{5.46}$$

$$\sum_{k=-\infty}^{\infty} s_{1k}(t) = S_C \cdot \sum_{k=-\infty}^{\infty} J_k(\Delta\varphi_1) \cdot J_0(\Delta\varphi_2) \cdot \cos[(\Omega + k\omega_1) \cdot t] \tag{5.47}$$

$$\sum_{i=-\infty}^{\infty} s_{2i}(t) = S_C \cdot \sum_{i=-\infty}^{\infty} J_i(\Delta\varphi_2) \cdot J_0(\Delta\varphi_1) \cdot \cos[(\Omega + i\omega_2) \cdot t] \tag{5.48}$$

Finally sidebands exist due to the combinations of ω_1 and ω_2:

$$\sum_{k=-\infty}^{\infty} \sum_{i=-\infty}^{\infty} s_{1k,2i}(t) = S_C \cdot \sum_{k=-\infty}^{\infty} \sum_{i=-\infty}^{\infty} J_k(\Delta\varphi_1) \cdot J_i(\Delta\varphi_2) \cdot \cos[(\Omega \pm k\omega_1 \pm i\omega_2) \cdot t] \tag{5.49}$$

For small arguments, that is, $\Delta\varphi_{1,2} << 1$, there is $J_1 < J_0$ and $J_{k,i} << J_0$ and $J_k \cdot J_i << J_{k,i}$. Consequently, the resulting spectrum can be approximated by the addition of the spectra of the original signals because all combination of ω_1 and ω_2 lead to significantly smaller amplitudes than the modulation products of the original signal frequencies. Hence, the influence of pulse shaping can be estimated by adding the spectra of single harmonics of the baseband signal. The same reasoning can be applied to the Fourier transform.

Gaussian-filtered MSK (GMSK) is a modulation scheme with widespread application in mobile radio communications. Therefore, it has been taken as an example of how to proceed if proper and complete mathematical evaluation of a modulated carrier signal really is needed. The same procedure can be applied to every type of modulation process to get the resulting signal and its representation in the time and frequency domains [27]. The Gaussian lowpass filter is characterized by its response $h_G(t)$ to a unity impulse or Dirac pulse:

$$h_G(t) = \frac{1}{\sqrt{2\pi} \cdot \sigma T_B} \cdot e^{-(t^2/2\sigma^2 T_B^2)} \quad \text{where} \quad \sigma = \frac{\sqrt{\ln 2}}{2\pi \cdot B_B T_B} \tag{5.50}$$

Here a rectangular impulse according to (5.11) with duration T_B and unit amplitude-time area represents a symbol or element of a longer message. The modulated signal in the time domain can be determined by the convolution of the message and the filter response:

$$g(t) = h_G(t) * \operatorname{rect}\left(\frac{t}{T_B}\right) \quad \text{with} \quad \int_{-\infty}^{\infty} g(t)\, dt = 1 \tag{5.51}$$

The next step is to add all of the filtered pulses one after the other to get the representation of the carrier phase in the time domain. The modulated frequency of the carrier can be derived from the phase by differentiation because $\omega = d\varphi/dt$:

$$\varphi(t) = \varphi_0 + \sum_{n=1}^{N} a_n \cdot \pi h \cdot \int_{-\infty}^{t-nT} g(\tau)\, d\tau \quad \text{with} \quad a_n \in \{+1, -1\} \tag{5.52}$$

If $h = 1/2$, then GMSK is the result; otherwise, it would be Gaussian FSK or GFSK. The sequence of symbols represents the complete message and the modulated carrier signal can now be written as follows:

$$s(t) = S_C \cdot \cos\left[\Omega t + \varphi_0 + \sum_{n=1}^{N} a_n \cdot \pi h \cdot \int_{-\infty}^{t-nT} g(\tau)\, d\tau\right] \tag{5.53}$$

The whole procedure looks somewhat complicated, but nowadays it can be performed on a PC. A Fourier transform according to (5.3) or (5.5) delivers the spectrum of the modulated carrier and the result is shown in Figure 5.9.

The graph shows only one side of the spectrum but of course it looks the same on both sides of the carrier frequency. The width of the modulation spectrum is determined by $B_B T_B$. The unfiltered digital baseband signal characterized by $B_B T_B = \infty$ is plain MSK, which exhibits a very large bandwidth. The diagram shows that the modulation bandwidth at –20 dBc is not affected very much by the premodulation filtering. Of course, there is a slight reduction due to the approach to the filter cutoff frequency. However, the situation is completely different at, for example, –60 dBc. There the modulation sidebands can be considerably reduced by appropriate filtering.

Besides Gaussian premodulation filtering or RC shaping, other methods have been investigated to achieve steeper spectrum slopes. Two of these are *tamed frequency modulation* (TFM) and *generalized TFM* (GTFM), which have a modulation index of $h = 1/2$ like GMSK. The specific two-stage filtering of GTFM permits

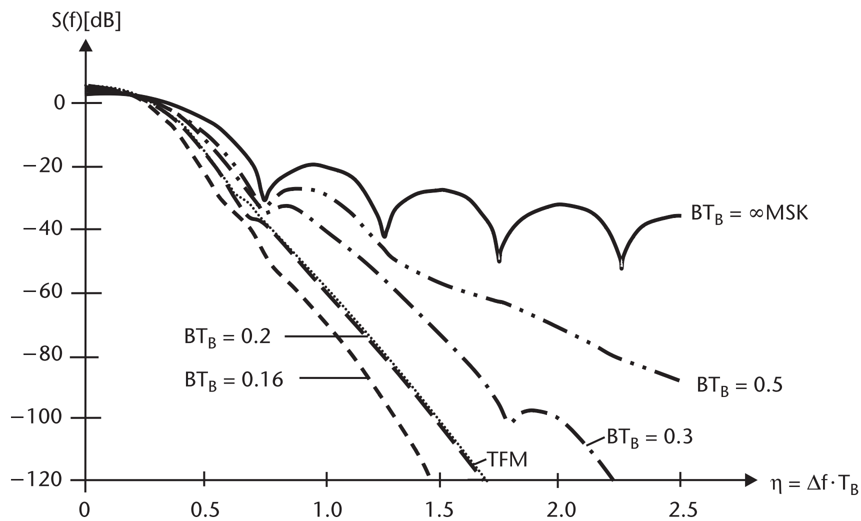


Figure 5.9 Modulation spectrum of GMSK. (*After:* CCIR, 1982.)

different combinations of parameters to get the desired spectral performance. Figure 5.9 shows that the spectra of TFM and GMSK with $B_B T_B = 0.2$ are very similar [1, 6, 12, 17, 28–31].

For those readers who are not familiar with radio equipment concepts, a brief explanation follows as to why the modulation bandwidth of the transmitter is restricted by premodulation filtering and not by filtering on the carrier frequency. The reason is simply that premodulation filtering requires a fixed baseband filter, whereas RF filtering needs a tunable filter with an extremely narrow bandwidth relative to the carrier, which must be shifted if another radio channel is accessed. Such filters are extremely expensive or virtually impossible to build. It is the same reason why channel filtering in the receiver is not performed on the received carrier frequency but on a fixed *intermediate frequency* (IF) or in the baseband if a direct conversion receiver is used.

5.4 Multilevel and Partial Response Modulation Schemes

Multilevel modulation and partial response schemes are other means to reduce the modulation bandwidth besides premodulation filtering. Higher level modulation with a lower symbol rate for a given user bit rate results in a more vulnerable modulation but also allows the application of a stronger error correction scheme. Within certain limits the result is an overall improvement as Ungerboeck showed a long time ago [32, 33]. Additionally, interesting facts concerning bandwidth and receiver performance for various FSK schemes with or without partial response have been presented by Aulin et al. [1, 6, 17, 24, 25, 34, 35].

Multilevel modulation means that, in the case of mFSK, m distinct frequencies are used instead of only two. This can be regarded also as one carrier frequency modulated with m distinct FSK levels. In principle, m can have any value, but preferably a power of 2 is used to obtain a simple relationship between the number of modulation levels and the number of bits one symbol can carry:

$$m = 2^q \quad \text{and} \quad q = \log_2 m \quad \text{where } q \geq 1 \text{ and an integer} \tag{5.54}$$

For any type of modulation, a simple relationship between bit and symbol energy E_B and E_S, bit and symbol rate R_B and R_S, and carrier power P_C is valid and there is also a self-explanatory relation between the baseband frequency f_B, the symbol rate R_S, and the symbol duration T_S:

$$P_C = R_B \cdot E_B = R_S \cdot E_S \quad \text{with} \quad R_B = qR_S \quad \text{and} \quad E_S = qE_B \tag{5.55}$$

$$f_B = R_S/2 = 1/2T_S = R_B/2q = 1/2qT_B = \omega/2\pi \tag{5.56}$$

For bandwidth calculations T_B and R_B have to be replaced by T_S and R_S in the equations for BFSK. Half the frequency difference between two adjacent frequency levels is ΔF_{step}, which in the case of BFSK becomes equal to the peak frequency deviation ΔF_C. The modulation index can be expressed in terms of peak and step frequency deviation. Likewise, the phase difference $\Delta\Phi$ between adjacent symbols with extreme modulation levels φ_m and φ_1, the carrier peak phase deviation $\Delta\varphi_C$, and the modulation index h are linked:

$$h = \frac{2\Delta F_C \cdot T_S}{m-1} = 2\Delta F_{\text{step}} \cdot T_S \tag{5.57}$$

$$\Delta\Phi = 2\Delta\varphi_C = |\varphi_{m,n} - \varphi_{1,n-1}| = (m-1) \cdot \pi h \tag{5.58}$$

One of the most important m-ary modulation schemes in mobile radio communications is four-level FSK, which combines sufficient sensitivity and CCR with an increased transmission rate. Additionally, the symbol duration is longer than in the BFSK case for the same data rate and, therefore, multipath propagation with long delays is less critical. Its benefit is a constant envelope and the freedom to choose any appropriate modulation index in the case of noncoherent demodulation. However, coherent demodulation provides a better sensitivity but needs good carrier recovery and a well-defined fixed modulation index is preferable, for example, $h = 1/3$ or $1/2$. If pulse shaping is performed by a Gaussian lowpass premodulation filter the result is Gaussian-filtered FSK or 4GFSK. If mGFSK is regarded, then (5.52) and (5.53) remain valid except for the set of logic symbols, which will become with the help of (5.54):

$$a_n \in \{\pm 1\} \xrightarrow{2 \to m} \{\pm 1, \pm 3 \ldots \pm(m-1)\} \tag{5.59}$$

Figure 5.10 shows an example of four-level FSK. In the drawing a Gaussian pulse shaping with $B_B T_B \approx 1/2$ is assumed. This looks similar to a raised cosine shaping with $n = 2$ (2RC) where n determines the baseline length of the pulse, which is nT_B or nT_S, respectively.

Each symbol can have one of four possible states 00 via 01 and 10 to 11 and therefore represents 2 bits (a *dibit*). However, if for instance the symbol 01 is confused with the adjacent 10 due to noise or interference, this means that 2 bits

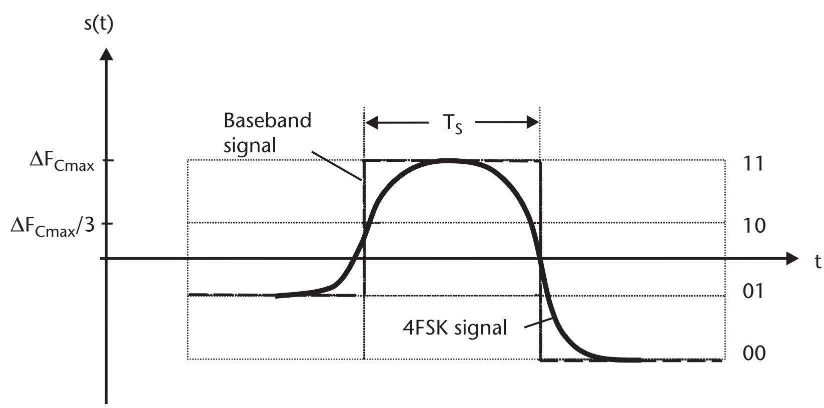


Figure 5.10 Pulse shaping for 4FSK.

are in error. This may also happen to modulation schemes with a higher number of levels. To avoid this problem, Gray coding is used in which adjacent levels differ by only 1 bit, and this is achieved by clever reordering between the modulation levels and the bit assignment [23]. Table 5.4 shows Gray codes for up to 16 modulation levels. Symbol No. 1 appears twice to show the problem more clearly for PSK schemes where both become adjacent symbols on a circle in contradiction to the linear sequence for FSK. In case of 4FSK, it is merely necessary to simply exchange levels 00 with 01 or 10 with 11 because in FSK schemes the outer level usually cannot be confused. In a four- or eight-state PSK scheme, however, this is a serious problem.

If the occupied bandwidth of mFSK signals is considered, then there are two contradicting relations provided that the modulation index and the bit rate are kept constant. One is that the bandwidth increases with the number m of modulation levels and the other is that it will be reduced because the symbol rate is decreasing. Which effect counts for more? To decide that, the bandwidth formula has to be rewritten appropriately. Since for our problem an approximation suffices, premodulation filtering with $B_B T_S = 1/2$ or 2RC pulse shaping is assumed and therefore the Carson formula with $\Delta\varphi_{\text{eff}} \approx h$ can be used. From (5.40) and (5.54) to (5.57) the desired relation, which delivers (5.40) for $q = 1$ meaning $m = 2$, can now be derived:

$$\frac{B_{m\text{FSK}}}{R_B} \approx \frac{1}{q} \cdot [(2^q - 1) \cdot h + 1] \quad \text{for} \quad h \geq 1 \tag{5.60}$$

However, now we are interested in relatively small modulation indices, and the formulas (4.21) and (4.26) from Chapter 4 that we used to calculate the modulation bandwidth B_M and the adjacent channel power bandwidth B_{ACh} do not work with sufficient accuracy. Hence, Table 5.5 has been calculated for the typical small modulation indices usually employed for digital modulation. Its

Table 5.4 Gray Coding

Symbol No.	*Binary Coded*	*Binary Bit Difference*	*Gray Coded*	*Gray Coded Bit Difference*
1	0000	—	0000	—
2	0001	1	0001	1
3	0010	2	0011	1
4	0011	1	0010	1
5	0100	3	0110	1
6	0101	1	0111	1
7	0110	2	0101	1
8	0111	1	0100	1
9	1000	4	1100	1
10	1001	1	1101	1
11	1010	2	1111	1
12	1011	1	1110	1
13	1100	3	1010	1
14	1101	1	1011	1
15	1110	2	1001	1
16	1111	1	1000	1
1	0000	4	0000	1

Table 5.5 Squared Bessel Functions of the First Kind and pth Order (in –dBc)

	$10 \cdot \log_{10}\{J_p^2(h)\}$ [–dBc]							
h	$p = 0$	1	2	3	4	5	6	7
0	≡ 0	∞	∞	∞	∞	∞	∞	∞
1/4	0.1	18.1	42.2	69.8	>80	>80	>80	>80
1/3	0.2	15.7	37.2	62.3	>80	>80	>80	>80
1/2	0.6	12.3	30.3	51.8	75.9	>80	>80	>80
2/3	1.0	10.0	25.4	44.4	66.0	>80	>80	>80
3/4	1.3	9.1	23.5	41.4	61.9	>80	>80	>80
1	2.3	7.1	18.8	34.2	52.1	72.0	>80	>80
5/4	3.8	5.8	15.3	28.7	44.6	62.6	>80	>80
4/3	4.4	5.5	14.4	27.1	42.5	59.8	78.8	>80
3/2	5.8	5.1	12.7	24.3	38.6	54.9	72.8	>80
5/3	7.6	4.8	11.3	21.8	35.2	50.5	67.5	>80
7/4	8.7	4.7	10.6	20.7	33.6	48.5	65.1	>80
2	13.0	4.8	9.0	17.8	29.4	43.0	58.4	75.1

evaluation gives a slightly modified linear approach with $a = 5/4$ and $b = 1$ for the normalized modulation bandwidth B_M. However, for the adjacent channel power bandwidth B_{ACh} a nonlinear approximation fits much better:

$$\frac{B_M}{R_B} \leq \frac{1}{q} \cdot \left[\frac{5}{4} \cdot (2^q - 1) \cdot h + 1\right] \quad \text{for} \quad 0 \leq h \leq 2 \tag{5.61}$$

$$\frac{B_{\text{ACP}}}{R_B} \leq \frac{1}{q} \cdot \left[\frac{5}{\sqrt{2}} \cdot (2^q - 1) \cdot \sqrt{h} + 1\right] \quad \text{for} \quad 0 \leq h \leq 2 \tag{5.62}$$

Now we can investigate the trade-off between an increasing number of modulation levels and the decreasing symbol rate in greater depth. The necessary normalized channel separation can now be calculated similar to the calculation for (4.25). For simplicity, transmitter and receiver frequency tolerances can be neglected because usually they are much smaller than the modulation bandwidth:

$$\frac{\Delta F_{\text{Ch}}}{R_B} \geq \frac{1}{2} \cdot [B_M + B_{\text{ACh}}] \tag{5.63}$$

At first a modulation index of $h = 1/2$ is widely used and a varying number of modulation levels are considered, for which the results are given in Table 5.6. From there, it follows that the modulation bandwidth is acceptable for $m = 1$ to 3 or BFSK to 8FSK, but unfortunately the adjacent power bandwidth increases quickly

Table 5.6 Modulation and Adjacent Channel Power Bandwidth for mFSK with $h = 1/2$

q	1	2	3	4	5	6
m	**2**	**4**	8	16	32	64
B_M/R_B	**1.62**	**1.44**	1.79	2.59	4.08	6.73
B_{ACP}/R_B	**3.50**	**4.25**	6.17	9.63	15.70	26.42
$\Delta F_{\text{Ch}}/R_B$	**2.56**	**2.85**	3.98	6.11	9.89	16.58

with an increasing number of modulation levels. BFSK and 4FSK provide the best compromise and, therefore, these modulation schemes have been chosen for many DPMR systems. Admittedly, the channel separation is slightly worse for 4FSK, but the longer symbols result in less vulnerability against multipath propagation distortions.

In Chapter 4 we showed that for hard clipping the unwanted phase deviation cannot exceed $\Delta\varphi_C = \pi/2$ as long as the wanted signal amplitude exceeds that of the interferer. If the phase deviation of the modulation is larger than $\pi/2$, then the static CCR must become 0 dB. Due to (5.39), therefore, $h \geq 3/2$ must be fulfilled. Now the question about the most beneficial mFSK modulation can be looked at again, but under the assumption of improved robustness against multipath interference; the results for $h = 3/2$ are listed in Table 5.7. As expected the bandwidth increases much more quickly with an increase of modulation levels as in the previous case. Again, BFSK and 4FSK show the best results but the higher levels become much worse.

It has been shown in Chapter 3 that for unaffected frequency economy an increased bandwidth might be compensated by an improved CCR. If a static CCR of 0 dB is assumed, then for realistic cases a dynamic CCR of 9 dB will result and with shadowing about 13 dB has to be assumed. Table 5.7 delivers a necessary channel separation of 50 kHz for 4FSK systems with h = 3/2 and a maximum possible bit rate of 10 kbit/s (that is the order of many DPMR systems) and with a $B_B T_S$ = 1/2 or 2RC shaping. With these assumptions the cluster size and the number of channels can be calculated from (3.99) and (3.100). The result is $N_C = 3.45$ (in practice, 4 should be used) and $N_I = 5.79$ channels/MHz. Comparing this result with all others shown in Table 3.11, this is close to values for some early DPMR systems, but this hypothetical system would be based on constant envelope modulation and very simple bit-by-bit demodulation and would under most conditions not necessarily need an equalizer. So it could be made cheaper than many other and more sophisticated DPMR systems, provided that the same market figures could be achieved. Also, partial response modulation would make it a little more complex but would allow a reduced channel separation of, say, 25 kHz and only demodulation and decoding would be more complex. But this additional expenditure would only affect the software with negligible influence on production costs.

Bandwidth reduction is possible also for PSK not only by premodulation filtering but by also selecting multilevel modulation. A good trade-off between bandwidth and vulnerability for PSK schemes again is the four-level approach as with FSK. *Quadrature PSK* (QPSK) has four phase states and the phase deviation and the modulation index are only half of that of BPSK:

Table 5.7 Modulation and Adjacent Channel Power Bandwidth for mFSK with h = 1/2

q	1	2	3	4	5	6
m	**2**	**4**	8	16	32	64
B_M/R_B	**2.88**	**3.31**	4.71	7.28	11.83	19.85
B_{ACP}/R_B	**5.33**	**7.00**	10.44	16.49	27.05	45.63
$\Delta F_{Ch}/R_B$	**4.11**	**5.16**	7.58	11.89	19.44	32.74

$$\varphi_n - \varphi_{n-1} = \begin{cases} 0 \\ \pm\pi/2 \\ \pm\pi \end{cases} \quad \text{and} \quad 2\Delta\varphi = \pi h \quad \text{with} \quad h = 1/2 \tag{5.64}$$

For ordinary mPSK schemes, all phase states are situated on the circumference of a circle and therefore the following relation results:

$$\Delta\Phi = \varphi_n - \varphi_{n-1} = \pi h = \frac{2\pi}{m} \to h = \frac{2}{m} \tag{5.65}$$

As with BPSK and also for QPSK, the transition from one symbol to the other is performed in the phase state diagram by the shortest route, meaning that sometimes the origin will be passed through. These hard phase jumps create a wide modulation spectrum. With premodulation filtering the phase transitions become softer but sometimes they will still pass through the origin of the phase constellation diagram, which results in large envelope fluctuations.

Thus, ordinary mPSK schemes do not possess a constant envelope and therefore have to be treated as linear modulation schemes. In fact, they can be interpreted as a combination of PSK and ASK. QPSK is therefore often called *quadrature ASK* (QASK) or *quadrature AM* (QAM). The benefit is a modulation bandwidth smaller than that of comparable constant envelope schemes, but the drawback is that the transmitter power amplifier must be extremely linear [1, 12]. In contrast, many types of mFSK and some kinds of mPSK exhibit a constant envelope characteristic similar to analog FM and PM. Like those, they occupy a somewhat larger modulation bandwidth than the corresponding linear modulation schemes due to the Bessel functions involved, but no linear transmitter power amplifier is needed.

Reducing the modulation bandwidth of GMSK or mGFSK by decreasing the premodulation filter cutoff frequency to $B_B T_S < 0.5$ leads to built-in ISI, which complicates the demodulation process. The pulse shaping can also be done differently, for example, by selection of a $\cos^2$ or raised cosine shape (nRC) with $n = 3$ to 4. Those systems are called *partial response systems* because every symbol's response falls not only into its own sampling time but also into those of the adjacent symbols [25]. On the other hand, the pulse shape and the amount to be added to the adjacent symbols is well known by definition and can be corrected for during the demodulation process. In these cases there is no longer the possibility of bit-by-bit demodulation, but with more sophisticated demodulation methods at least the sensitivity can be nearly preserved but CCR may be somewhat reduced. Hence, once again for the design engineer it is no easy task to find the best trade-off between the different parameters.

For practical systems utilizing GMSK, $0.25 \le B_B T_B \le 0.5$ is chosen for the most part. For DECT it is 0.5, meaning that there is almost no ISI, which permits simple noncoherent bit-by-bit detection. In contrast, the GSM system employs $B_B T_B = 0.3$. A similar example is the 4GFSK of DIIS with $B_B T_S = 0.28$ and $h = 1/3$. With a value of 0.25 TETRAPOL employs GMSK with the smallest $B_B T_B$.

5.5 Sophisticated PSK Techniques

What is needed is a compromise between constant envelope modulation and linear schemes to combine the robustness against nonlinear distortions of the first with the small modulation bandwidth of the latter. Different approaches are used to find a trade-off that results in an acceptable expenditure.

OQPSK is a specific variation of QPSK. It can be composed of two independent BPSK schemes such as ordinary QPSK, but for OQPSK both BPSK schemes are shifted by half a symbol duration relative to each other. This avoids the nulls that simple QPSK exhibits and, therefore, the amplitude variations diminish. The smaller the amplitude variation, the smaller the transmitter linearity requirement.

In $\pi/4$-shift QPSK the four reference phase states are shifted from symbol to symbol by $\pi/4$. This means that eight phase states exist, for even symbols only four of them are valid and for odd symbols the other four. Hence, the phase change never passes through the origin of the phase constellation diagram. This type of modulation was proposed as early as 1962 by P. A. Baker [19, 21, 22, 36, 37].

If a fixed relation to a reference phase is not needed, then a differential modulation would be appropriate [38]. This means that the symbols are not transmitted, only their differences. Hence, there is no need to recover a reference phase in the receiver; however, each symbol error destroys 2 bits. Thus, the bit error rate is doubled.

The principle of differential coding is shown in Figure 5.11. Differential coding can be performed by a simple logic function between two adjacent bits. It is the "exclusive or" function (XOR), which is equivalent to a binary addition or subtraction:

$$
\begin{aligned}
&\text{Original bit stream:} && a_n \in \{0, 1\} \\
&\text{Transmitted bit stream:} && b_n = a_n \oplus b_{n-1} \\
&\text{Decoded bit stream:} && c_n = b_n \oplus b_{n-1}
\end{aligned}
\tag{5.66}
$$

In the receiver the original bit stream is reconstructed by a second XOR operation between adjacent bits. The proof that this works properly is simple:

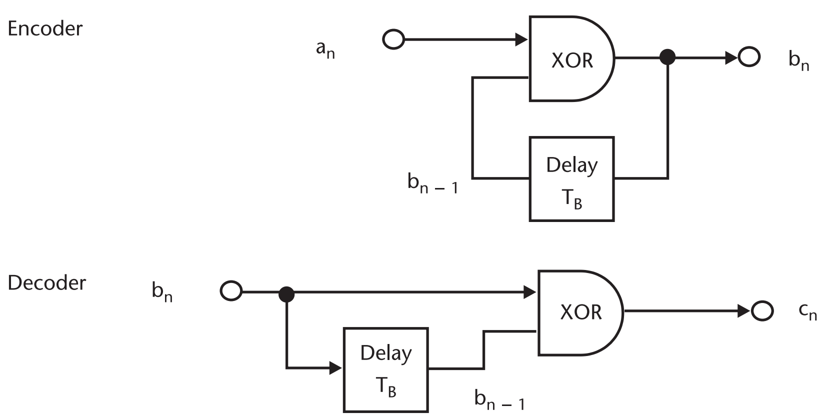


Figure 5.11 Differential coding.

$$c_n = (a_n \oplus b_{n-1}) \oplus b_{n-1} = a_n \oplus (b_{n-1} \oplus b_{n-1}) = a_n \qquad (5.67)$$

In Table 5.8 the coding, transmission, and decoding of a certain bit sequence a_n are shown. The value b_n of the first bit can be chosen arbitrarily because there is no preceding bit b_{n-1}. It is assumed that during the transmission two single-bit errors occur (shown in bold italic in the table) producing the faulty sequence b_n^*, which is decoded to the faulty sequence c_n^* (instead of c_n) containing two *double*-bit errors instead of two *single*-bit errors.

The differential coding of QPSK states together with the $\pi/4$-shift between the even and the odd symbols is called *π/4-shift DQPSK* or shorter *π/4-DQPSK*. This is the modulation scheme employed by the American and Japanese digital cellular systems D-AMPS and JDC, respectively, and by TETRA. Of course, additional pulse shaping is necessary. In TETRA this is achieved by premodulation filtering using a RRC filter with a roll-off factor of $r_O = 0.35$. The drawings in Figure 5.12 show the composition of a $\pi/4$-DQPSK signal from the eight phase states where the phase transitions do not pass the origin of the phase constellation diagram [6, 12, 17, 19, 21, 22, 34, 35, 37].

It is interesting to compare the modulation bandwidth of linear $\pi/4$-shift DQPSK with that of 4GFSK exhibiting a constant envelope, that is, those of TETRA and DIIS. Figure 5.13 shows that the first exhibits a much steeper decrease of the

Table 5.8 Example for Error Progress with Differential Coding

n	0	1	2	3	4	5	6	7	8	9	10	11	12
a_n	—	0	***1***[1]	0[1]	0	1	1	0	0	0	***1***[1]	1[1]	1
b_n	0[2]	0	1	1	1	0	1	1	1	1	0	1	0
c_n	—	0	***1***[1]	0[1]	0	1	1	0	0	0	***1***[1]	1[1]	1
b_n^*	0[2]	0	***0***[3]	1	1	0	1	1	1	1	***1***[3]	1	0
c_n^*	—	0	***0***[1,4]	***1***[1,4]	0	1	1	0	0	0	***0***[1,4]	***0***[1,4]	1

[1]Bold italic bits show the critical parts of the bit sequences.
[2]Arbitrarily chosen.
[3]Transmission error $b_n^* \neq b_n$.
[4]Double-bit errors in c_n^* after decoding.

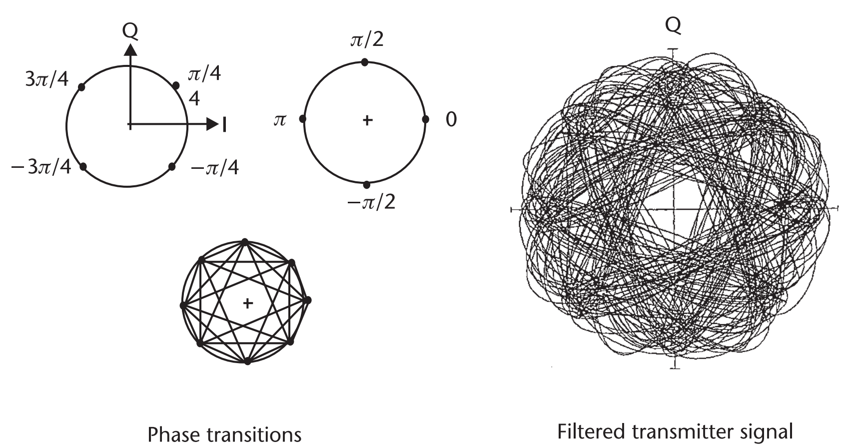


Figure 5.12 The phase constellation diagram of $\pi/4$-shift DQPSK.

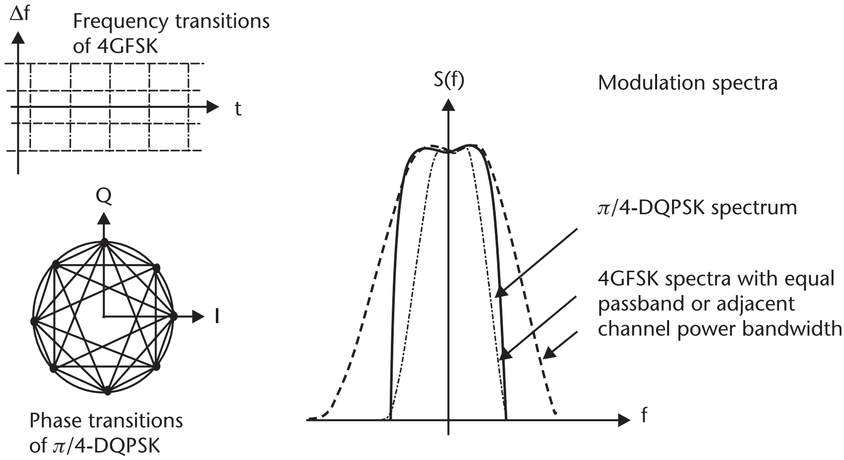
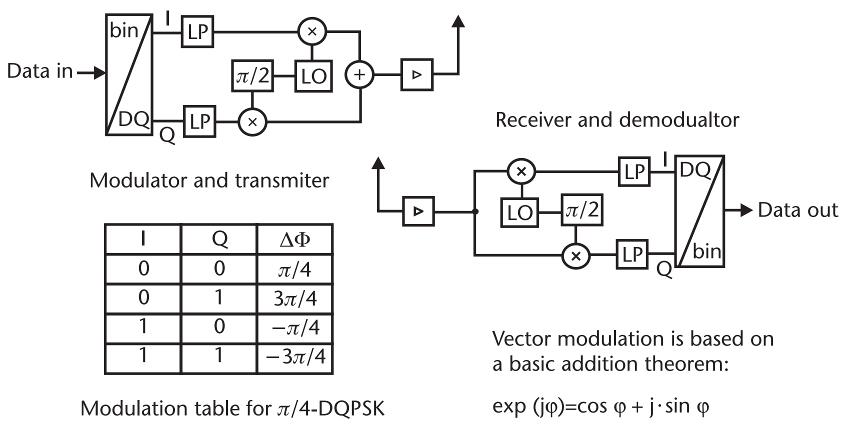


Figure 5.13 Modulation bandwidth comparison of $\pi/4$-shift DQPSK and 4GFSK.

modulation sidebands than 4GFSK for the same bit rate, which results in an increased adjacent channel power level for the latter. If the same adjacent power limit has to be met, then the bit rate of 4GFKS has to be lower than that of $\pi/4$-shift DQPSK, provided the premodulation filtering in all cases is comparable.

Every instant of a sinusoidal signal can be defined in the complex plane by its magnitude and phase. Likewise it can be composed from a real and an imaginary part. Hence, a modulator can be built by shifting the RF carrier signal by 90° or $\pi/2$, multiplying the original and the shifted carrier signal by the wanted amplitude and adding both. This is equivalent to the addition of two vectors and, therefore, this method is called vector or IQ modulation where I is the in-phase and Q the quadrature component of the complex signal. Figure 5.14 shows how such a vector modulator works. For TETRA modulation a table is given that shows the relation between the phase states and the I and Q signals.

Because a vector modulator can control the amplitude and phase of a signal simultaneously, it can be used for any kind of amplitude, frequency, or phase



I	Q	ΔΦ
0	0	$\pi/4$
0	1	$3\pi/4$
1	0	$-\pi/4$
1	1	$-3\pi/4$

Modulation table for $\pi/4$-DQPSK

Figure 5.14 IQ or vector modulation.

modulation in any combination and is therefore well suited to all kinds of analog and digital modulation. Moreover, such a device can also be used as a demodulator to reconstruct the I and Q components of the received signal. For this purpose specific synchronization procedures requiring dedicated synchronization circuitry are used. Coherent demodulation delivers the best results, but perfect synchronization is the necessary precondition that requires us to recover the frequency and phase of the unmodulated carrier with negligible error. However, the performance of coherent demodulation drops dramatically with increasing phase or frequency error. On the other hand, synchronization is always based on clock evaluation of a certain number of symbols. In a noisy channel, each single symbol may show a significant random phase jitter, but by averaging over a number of symbols the jitter is considerably reduced. During the detection process, symbol impairments have therefore much more effect than sample clock errors.

An ordinary noncoherent frequency or phase demodulator is usually based on a simple frequency-to-voltage conversion as is done in conventional FM receivers by a discriminator. The performance is roughly 2 dB worse compared to coherent detection when there is no frequency error. However, even large frequency errors introduce only a small degradation as long as the spectrum of the received signal is not impaired by one of the IF filter slopes. In practice, the break-even point is at maximum of several hundred hertz, which is on the same order of a possible Doppler shift. Hence, under unfavorable propagation conditions, including high and fast changing Doppler shift, a noncoherent demodulator might be preferable to coherent demodulation [38].

Many other PSK modulation schemes have been designed for various purposes and the diagrams in Figure 5.15 show their phase states and phase transitions. We review them briefly because some of them are also used in mobile radio communications.

QAM schemes with more than four states are not commonly used in mobile radio systems because mQAM provides a small modulation bandwidth with high bit rates at the expense of reduced receiver sensitivity and CCR. For mobile applica-

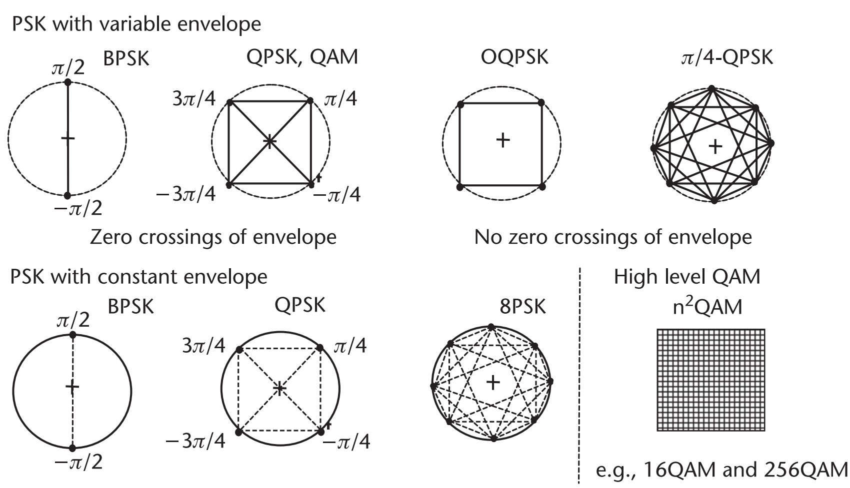


Figure 5.15 Different kinds of PSK.

tions, therefore, the number of modulation levels has to be restricted and currently Motorola's iDEN system is the only example for $m > 4$ (16QAM). However, in fixed radio links mQAM schemes with $m = 256$ and even higher levels permit very high bit rates because stationary applications are not impaired by severe and rapidly changing multipath conditions and barely sufficient receiver input power.

The mPSK with a large number of levels is also not very appropriate for mobile radio applications. One of the latest applications of 8PSK is EDGE for GSM, which offers a reasonable trade-off between the benefits and drawbacks of mPSK when the channel properties are not too severe. EDGE uses a $3\pi/8$-shift between even and odd symbols to avoid zero crossings in the phase constellation diagram to reduce the linearity requirements, similar to TETRA with its $\pi/4$-shift. The bit rate is tripled compared to binary GMSK modulation. However, due to the reduced phase difference between adjacent symbols ($\pi/4$ instead of $\pi/2$), this modulation is more vulnerable and good propagation conditions are necessary. If it fails under bad conditions the modulation is switched back to GMSK.

A specific approach to digital radio transmission is subcarrier modulation. One of the most widespread subcarrier modulation schemes is the 1,200-bit/s system defined, for example, in MPT 1327 and ETS 300 230 (BIIS 1200). It employs switching between 1,200 and 1,800 Hz at the zero crossings. Thus, it is a CP scheme comprising one period of 1,200 Hz or one and a half periods at 1,800 Hz per symbol where the latter represents the logic "0." Therefore, it has been called *fast FSK* or *FFSK* but in fact it is plain MSK because the subcarrier frequency deviation $\Delta F_{SC} = 300$ Hz is a quarter of the bit rate $R_B = 1{,}200$ bit/s. Figure 5.16 visualizes the double modulation process needed.

At first, the digital information is modulated onto a subcarrier creating MSK with two sidebands. This subcarrier modulation fits into the AF passband of analog mobile radios. Then the whole passband together with this digital subcarrier signal is modulated onto the RF carrier by analog FM or PM, again creating a number of sidebands. The total modulation spectrum around the RF carrier looks even much more complicated than that shown in Figure 5.16 where the higher sidebands of the second modulation process have been suppressed for clarity.

Assuming $B_B T_B \approx 0.5$ due to the AF lowpass filter (in reality, it is often much higher), the occupied modulation bandwidth is $B_{SC} \approx 2 \cdot (\Delta F_{SC} + R_B/2) = 1.8$ kHz symmetrically to the subcarrier frequency $f_{SC} = 1.5$ kHz. Modulating this subcarrier signal by FM onto an RF carrier with a peak deviation of 2.0 kHz at a 12.5-kHz channel separation the final modulation bandwidth after this double modulation process is $B_{M\,SC} \approx 8.8$ kHz. A direct carrier modulation with the same peak deviation

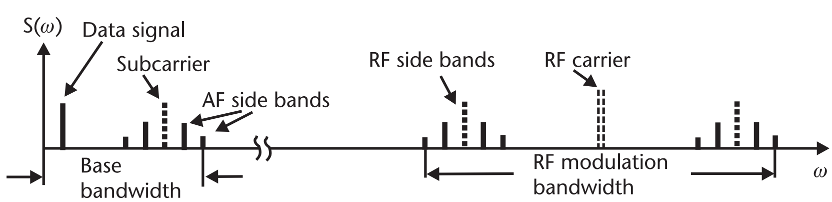


Figure 5.16 Indirect or subcarrier modulation.

would deliver for $B_B T_B = 0.5$ a bit rate of $R_B = B_M - 2\Delta F_C = 4.8$ kbit/s, which is four times greater.

Why has this slow and inefficient modulation scheme system been used so much? In fact, it is still used currently and there is good justification: It is applicable to existing analog mobile radio systems. Hence, it has been possible to introduce digital transmission into the analog world without the need to change the basic radio design and to apply it retroactively to existing radios. Additionally, this modulation can be transmitted via wire lines to repeaters and remotely controlled base stations.

Now some conclusions from this review of the various modulation schemes suited to mobile radio applications can be drawn. Digital modulation for PMR applications should provide as high a bit rate as possible in narrow channels and it should exhibit good sensitivity and CCR as well as high robustness against multipath distortion. Only a few modulation schemes fulfill these stringent requirements. Among these CP-BFSK, GMSK, $\pi/4$-DQPSK, 4GFSK, and CP-4FSK(2RC) seem to be best suited. Therefore, most of the contemporary digital PMR systems are based on one of these modulation schemes.

5.6 Sensitivity and Cochannel Rejection of Digital Receivers

The signal-to-noise ratio S/N (in terms of voltage) or the signal-to-noise power ratio P_S/P_N usually regarded in the baseband should be replaced for digital transmission by a more appropriate measure [9]. It can be expressed by the ratio of the bit energy to the noise power density E_B/N_0 within the baseband width B_B:

$$\left(\frac{\mathrm{S}}{\mathrm{N}}\right)^2 = \frac{P_S}{P_N} = \frac{E_B \cdot R_B}{N_0 \cdot B_B} \tag{5.68}$$

Introducing this into the relation for the dynamic sensitivity of analog receivers where the AF bandwidth B_{AF} is now called the baseband width B_B results in

$$\begin{aligned} p_{\mathrm{Rx\,digital\,min}} = {} & 10 \cdot \log_{10} N_0 + 10 \cdot \log_{10} B_B + k \cdot \sigma_F + 10 \cdot \log_{10} \frac{E_B \cdot R_B}{N_0 \cdot B_B} \\ & + 10 \log_{10} F + g_M \end{aligned} \tag{5.69}$$

All implementation losses are embodied in the noise figure F and g_M is the modulation gain as defined in Chapter 4 for analog receivers. It accounts for the difference between RF or IF and AF or baseband width as well as for the fact that the average carrier power is different from the real RF signal power. On the other hand, the calculation is often done at the demodulator input based on the average received RF signal power P_R and the equivalent IF or RF noise bandwidth B_{RF}:

$$\frac{P_R}{P_N} = \frac{E_B \cdot R_B}{N_0 \cdot B_{RF}} \tag{5.70}$$

The modulation gain, if any, then appears indirectly in the E_B/N_0 calculation where the result depends on the type of detection process involved. Moreover, in many simulation results the ratio E_B/N_0 includes fading and modulation characteristics and is given as $(E_B/N_0)_{\text{sim}}$, so with additional reordering and getting rid of B_B the formula looks much simpler:

$$p_{\text{Rx digital min}} = 10 \cdot \log_{10} N_0 + 10 \cdot \log_{10} R_B + 10 \cdot \log_{10} \left(\frac{E_B}{N_0}\right)_{\text{sim}} + 10 \cdot \log_{10} F \tag{5.71}$$

This equation as it stands can also be applied to multilevel modulation; alternatively, E_B can be replaced by E_S and R_B by R_S.

Recall that noise figure F is on the order of 6–10 dB and $10 \cdot \log_{10} N_0 =$ –174 dBm/Hz. The quality and robustness of a digital modulation scheme depends on its E_B/N_0 or E_S/N_0 ratio at a given *bit error rate* (BER) or *symbol error rate* (SER). The smaller this ratio at a certain error rate, the better the performance of the modulation in question. However, the calculation of the relationship between E_B/N_0 and BER or the bit error probability p_B or the symbol error probability p_S in many cases is difficult, in particular for sophisticated modulation schemes. A simple case gives better insight of how to proceed [14].

A bipolar binary signal $s_B(t)$ in the baseband with the two logic states $S_1 = S_B$ and $S_0 = -S_B$ and a threshold exactly in the middle between them at $S_T = 0$ is assumed. This signal may exhibit any transition between S_B and $-S_B$ as Figure 5.17 shows. Further there should be no ISI and all different signal states should be equiprobable. These assumptions are kept also for further consideration.

Sampling this signal over time and superposing all of the symbol transitions results in the so-called "eye diagram" [10, 29]. To detect what has been transmitted, $s_B(t)$ will be sampled in the symbol centers at t_n, t_{n+1}, t_{n+2} and so forth on the assumption that perfect synchronization has been achieved. The latter often is done by fairly complex *phase lock loop* (PLL) synchronization circuits. For the different

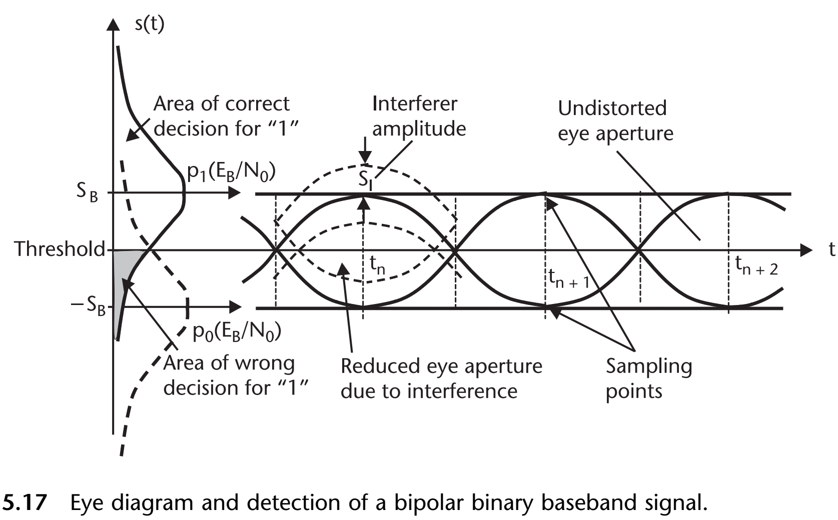


Figure 5.17 Eye diagram and detection of a bipolar binary baseband signal.

modulation schemes, various particularly well-suited synchronization principles have been developed but there is not enough space here to pursue this highly specialized topic further. With a fixed threshold at $S_T = 0$ the result of the detection process is obvious:

$$s_B(t_n) = \begin{cases} >0 & \text{equivalent to logic "1"} \\ <0 & \text{equivalent to logic "0"} \end{cases} \tag{5.72}$$

If an interfering signal $s_I(t)$ is added, errors can occur only if $S_I > S_B$, which means that if a small but increasing interfering signal is added, the eye aperture becomes smaller and for $S_I > S_B$ the eyes will close, provided that the interfering signal has a different frequency or a varying phase relation to the desired signal. Closed eyes indicate that transmission errors have occurred because the samples taken will be sometimes on the "wrong" side of the threshold. (Pattern recognition and soft decision provide the possibilities of avoiding errors during symbol detection even then.)

If the interfering signal is noisy, then the result of the detection process depends on the temporary amplitude of the interfering signal and its phase. In the case of *additive white Gaussian noise* (AWGN) the interfering signal in such a channel exhibits a Gaussian or normal distribution. Its standard deviation σ_N is directly related to average noise power density N_0 and the baseband width $B_B = 1/2T_B$ and therefore with the average noise power P_N and the average noise amplitude S_N:

$$\sigma_N^2 = P_N = N_0 \cdot B_B \tag{5.73}$$

$$S_N = \sigma_N \cdot \sqrt{2} = \sqrt{2N_0 \cdot B_B} = \sqrt{N_0 / T_B} \tag{5.74}$$

For the desired signal the relation between bit energy E_B, bit duration T_B, and signal amplitude S_B for a sinusoidal can easily be expressed in terms of signal power P_S:

$$S_B = \sqrt{2E_B / T_B} = \sqrt{2P_S} \tag{5.75}$$

From Figure 5.17 the bit error probability can now be derived with the help of the normal or Gaussian probability density function $p(\mu, \sigma, x)$ for a standard deviation σ and a mean of μ:

$$p(\mu, \sigma, x) \equiv \frac{1}{\sigma\sqrt{2\pi}} \cdot e^{-\frac{(x-\mu)^2}{2\sigma^2}} \tag{5.76}$$

This function delivers the probability that a particular event as a function of x will occur exactly at x. The cumulative Gaussian probability function $P(\mu, \sigma, x)$ or normal distribution describes the probability that this event will occur somewhere in the interval $-\infty \le t \le x$. Of course, the probability that it will occur in the complete interval $-\infty \le t \le \infty$ must be 1 [2]:

$$P(\mu, \sigma, x) \equiv \int_{-\infty}^{x} p(\mu, \sigma, x) = \frac{1}{\sigma\sqrt{2\pi}} \cdot \int_{-\infty}^{x} e^{-\frac{(t-\mu)^2}{2\sigma^2}} dt \tag{5.77}$$

$$P(\mu, \sigma, \infty) = \frac{1}{\sigma\sqrt{2\pi}} \cdot \int_{-\infty}^{\infty} e^{-\frac{(t-\mu)^2}{2\sigma^2}} dt = 1 \quad \text{because} \int_{0}^{\infty} e^{-a^2x^2} dx = \frac{\sqrt{\pi}}{2a} \tag{5.78}$$

Usually a normalization with the mean $\mu = 0$ is used because probabilities are often merely looked up from normalized tables. For practical reasons the complement $Q(\sigma, x)$ with $P(\sigma, x) + Q(\sigma, x) = 1$ has also been defined and is often needed:

$$P(\sigma, x) = \frac{1}{\sigma\sqrt{2\pi}} \cdot \int_{-\infty}^{x} e^{-t^2/2\sigma^2} dt \tag{5.79}$$

$$Q(\sigma, x) \equiv 1 - P(\sigma, x) = \frac{1}{\sigma\sqrt{2\pi}} \cdot \int_{x}^{\infty} e^{-t^2/2\sigma^2} dt \tag{5.80}$$

In a similar way the error function $\operatorname{erf}(x)$ and the complementary error function $\operatorname{erfc}(x)$ have been defined:

$$\operatorname{erf}(x) \equiv \frac{1}{\sqrt{\pi}} \cdot \int_{-x}^{x} e^{-t^2} dt = \frac{2}{\sqrt{\pi}} \cdot \int_{0}^{x} e^{-t^2} dt \tag{5.81}$$

$$\operatorname{erfc}(x) \equiv 1 - \operatorname{erf}(x) = \frac{2}{\sqrt{\pi}} \cdot \int_{x}^{\infty} e^{-t^2} dt \tag{5.82}$$

If in (5.79) and (5.80) $\sigma = 1$ is set and if the substitution $u = t/\sqrt{2}$ is used, some simple relationships between the P function and the error function as well as for the Q function and the complementary error function result:

$$\operatorname{erf}(x) = 2P\left(x\sqrt{2}\right) - 1 \quad \text{or} \quad P(x) = \frac{1}{2} \cdot \left[1 + \operatorname{erf}\left(\frac{x}{\sqrt{2}}\right)\right] \tag{5.83}$$

$$\operatorname{erfc}(x) = 2Q\left(x\sqrt{2}\right) \quad \text{or} \quad Q(x) = \frac{1}{2} \cdot \operatorname{erfc}\left(\frac{x}{\sqrt{2}}\right) \tag{5.84}$$

All of these probability relationships have been reviewed because in the literature different definitions and representations are in use and the comparison of results presented by different authors may be difficult [1, 2, 9, 12, 20].

Figure 5.18 shows the functions $\exp(-x)$ and $\exp(-x^2)$ and the close relationships to the other probability functions. For $P(x)$, $Q(x)$, $\mathrm{erf}(x)$, and $\mathrm{erfc}(x)$, the intervals are given over which these functions have to be integrated but no plots of the functions. Therefore, it is useful to look back to Figure 3.9 in Chapter 3. For large x a good approximation for $Q(x)$ has been found [9, 13] from which $\mathrm{erfc}(x)$ can be approximated with (5.84):

$$Q(x) \approx \frac{1}{x\sqrt{2\pi}} \cdot e^{-x^2/2} \quad \text{for } x > 3$$

or (5.85)

$$\mathrm{erfc}(x) \approx \frac{1}{x\sqrt{\pi}} \cdot e^{-x^2} \quad \text{for } x > 4$$

Most of the functions needed for error probability calculations can be looked up in tables or calculated with a scientific pocket calculator or PC. Rough calculations for some values are given in Table 5.9. Values below 10^{-5} have been suppressed and those above 0.9999 have been set to 1 because their exact size is not of interest for our investigations.

Returning to Figure 5.17 the error probability for false detection of a 0 instead of a 1 and vice versa can now be calculated. Due to the symmetry of the system, the probability for false detection of both logic states during the sampling instants t_n is equal and, therefore, the total bit error probability is

$$p_{B\,\mathrm{bip}} = \frac{1}{\sigma\sqrt{2\pi}} \cdot \int_{-\infty}^{-S_B} e^{-\frac{1}{2}\left[\frac{s(t_n)}{\sigma}\right]^2} ds = \frac{1}{\sigma\sqrt{2\pi}} \cdot \int_{S_B}^{\infty} e^{-\frac{1}{2}\left[\frac{s(t_n)}{\sigma}\right]^2} ds \qquad (5.86)$$

With the amplitudes S_B and S_N of the desired signal and the noise as calculated above, we get from (5.74) to (5.77) the error probability $p_{B\,\mathrm{bip}}$:

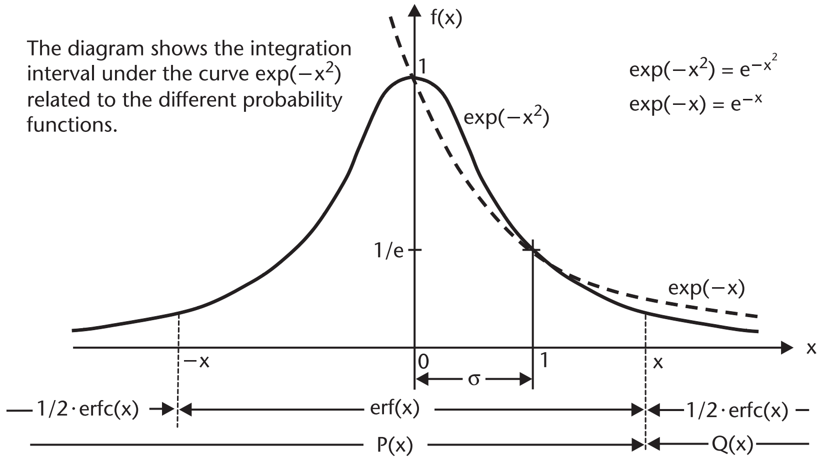


Figure 5.18 The relations between $\exp(-x)$, $\exp(-x^2)$, $P(x)$, $Q(x)$, $\mathrm{erf}(x)$, and $\mathrm{erfc}(x)$.

Table 5.9 The Functions $\exp(-x)$, $\exp(-x^2)$, $P(x)$, $Q(x)$, $\mathrm{erf}(x)$, and $\mathrm{erfc}(x)$

x	$exp(-x)$	$exp(-x^2)$	$P(x)$	$Q(x)$	$erf(x)$	$erfc(x)$
0.0	1.0000	1.0000	0.5000	0.5000	0.0000	1.0000
0.1	0.9048	0.9900	0.5398	0.4602	0.1125	0.8875
0.2	0.8187	0.9608	0.5793	0.4207	0.2227	0.7773
0.3	0.7408	0.9139	0.6179	0.3821	0.3286	0.6714
0.4	0.6703	0.8521	0.6554	0.3446	0.4284	0.5716
0.5	0.6065	0.7788	0.6915	0.3085	0.5205	0.4795
0.6	0.5488	0.6977	0.7257	0.2743	0.6039	0.3961
0.7	0.4966	0.6126	0.7580	0.2420	0.6778	0.3222
0.8	0.4493	0.5273	0.7881	0.2119	0.7421	0.2579
0.9	0.4066	0.4449	0.8159	0.1841	0.7969	0.2031
1.0	0.3679	0.3679	0.8413	0.1587	0.8427	0.1573
1.2	0.3012	0.2369	0.8849	0.1151	0.9103	0.0897
1.4	0.2466	0.1409	0.9192	0.0808	0.9523	0.0477
1.6	0.2019	0.0773	0.9452	0.0548	0.9763	0.0237
1.8	0.1653	0.0392	0.9641	0.0359	0.9891	0.0109
2.0	0.1353	0.0183	0.9772	0.0228	0.9953	$4.7 \cdot 10^{-3}$
2.2	0.1108	$7.9 \cdot 10^{-3}$	0.9861	0.0139	0.9981	$1.9 \cdot 10^{-3}$
2.4	0.0907	$3.2 \cdot 10^{-3}$	0.9918	$8.2 \cdot 10^{-3}$	0.9993	$6.9 \cdot 10^{-4}$
2.6	0.0743	$1.2 \cdot 10^{-3}$	0.9953	$4.7 \cdot 10^{-3}$	0.9998	$2.4 \cdot 10^{-4}$
2.8	0.0608	$3.9 \cdot 10^{-4}$	0.9974	$2.6 \cdot 10^{-3}$	0.9999	$7.5 \cdot 10^{-5}$
3.0	0.0498	$1.2 \cdot 10^{-4}$	0.9987	$1.3 \cdot 10^{-3}$	≈ 1	$2.2 \cdot 10^{-5}$
3.2	0.0408	$3.6 \cdot 10^{-5}$	0.9993	$6.9 \cdot 10^{-4}$	≈ 1	≈ 0
3.4	0.0334	$1.0 \cdot 10^{-5}$	0.9997	$3.4 \cdot 10^{-4}$	≈ 1	≈ 0
3.6	0.0273	≈ 0	0.9998	$1.6 \cdot 10^{-4}$	≈ 1	≈ 0
3.8	0.0224	≈ 0	0.9999	$7.2 \cdot 10^{-5}$	≈ 1	≈ 0
4.0	0.0183	≈ 0	≈ 1	$3.2 \cdot 10^{-5}$	≈ 1	≈ 0
4.5	0.0111	≈ 0	≈ 1	≈ 0	≈ 1	≈ 0
5.0	0.0067	≈ 0	≈ 1	≈ 0	≈ 1	≈ 0

$$p_{B\,\mathrm{bip}} = 1 - P\left(\frac{\sqrt{2E_B/T_B}}{\sqrt{N_0/T_B}}\right) = Q\left(\sqrt{2E_B/N_0}\right) = \frac{1}{2} \cdot \mathrm{erfc}\left(\sqrt{E_B/N_0}\right) \tag{5.87}$$

This result is valid for a simple bipolar binary baseband signal in an AWGN channel with perfect synchronization and a fixed threshold. If a unipolar signal is used in the baseband, then the logic states are S_B and 0 with the threshold at $S_B/2$. Hence, E_B/N_0 then has to be replaced in the formula above by $E_B/2N_0$ to get $p_{B\,\mathrm{uni}}$.

The bit-by-bit decision at a fixed threshold is called *hard decision*. Improved E_B/N_0 ratios can be achieved if bit-by-bit detection with hard decision is not performed on its own. This is achieved during detection by taking adjacent bits into account and using optimized detection algorithms. Consequently, this approach is called soft decision. In practice, for many modulation schemes, the calculation is much more complicated than shown above or cannot be done rigorously so only approximations are possible. Therefore, only the results of some examples of major importance are given below [1, 6, 9, 12, 14, 16, 18, 19, 21, 23, 39].

It has been shown that for many modulation schemes with coherent detection, the bit and symbol error probability obeys a similar law (A) as we have found for baseband signals where a and b are modulation-dependent constants as shown later in Table 5.10:

Table 5.10 Error Probability and Sensitivity in the AWGN Channel for BER = 1%

Modulation	*Parameter*	*Detection*	*Formula*	*a*	*b*	E_B/N_0 *at* $p_B = 1\%$ *[dB]*
Baseband	Bipolar	Coherent	A	0.5	1.0	4.3
(binary)	Unipolar	Coherent	A	0.5	0.5	7.4
BASK[1]	$S_0 = 0$	Coherent	A	0.5	0.25	10.3
	$S_1 = S_C$	Noncoherent	B	0.5	0.125	15.0
BFSK[2]	$h = 1$ (antipodal)	Coherent	A	0.5	1.0	4.3
	$h = 0.72$	Coherent	A	0.5	0.61	6.5
	$h = 0.5$ (orthogonal)	Coherent	A	0.5	0.5	7.4
		Noncoherent	B	0.5	0.5	8.9
GMSK[2]	$B_BT_B = \infty$[3] $h = 0.5$	Coherent	A	≈ 0.5	0.85	5.1
	$B_BT_B = 0.25$[4] $h = 0.5$	Coherent	A	≈ 0.5	0.68	6.0
4FSK[2,5]	$h = 0.5$	Coherent	A	≤ 1.5	1.0	5.7
		Noncoherent	B	≤ 1.5	1.0	7.0
mFSK[2,5]	$h = 0.5$	Coherent	A	$\frac{1}{2} \cdot (m-1)$	$q/2$	—[6]
		Noncoherent	B	$\frac{1}{2} \cdot (m-1)$	$q/2$	—[6]
BPSK	$h = 1$	Coherent	A	0.5	1.0	4.3
		Noncoherent	B	0.5	1.0	5.9
DBPSK	$h = 1$	Coherent	A	≈ 1.0	1.0	5.2
		Noncoherent	B	0.5	1.0	5.9
QPSK[5,7]	$h = 0.5$	Coherent	A	0.5	1.0	4.3
DQPSK[5,8]	$h = 0.5$	Coherent	B	≈ 1.0	1.0	5.2
		Noncoherent	B	0.5	1.0	5.9
mPSK[5,8,9]	$h = 2/m$	Coherent	A	$\approx 1/q$	$q \cdot \sin^2(\pi/m)$	—[6]
mDPSK[5,9]	$h = 2/m$	Coherent	A	$\approx 1/q$	$q \cdot \sin^2(\pi/m\sqrt{2})$	—[6]

[1]Binary ASK with $m = 100\%$.
[2]Constant envelope.
[3]MSK.
[4]Partial response for $BT < 0.5$.
[5]Gray coded.
[6]Depends on m.
[7]Also OQPSK and $\pi/4$-QPSK.
[8]Also $\pi/4$-DQPSK.
[9]For large E_B/N_0.

$$p_{B,S} = a \cdot \operatorname{erfc}\left(\sqrt{bE_{B,S}/N_0}\right) \qquad \text{(A)} \tag{5.88}$$

Figure 5.19 suggests how to derive the error probability for coherent detection in the general case [7, 9]. If the two signal vectors $s_1(t, \varphi_1)$ and $s_2(t, \varphi_2)$ have equal amplitudes S_B and frequencies $f = \omega/2\pi$ but a phase difference of $\Delta\Phi = |\varphi_1 - \varphi_2|$, then their grade of correlation according to (3.59) and (3.60) is $\rho = \cos(\Delta\Phi)$. Hence, with their vector difference $2D_B$ and noise amplitude S_N we find the bit or symbol error probability $p_{B,S\rho}$:

$$p_{B,S\rho} = 1 - P(D_B/S_N) = Q(D_B/S_N) \tag{5.89}$$

For the triangle in Figure 5.19 the cosine rule is applicable [2]:

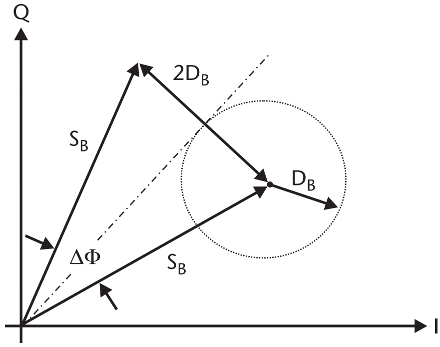


Figure 5.19 Coherent detection for arbitrary phase difference.

$$4D_B^2 = 2S_B^2 - 2S_B^2 \cdot \cos(\Delta\Phi) \rightarrow D_B^2 = \frac{S_B^2}{2} \cdot (1 - \rho)$$

because $a^2 = b^2 + c^2 - 2bc \cdot \cos\alpha$ where α is the angle opposite to side a in the triangle. Then

$$S_B = \sqrt{2E_B/T_B} \quad \text{and} \quad S_N = \sqrt{N_0/T_B}$$

deliver with $\sqrt{(1 - \cos x)/2} = \sin(x/2)$ [2]:

$$p_{B,S\rho} = \frac{1}{2} \cdot \mathrm{erfc}\left(\sqrt{\frac{E_{B,S}}{2N_0} \cdot (1 - \rho)}\right) = \frac{1}{2} \cdot \mathrm{erfc}\left(\sqrt{\frac{E_{B,S}}{N_0} \cdot \sin^2\left(\frac{\Delta\Phi}{2}\right)}\right) \tag{5.90}$$

Now we can do the calculation for any phase difference. For antipodal signals with a correlation of $\rho = -1$, the situation is similar to the baseband case with bipolar signals. For instance, this is true for BPSK with a phase difference of π between the logic states. If, however, orthogonal signals are used as with MSK, then the E_B/N_0 ratio has to be divided by 2 (meaning $\sqrt{2}$ for the amplitudes) because the signals are orthogonal with a phase difference of $\pi/2$ and therefore $\rho = 0$. However, for multilevel schemes there is a certain probability that instead of the erroneous detection of an adjacent symbol, another one is detected that has a larger distance and causes a multibit error. Therefore, the error statistics become more complicated and sometimes no solution in closed form is available. In these cases, approximations have been found the accuracy of which suffices for most practical purposes.

For noncoherent envelope detection in many cases another common law (B) is valid. The derivation of this result is more complicated than in the case of coherent demodulation due to the nonlinearity of the envelope detector and therefore it is given without proof:

$$p_{B,S} = a \cdot e^{-bE_{B,S}/N_0} \qquad \text{(B)} \tag{5.91}$$

The reason why most of the coherent and noncoherent detection processes are governed by only two basic laws is that more or less the same underlying failure mechanisms are present.

To calculate the error probability, the linear energy ratio is needed. If this ratio is given in x dB, then it can easily be converted into a linear format because $E_{B,S}/N_0 = 10^{x/10}$. The results given in Table 5.10 are valid for the unprotected, that is, uncoded AWGN channel with $p_B = 1\%$. This is often used as criterion for digital receiver sensitivity and transparent or unprotected transmission. The constants a and b are different for the various types of modulation, but they can be taken for both formulas (5.88) and (5.91) from Table 5.10. Both allow primarily the calculation of symbol error probability p_S dependent on the E_S/N_0 ratio for m-ary modulation. Because m can have any value in Table 5.10, no examples have been given for $m > 4$.

However, in Table 5.10 bit error probability p_B is always given, *not* the symbol error probability, because the end user is only interested in the BER of the user bit rate. To determine the bit error probability p_B for m-ary schemes E_B/N_0 can be derived from E_S/N_0 according to (5.55) and p_B must be calculated from p_S, which is more difficult.

For many mFSK schemes, in the case of an error any of the other symbols can be selected, resulting in the following relation between the symbol and the bit error probability [9]:

$$p_B = \frac{m/2}{m-1} \cdot p_S \xrightarrow{m>>2} \frac{p_S}{2} \tag{5.92}$$

For Gray-coded mPSK, multiple bit errors per symbol are much less likely than single-bit errors:

$$p_B \approx \frac{p_S}{\log_2 m} = \frac{p_S}{q} \tag{5.93}$$

From the general formulas for the error probability of mFSK and mPSK, specific cases can be derived, for example, for $m = 4$. However, some care is necessary. If, for instance, BPSK is regarded then the derivation from mPSK would result in twice the error probability. Generally in mPSK each symbol has two neighbors that contribute to the errors in the presence of noise. However, for BFSK the upper and lower adjacent signals are identical and should therefore not be counted twice. Hence, p_B as derived from mPSK has additionally to be divided by 2 and therefore BPSK and QPSK have the same bit error probability. Assuming that coherent QPSK detection handles I and Q channels independently, and therefore both must show the same performance as BPSK, leads to the same conclusion. The consequence is that QPSK allows double the bit rate in comparison to BPSK for a given receiving power. Moreover, the bandwidth of QPSK is slightly reduced because the phase difference of some of the phase transitions is smaller than with BPSK.

Another specific point is the derivation of the error probability of 4FSK from that of mFSK, which gives the result shown in Table 5.10. In practice, however, there are some methods for improvement that lead to better results, and this is also true for some other modulation schemes. For noncoherent detection of multilevel modulation schemes, the same principal problem exists as for coherent detection: The failure mechanisms may not only impair the adjacent symbol levels but also others, albeit with reduced error probability.

There is not enough space here to treat the error performance of the different modulation schemes in detail, but Figure 5.20 should give the reader a basic idea of how the different unprotected modulation schemes perform in the AWGN channel and what additional impairments are caused by Rayleigh fading.

5.7 Channel Access and Transmitter Keying

There are several types of channel access: *frequency division multiple access* (FDMA), *time division multiple access* (TDMA), *code division multiple access* (CDMA), and *orthogonal frequency division multiplexing* (OFDM). The latter two are sometimes confused with modulation methods. All of these access methods have different benefits and drawbacks and therefore are not equally suitable in all applications [6, 12, 15, 20, 40–43].

In PMR the classic access method is the use of a single carrier frequency per communication channel. FDMA is well suited for radio communication with wide coverage and low traffic density. The channels are several kilohertz to some tens of kilohertz separated from each other, but rarely less than 10 kHz.

In the case of medium to high traffic density it is economically more feasible to put several communication channels onto one carrier frequency. TDMA saves radio hardware at the base station and eases the coupling of a high number of communication channels to one single antenna system. The channel separation is on the order of tens or even hundreds of kilohertz or more, for example, 200 kHz for GSM. Figure 5.21 shows the difference between FDMA and TDMA using the example of four-channel usage as in the case of DIIS and TETRA, respectively.

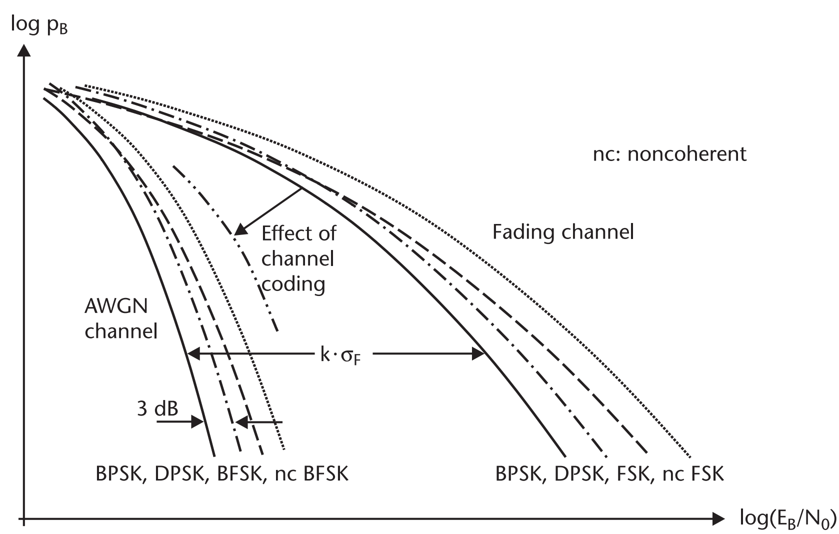


Figure 5.20 Rough comparison of different modulation schemes.

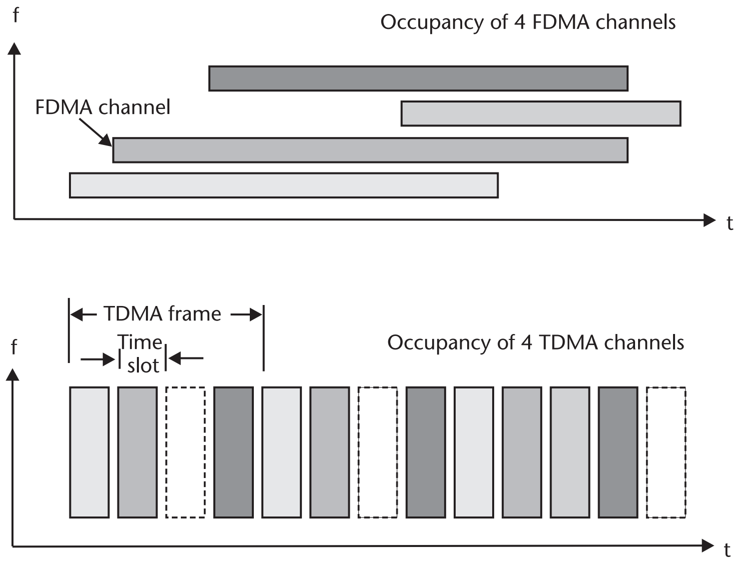


Figure 5.21 FDMA and TDMA transmission.

Most TDMA systems for high traffic capacity do not carry all of the communication channels on only one frequency. Several carriers, each providing a given number of communication channels, are usually used. Hence, these schemes are better called *FD/TDMA* [41]. Interchanging the carrier frequency during transmission from time slot to time slot in such a FD/TDMA system is called *slow frequency hopping* (SFH) [6, 44]. It is used to achieve a more homogeneous distribution of transmission errors, which eases error detection and correction considerably, in particular, for nonmoving or only slowly moving mobiles. Otherwise, these might stay in a fading null for a period of time, which is not a problem for fast moving mobiles because these pass through the local fading pattern fast enough to get a better correctable error distribution. However, SFH only makes sense if enough different carrier frequencies are available that can be distributed over a sufficiently large bandwidth as in the case of GSM.

The third important access method is CDMA. *Frequency-hopping spread spectrum* (FH-CDMA) is similar to SHF insofar as the frequency is changed during the transmission of a message [6, 20, 42, 43]. However, this is not done on a time slot basis but several times per bit as Figure 5.22 shows. For better clarity it is assumed that only one communication channel is occupied and that every bit comprises only four pieces called *chips* that are spread by a certain sequence over the system bandwidth B_{Syst}. All of the transmitting power, therefore, is distributed over a wide bandwidth resulting in a very low spectral power density.

Another method called *direct sequence CDMA* (DS-CDMA) spreads the signal energy over the whole band by multiplying the modulated carrier with a binary spreading sequence as shown in Figure 5.23.

In the receiver the spreading is removed by *despreading,* which is the inverse of the spreading process. The result is good resistance against fading and interferers. The ratio of modulation bandwidth B_M to system bandwidth B_{Syst} is called the

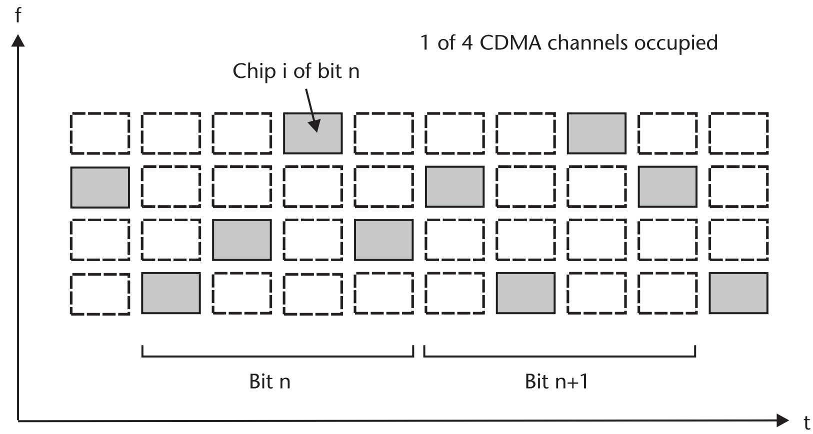


Figure 5.22 The FH-CDMA transmission principle.

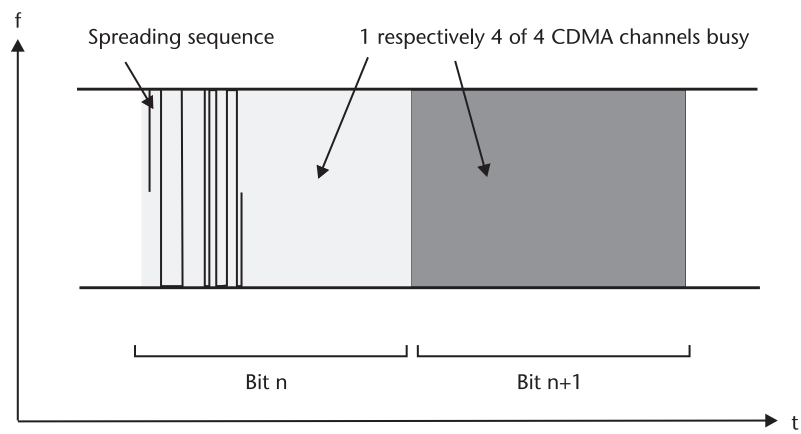


Figure 5.23 DS-CDMA transmission.

spreading factor and that determines the possible maximum number $n_{\max}$ of communication channels in a CDMA system. The spreading factor can also be defined as the ratio of the chip rate R_C to bit rate R_B or as the ratio of bit duration T_B to chip duration T_C:

$$n_{\max} = \frac{B_{\text{Syst}}}{B_M} = \frac{R_C}{R_B} = \frac{T_B}{T_C} \tag{5.94}$$

CDMA can also be combined with a multilevel modulation scheme, but then of course T_B has to be replaced by the symbol duration T_S. The processing gain g_P depends on the spreading factor. It is the gain in the ratio between a CDMA signal and a fixed frequency interferer obtained from the signal spreading:

$$g_P = 10 \cdot \log_{10} n_{\max} \tag{5.95}$$

The P_C/P_I ratio is considerably improved by CDMA compared to the P_C/P_N ratio where P_C is the carrier power, P_I is the interferer power, and P_N is the noise-like proportion of the interferer power falling into the modulation bandwidth after despreading. The larger the ratio B_M/B_{Syst} the better the suppression of noise and interferers:

$$10 \cdot \log_{10}\left(\frac{P_C}{P_N}\right) = 10 \cdot \log_{10}\left(\frac{P_C}{P_I}\right) + g_P \tag{5.96}$$

With CDMA a single-frequency or narrowband interferer within the transmission band is spread over the whole band. That means that only a very small portion of its power falls into the receiver channel filter after despreading as Figure 5.24 shows. If, for instance, $10 \cdot \log_{10}(P_C/P_I) = -15$ dB and $g_P = 25$ dB then $10 \cdot \log_{10}(P_C/P_I) = 10$ dB is the result. Hence, a fixed frequency interferer cannot for the most part block a CDMA transmission even if its power is much higher. That is why this method initially was invented for military applications as a means to nullify the effects of fixed frequency jammers.

Because CDMA systems are designed to cope with severe fading and multipath propagation conditions, they have no problems with bad multipath scenarios. However, it is of major importance to use spreading codes with low cross-correlation. If the codes are different, the carrier frequency can, in principle, be reused in adjacent cells. This and the extraordinarily good interference suppression are the reasons for the superior frequency economy claimed by CDMA. However, in practice, a certain capacity reduction has to be taken into account as already mentioned in Chapter 3, and several other effects also limit the usability of CDMA for PMR.

At first the uplink power must be controlled with an accuracy of around 1 dB or better; otherwise, the number of CDMA channels that can be used simultaneously will be reduced. This is no easy task when the dynamic range of a mobile radio communication system is on the order of 80–100 dB even for CDMA systems, which use very fast power control mechanisms with refresh rates of up to 2 kHz

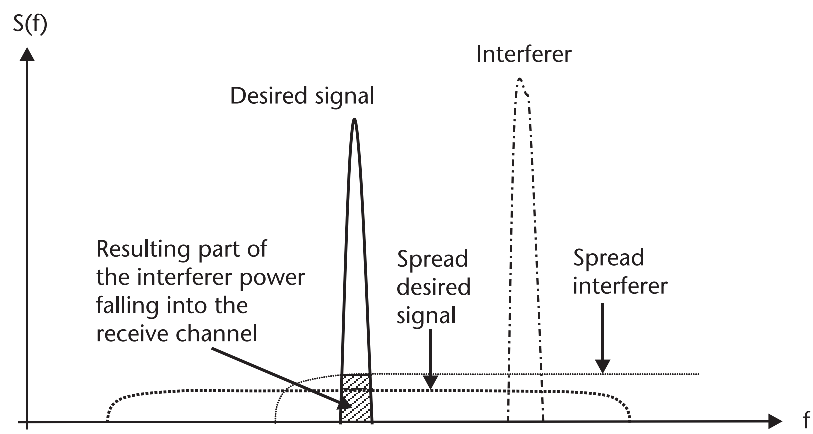


Figure 5.24 Interference rejection with CDMA.

[43]. Secondly, proper synchronization is a precondition for successful operation of a CDMA system and, in particular, initial synchronization is a tough problem.

High-capacity CDMA systems are very cost effective because they minimize the need for base station radio hardware provided that powerful chips are available for all of the required digital operations. In particular, for the mobiles these chips must be at an acceptable price and have a low power consumption. However, in PMR there are few systems that need to have a high number of users and produce a high traffic density. Therefore, the minimum bandwidth necessary for a well-designed CDMA system being at least 1 MHz can hardly be justified and thus CDMA is not attractive for most contemporary PMR applications, whereas for cordless telephones or cellular systems it is an important alternative.

Very often is has been argued that CDMA provides superior capacity compared to FDMA, TDMA, and FD/TDMA systems. However, it has been shown that the outcome of a fair comparison always leads to the same total transmission capacity [44], which is near that of the principal boundary found by Shannon. This defines the capacity limit to which all systems tend.

Lately OFDM has come into use. For *digital audio broadcasting* (DAB) and digital TV, it employs hundreds of separate carriers modulated by OQPSK at low bit rates. This bundle of carriers is transmitted and processed in the receiver as a whole. This allows the transmission of very high bit rates split into small portions and transmitted in parallel [6]. Each of these single carriers is modulated with symbols of relatively long duration and therefore echoes due to multipath propagation do not degrade the quality of reception. How can such complexity be afforded? The RF and signal processing hardware is not composed from a large number of mixers, multipliers, phase shifters, filters, adder stages, and so on, but instead employs only a single DSP that performs all of the necessary operations. However, as for linear modulation schemes, a linear transmitter power amplifier is needed to avoid signal distortions caused by intermodulation products. OFDM is also feasible for high bit rates and mobile radio transmission as several studies (for example, in the European ACTS research program) have proven. However, until now no operational high-speed mobile systems have been based on OFDM.

In FDMA systems carrier keying occurs only at the beginning and end of a message transmission that lasts at least several seconds and, hence, it occurs relatively seldom compared to pulse interference from ignition systems or similar interference sources. The situation is quite different in TDMA systems where the carrier is periodically keyed on and off many times during one message. According to the time slot duration, the keying frequency in many TDMA systems is somewhere between 20 and 2,000 Hz, and in FH-CDMA it is equal to the chip rate, which is a multiple of the gross bit rate. These considerations make clear that carrier keying may become a problem if the generated switching spectra are much wider than the modulation or spreading bandwidth [45, 46]. For sure this is the case if carrier keying is done by hard switching as Figure 5.25 shows.

This example shows clearly that hard switching of a carrier frequency produces a switching spectrum that exceeds the modulation bandwidth. However, if the envelope slopes during carrier switching on or off are not too steep and are adequately shaped, then the switching spectrum will be masked by the modulation spectrum [45].

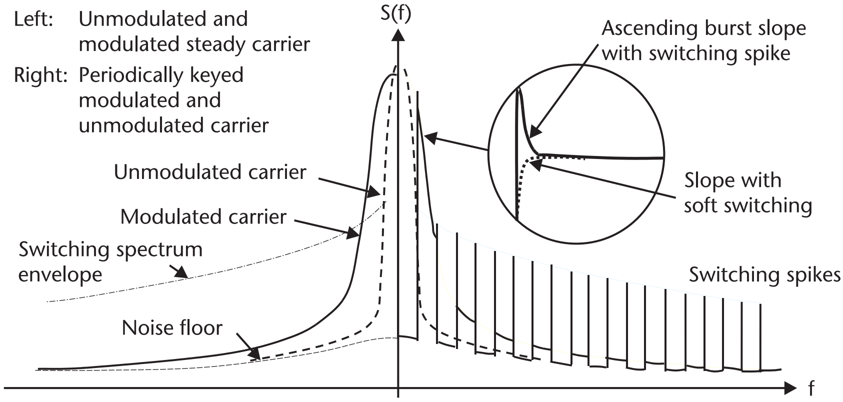


Figure 5.25 Switching spectrum of a RF carrier.

Soft keying of a radio frequency can be treated as amplitude modulated by a sinusoidal function with $m = 1$. The duration T_K of such a $\cos^2$ slope, therefore, equals half the period of the AM with $\Delta f_K = \pm 1/2T_K$ determining the width of the switching spectrum:

$$s(t) = S_0 \cdot [1 + m(t)] \cdot \cos(\Omega t) \quad \text{with} \quad m(t) = \frac{1}{2} \cdot \left[1 - \cos\left(\frac{\pi t}{T_K}\right)\right] \tag{5.97}$$

$$B_K \approx 2 \cdot \Delta f_K = 1/T_K \tag{5.98}$$

This bandwidth is comparable to that of an ASK signal with $B_M \approx R_S = 1/T_S$. To ensure that the switching bandwidth does not exceed the modulation bandwidth, $T_K > T_S$ must be fulfilled. Obviously a longer keying time produces a smaller switching spectrum. Hence, for properly shaped on and off keying, the necessary duration of the switching slope is somewhat greater than the symbol duration. In practice, a longer duration provides a margin for tolerances and nonlinear effects [46]:

$$T_K \geq 2 \ldots 3T_S \tag{5.99}$$

Because soft keying is not fully equivalent to steady AM, additional harmonics of these two side frequencies also appear but these decrease rapidly with increasing distance from the carrier. If, however, the switching spectrum in the region of –60 dBc or below is of interest, a more accurate analysis is necessary. For ASK with $m = 1$, $S_0 = 1$, and a $\cos^2$ pulse being the envelope of the carrier frequency, the symbol and carrier keying slope are identical and so the switching and modulation spectrum are the same. Hence, the voltage density spectrum around the carrier frequency is

$$S_{\mathrm{RC}}(\Omega \pm \Delta\omega) = \frac{\pi^2 T_K \cdot \mathrm{si}\,[(\Omega \pm \Delta\omega)T_K]}{2 \cdot [\pi^2 - (\pm \Delta\omega)^2 T_K^2]} \tag{5.100}$$

With $m = 1$ we get a factor of 1/2 for the sidebands and from the Fourier integral over 0 to ∞ instead of $-\infty$ to ∞ (real frequencies only), we get a factor of 2, which cancels out. With the normalization $\eta = \Delta\omega T_K/2\pi$ where $\Delta\omega = 2\pi\Delta f$ is the difference to the carrier frequency and referring to $S_{RC}(0)$, this expression can be simplified considerably. If we compare the flat passband voltage density with the decreasing density in the adjacent channel, the adjacent power level in decibels can be estimated for the worst case:

$$\left|\frac{S_{RC}(\eta)}{S_{RC}(0)}\right| = \left|\frac{\mathrm{si}(2\pi\eta)}{[1-(2\eta)^2]}\right| < \frac{1}{|\pi\cdot(2\eta)^3|} \quad \text{for} \quad \eta = \Delta f T_K >> 1 \tag{5.101}$$

$$p_{ACh} \le -20\cdot\log_{10}(8\pi) - 60\cdot\log_{10}(\eta) = -28 - 60\cdot\log_{10}(\Delta f T_K) < -60 \text{ dB} \tag{5.102}$$

Hence, for $\Delta f_{ACh}T_K \ge 3.4$ the attenuation of the switching and modulation spectrum of ASK exceeds 60 dB. This is also valid for FSK as an approximation for a very short burst that could hardly carry reasonable modulation content. Hence, as a model case for a very long burst the RC ramp is preferable. It exhibits an even smaller spectrum away from the carrier as the comparison with the $\cos^2$ pulse shows:

$$\left|\frac{\max[S_{\epsilon RC}(\eta)]}{\max[S_{RC}(\eta)]}\right| \xrightarrow{\eta>>1} \approx \frac{1}{2} \tag{5.103}$$

because $\max|\sin x| = \max|\cos x| = 1$ [3]. Here we have assumed that the duration of the RC ramp is T_K so that the ascending slope has equal duration in both cases and we get

$$p_{ACh} = -20\cdot\log_{10}(16\pi) - 60\cdot\log_{10}(\eta) = -34 - 60\cdot\log_{10}(\Delta f T_K) < -60 \text{ dB} \tag{5.104}$$

To meet this condition $\Delta f_{ACh}\cdot T_K \approx 2.8$ is needed.

For FSK and constant envelope the keying slopes have to be shaped by the transmitter power control circuit, which can be done even with many standard class C amplifiers by means of appropriate input power control. Assuming, for example, GMSK with $B_B T_B = 0.5$, (5.62) delivers with $\Delta f_{ACh} \approx B_{ACh}/2$ and $T_B = T_S$:

$$\Delta f_{ACh} \ge \frac{1.75}{T_S} \tag{5.105}$$

Hence, we finally get:

$$T_K \ge \frac{2.8}{1.75}\cdot T_S = 1.6\cdot T_S \tag{5.106}$$

Imperfections will worsen the situation somewhat, thus $T_K \geq 2T_S$ is advisable. All of these results are a pretty good match to the rough estimate of (5.99) where the soft switching was compared with steady AM. In the case for which the adjacent power level p_{ACh} has to be determined precisely, then the power falling into the band between the two corner frequencies of that channel has to be calculated:

$$10 \cdot \log_{10} \frac{P_{\mathrm{ACh}}}{P_C} = 20 \cdot \log_{10} \left[\frac{2}{S(0)} \cdot \int_{\omega_L}^{\omega_U} S(\omega) \cdot H(\omega)\, d\omega \right] \qquad (5.107)$$

The factor 2 comes again from the replacement of the double-sided Fourier integral by a single-sided one. If the transfer function $H(\omega)$ of the channel filter is not known exactly, then it can be replaced by a constant between the lower and upper corner frequencies ω_L and ω_U, disregarding its phase characteristic. This usually results in a worst case estimation. If this is sufficient, it is not necessary to perform the complete calculation with a real channel filter characteristic.

A specific problem is the first access of a mobile to a radio channel. For call setup in trunked systems, the mobiles predominantly have to access a predefined channel. Different methods to do this have been developed. The channel access attempts usually occur randomly distributed in time and are short compared to the messages sent during an ongoing call [6, 9, 15, 18, 36, 42, 47]. The simplest method is unconditional direct random access called the *ALOHA* method. This can result in signal collisions due to an overlap in time where unfortunately a partial time overlap has the same result as a total overlap: Overlapping messages are destroyed and have to be repeated. The throughput therefore is limited to $1/2e$ or roughly 18%. A reduction of the number of collisions is possible if the call attempts are only allowed in regular time slots and never cross the time slot boundaries. The maximum throughput then doubles to $1/e$ or about 37%.

In radio systems a collision may not necessarily mean the loss of the colliding messages. If the level difference is greater than the CCR, then the stronger signal will be correctly received and processed by the base station while only the weaker gets lost and has to be repeated. This capture effect improves the throughput compared to systems without capture. For slotted ALOHA with capture the throughput therefore can exceed one-half or 50% [47].

There are more refined methods that aim to reduce the loss due to collisions even further. One is *carrier sense multiple access* (CSMA), in which the channel can only be accessed if it is unoccupied. This is verified by checking the channel beforehand, which can be done simply by testing the presence of a carrier. Advanced systems use more sophisticated methods such as *digital sense multiple access* (DSMA). But even then collisions cannot be totally avoided. Hence, the collision probability is lowered but it is not zero. There are several variations on this principle, and the throughput can be increased to the region between 50% and 80% depending on the method chosen. In any of these methods after a collision and loss of a message, a retry is initiated with a randomly distributed delay to avoid a regular mutual blocking of several stations trying to access the channel repeatedly at the same time. The loss of a message is usually detected by the initiating mobile because it gets no acknowledgment. Even more advanced methods for channel access and collision resolving have been developed, for instance, access only in predefined or reserved time slots, but this is another issue we cannot pursue further in this book.

5.8 Fundamental Principles of Error Correction and Channel Equalization

More than half a century ago, Claude E. Shannon dealt with the question of the maximum information volume that can be transmitted free of errors over a given AWGN channel. One of his most important findings was that the maximum channel capacity C in bits per second depends on the signal power P_S, the noise power P_N, and the channel bandwidth B_{Ch} [5, 6, 12, 15, 16, 19, 23, 31, 39, 47]:

$$C = B_{\mathrm{Ch}} \cdot \log_2\left(1 + \frac{P_S}{P_N}\right) = 3.322 \cdot B_{\mathrm{Ch}} \cdot \log_{10}\left(1 + \frac{P_S}{P_N}\right) \tag{5.108}$$

The remarkable consequence is that even in the case of $P_S < P_N$ error-free transmission up to a certain bit rate is possible. Recalling (5.27) and (5.68) we get for the Nyquist channel with $R_B \leq C$ immediately:

$$R_B \leq B_{\mathrm{Ch}} \cdot \log_2\left(1 + \frac{E_B R_B}{N_0 B_{\mathrm{Ch}}}\right) \xrightarrow{R_B \to 2B_{\mathrm{Ch}}} B_{\mathrm{Ch}} \cdot \log_2\left(1 + \frac{2E_B}{N_0}\right) \tag{5.109}$$

Shannon demonstrated the limit, but he did not say how to reach it. Researchers have finally found that in practical systems the maximum channel capacity cannot be reached even if in the last decades significant progress has been made by applying clever modulation and coding methods. The equation is valid for the baseband channel and remains unchanged for RF channels with SSB modulation. For other types of modulation and in the presence of fading and shadowing, the situation is more complicated as the considerations concerning digital receiver sensitivity have shown [see (5.69) to (5.71)].

If the channel bandwidth has to be changed, say, it has to be divided by k, an interesting conclusion can be drawn from (5.108) after a simple modification if the same bit rate as before is needed:

$$R_{B\max} = \frac{B_{\mathrm{Ch}}}{k} \cdot \log_2\left(1 + \frac{P_S}{P_N}\right)^k \tag{5.110}$$

For $P_S >> P_N$ and with $P_N \to N_0 B_{\mathrm{ch}}/k$ we finally get the following:

$$R_{B\max} \xrightarrow{P_S >> P_N} \frac{B_{\mathrm{Ch}}}{k} \cdot \log_2\left(\frac{P_S}{N_0 B_{\mathrm{Ch}}/k}\right)^k = B_{\mathrm{Ch}} \cdot \log_2\left(\frac{P_S}{N_0 B_{\mathrm{Ch}}/k}\right) \tag{5.111}$$

This means, for instance, for $k = 4$ to go to a quarter of the bandwidth, causing a reduction of P_N to a quarter, and to allow P_S also to be reduced to a quarter. However, if the original channel was a binary one, then according to (5.27) we would only get a quarter of the original bit rate and consequently the wish to retain the original bit rate forces us to use an appropriate multilevel transmission scheme. If $P_S >> P_N$ is not fulfilled, then the result is a little bit different but the tendency and the consequences are the same.

Another important conclusion can be drawn from (5.108) if we look for the total number of bits $N_B = R_{B\max} \cdot T_M$ to be transmitted during the message duration T_M:

$$N_B = B_{\mathrm{Ch}} T_M \cdot \log_2 \left(1 + \frac{P_S T_M}{N_0 B_{\mathrm{Ch}} T_M} \right) = B_{\mathrm{Ch}} T_M \cdot \log_2 \left(1 + \frac{E_M}{E_N} \right) \tag{5.112}$$

Hence, a simple rule for the total message energy E_M can be derived therefrom for $B_{\mathrm{Ch}} \cdot T_M = \text{const}$. Since the total noise energy E_N remains unchanged this must also be true for the total message energy $E_M = P_S T_M = E_B R_B T_M = \text{const}$. Let's assume that we go to 10% of the original bandwidth, but we extend the message duration by a factor of 10, which results in an unchanged total number of transmitted bits. Then we find that the total message energy and the noise energy remain constant, which means in our case 10% of the original signal power P_S for 10 times the original duration.

Because the radio channel is vulnerable to interference, it is necessary to protect the messages to be transmitted. In the case of digital transmission, a powerful method is the use of channel coding to provide protection against transmission errors. Here the very basic principles of equalizing and error protection are briefly reviewed.

PMR systems are narrowband systems, which means that primarily flat or nonselective fading is experienced. Therefore, equalizing the radio channel is very often unnecessary. However, if required, such an equalizer negates the characteristics and distortions of the mobile radio channel, but this requires constant tracking of the varying channel properties. In practice, this is done in TDMA systems by transmitting a known training sequence in each time slot together with the payload and some control information. Once such a burst has been received, mapping from the known sequence to the received and distorted one can be calculated. Applying the inverse function to the payload allows the elimination of those transmission errors caused by channel distortion. Of course, short impulsive transmission errors such as ignition noise cannot be corrected by this method. These are subject to error detection and correction [6, 8, 12, 14, 15, 35].

The principle of error correction is based on additional check bits being added to a given set of code words consisting of a certain number of bits [48–50]. These are called the *redundancy bits*. The resulting code words differ in a number of bits, which is called the *Hamming distance* d_{H}. Figure 5.26 shows clearly the conditions under which a false code word can be detected and corrected. If it is, however, on the border between two valid code words it can be detected as faulty, but without any possibility of correction.

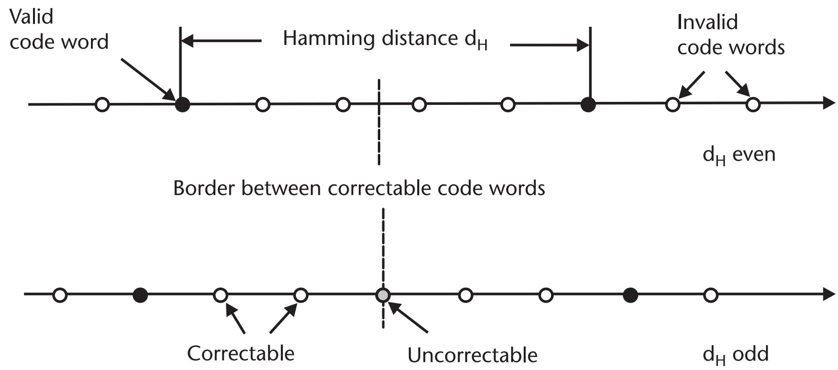


Figure 5.26 The principle of error correction.

A false code word can be detected by comparing it to all valid code words. This can be made easier by the introduction of certain checks such as simple parity check. There is a fixed relation between the Hamming distance d_H of a code and the number of detectable false bits F_D:

$$F_D = d_\mathrm{H} - 1 \tag{5.113}$$

The principle of correction of a faulty code word is simply to replace it with this valid one, which has the smallest Hamming distance. The number of faulty bits F_C that can be corrected is also determined by the Hamming distance:

$$F_C = \begin{cases} (d_\mathrm{H} - 2)/2 & \text{for } d_\mathrm{H} \text{ even} \\ (d_\mathrm{H} - 1)/2 & \text{for } d_\mathrm{H} \text{ odd} \end{cases} \tag{5.114}$$

However, under certain conditions the correction capability of a code breaks down. This obviously happens when the number of errors exceeds half the Hamming distance between valid code words because the replacement by the nearest code word match will fail. Under high errors conditions, there are only a few options for improvement. One is to enlarge the Hamming distance, which means that the useful bit rate decreases if the gross bit rate remains unchanged. Another possibility is to increase the received power by, for example, increasing the transmit power.

The code rate r_C is defined as the percentage of useful bits in a message. It is the ratio of the number of information bits n_Inf to the total number of bits n_Tot and the number of redundant bits n_Red is simply their difference:

$$r_C = \frac{n_\mathrm{Inf}}{n_\mathrm{Tot}} = \frac{n_\mathrm{Inf}}{n_\mathrm{Inf} + n_\mathrm{Red}} \leq 1 \tag{5.115}$$

The consequence is obvious: The lower the code rate, the higher the error correction capability of a code. For very long code words, there is a simple relation between the code rate and the number of correctable errors:

$$F_C \xrightarrow{n_{\mathrm{Tot}} \to \infty} \leq \frac{n_{\mathrm{Tot}}}{2} \cdot (1 - r_C) \tag{5.116}$$

This relation gives an upper bound that is only valid under the assumption that the architecture of the code is suited to the error pattern encountered, which depends on the type of channel and its error mechanisms. Hence, different families of codes are optimized for different purposes and different channel types. There is not enough space in this book to deal in-depth with coding theory; therefore, only the very basic characteristics are summarized.

The major categories for error correction are block codes and convolutional codes. For block codes the information code words are converted into code words of fixed length by adding a fixed number of redundant bits to each original code word. In convolutional codes, the bits of the original bit stream are linked across a fixed distance with one or more adjacent bits, creating therefrom additional redundancy that is added to the bit stream. Both families comprise a large number of different codes with very differing properties.

The mobile radio channel's properties vary considerably and quickly, depending on the environment and the movement of the mobiles. The major reason for the bursty nature of errors in mobile radio channels is that the mobiles from time to time pass through fading nulls where a number of consecutive bits is always lost. One consequence is that none of the simple coding schemes is perfectly matched to the requirements of the mobile radio channel. A well-known practice, therefore, is to use a combination of coding principles. To combat long error bursts, codes with specific burst error correction capabilities or interleaving in combination with block codes can be used [15, 35, 42, 48, 49, 51]. For the latter a number of relatively short consecutive code words are taken as a block. All the code words are written into consecutive lines of a matrix. Then they are read one after the other from the columns of the matrix and transmitted over the radio channel. In the receiver the inverse process is performed. Thus, every error burst is spread into a number of single-bit or symbol errors as Figure 5.27 shows. The correction capability of a block code is only a few bits per code word, but the error bursts occurring during transmission in the radio channel are much longer. Interleaving

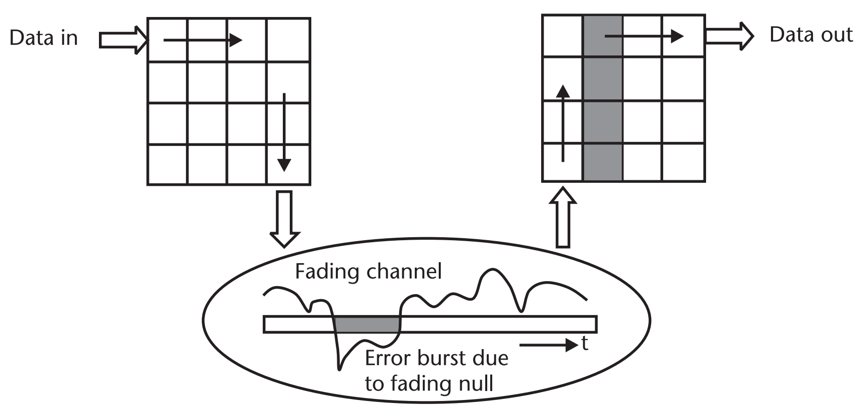


Figure 5.27 Error reordering by interleaving.

causes the burst errors to be more evenly distributed over many more code words but with very few errors per individual code word. Thus, almost all of the affected code words can be corrected and none are lost.

Codes are not only needed for error detection and error protection but also for synchronization. At first, carrier frequency and data clock frequency synchronization have to be achieved by the receiver before proper symbol detection is possible. Therefore, specific codes are additionally embedded in the messages that have outstanding autocorrelation properties to avoid false detection. An example of such codes that have been used for a long time specifically for synchronization are the Barker codes [19, 51]. Detecting such a synchronization sequence allows the proper identification of the beginning of the message and retrieval of its contents. In principle, autocorrelation methods and appropriate codes can also be used to perform clock and frame synchronization simultaneously.

Coding theory and data compression are living working areas where new codes are constantly being created to meet the requirements of new applications. In the past few decades significant progress has been made in both areas and many of the features of contemporary digital radio transmission systems would be impossible without these achievements.

5.9 A Short Introduction to Speech Encoding

Because we cannot afford to waste transmission capacity and bandwidth, it is necessary to apply data compression wherever possible. Text data, pictures, video streams, and music have very differing properties and therefore different compression principles have to be applied. In principle, lossless or lossy compression techniques can be used. The latter usually permit better compression ratios at the expense of quality. Using them leads to a trade-off between compression rate and quality. Here only speech coding as one of the most interesting examples is reviewed.

Digital speech transmission on ordinary PMR channels with adequate speech quality and sufficient error protection was neither technically nor economically feasible before around 1990. The required net data rate, at that time around 16 kbit/s, would have demanded at least 20 kbit/s to allow for some error protection and could have been realized on 20- or 25-kHz channels with reasonable expenditure. However, at a channel separation of 12.5 kHz and below this was not feasible.

Not until new modulation methods and voice coders, with reduced bit rates but very good intelligibility and nearly unaffected speech quality, were developed was the door pushed wide open for the broad introduction of digital technology in mobile radio transmission. In public mobile radio systems, the transition has already been made with the introduction of the GSM system at the beginning of the 1990s while in PMR it really started about a decade later with different proprietary DPMR systems and the APCO 25 and TETRA standards. However, it is a very difficult and complex field, so only the basic principles will be reviewed here [14, 52–56].

In telephony, speech is digitized and transmitted at 64 kbit/s to give good quality using *pulse code modulation* (PCM). Unfortunately, this bit rate results in too large a modulation bandwidth for all modulation schemes suitable for mobile radio channels. For mobile radio communications, only 10% or less of the PCM speech bit rate is acceptable. During the past two decades, considerable progress has been made in reducing the bit rate while maintaining intelligibility and the quality of speech. This has been achieved by thorough studies of human speech synthesis and speech processing in the human brain, which led to very sophisticated methods of removing redundancy. However, the lower the desired bit rate, the more complex the task because a number of serious problems have to be solved as shown in Table 5.11.

The simplest method of speech coding is waveform coding: a straightforward analog-to-digital conversion. However, for sufficient speech quality as in PCM a relatively high resolution and sampling frequency are needed, which results in a high bit rate. Without increasing the bit rate, a considerable quality gain can be achieved by introducing a nonlinear sampling characteristic for level compression. The two slightly different nonlinear characteristics are called the μ-law and A-law, both of which deliver a speech quality equivalent to linear sampling with a resolution of 13 bits per sample instead of only eight [15, 21, 52]. At a coding rate of 64 kbit/s, this delivers the PCM speech signal employed in modern fixed-line telephony systems.

Several simple methods are available for reducing the bit rate to 32 or 16 kbit/s, but the quality decreases in parallel with the bit rate. For most of the mobile radio applications, even that modulation bandwidth would be too high. For DECT, *adaptive differential PCM* (ADPCM) at 32 kbit/s has been selected because at short coverage range a higher bandwidth is affordable. Because ADPCM is a simple and fast method, it meets all of the technical and economical requirements imposed by

Table 5.11 Difficulties with Speech Encoding

Property	*Remark*
Intelligibility	Good intelligibility even in the case of severe channel impairment is needed for most applications.
Speaker recognition	This is indispensable for certain applications, for example, for the police.
Speech quality	For the inexperienced user, any degradation compared to fixed-line telephony is hardly acceptable.
Tandem encoding	Often the occurrence of two consecutive speech encoding and decoding processes cannot be avoided. The three issues above should not be remarkably affected.
Suppression of environmental noise	This feature is required to get an improved speech clarity even in noisy environments.
Recognition of environmental noise	Sometimes environmental noise must be clearly identifiable; for example, policemen might need to distinguish a shot from a bang caused by a backfire. This clearly contradicts the preceding point.
Robustness against transmission errors	Faulty and uncorrected bits should not cause annoying audible tones or other undesirable acoustic events.
Processing time and transmission delay	If these become too large, then the flow of communication is impaired. In particular, tandem hops with separate encoding and decoding are very critical in this respect because they may double the overall delay time.
Restricted computational power	Excessive need for computational power requires high power consumption, leading to unacceptably short battery duty cycles in portables.

cheap digital cordless telephone systems. Further transmission rate reductions with methods such as ADPCM or *delta modulation* (DM) to 16 kbit/s and below delivers an unacceptable speech quality.

To achieve considerably lower bit rates together with good speech quality, parametric coding is a much more powerful approach. Only a limited number of parameters are used to describe speech with sufficient accuracy. These parameters vary only slowly. Hence, instead of digitizing and transmitting sample after sample, it is easier to split the signal into intervals of usually 20–30 ms to measure its main parameters and to transmit only the results.

The very first vocoders were very simple realizations of this principle. They split the voice band with the help of a filter bank, transmitted only the amplitude of the individual bands, and used this to control the amplitude of a set of generators on center frequencies of the speech analysis bands to reconstruct the speech signal. Thus, speech parameters such as pitch frequency, some of its harmonics, noise characteristics (for example, "color"), and the power level are transmitted. Because the human ear does not evaluate the phase of a tone—only phase differences to locate the source—and does exhibit additional masking effects, certain parameters need not be transmitted. Moreover, it does not matter very much if there are small differences between the original pitch frequency and its harmonics and the generator frequencies. All of this permits a remarkable reduction of the bit rate without sacrificing too much of the speech quality.

Knowing the set of parameters used for the previous interval, the parameters of the following interval can be predicted. This *linear prediction coding* (LPC) permits a further reduction of the information to be transmitted. Comparison of the prediction with the actual signal allows the evaluation of the prediction error. Because the prediction can also be made on the receiving side, only this prediction error needs to be transmitted. The GSM full-rate coder uses an advanced method of this kind, called *regular pulse excitation-long term prediction* (RPE-LTP), which provides nearly PCM quality.

Another powerful method is the use of code books. As before, the speech is split into samples from which the speech parameters are extracted. Samples of parameters are treated as standardized patterns, which are replaced by codes, for example, simple numbers. This is possible because only a limited number of such patterns are necessary to reconstruct speech with sufficient quality. Because the codes need less information capacity compared to the speech parameter set characterizing one block of speech, a further reduction of the transmission rate results without affecting speech intelligibility and speech quality too much. For example, the code book principle is used for TETRA.

Much work is currently being done to achieve a further reduction of the necessary bit rate while maintaining good speech quality. Today bit rates from 2 to 4 kbit/s are feasible with astonishing quality. The latest approach uses multirate speech encoders. These provide high speech quality under good receiving conditions; however, under bad propagation conditions, they provide a reduced rate with slightly lower quality but better error protection. Instead of a breakdown of the link, slightly reduced speech quality results temporarily with a negligible effect on intelligibility when the bit error rate increases due to channel distortion. In many cases this is hardly noticeable.

It is interesting to compare the speech quality of the GSM full-rate speech encoder at 13 kbit/s with that of TETRA running at about 4.6 kbit/s. At roughly one-third of the bit rate, the TETRA speech coder performs nearly as well as the GSM coder. The latter has been rated at a *mean opinion score* (MOS) of slightly below 4, whereas the TETRA coder has been rated at better than 3.6. MOS is a subjective measuring result obtained from the judgment of a sufficient number of persons having compared uncoded speech samples with coded ones. MOS = 4 means excellent quality with imperceptible impairment; MOS = 3 means good quality with just perceptible impairment, but not annoying. This demonstrates the progress achieved in speech coding in somewhat less than one decade [56, 57]. For DIIS a dual-rate speech codec is envisaged running at roughly 2 or 4 kbit/s depending on the channel quality. At the higher bit rate it is comparable with the quality of the TETRA codec.

In comparing analog to digital speech transmission, an important feature is shown in Figure 5.28 as was pointed out a long time ago [34, 46, 58]. Note that the diagram has to be read keeping in mind that the horizontal axis exhibits a logarithmic scale. Very near the base station, a well-designed analog system will provide superior speech quality, but due to environmental noise this will not be noticed in most cases. For about two-thirds of the coverage distance or 90% of the covered surface, however, digital speech shows better quality. Unfortunately, digital speech quality decreases rapidly beyond the coverage limit, whereas analog speech shows a much more graceful degradation. With multirate coders, graceful degradation can also be achieved, but these will only be implemented in future systems. In conclusion, we can state that overall digital speech quality is now superior to analog even at a fixed coding rate.

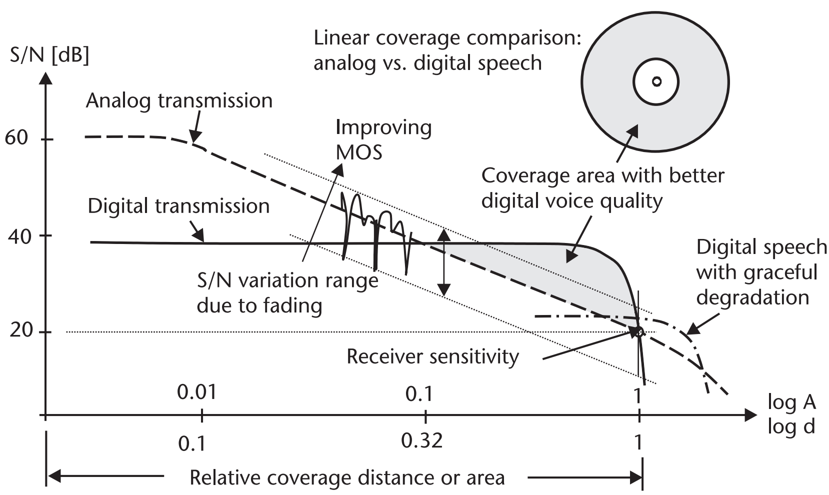


Figure 5.28 Comparison of analog and digital voice transmission.

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CHAPTER 6

An Overview of TETRA

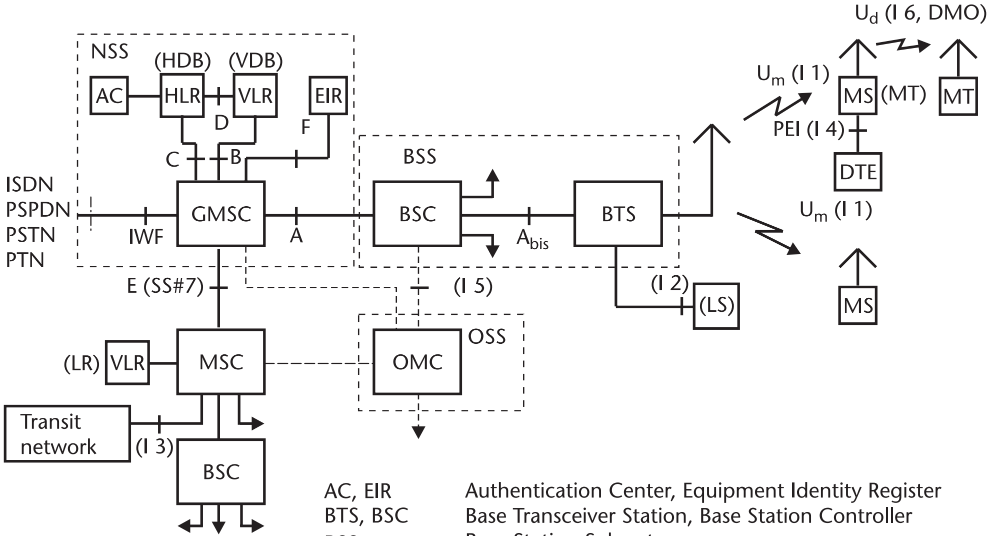
Many public and large organizational professional users of PMR systems are contractually obliged to select a standardized system, or they may do so to ensure a broader choice of products from different manufacturers. TETRA is a true European digital *standard* for trunked PMR systems as opposed to other *proprietary* digital PMR systems. It is intended primarily for professional use and will be introduced in all European countries.

Many TETRA systems are already in operation or are being rolled out now in the early 2000s. Moreover there are strong indications that TETRA is another standard for mobile communication that is going to spread across the world as did GSM. Therefore, we now look at this system in greater detail. However, it is not our intention to give a complete description of TETRA but to introduce its main capabilities to the reader and to explain in particular the basics of the air interface because this is always the bottleneck in every digital PMR system.

6.1 General Structure of Modern Digital Mobile Radio Systems

The basic structure of all modern digital radio communication systems is very similar as can be shown by comparing GSM and TETRA. Their comprehensive sets of features can only be handled by a clear and systematic system architecture. Figure 6.1 shows the general structure of modern digital mobile radio communication systems [1, 2]. This structure was created for the GSM system as the mobile extension of the *Integrated Services Digital Network* (ISDN). Later it was modified somewhat to tailor it to the needs of TETRA. To be honest, GSM was not the original starting point. Similar but simpler architectures had been introduced much earlier with NMT and, even before that, the fundamentals of cellular telephony were developed in the AMPS system.

Unfortunately, the network elements in GSM and TETRA have different names even if they have the same or similar tasks. The list in Figure 6.1 gives the names of the different GSM network elements from which their main tasks can be derived. They also apply to TETRA except where a different name appears in brackets. The standardized interfaces between GSM network elements are also shown in Figure 6.1. Only at these interfaces can network elements from different manufacturers be connected to each other without endangering the proper operation of the whole system. Despite the similar network structures of GSM and TETRA, some of the standardized interfaces are located between different network elements. One



Denominations according to GSM, TETRA names in brackets

AC, EIR	Authentication Center, Equipment Identity Register
BTS, BSC	Base Transceiver Station, Base Station Controller
BSS	Base Station Subsystem
DTE, LS	Data Terminal Equipment, Line Station
GMSC, IWF	Gateway Mobile Switching Center, Interworking Function
HDB, VDB	Home Data Base, Visitor Data Base
HLR, VLR	Home Location Register, Visitors Location Register
ISDN, SS#7	Integrated Services Digital Network, Signaling System #7
LSC, MSC	Local Switching Center, Mobile (Main) Switching Center
MS, MT	Mobile Station, Mobile Terminal
NSS, LR	Network and Switching Subsystem, Location Register
OMC, OSS	Operation & Maintenance Center, Operation Subsystem
PEI	Peripheral Equipment Interface
PSPDN	Packet Switched Public Data Network
PSTN	Public Switched Telephone Network
PTN	Private Telephone Network (PABX, LAN, DECT)

Figure 6.1 General structure of digital mobile radio communication systems.

interface, however, is unique to TETRA and this is the air interface for *direct mode operation* (DMO), which simply does not exist in GSM.

Specifying a complex mobile radio communication system such as TETRA is a very comprehensive task. All of the different elements have to be dealt with separately, and their interfaces have to be defined very accurately to ensure correct operation, particularly if network components from different manufacturers have to be interconnected. Therefore, six interfaces, I 1 to I 6, were specified for TETRA. For the user only the air interfaces I 1 or U_m between BS and MT for *trunked mode operation* (TMO), I 6 or U_d for DMO between mobiles and the versatile *peripheral equipment interface* (PEI) I 4 are directly accessible. The PEI connects the radio terminals for data transmission with laptops, dedicated data terminals, and so on. Various terminals are also linked to a mobile radio system, for example, mobile stations, remote telephone sets via fixed lines, and dispatcher stations in *operating and maintenance centers* (OMCs). Therefore, the wire-line interfaces

I 2 between *line station* (LS) LS and BS and I 5 between BSC and OMC have been created. However, I 2 and I 5 and the line station were never standardized, so all interfaces within the *switching and management infrastructure* (SMI) are proprietary.

If a communication runs from one radio terminal to another via two different TETRA networks, then the two networks are linked by another wire-line interface I 3 serving as an *intersystem interface* (ISI). If both TETRA networks are remote and have no common ISI, then the communication may use the PSTN or the ISDN or another network as a backbone. If several TETRA networks are interconnected, then the whole the structure will become extremely complex. Each single TETRA network may be comprised of a number of BS controllers, each of them connected to several base stations. Local or remote dispatchers will control the operation of mobiles in their systems. Additional tasks have to be performed over and above mobility management, for example, call and encryption key management and billing. Also there are gateways to the PSTN, ISDN, and PDN as well as connections to system engineering, operation, and maintenance management terminals and networks. However, note that complex modern network architectures are increasingly based on packet transmission and IP network structures instead of a circuit switched topology.

6.2 The ISO-OSI Reference Model

Complex hardware or software systems have to be well and cleanly structured; otherwise, testing, debugging, operation, maintenance, and further development of such systems would be impossible. This is why the layered *International Standardization Organization—Open Systems Interconnection* (ISO-OSI) reference model was invented to give guidance to the creation and operation of terrestrial telephone systems. In principle, the same approach suffices for mobile communication systems. Figure 6.2 shows the layered structure of a modern digital communi-

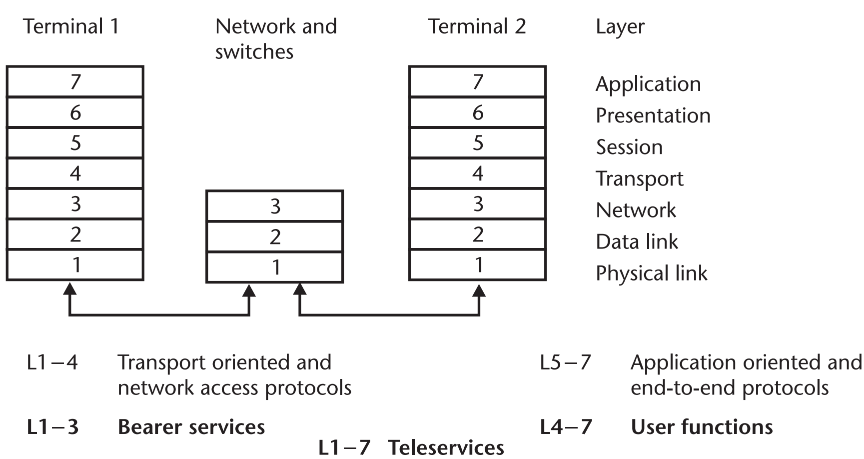


Figure 6.2 The ISO-OSI reference model.

cation system applied to the case for which end-to-end communication is ongoing between two terminals.

Inside the network only the lower three layers are needed to set up, maintain, and release a communication link. The upper layers are required in the terminals to interface applications and control sessions. However, due to the need for mobility management of the subscribers and other new tasks, which were not present in conventional terrestrial wire-line telephone systems, the complexity of mobile radio communication systems is considerably greater. The detailed explanations of the tasks of each of the first three layers in Table 6.1 correspond to TETRA [3–6]. The specific tasks of the different layers and how layers 2 and 3 are subdivided to guarantee smooth cooperation by the different protocol entities required to perform the particular tasks are also shown.

For the definition of the different services, the distinction between bearer services and teleservices as shown in Figure 6.3 has to be made. A bearer service offers a certain technical capability between two interfaces of a communication link. It comprises only layers 1 to 3 of the OSI reference model and provides a means to carry various teleservices, which are accessible at the *man/machine interface* (MMI). The latter are based on bearer services comprising network functions in layers 1 to 3 plus additional user functions in layers 4 to 7. An example for a bearer service is circuit mode data, which can carry different data services, whereas individual data calls or a data group calls are examples of teleservices.

6.3 TETRA Standardization

In the beginning within ETSI a subcommittee of the *Technical Committee Radio Equipment and Systems* (TC RES), called STC RES 06 and inaugurated in 1987, was responsible for the standardization of a new digital PMR system called *Mobile Digital Trunked Radio System* (MDTRS) but later this clumsy name and the tongue-tying acronym were replaced by TETRA. In the 1990s, the entire ETSI structure was reorganized and Project TETRA (EP TETRA) was formed to carry on the TETRA standardization. Similarly GSM became EP GSM. A liaison group is responsible for relations with other ETSI bodies, and specific limited tasks with clear objectives are performed by *project teams* (PTs). However, the main part of the TETRA standardization work is done within working groups of EP TETRA.

TETRA is called a standard, however, in reality it is a bundle of five separate core standards. Each standard contains several parts, because such a comprehensive system is much too complex to be put into one single specification. The standard EN 300 392 called *TETRA Voice + Data* (V+D) specifies the basic parts for voice and data transmission on radio channels with a separation of 25-kHz employing the TDMA principle with four traffic channels per carrier frequency. For packet data transmission ETS 300 393 has been specified (*packet data optimized*, PDO). Defining a complex system is one thing but thorough conformance testing is another. Therefore, it has been specified in a separate standard EN 300 394. Speech coding is separately specified in EN 300 395 because the achieved data compression rate needs a complex algorithm called *algebraic code excited linear prediction* (ACELP) with a high level of computational complexity. Lastly there are specific

Table 6.1 Description of ISO-OSI Layer Tasks for TETRA

Layer	*Denomination*	*Short Description*
7	Application layer	User software and network services such as voice transmission, file transfer, and e-mail
6	Presentation layer	Data representation including end-to-end encryption and format conversion
5	Session layer	Network addressing, session management, dialog control, and so on
4	Transport layer	End-to-end transmission control, GoS, and so on
3B[1]	Upper network layer (subnetwork access functions or SNAFs)	*Mobility management* (MM) with location area selection, registration, authentication, user attachment, and network selection *Circuit mode control entity* (CMCE) with connectionless short data service (SDS) handling, call control (CC) including addressing and control of supplementary services (SS) *Packet data* (PD) handling for packet data services *Packet Data Protocol* (PDP) for IPv4 and IPv6 using the Sub-Network Dependent Communication Protocol (SNDCP)
3A[1]	Lower network layer (MS/BS link)	*MS/BS link control* (MLE/MLB) with protocol discrimination, MS/BS connection management, identity management, QoS selection and mobility within a registered area, broadcast information handling
2B[2]	Connection control layer (logical link control or LLC)	Point-to-point logical link handling and advanced links for improved QoS on demand with exchange of control and user data, scheduling of data transmission, error measurements and retransmissions (ARQ), flow control and acknowledgment of received data, logical channel allocation
2A[2]	Access control layer (medium access control or MAC)	Interleaving/deinterleaving over $N = 1$, 4, or 8 blocks, channel encoding/decoding with FEC and CRC, radio channel access control with frame synchronization, random access procedure, fragmentation of service and data units, multiplexing of logical channels and multiframe building and synchronization, radio resource management with bit and block error (BER and BLER) measurements, path loss calculation and scanning of adjacent cells, address management for individual, group, and broadcast calls, power control management, radio path establishment, radio resource allocation (frequency and slot), buffering of control and user information, and circuit mode speech and data applications
1	Physical layer (PL for bit transfer)	Modulation/demodulation, Tx/Rx switching, frequency/channel setting, radio signal strength (RSSI) evaluation, frequency correction, power control (MS only), symbol and burst synchronization, burst building, flag encoding/decoding, scrambling

[1]Layer 3 is subdivided into layers 3A and 3B, C-plane only.
[2]Layer 2 is subdivided into layers 2A and 2B.

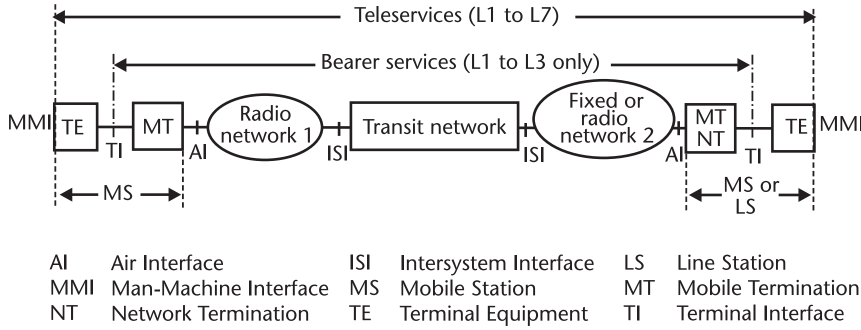


Figure 6.3 Definition of bearer services and teleservices.

problems related to TDMA in the case of direct mobile-to-mobile communication. All of the necessary means have therefore been specified in the TETRA DMO standard EN 300 396 [5, 7–10]. All ETSI standards were formerly called *ETSI Telecommunication Standards* (ETSs), but most of them have now been transferred to *European Norms* (ENs).

Additionally a sound introduction and comprehensive description of TETRA has been compiled in Parts 1 to 5 of an ETSI technical report called the *Designer's Guide*, ETR 300-1 to 300-5 [6, 11–14]. Complementing the technical and regulatory requirements, specifications and information concerning MMI, *subscriber identity module* (SIM) cards and interface, emergency access, repeaters, managed DMO, security aspects, EMC properties, essential requirements under the R&TTE Directive, and many other issues have been published in various additional ENs, *Technical Basis for Regulations* (TBRs), and ETRs. The whole bundle is in excess of some 200 different documents.

Originally two additional TETRA variations were planned for V+D and DMO, both for 12.5-kHz channels. However, these were skipped after the decision was made that separate frequency bands with 25-kHz carrier spacing and sufficient bandwidth for TETRA would be made available in Europe [15, 16]. Finally, a quasi-FDMA solution (or better described as a single time slot TDMA version) called PMR 6 was proposed with a channel separation of 6.25 kHz, but this has not been pursued, although it would have covered all three main channel separations in use in Europe because it would have fitted nicely into the 12.5- and 25-kHz raster, and three of these systems could also share one 20-kHz channel, which is necessary, for example, in Germany and some neighboring countries [2]. For the second phase of the APCO 25 project in the United States, 6.25-, 12.5-, and 25-kHz versions of TETRA have also been considered.

To promote TETRA internationally and to perform additional tasks outside the scope of ETSI, the TETRA MoU Group was created in December 1994. There coordination of various manufacturers' activities takes place, for example, for conformance testing and type approval issues. It is also a forum for regulators and user groups to address specific TETRA issues and to exchange operational experiences.

One of the main goals of the TETRA MoU Group is to ensure that products from different suppliers are compatible and that TETRA is a real multivendor standard. Over a period of time this has been proven by many tests where mobiles and network elements from different sources have successfully worked in mixed systems. Early field tests in the Channel Islands on Jersey and elsewhere in the second half of the 1990s have shown that the TETRA technology is on the right track even if the full functionality was not available in these trials [17]. The results of the pilot system tests in Berlin during 1997 and 1998 and in Aachen from 2001 to 2003 are of paramount importance for public safety forces in Belgium, Germany, and the Netherlands where a major goal is to achieve smooth and efficient trans-border cooperation. The success or failure of this three-country field trial will be decisive for the further strategy of all parties involved in the Schengen treaty. In addition to these tests, a considerable number of big TETRA networks for civil applications and PAMR have begun operation. A large PAMR system in Austria and several large public safety systems like VIRVE in Finland, ASTRID in Belgium, and ADONIS in Austria are some important examples. These have clearly shown that TETRA fulfills the user expectations and considerably exceeds the features of older digital systems.

6.4 General Properties and the Physical Layer of TETRA V+D

To understand how a TETRA network performs its numerous tasks, we now discuss the structure of the lower three layers and the services offered by TETRA in more detail. The TETRA protocol stack is subdivided into two main parts: the user plane and the control plane. The first is responsible for the transport of the user data, while the second is needed to control the communication link as Figure 6.4 shows [1, 5, 6, 11].

The control plane is much more complex than the user plane and comprises various functions located in layers 2B and 3. The user data, however, accesses layer 2A directly. For both planes, only layer 1 and layer 2A are identical. These two layers have no knowledge of the data or signaling content they are handling.

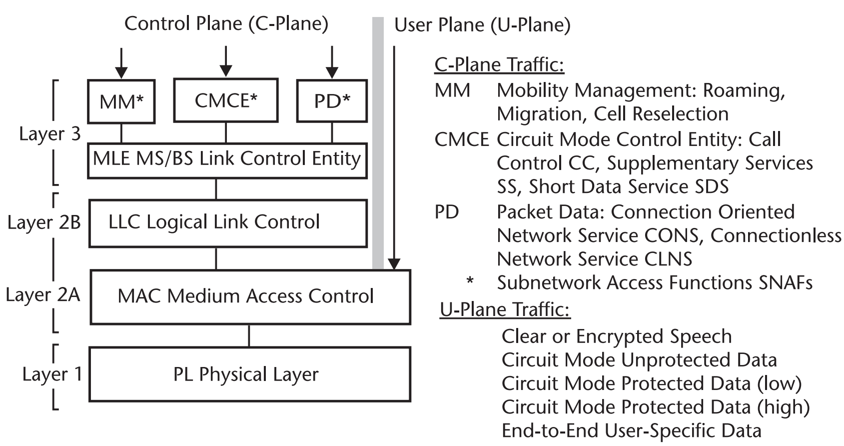


Figure 6.4 The protocol stack of TETRA V+D (TMO).

Initially, TETRA was envisaged to operate in the 400-MHz range, but it was soon realized that all of the PMR bands in Europe were very crowded, it would have been a major problem to find sufficient bandwidth for TETRA allocations, and the different European national administrations had major difficulties to define common frequency bands for TETRA. Hence, TETRA has to match the properties of all frequencies from 150 to 900 MHz and even higher. Eventually CEPT ERC (now CEPT ECC) fixed in its Recommendation T/R 22-05 the general frequency allocations for digital trunked radio systems in the CEPT countries (see Table 6.2). Much more difficult to achieve were the decisions on DMO for TETRA Emergency

Table 6.2 Basic Properties of the TETRA V+D Air Interface

Property	*Description or Remark*
Frequency bands[1]	*Public safety*[2]*:* 380–385 paired with 390–395 MHz DMO[3]: 380–380.15 and 390–390.15 MHz *Civil applications:* 385–390 MHz paired with 395–400 MHz[2] 410–430 MHz, 450–470 MHz, 870–888 MHz paired with 915–933 MHz[4] DMO[1]: 445.2–445.3 MHz (simplex mode) Various applications: 806–825 MHz paired with 851–870 MHz[5]
Carrier separation	25 kHz
Duplex separation	10/45 MHz (400/900-MHz range)
Mode of operation	Simplex (DMO), half-duplex and duplex (TMO)
Channel access	4TDMA, time slot concatenation possible
Modulation	$\pi/4$-DQPSK
Modulation rate	18 ksymbols/s equal to 36 kbit/s
User data rates	(1 to 4) × 7.2 kbit/s transparent (unprotected) (1 to 4) × 4.8 kbit/s good error protection (1 to 4) × 2.4 kbit/s strong error protection
Speech coding	ACELP[6] at 4.567 kbit/s, 7.2 kbit/s including FEC
Transmitter power and power control range[7]	BS: 28–46 dBm MS: 25–45 dBm HP: 15–35 dBm
Adjacent channel power	–60 dBc (–55 dBc for $P_T \leq 30$ dBm and above 700 MHz)[7]
Receiver sensitivity[8]	BS: –106 dBm typical with Rayleigh fading MS: –105 dBm typical with Rayleigh fading HP: –103 dBm typical with Rayleigh fading
Cochannel rejection	≥–19 dB with Rayleigh fading
Adjacent channel rejection	≥57 dB with Rayleigh fading
Symbol duration	55.6 μs or 2 bits
Time slot duration	14.167 ms or 510 bits
Frame duration	56.67 ms or 4 time slots
Multiframe duration	1.02 seconds or 18 frames
Hyperframe duration	61.2 seconds or 60 multiframes
Call setup time	<300 ms for two mobiles within one cell[9]
Signal delay	<200 ms including speech processing delay

[1]In Europe.
[2]In European NATO countries.
[3]National deviations permissible.
[4]Seldom in Europe.
[5]Non-European countries only.
[6]*Algebraic code excited linear prediction.*
[7]See Table 6.3, MS power classes 4 and below.
[8]See Table 6.4.
[9]This can be <500 ms nationwide with modern IP-based procedures.

and the allocation of a Trans-European DMO band for TETRA Civil, which came as late as in 2001. The latter has been allocated between the two main European civil TMO bands so that TETRA terminals in both bands can use it throughout Europe. For unrestricted licensing and free circulation of TETRA terminals in Europe, further regulatory provisions were necessary [6, 15, 16, 18–23].

Additionally in major countries of Europe parts of existing military frequency allocations could be freed. The introduction of TETRA within this assignment area calls for some stringent preconditions because it remains shared spectrum. Only parts of the so-called "NATO band" between 380 and 400 MHz could be used and then only for public safety, at least in many countries, and there must be only one *common* standard. At present, the situation looks somewhat different because in many European countries TETRA will be introduced for this purpose, but other countries prefer TETRAPOL.

For civil TETRA applications in most of the European countries part of the band from 410 to 430 MHz with a duplex separation of 10 MHz has been made available. Additionally, there will be TETRA allocations with 10-MHz duplex separation in the region between 450 and 470 MHz, which is currently allocated to analog cellular systems (NMT-450 and C 450). Some of these have already been closed or will be shut down in the near future.

Many European countries are short of spectrum in the 400-MHz range but are able to offer frequencies in the 900-MHz region with a duplex separation of 45 MHz, in particular, 870–876 MHz paired with 915–921 MHz, where the upper band borders were extended later to 888 and 933 MHz. Many overseas markets are also suffering from spectrum shortages in the lower bands and are therefore pushing TETRA development in higher bands, for instance, at 806–825 MHz paired with 851–870 MHz. In addition to the frequency ranges, Table 6.2 also lists the basic properties of the TETRA V+D air interface.

TETRA employs four-level modulation with a symbol rate of 18 ksymbols/s resulting in a modulation bit rate of 36 kbit/s. Hence, each of the four time slots of a frame provides a gross bit rate of 9 kbit/s. This means that without synchronization and burst building overhead, an unprotected bit rate of 7.2 kbit/s per traffic channel can be provided. If this transparent channel is used for data, the onus for error protection is placed on the user. Within the standard, good error protection at 4.8 kbit/s is available and for specific cases needing strong error protection a bit rate of 2.4 kbit/s can be provided. All time slots of a frame can be used separately, which means that up to four independent traffic channels can be established per carrier. However, on demand, two, three, or four time slots can be combined and assigned to a single user if a higher transmission capacity is needed. Moreover it is possible for one user to transmit voice and data *simultaneously*.

TETRA is extremely flexible and offers many functions that are not always needed. In many cases the capabilities can be reduced without any noticeable drawback. For this purpose, different classes have been created for particular functions of mobiles. This concerns transmitting power, receiver classes, multislot operation, duplex capabilities, support of concurrent calls, and end-to-end encryption. The reason is to reduce the expenditure by deleting unnecessary functions in networks where they are not needed. For the same reason there are also different power classes for BS transmitters.

TETRA V+D employs MS power control and for each call the power is held within given limits but reduced to the necessary minimum. Unlike most cellular networks, the mobile makes the decision regarding which power level is to be used based on BER and the field strength of the received signal (that is, the *received signal strength indicator,* or RSSI). These criteria are also used for handover decisions, in combination with broadcast information from the base station. Several power classes have been defined as shown in Table 6.3.

According to the distance from the base station, the peak power at mobiles can be varied dynamically between 15 and 45 dBm in steps of 5 dB. The BS transmitting power can be varied between 28 and 46 dBm in steps of 2 dB and is fixed at a level individually optimized for each cell. In particular, the highest BS power levels are only needed for systems with maximum coverage. In cases where high traffic density is needed, sometimes the cell size will be limited by the maximum traffic and not by the maximum coverage distance. Then lower peak power suffices, and cheaper power amplifiers with reduced peak power can be installed at the base stations.

For mobiles the limiting factor in many cases is the battery capacity. This is not so much the problem with vehicle-borne mobiles but with handhelds, which otherwise become too heavy and too expensive. Penultimately the transmitting power of handhelds should be limited to avoid possible health hazards, but this specific topic will be further discussed in Chapter 9. For all of these reasons, handheld transmitting power usually does not exceed about 6W. Last but not least, the use of power control reduces the overall interference within a TETRA system and also in adjacent frequency bands.

Different receiver classes are optimized for different environments and propagation conditions. Specifically, class E receivers are needed for the downlink of simulcast systems exhibiting large delay spread up to 100 μs. For less demanding propagation conditions, the simpler and cheaper class A or B receiver design suffices. The first one is for urban, hilly, or mountainous terrain with vehicle speeds up to 200 km/hr, whereas the latter is for urban and built-up areas with vehicle speeds up to 50 km/hr.

Table 6.3 Power Classes of TETRA V+D

Power Class	*MS Peak Power*[1] *[dBm]*	*[W]*	*BS Peak Power* *[dBm]*	*[W]*
1[2]	45	30	46	40
2[1]	40	10	44	25
3[1]	35	3	42	15
4[1]	30	1	40	10
5[1]	25	0.3	38	6.3
6	—	—	36	4
7	—	—	34	2.5
8	—	—	32	1.6
9	—	—	30	1
10	—	—	28	0.6

[1]Additional classes 2L to 5L defined for DMO where each L class is 2.5 dB below the corresponding main class.
[2]Not defined for mobiles in DMO.

The sensitivity of TETRA receivers depends on the receiver class, the propagation conditions, and the bearer service in question. Table 6.4 gives a brief summary for the worst case, which is essentially HT200. The table must be interpreted as the given BER being the maximum value for the worst receiver class at the reference sensitivity. This means that, in practice, the BER at the reference sensitivity level is usually well below that limit.

Several typical propagation profiles have been selected for test purposes. These are static, TU50, BU50, HT50, HT200, RA 200, and EQ200. Of course, static means AWGN conditions, whereas TU is typical urban, BU is bad urban, HT means hilly terrain, and RA is rural area. These profiles depend not only on the environment but also on the mobile speed denominated by the associated number. Hence, BU50 means bad urban conditions at 50 km/hr. EQ200 is an artificial test case for equalizers that presents a signal to the receiver that is composed of six independently fading signals with equal average power but evenly distributed over a total delay spread of 100 μs.

Multislot operation that is dependent on the number of slots to be bundled requires duplex capabilities and both have a major influence on the design and production costs of TETRA mobiles. Table 6.5 shows the various possibilities but what really is available also depends on the infrastructure and the services offered.

Before the details of the structure of the physical layer of the TETRA V+D air interface are explained, the basic structure of the physical transmission channel on the bit level should be discussed. The TETRA modulation is $\pi/4$-DQPSK, and one TETRA carrier occupies a modulation bandwidth that fits into a (channel or better) carrier separation of 25 kHz.

At the transmitter the incoming bit stream is differentially encoded and mapped on to I and Q channels. In both channels premodulation filtering is applied to create the appropriate shape for every symbol. Then both channels are fed into a vector modulator that delivers the modulated four-state RF carrier signal ready for transmission (see also Figures 5.12 and 5.14). The channel filtering in TETRA follows a *raised cosine* (RC) characteristic with a roll-off factor of $r_O = 0.35$. Half the filter is incorporated in the transmitter and the other half in the receiver for optimal sensitivity. Hence, both are *square root RC* (SRRC) filters that together

Table 6.4 Dynamic Reference Sensitivity of TETRA Receivers

Channel Type[1]	*BER [%]*	*Receiving Reference Sensitivity Level [dBm]*[2] *MS*[3]	*BS*[4]
TCH/7.2	≤4.0	–103	–106
TCH/4.8, $N = 1$	≤4.0	–103	–106
TCH/4.8, $N = 4$	≤3.3	–103	–106
TCH/4.8, $N = 8$	≤3.0	–103	–106
TCH/2.4, $N = 1$	≤1.1	–103	–106
TCH/2.4, $N = 4$	≤0.4	–103	–106
TCH/2.4, $N = 8$	≤0.13	–103	–106

[1]TCH/*x.y* means traffic channel for *x.y* kbit/s user bit rate and N is the number of interleaved time slots.
[2]For dynamic conditions except EQ200.
[3]Noise figure 9.4 dB.
[4]Noise figure 6.4 dB.

Table 6.5 MS Duplex, Multislot Operation, and Encryption Properties

Property	*Description and Characteristics*
Single-slot operation	This is mandatory for all mobiles on nonoverlapping and not adjacent time slots, Rx switching time ≤1 time slot; hence, simplex in two nonconsecutive time slots on one carrier (DMO) and half-duplex and TDD on 1 time slot pair on two carriers (uplink and downlink, TMO) is possible.
Multislot operation	This is optional; transmission and reception are possible simultaneously. Hence, all four time slots on one carrier or a pair of carriers (uplink and downlink) can be used. Extended bit rates up to 28.8 kbit/s are possible even in duplex mode and up to four concurrent calls can be handled.
End-to-end encryption	An option that only impacts the U-plane (provided that external encryption devices are used and no stealing mechanisms or other TETRA functions invoking the C-plane are needed).

form an RC channel filter. This filter enlarges the bandwidth of a si-pulse by 35% compared to a lowpass filter with infinite steepness but it suppresses very well pulse ringing and ISI between adjacent symbols as already shown in Figure 5.6. Thus, the degradation due to tolerances and propagation impairments is considerably reduced without enlarging the modulation bandwidth excessively.

The modulated RF signal then passes through the transmission channel where it is impaired by multipath propagation and interference from other RF sources. At the receiver the incoming signal is applied to a coherent vector demodulator. It is necessary to recover the precise frequency and phase of the carrier independently. This is performed by separate synchronization circuitry. Even under unfavorable reception conditions, the recovery of the carrier is possible with greater precision than detection of each single symbol because synchronization is based on a chain of appropriate symbols where averaging reduces phase and frequency errors and symbol jitter. After recovering the I and Q channels, filtering with another SRRC filter (the second half of the channel RC filter) is applied to reduce noise and to optimize the overall response of the TETRA channel. The I and Q symbols are then evaluated by decision circuitry and the original bit stream reconstructed.

The transmission of a TETRA signal requires the sending of a lot of control information in addition to user data. The information for the different control purposes is put into various logical channels. The user may wish to transmit speech or data and the latter may be at different data rates, formats, and error protection. Different logical channels are used for all of these different formats. The logical channels are error protected by different coding schemes tailored to the needs of the specific data format or control function. Moreover, there are different coding procedures, which are applied sequentially. As mentioned earlier, such multilayered coding provides improved protection against transmission errors even under fast and widely changing propagation conditions.

In the transmitter at first block coding is applied to the different logical channels carrying user data or signaling information at different speeds. These channels are described later in more detail. Then a *rate-compatible punctured convolutional coding* (RCPC) is performed and followed by an interleaving scheme over $N = 1$, 4, or 8 blocks to improve protection against long deep fades. The convolutional coding permits *forward error correction* (FEC) while the block coding detects any residual errors in the message if the FEC has failed to correct them. Finally, a scrambling procedure is applied.

There are two exceptions to this general scheme. The *access assignment channel* (AACH) is only protected by a shortened *Reed-Muller* (RM) code and the transparent 7.2-kbit/s channel TCH/7.2 (traffic channel with 7.2 kbit/s) has no protection at all. However, scrambling is applied to both. After the different coding and scrambling stages, multiplexing, burst building, and modulation are performed to produce the final transmitter signal. As shown in Figure 6.5, in the receiver these operations are performed in the reverse order: demodulation, burst disassembling, demultiplexing, decoding, and error correction of the different logical channels.

For speech coding a different scheme is used. The data bits from two 30-ms speech frames generated by the ACELP encoder at 4.6 kbit/s are assembled into one 14.17-ms transmitter burst. All bits are reordered into three classes 0 to 2 according to their importance and their impact on the speech quality should errors occur. Class 0 bits are transmitted without additional coding, whereas class 1 bits are RCPC encoded at 2/3 rate and the most critical class 2 bits are RCPC encoded with 4/9 rate with additional tail and CRC encoding.

One specific problem is related to this structure. Because some of the channels are not transmitted very frequently, some time delay is always experienced, which means that buffering of all received data is needed and that consequently there is always a certain amount of latency from when an event occurs until the transmission of associated control information. However, most of these delays are small enough that they are hardly noticeable.

The very basic elements of the TETRA frame structure are the four-level $\pi/4$-DQPSK symbols. The modulation bit rate of 36 kbit/s is equivalent to a

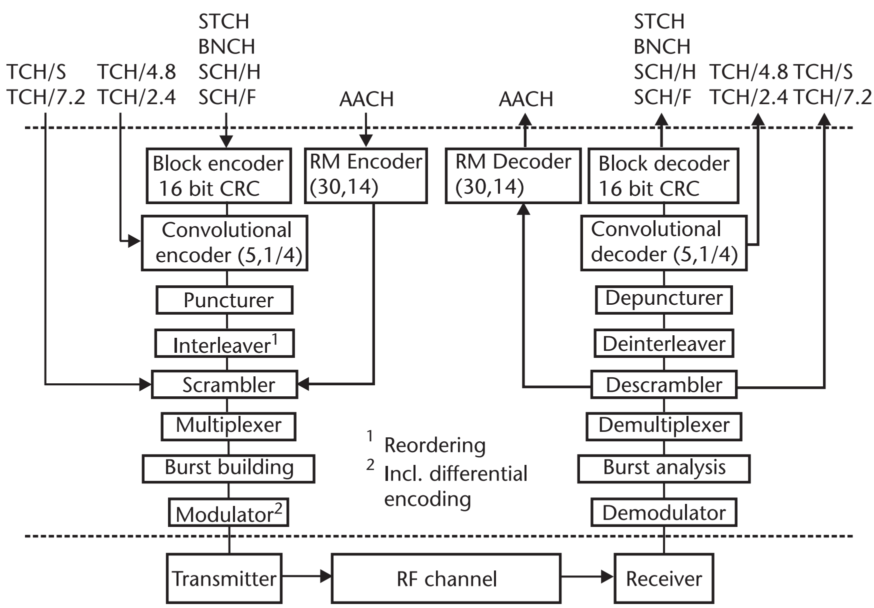


Figure 6.5 The different steps of coding in the complete TETRA channel.

symbol rate of 18 ksymbols/s, which results in a symbol duration of 55.56 μs. The transmission of information is performed in packets of symbols called *bursts*. There are five different basic burst formats serving different purposes and Figure 6.6 shows their structure.

The main features can be explained by looking at the normal MS to BS burst. In the center of each burst, there is a known training sequence comprised of 22 bits that is used for synchronization and the evaluation of the ever-changing transmission channel properties. (TETRA does not usually require equalization but the standard does specify such a mobile for use in difficult conditions.) At the middle of the burst, the channel is nearly stable but with increasing distance from the training sequence the channel synchronization accuracy gets worse and more errors may occur. However, compared to the case with the training sequence at the beginning of the burst the maximum departure of the channel properties compared to the time interval occupied by the training sequence would be only half. The useful information is carried in two blocks of 216 bits each in *subslot numbers 1 and 2* (SSN1 and SSN2) before and after the training sequence and is protected by appropriate coding. At the beginning of the burst, there are the tail bits from the coding plus some bits for ramping and linearization and at the end a second group of tail bits from the coding is needed. Between the bursts there is a gap to

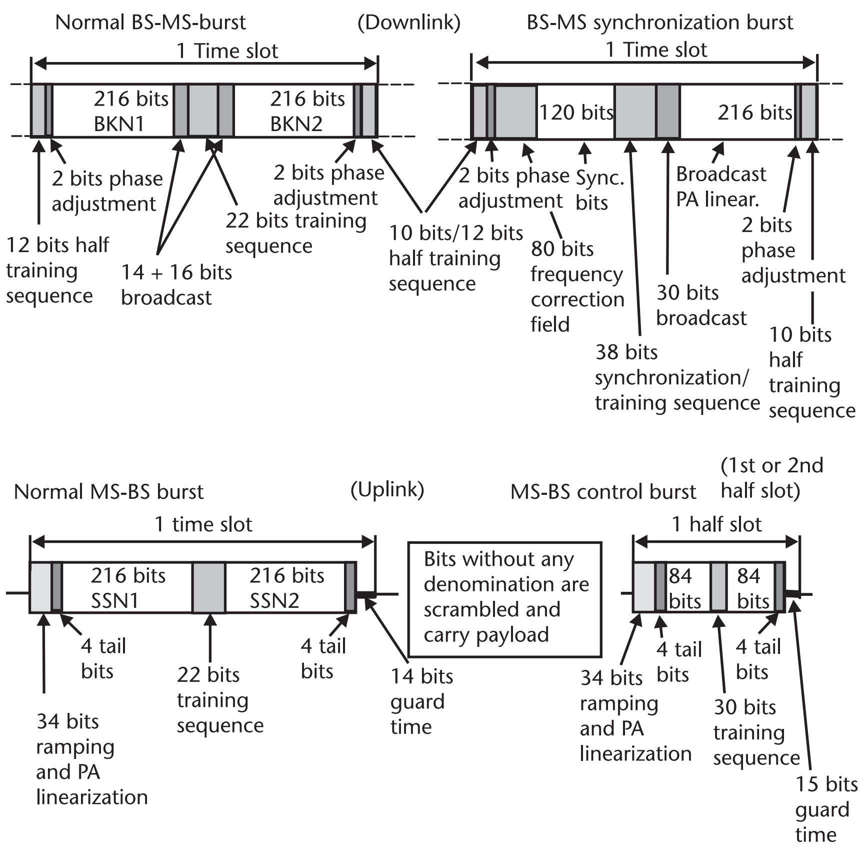


Figure 6.6 The burst structure of TETRA V+D.

compensate for the different burst delay times from mobiles at differing distances from the base station.

The structure of the BS to MS bursts is very similar. Again the useful information is appropriately coded and transmitted in two *blocks with numbers 1 and 2* (BKN1 and BKN2) of 216 bits each. However, the gaps between the different bursts are replaced by additional training sequences to improve synchronization (and if also applied channel equalization) in the mobiles. Additionally, there are is a *broadcast block* (BBK) belonging to the AACH, of which one-half (precisely 14 bits) is sent just before the training sequence and the other (16 bits) after. Moreover some bits are needed for phase adjustment.

An example for one of the additional bursts needed for other purposes such as channel access is the synchronization burst transmitted from the base station to mobiles in order to provide improved frequency synchronization. It contains a specific training sequence that is much longer than in the normal case and has a duration of 38 bits. In addition, there is an 80-bit frequency correction field. Also there is a broadcast block (BKN1) to distribute system information, and the second block (BKN2) is available for additional information transfer or PA linearization. This burst is used for synchronization in those cases where problems may occur should a mobile unit's reference oscillator drift off frequency. It enables software-controlled frequency correction to be made in the mobile.

For specific purposes such as call setup a time slot may be subdivided in two half slots (SSN1 and SSN2) as needed for random and reserved access of the mobiles on the uplink via the *main control channel* (MCCH). At least one-half slot downlink and uplink is available in each slot of frame 18 for access control. Table 6.6 lists the main types of bursts and provides a brief summary of their properties and purposes.

One time slot has a duration of 255 symbols or 510 bits and therefore lasts 14.17 ms. Four time slots comprise a frame having a duration of nearly 57 ms. A frame carries four time slots belonging to four independent traffic channels. The first slot of one of a bundle of several duplex carrier frequency pairs of a base station is dedicated to the MCCH. A multiframe has a length of about 1 second and is composed of 18 frames where the last frame is the control frame, which is devoted to signaling and control functions, in particular for continuously sent broadcast information for synchronization and system information. Moreover, at least one-half of a slot on downlink and uplink is available in each slot of the

Table 6.6 Basic Burst Formats of TETRA V+D

Type	*Purpose and Direction*	*Format and Specific Task*
a	Uplink control burst	Half slots for random and reserved access
b	Uplink normal burst	Normal MS-to-BS full slot format burst after initial system access during ongoing calls
c	Uplink linearization burst	Empty full slot format, for MS transmitter *power amplifier* (PA) linearization, needed only while tuning to a new carrier frequency
d	Downlink normal burst	Normal full slot format BS to MS burst during ongoing calls
e	Downlink synchronization burst	Full slot format burst for MS synchronization and for RF fine tuning

control frame for access control. Eventually 60 multiframes compose a hyperframe that lasts a little longer than 1 minute. This complex frame structure, which is shown in Figure 6.7, is needed to accommodate all of the different transmission, signaling, encryption, and control functions needed to provide the extraordinary flexibility of TETRA V+D.

In TMO different carriers are used for uplink and downlink. Time slots with the same number belonging to the same duplex traffic channel are shifted relative to each other by two time slots. Therefore, duplex traffic is possible without the need for a duplex filter in the mobile station; a simple antenna switch suffices. Moreover, there is enough time to switch from transmit to receive or vice versa and a single-loop *phase-locked loop* (PLL) synthesizer can manage this easily.

Of course, the situation is different if two time slots are assigned to one user; in such a case, the switching time becomes crucial. Assigning three or four time slots in full-duplex mode to one user requires specific mobile stations that employ a duplex filter instead of an antenna switch. There are additional differences in design because receiver and transmitter may be working simultaneously some of the time. Hence, such a mobile is more complicated and its manufacture is more costly. (By the same reasoning, a similar problem occurs in the latest GSM variations if too many carrier time slots are assigned to a single user to provide higher bit rates, for example, in HSCSD.)

In a frame the four time slots can be assigned independently to one or more users. For example, time slots 1 and 2 are assigned to a communication providing voice and data transmission simultaneously for the same user. Time slot 3 is assigned to a voice communication for another user, and time slot 4 is carrying data for yet a third. Note that not all of the counterparts of the participants in the active communication links shown necessarily have to be located in the same cell.

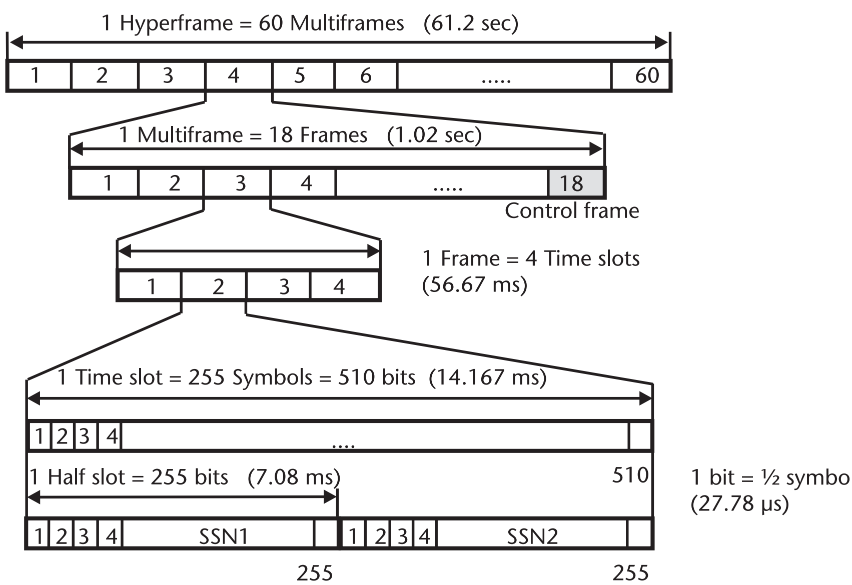


Figure 6.7 The frame structure of TETRA V+D.

Moreover, in TETRA uplink and downlink channels may be assigned by the BS independently to mobiles via the *access assignment channel* (AACH) as the following examples illustrate [6]:

1. Circuit mode call on downlink and another one on uplink for different mobiles;
2. Circuit mode call on downlink and assigned *secondary control channel* (SCCH) on uplink, or vice versa;
3. *Main control channel* (MCCH) on downlink (always slot 1) and circuit mode call on the corresponding slot 1 of uplink, or vice versa.

As one example of the various procedures used to initiate or release a transmission, the individual (mobile-to-mobile) voice call setup procedure is briefly described with the help of Figure 6.8. The first unsolicited channel access in performed by the mobiles with a slotted ALOHA procedure to resolve collisions. Afterwards solicited access always takes place in slots reserved by the base station where no collisions can happen. Hence, mobile station MS 1 initiates the call by sending a random access single burst *u-setup* message in a half slot in time slot 1 on the uplink. (Remember that time slot 1 on the primary carrier is always dedicated as the control channel.) This is acknowledged by the BS with a *d-call proceeding* message in slot 1 of the downlink and simultaneously mobile MS 2 is called with a *d-setup* message. MS 2 replies with a *u-connect* message in slot 1 of the next frame of the uplink. The BS then assigns a traffic channel on the same or another carrier via the *d-connect* message to MS 1 on the downlink in slot 1 of its next frame and simultaneously via a *d-connect ack* message to MS 2. Any slot other than slot 1 may be used for the traffic channel. Both mobiles then linearize their transmitters in the first slot of the assigned TCH on the uplink. In the downlink return slot the BS grants transmitting permission to MS 1, which may start its transmission in the following uplink frame. The whole call setup procedure needs 16 slots or somewhat less than 230 ms in the best case but does not exceed

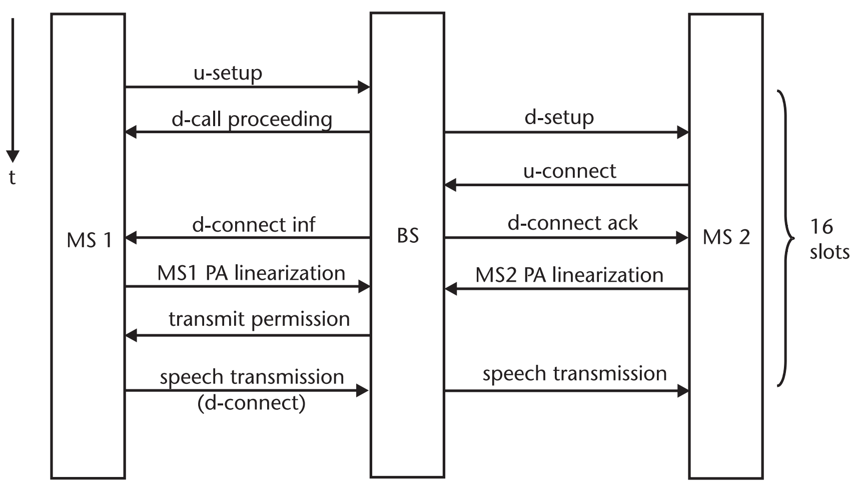


Figure 6.8 Setup of an individual call for speech transmission in TETRA V+D.

300 ms under bad propagation conditions if both mobiles are in the same cell. Of course, the call setup time is longer if both mobiles are in different cells or even in different TETRA networks, but with IP-based signaling methods modern TETRA system design allows call setup in less than 500 ms.

Similar procedures are used for the setup of group and data calls. Group calls are quicker to set up because intermediate acknowledgments are not used. Other procedures are needed for channel release and for the various signaling purposes. The situation becomes more complicated if authentication, encryption, and other additional functions are involved.

TETRA employs trunking to reduce the total occupied spectrum. In detail TETRA supports three different trunking methods. The simplest method is *message trunking*, where for the whole call, which consists of several transactions, one channel comprising one or more time slots on one carrier frequency is allocated. Gaps in data or speech transmission remain unused, which means that their capacity is wasted. In contrast to that, *transmission trunking* allocates to each transaction (each "over") a new channel during a call [24]. The related signaling takes place on the control channel. This method ensures efficient use of the available transmission capacity but needs a system with short reaction times. However, the drawback is that in heavily loaded systems or systems with only a few channels noticeable access delays may occur and parts of the conversation may need to be repeated. This and the need for increased signaling devour parts of the achieved capacity gain. Therefore, a good compromise is *quasi-transmission trunking*, in which transmission trunking is performed but with channel release that is delayed for a short time at the end of each transaction. During this so-called "hang time," the channel remains reserved for the next transaction. If the channel is released, then further signaling is performed via the control channel. This saves less channel capacity than full transmission trunking but reduces the risk of too many conversation breaks when capacity shortages occur. TETRA supports discontinuous transmission, which allows several base stations in low traffic density areas and only when limited spectrum is available to share a common carrier frequency. In this case transmission trunking would not be viable and it might be prudent to invoke message trunking.

As the reader may have already noticed, all signaling and traffic runs over dedicated logical channels should not be confused with the concept of physical channels. Simply said, different functions are grouped separately in logical channels that may be mapped on one single or several separate physical channels. This concept is extensively used in GSM and has also been adopted for TETRA. Because these logical channels have already been mentioned several times, a brief summary seems helpful to illustrate the fundamentals of this concept. Table 6.7 gives an overview and shows the relationships between the different operational modes (or states) and logical channels.

The main types of logical channels in the lower *medium access control* (MAC) layer are the *signaling channel* (SCH, uplink and downlink), *access assignment channel* (AACH, downlink), *broadcast synchronization channel* (BSCH, downlink), *stealing channel* (STCH, uplink and downlink), *broadcast network channel* (BNCH, downlink), and *common linearization channel* (CLCH, mainly uplink, discontinuous BS may also need to linearize).

Table 6.7 Basic Concept of Logical Channel Use in TETRA V+D

MS State	*Downlink Activity*	*Uplink Activity*	*Remark*
Idle mode	MCCH	None	MS is registered but not busy and monitors MCCH and all other signaling and measures adjacent channel field strength (RSSI) for cell reselection.
Energy economy mode	MCCH	None	In idle mode an MS may sleep for a certain number of frames and awake regularly to monitor predefined control slots to save battery capacity. Seven sleep rations from 1:1 to 1:359 are available.
Common signaling and packet mode	FACCH, SACCH	FACCH, SACCH	*Fast and slow associated control channels* (FACCH in frames 1 to 17, SACCH in frame 18) support all signaling and packet data transfer and fast signaling during call setup and so on; no *stealing channel* (STCH) is needed.
Traffic mode for voice or data transmission	TCH, SACCH	TCH, STCH, SACCH	MS is assigned to a TCH during circuit mode voice or data transmission. Signaling runs via SACCH or STCH if SACCH capacity or timing is not appropriate.
Transmission trunking mode	MCCH	MCCH	Ordinary signaling takes place on the MCCH.

The signaling between the upper MAC layer and the higher layers is performed by the *common control channel* (CCCH), comprising the *main control channel* (MCCH) and *extended control channel* (ECCH), and the ACCH, comprising the *fast associated control channel* (FACCH), *slow associated control channel* (SACCH), and STCH. If MCCH capacity is low, two types of *secondary control channels* (SCCHs) may be established: a common SCCH for a subset of mobiles or an assigned SCCH that is allocated to certain mobiles after a paging message has been sent to them or after an initial random access by the mobiles. The *broadcast common control channel* (BCCH) provides system information on the downlink and comprises the *broadcast synchronization channel* (BSCH) and the *broadcast network channel* (BNCH). Finally circuit mode traffic information running via the U-plane uses one of the different TCHs, that is, the *speech traffic channel* (TCH/S) or one of the speech or data traffic channels with different speeds and protections (TCH/2.4, TCH/4.8, and TCH/7.2).

A few short remarks should explain some very specific logical channels and their purpose. If a mobile transmitter is keyed on, some time is required for the linearization circuitry to remove nonlinear PA distortion. This is done during a specific burst period assigned to the linearization channel. During information transfer (speech or data) in traffic mode, the capacity of the SACCH does not always suffice or the latency is unacceptable for certain purposes. Then a small part of the traffic capacity is stolen and used for the STCH in a manner similar to they way GSM overcomes this bottleneck. The clever trick is that a small loss of information capacity in most data situations does not matter because the FEC mechanism of interleaving over several slots reconstructs the missing information similar to loss during a short fading null. In the case of voice, a time slot may be lost but this only slightly degrades the overall speech quality.

Further details of the signaling, the precise structure of all signaling and traffic channels, and the details of the protocol may be retrieved from one of the references and in particular from the *Designer's Guide,* which can be downloaded from ETSI (http://www.etsi.org).

6.5 TETRA V+D Services and Key Features

TETRA provides speech transmission and different alternatives for data transmission. Initially, three different air interfaces were specified: TETRA V+D (TMO), TETRA DMO, and TETRA *packet data optimized* (PDO). The choice depends on the intentions of the user, his tasks, and how he can best fulfill them. Moreover, numerous services are available that are well suited to different tasks but dependent on the type of equipment. Of course, high-speed data using, for example, four time slots in duplex mode is only possible with specific duplex equipment. To begin our discussion, we turn to the TETRA V+D services [1, 5, 6].

TETRA V+D provides a number of basic services that comprise its bearer services and teleservices for speech and data transmission. Additionally, two types of *supplementary services* (SSs) are available: PMR-type and telephone-type SSs. The four types of bearer services are listed in Table 6.8. They can be run in five different modes where the latter may have in some cases up to four different transmission speeds. The resulting variety of combinations meets nearly every possible requirement.

In circuit mode transparent data transmission at 7.2 kbit/s per time slot is available without any error protection. In circuit mode users can also transmit data with good protection at 4.8 kbit/s or with even stronger protection at 2.4 kbit/s per time slot. Putting time slots together means getting up to four times these bit rates. Finally packet data can be transmitted in a *connection oriented network service* (CONS) mode or in a connectionless mode, called *connectionless network service* (CLNS). Besides these bearer services, TETRA provides the different types of teleservices shown in Table 6.9. Again a comprehensive number of combinations is possible.

One of the most important teleservices is circuit mode ACELP encoded duplex voice [1, 9, 25]. A specific problem for voice transmission is that frame 18 of every multiframe is a control frame. To make space for it, the coded voice packets have

Table 6.8 Bearer Services Offered by TETRA V+D

Type of Bearer Service	*Mode*	*Properties*
Individual call	Circuit mode unprotected data	Transparent 7.2, 14.4, 21.6, and 28.8 kbit/s without any error protection
Group call	Circuit mode protected data	4.8, 9.6, 14.4, and 19.2 kbit/s with good error protection
Acknowledged group call	Circuit mode protected data	2.4, 4.8, 7.2, and 9.6 kbit/s with strong error protection
Broadcast call	Packet connection oriented data	No fixed steady connection
	Connectionless packet data	No fixed connection at all

Table 6.9 Teleservices Offered by TETRA V+D

Type of Teleservice[1]	*Underlying Bearer Service*[2]
Speech	Individual call (point-to-point)
Data User defined[3]	Group call (point-to-multipoint)
	Acknowledged group call (point-to-multipoint)
	Broadcast call (point-to-multipoint)

[1]Service may run clear or encrypted.
[2]Bearer services as specified in Table 6.8.
[3]Nonstandardized user services.

to be temporarily buffered and compressed prior to transmission and then afterward expanded again, which causes some additional delay on top of all other voice processing times. For most data services, this problem is of less importance because usually there are no stringent real-time requirements.

The ACELP algorithm is one of the powerful modern methods available to provide very good speech quality even under bad propagation conditions as has already been pointed out in Chapter 5. Nevertheless provision has been made in TETRA to replace the current ACELP coder by another type that may provide better speech quality in the future. Therefore, TETRA has been designed to handle in total up to four different speech coders. In particular *adaptive multirate* (AMR) coders for enhanced telephone quality are under consideration and have been proposed recently for several fixed and mobile telecommunication systems. These offer the feature that in channels with low BER extraordinarily good speech quality can be achieved at a comparably low bit rate. If the BER temporarily increases, the transmission rate is reduced for that time interval and the freed transmission capacity is used to increase the redundancy. Thus, a stronger error correction scheme with reduced speech quality but near good intelligibility can be used until the channel quality returns to its former quality level. This, of course, requires feedback to the sending end to indicate the quality of the transmission path.

Because TETRA offers transparent channels there is the possibility to define additional user-specific services that have not been standardized. Of course, such a particular service can only be used in networks where they are offered and by users whose equipment is appropriately qualified. This feature makes TETRA particularly future-proof. For normal applications a comprehensive group of data services with different predefined and flexible formats as listed in Table 6.10 belongs to the family of TETRA teleservices [26].

Status messages are predefined and can be used instead of lengthy voice routine messages. Their exchange is much faster and reliable than interrogating all members of a group by voice calls. There are 16 user-defined bits from which up to 15 may be used, which is equivalent to more than 32,000 different messages. Status message No. 0 is reserved for emergency calls. However, sometimes only smaller subsets are used, for example, for the PSRCS system in the United Kingdom.

Another important feature of TETRA is SDS, which offers three fixed formats with a length of 2, 4, or 8 bytes or characters and an additional flexible format with up to 255 characters. Besides status messages SDS offers another possibility to exchange information in a very efficient way. It is comparable to the SMS in GSM but instead of the 160- by 7-bit character limit, the SDS allows considerably

Table 6.10 Predefined Data Services Offered by TETRA V+D

Data Service	*Information Format*	*Characteristics*
Status	16 bits	Individual and group addressing Standardized messages and application-defined messages
SDS[1]	SDS1: 16 bits SDS2: 32 bits SDS3: 64 bits SDS4[2]: 1–2,047 bits	Individual and group addressing Text messaging and application-defined use with up to 255 characters per SDS message
TETRA-IP[3]	Standard TCP/IP protocol WAP	Flexible for various applications TETRA-specific adaptation for about 200–500 bytes suited to intranet, Internet, database access, and so on

[1]Short Data Service (SDS).
[2]Only format usable for independent applications.
[3]Adaptation of new services possible with low additional expenditure.

longer messages that can be sent directly from one subscriber to another, to a group, or to all users of a system without the need for a message service center. We should stress that status messages and SDS are very efficient in terms of transmission capacity and very fast because they are delivered in real time.

Packet data transmission is becoming increasingly important because it permits mobile Internet access as well as many other applications. This type of transmission occupies the channel only if there are data packets to be transmitted. To maintain a packet data link, it is not necessary to keep a line connection open at all times. Hence, only the amount of transmission capacity really needed is occupied. This is of interest, for example, for the billing of mobile Internet access. Circuit mode data transmission preferably is used when there is a need for uninterrupted data such as with video surveillance and similar real-time tasks.

There are countless applications for data transmission in TETRA systems. One important case is automatic vehicle location updating (aided by the GPS system and using status messages), for example, for control of big fleets in public transportation systems and in particular for trams and buses. The central computer of such a system may have knowledge of all schedules and it may be allowed to influence traffic lights in such a way that a delayed vehicle could be accelerated back onto its timetable. Other examples of data transmission in TETRA are:

- Workforce management;
- Transmission of instructions and commands;
- Mobile control and monitoring;
- Database interrogation, inquiries, and information retrieval;
- Information distribution;
- Data file transfer;
- Fax, e-mail, and voice mail;
- Transmission of maps, fingerprints, photos, and images;
- Surveillance of dangerous goods transportation;
- Alarms of any kind;
- Multilevel priority and emergency calls;

- Slow scan video surveillance;
- Telemetry and many more.

The most important point is that the whole TETRA structure and protocol stack has been designed to transmit digital voice as well as data. Unlike GSM, which was designed primarily for voice with a data capability, TETRA was designed right from the beginning to carry voice as well as circuit mode, packet mode, and IP data. It provides point-to-point services and typical PMR point-to-multipoint services from the base station to mobiles, mobile to mobile, and mobile to a base station. An example is the case of a broadcast message, which by definition is only transmitted in one direction in a point-to-multipoint mode. Most of the services can be used with or without acknowledgment. This is of particular interest for group calls. On one hand, to collect all acknowledgments from a large group may be too time consuming and is often not necessary. On the other hand, cases exist where it is of paramount importance that a message or a call be confirmed by *all* members of the group who have received it. This often occurs in public safety systems.

The high flexibility of TETRA becomes visible in particular when the supplementary services are reviewed. Typical PMR-type supplementary services are listed in Table 6.11. These comprise different levels and types of priority calls, different control and operational functions in group calls, and some dispatcher-related functions. A specific TETRA dispatcher feature is ambiance listening in which, unknown to the user, the dispatcher may remotely activate the mobile transmitter and listen in.

Other functions and services are offered such as restriction to a specific area of operation, short number dialing, identification of the caller, and dynamic assignment and reassignment of group numbers. Of particular interest is late entry, which means that a newcomer may join a group communication at any time. The possibility of reorganization of group sizes and structures during operation is another very useful feature of TETRA that was not available in most of the previous conventional PMR systems nor is it yet in GSM.

Typical telephone-type supplementary services are also available in TETRA that are devoted to call handling, identification of callers, redirection of calls, and

Table 6.11 PMR-Type Supplementary Services Offered by TETRA V+D

Supplementary Service	*Remark*
Access priority, preemptive priority call, priority call	TETRA provides different types and levels of priority
Include call, late entry, transfer of control	The ability to join a running communication; group control can be handed over to another party
Call authorized by dispatcher, ambiance listening, discreet listening	Mobile calls need prior authorization; ambiance and discreet listening need justification by law in many countries
Area selection	Restriction of operation to parts of a TETRA service area
Short number addressing	Replacement of often-used numbers by short numbers
Talking party identification	Notification of who is communicating
Dynamic group number assignment and reassignment	Group structures can be changed during operation

so on; others are needed for call barring and the evaluation of charges. Most of them are well known from modern digital terrestrial telephone systems. All of the supplementary services are supported on interfaces I 1 to I 4 with the exception of discreet listening, which is not supported at I 3, the ISI. Table 6.12 shows a list of telephone-type supplementary services.

All of the different kinds of calls for voice transmission are similarly available for data transmission. The simplest example is an ordinary data call. Today an increasing number of applications use the Internet and therefore it is reasonable to use the IP format for this and other types of transmission because a unified approach for all messaging is feasible. Using the Internet transmission format and protocol (TCP/IP), TETRA-IP permits not only the interrogation of databases and the exchange of e-mails via the Internet but also access to the *World Wide Web* (WWW) as well as to file servers (FTP). Of course, intranet access is also possible and modern TETRA networks have implemented this feature. Recently it has turned out that the Internet address space as defined by the current *Internet Protocol* (IPv4) no longer suffices. In particular, the fast growth of mobile networks has increased this problem. However, a solution is already in place by the definition of IPv6. Hence, fast transition will be possible from IPv4 to IPv6, mainly triggered by the mobile networks [26–29].

Access to TETRA services is easy. An individual call is performed by simply dialing the number of the called party. In the case of a group call, the group number has to be dialed or call setup can be achieved by pressing the PTT for connection to a previously defined group. Of course, it is also possible to address the members of a group separately. A broadcast call addresses a whole group, but it is a directional transmission only. This means that information is distributed to the group without a conversation being initiated.

Due to the complexity of TETRA networks, dialing and addressing are specific issues that cannot be discussed here in depth. We should mention, though, that the address format is comprehensive and flexible enough to meet all reasonable requirements. The whole address space comprises 48 bits, where 10 bits indicate the *mobile country code* (MCC), another 14 bits are available for the *mobile*

Table 6.12 Telephone-Type Supplementary Services Offered by TETRA V+D

Supplementary Service	*Remark*
List search call	A list of calls is worked through
Call forwarding: unconditional, busy, no reply, or not reachable	Call forwarding can be made dependent on different conditions
Call barring of incoming calls or outgoing calls	Specification of restricted mobile use
Call: waiting, hold, retention, or report	Different call processing and information on terminated calls
Calling/connected	
Line identity presentation	Identity indication of calling or connected subscriber number
Line identity restriction	Suppression of identity of calling or connected subscriber number
Call completion: to busy subscriber or no reply	Call completion dependent on different conditions
Advice of charge	Presentation of billing information

network code (MNC), and the last 24 bits are devoted to the *short subscriber identity* (SSI). The first two address fields together are also called the *mobile network ID* [30].

Previously, a radio was an isolated communication tool providing the man/machine interface via buttons, lamps, or LEDs and perhaps a display. Besides a very simple air interface, there was only an acoustic interface comprising a microphone and a loudspeaker or a handset. Sometimes additional features like the control of a horn were provided and in the early days of data transmission systems sometimes small printers or mobile fax machines were connected to vehicle-borne radios. Today TETRA offers many additional features, including data transmission and cellular telephony from a small handset.

The MMI has been modernized and provides better display capabilities, but it is not totally new and the acoustic interface has also changed little. Data transmission has added another interface, the PEI, to connect the radio to new accessories such as a laptop, a PDA, or other mobile data processing equipment. In the future it is envisaged that a Bluetooth wireless digital data interface will be incorporated in many TETRA terminals. Another useful addition is the *Wireless Application Protocol* (WAP), which allows easy access to WAP-enabled Internet Web sites with the very limited resolution of mobile radio displays. Instead of HTML WAP uses a derivative of XML called WML to transfer limited content from Web sites or databases. Having been introduced for mobile Internet access via GSM phones, WAP will also be used for TETRA and other modern digital PMR systems.

Last but not least we want to stress that TETRA offers the user full mobility, which means handover and roaming. In TETRA handover is called *cell reselection and migration* while roaming is equivalent to a change of location area in GSM. This has very seldom been the case in conventional PMR systems, and even modern digital PMR systems do not always offer both functions. If they are available, they are usually available only in a limited form.

Some additional key features of TETRA should be mentioned. First, authentication procedures not only of users and terminals are available, but also for the infrastructure. Second, lost or stolen equipment can be remotely disabled. Third TETRA offers encryption on the air interface as well as end to end. Finally, TETRA offers *over-the-air rekeying* (OTAR), which enables the user to exchange the encryption keys protected by a specific sealing procedure over the air without the need to handle a specific piece of hardware such as a SIM card, which considerably reduces the risk of misuse of encryption keys [31, 32].

Today the necessary tools are available to perform comprehensive simulations to investigate how a radio system will work under different propagation conditions. The simulations are able to replace to a great extent time consuming and costly measurement campaigns. Thus, besides the early field trials such as those in Jersey in 1995, comprehensive simulations have been conducted to evaluate the performance of TETRA. The results are to be found in Part 2 of the TETRA *Designer's Guide* [11].

Regarding the TETRA standard, one important conclusion can be drawn: The TETRA air interfaces and those of other contemporary digital PMR systems are much more powerful and more flexible than those of earlier systems but the user is unaware of their increased complexity.

6.6 TETRA DMO

In addition to TETRA V+D, some variations of TETRA are available that are devoted to specific tasks. The most important of them is TETRA direct mode operation because this provides direct radio links between mobile stations without the support of an infrastructure and is of major importance for many PMR applications. The standardization of TETRA DMO is still not complete and, therefore, small changes may occur in the future if deemed necessary.

The main difference between TETRA V+D (TMO) and TETRA DMO is that in TETRA V+D all mobile-to-mobile links are routed via a base station from which they may also be switched through to a PABX or the terrestrial telephone network. In TETRA DMO, however, a radio link can be set up directly between two mobile stations without the involvement of a base station. Hence, in this case no connection is possible to the terrestrial telephone network. Nevertheless, links to a TETRA network can be made simultaneously if coverage is provided to the mobile in question. DMO takes place in simplex or half-duplex mode on one or two carrier frequencies with a variable separation of only a few channels, while the communication between mobiles and the base station in TMO is *always* performed on two different frequencies for uplink and downlink with a *fixed* duplex separation of several megahertz. Moreover, flexible repeater and gateway configurations enlarge DMO coverage and permit access to TMO and fixed networks such as the ISDN [10, 12, 33–35].

DMO is of major importance for many PMR applications, in particular, in all situations where the members of a mobile user group need to have fast and direct contact or where no infrastructure is available or where access to it is impossible for whatever reason. DMO can be used within the coverage area of the TETRA infrastructure or outside for direct communication between the members of a group. There are many situations where use of DMO is inevitable, for example, for public safety tasks. This may be a tactical police operation or a catastrophe to be resolved by a fire brigade. Unlike FDMA, the TDMA principle offers additional possibilities such as dual watch for DMO by the use of time slots not needed for ongoing communication.

TETRA DMO also provides a good means for implementing transborder services. A TETRA mobile traveling into another country could try to get in touch with other TETRA mobiles by calling them via DMO. To do this a common DMO frequency allocation has recently been achieved across Europe. In other regions of the world, users may experience problems finding a common DMO band in different countries.

Compared to TETRA V+D, the DMO protocol stack has a simpler structure. Due to the lack of involvement of an infrastructure, many TETRA V+D functions are not necessary or not available; for example, there are only a limited mobility management and a very restricted number of supplementary services. However, similar to TETRA V+D, the TETRA DMO stack consists of a control plane and a user plane. Again layer 1 and 2 are used by both planes in common, as Figure 6.9 shows.

The burst structure of TETRA DMO is similar to that of TETRA V+D, however, as Figure 6.10 shows, some modifications have been made. Of course, there are

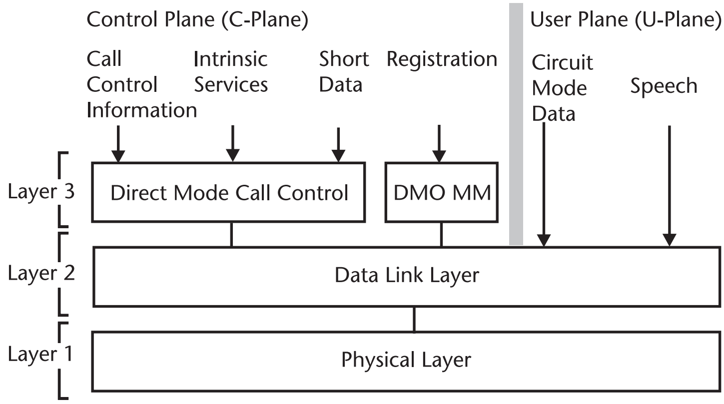


Figure 6.9 TETRA DMO protocol stack.

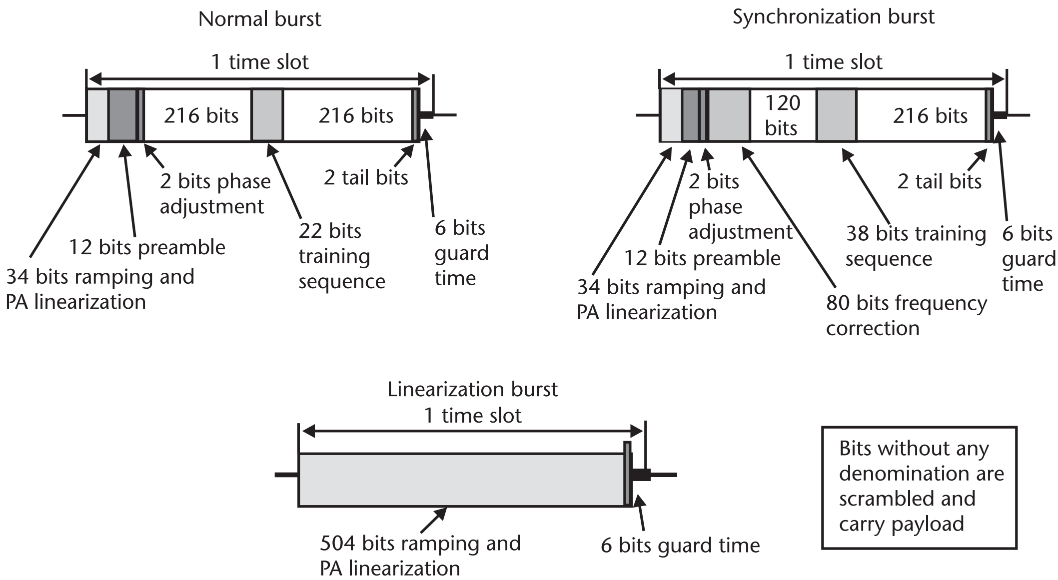


Figure 6.10 TETRA DMO burst structure.

no DMO bursts to be sent to and from base stations and therefore the number of different bursts has been reduced. On the other hand, a specific linearization burst has been added to cope with the more difficult propagation conditions associated with the direct mode.

The basic frame structure of TETRA DMO is similar to that of TETRA V+D. However, in total it is less complicated because the hyperframe has been omitted as can be seen in Figure 6.11.

As in TMO each TETRA DMO carrier provides four time slots. Thus, in principle, four basic modes of operation would be feasible, namely, one or two simplex or TDD channels per carrier. However, in its simple mode DMO provides one communication channel in each direction in two of the four time slots per frame and carrier. The two corresponding time slots are separated by one time

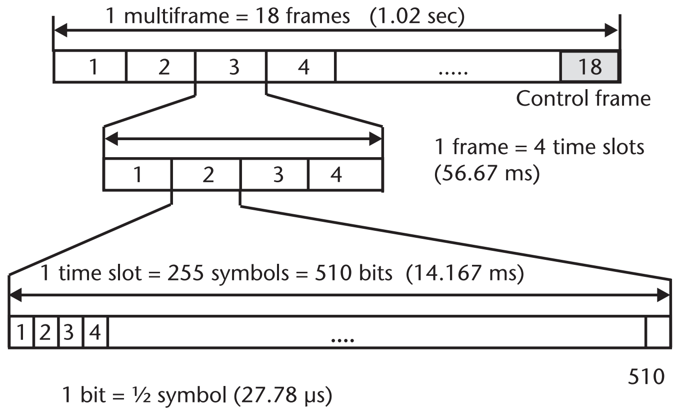


Figure 6.11 TETRA DMO frame structure.

slot so the Tx/Rx switching is not critical. The communication is organized in such a way that the master initializes the communication and the slave is given the ability to preempt. During a call the transmitting party always acts as the master and controls the communication.

Because only one or two time slots per frame are used, additional transmission capacity is available and therefore in a frequency efficient mode the remaining two slots can be used similarly for another call. The two channels are interleaved in such a way that slots 1 and 3 are used for the first channel, and slots 2 and 4 are assigned to the other. In simple mode 50% of the channel capacity is sacrificed, but this avoids difficult synchronization problems caused by the fact that sometimes not all of the four parties involved in frequency efficient mode on one single carrier frequency have common coverage.

Note that TDD operation has never been specified for DMO; instead, the reverse channel is used for signaling, channel maintenance, preemption, and late entry. Simplex operation (on one or two channels) is well suited to open channel applications where one party is speaking and the others are listening, whereas for individual calls both simplex and duplex operation are feasible.

After reviewing the "mechanics" of DMO, further explanations will show the extraordinary flexibility resulting from the combination of DMO and the TDMA principle. Basically, in TETRA DMO five fundamental operational modes are available:

1. Standard "back-to-back" operation in simplex mode;
2. Extended range via DMO repeater;
3. *Dual watch* (DW) meaning periodic scanning of free DMO carriers and TMO control channels;
4. Continuous connection via a DMO *gateway* (GW);
5. Continuous connection via a DMO gateway repeater.

The first and simplest case is a direct mobile-to-mobile link without any additional functionality, which suffers from the limited coverage of direct links between mobiles due to the low height of their antennas. In the second case the introduction of a repeater, which may be a vehicle-borne device, offers the benefit of considerable coverage extension. In particular, the radius of action can be substantially improved for handhelds, especially if the repeater is located at an elevated location.

The other three modes serve communication between DMO and TMO terminals and network links. *Full DW* allows simultaneous monitoring of a TMO control channel while being actively engaged in a DMO communication, and *idle DW* means monitoring a TMO control channel and DMO carriers simultaneously while *not* being engaged in a call. Full DW allows the user to interrupt an ongoing DMO communication by a DMO or TMO call with higher priority. A mobile may also communicate simultaneously during a DMO call via a gateway into the main infrastructure, for example, with a control center. A DMO gateway allows a continuous relay between DMO and TMO. Finally, repeaters and gateways can be combined. However, DW is not available in the frequency efficient mode of DMO. Figure 6.12 illustrates some of the basic DMO configurations.

Besides individual calls in DMO, group calls can also be made. A specific kind of group operation in DMO is the *managed DMO* (MDMO). There is no restriction if the channel is unoccupied; otherwise, an MDMO terminal can only transmit after it has received an authorizing command generated by a mobile in touch with the TMO infrastructure, by a DMO gateway or a repeater/gateway. Hence, channel access and usage are only permitted if the channel is free from interference. Specifically in nonharmonized spectrum MDMO potentially supports removal of interference from roaming DMO terminals, which is a specific problem frequently encountered in Europe.

Besides dual watch and repeater use, links via a DMO gateway to a network can be made. Moreover, mobiles can communicate via a repeater or a gateway to each other and to the network. And last but not least there is the *managed* DMO via repeater or gateway with a connection to the network. Figure 6.13 gives some

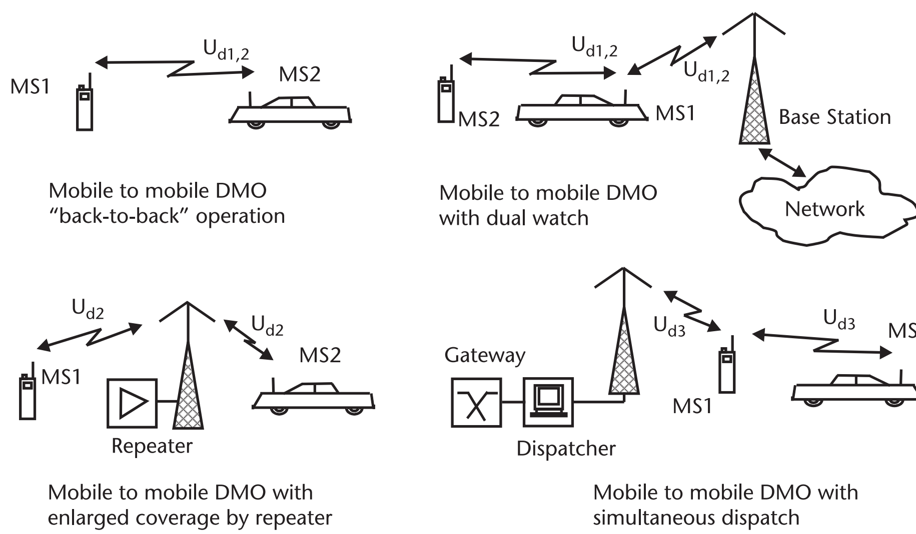


Figure 6.12 Some basic DMO configurations.

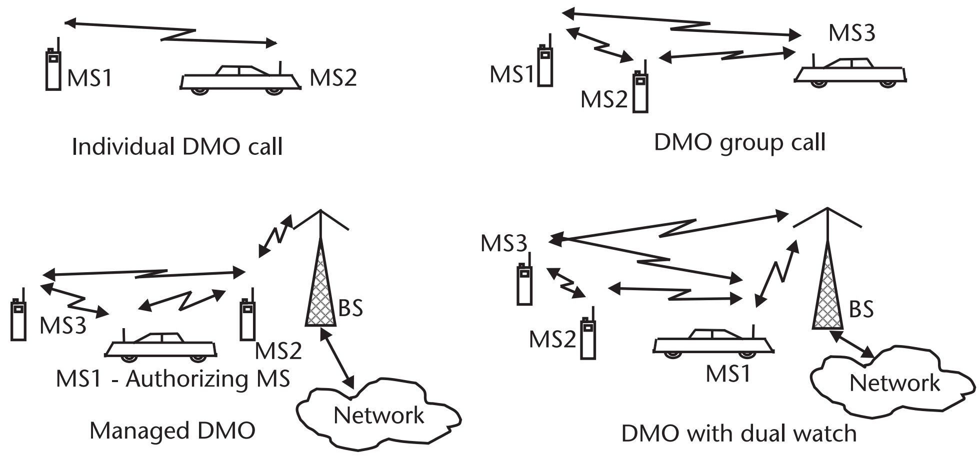


Figure 6.13 Examples of direct mode group operation.

examples for group operation in direct mode. In DMO different kinds of repeaters and gateways or combinations of both can be used and in Figure 6.14 some examples for more advanced DMO applications are shown.

DMO may operate in many different scenarios. The members of a group communicating on a DMO channel may stay within the coverage of a fixed network such as group A in Figure 6.15 or be outside like group C. There might also be other radio terminals inside the coverage area but not involved in the current communication. We must also consider a third DMO group, group B, that is partly inside and partly outside the TETRA network coverage area. The same scenarios as discussed above may occur in cases where a DMO repeater or a gateway is used. Which services are possible for a certain DMO group depends on whether the group has full or partial coverage from the network or none at all. In the latter

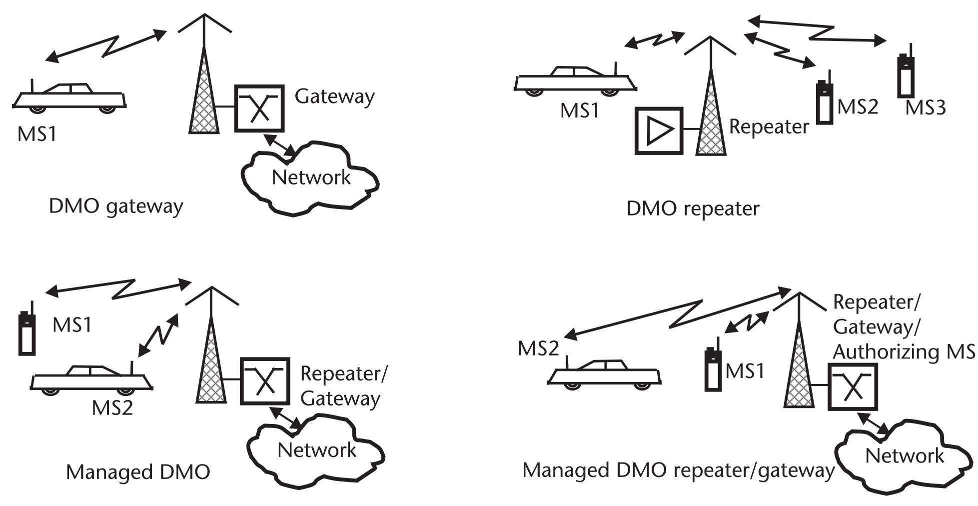


Figure 6.14 Use of repeaters and gateways in DMO.

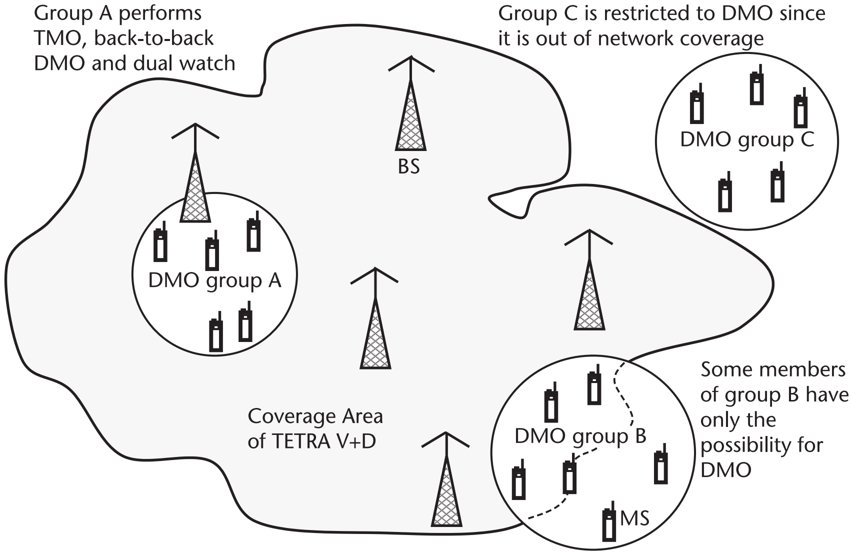


Figure 6.15 TETRA DMO and TMO in mixed scenarios.

case, the gateway itself must have a connection to the network; otherwise, no network access is possible. DMO may also be combined with the dual watch function in all of these situations. Figure 6.16 shows more clearly how DMO groups may operate in a TMO service area where additional repeaters and gateways are available.

Due to the operational requirements and the flexibility of TDMA, different types of repeaters are available in DMO and it is useful to have a closer look at their characteristics and capabilities. Therefore, at first we should return to back-

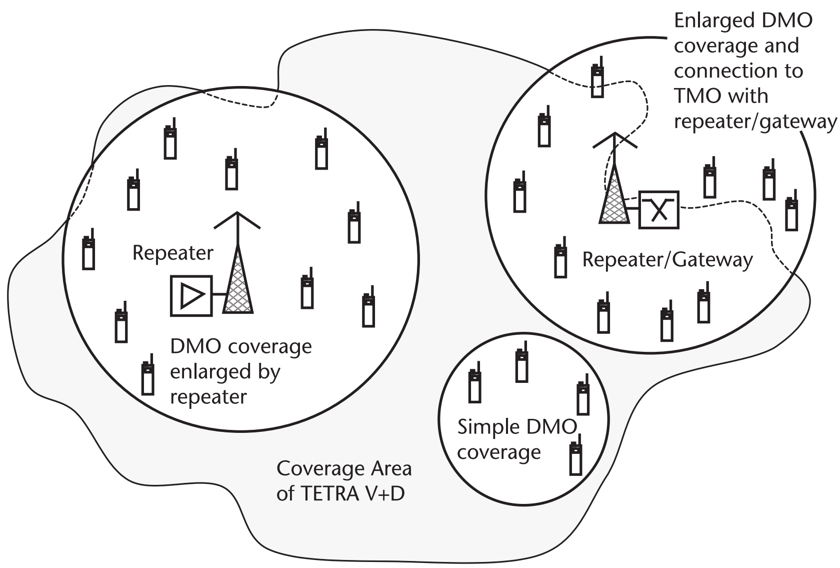


Figure 6.16 Mixed DMO and TMO scenarios with repeaters and gateways.

to-back operation in normal or frequency efficient DMO. As shown in Figure 6.17, two or four time slots of one carrier are needed.

The DMO repeaters are not simple RF amplification devices; they all are regenerative, which means that all received bursts are decoded, corrected if necessary, and reencoded again before retransmission. The necessary processing time causes a considerable delay of several time slots because one complete time slot is needed for reception and the signal processing and retransmission also need additional time. Furthermore, the retransmission must not lead to timing contradictions in the different DMO scenarios.

The simplest DMO Repeater is that of Type 1A, which supports one simplex call in two time slots per frame on a single carrier in simplex mode with three slots of delay. The other two slots are needed for preemption of the receiving station with five slots of delay. The frame numbering of the slave is shifted by one slot, for example, master slot 1 is slave slot 2. This type of repeater is suited to support of single vehicle operation in a master/slave configuration and Figure 6.18 shows schematically how it works.

A Type 1B DMO Repeater works in a similar way to Type 1A but on two carriers. It is therefore more complex and more versatile—and also more expen-

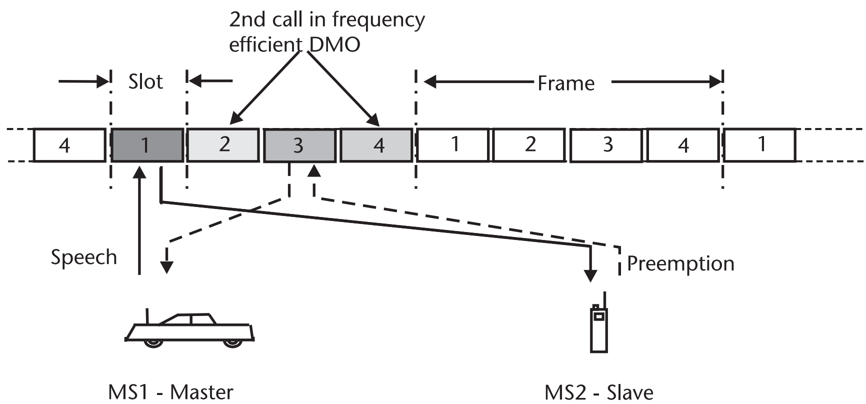


Figure 6.17 "Back-to-back" DMO.

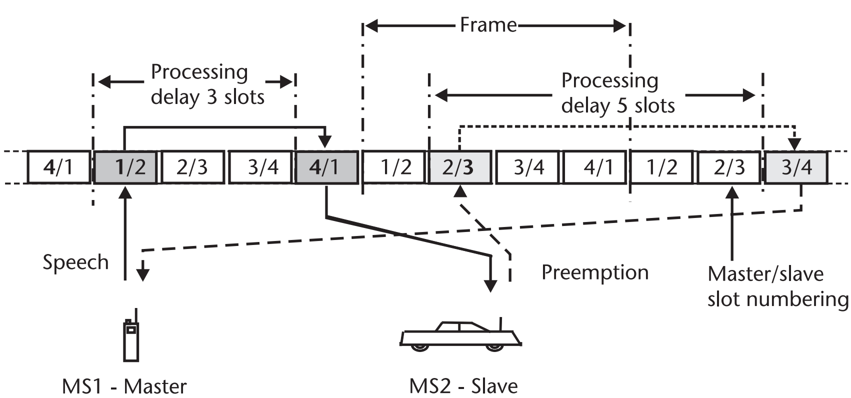


Figure 6.18 DMO via Repeater of Type 1A.

sive—but it reduces the potential for RF interference. It not only supports one call in two time slots as shown in Figure 6.19, but also allows multiple repeater configurations.

As shown in Figure 6.20, Type 2 DMO Repeaters also support dual carrier operation but they can carry two simultaneous calls. For the second call only the forward direction is shown for better clarity, but the receiving station is also capable of preemption. This type of repeater exhibits a delay of four slots for both speech and preemption and supports power control and multiple repeater configurations.

Finally, there are DMO gateways carrying one call per carrier. They operate on three frequencies, one pair for TMO in the network connection and an additional one for the DMO link to the mobiles. This flexibility can be further enhanced because the functionality of repeaters and gateways can be combined to DMO repeater gateways operating as talk-through repeaters to interconnect DMO and TMO. They use the DMO air interface with one or two carriers per call acting then as Type 1A or 1B Repeaters. They can operate automatically or can be controlled by an operator. Hence, these devices extend DMO coverage and additionally permit DMO users to have TMO access whenever they wish. The extraordinary flexibility of all possible DMO variations becomes clearer by looking at Table 6.13 where the different DMO equipment types are shown.

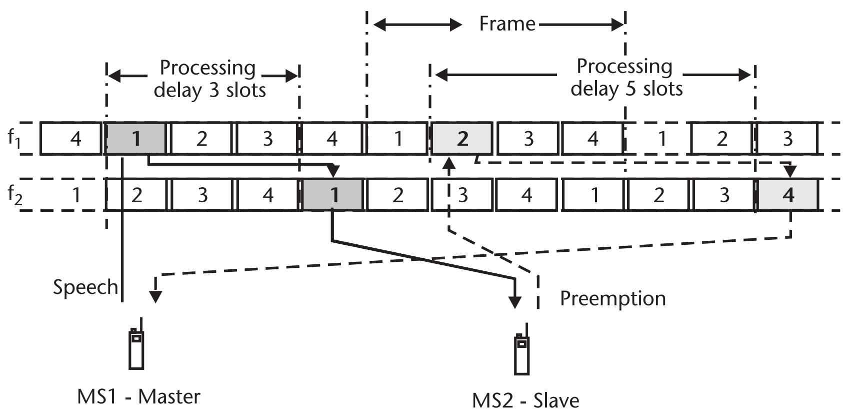


Figure 6.19 DMO via Repeater of Type 1B.

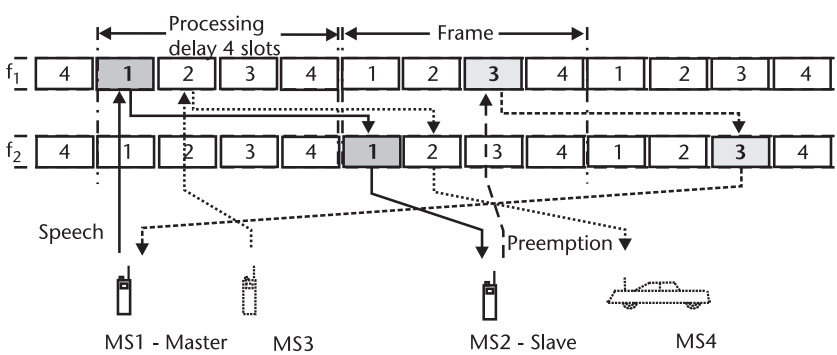


Figure 6.20 DMO via Repeater of Type 2.

Table 6.13 TETRA DMO Equipment Types and Properties

DMO Types	*Number of Channels*	*Number of Calls*	*DMO Air Interface*
MS-to-MS normal mode	1	1	U_{d1}[1]
MS-to-MS frequency efficient mode	1	2	U_{d2}[2]
Repeater Type 1A	1	1	U_{d2}[2]
Repeater Type 1B	2	1	U_{d2}[2]
Repeater Type 2	2	2	U_{d2}[2]
DMO gateway	1	1	U_{d3}[3]
Type 1A repeater/gateway	1	1	U_{d3}[3]
Type 1B repeater/gateway	2	1	U_{d3}[3]

[1]Identical to ordinary DMO air interface I 6.
[2]Air interface for DMO via repeater or I 6″.
[3]Air interface for gateway to TMO or I 6′.

In DMO power control is also available. All mobile terminals try to use the lowest of several burst peak power levels that will provide sufficient transmission quality. In reducing the power level to the needed minimum, it is possible to reduce the overall interference level within a complex TETRA network. As with TMO, DMO also has several MS power classes defined as indicated earlier in Table 6.3. The power class determines the peak power to be used for transmission if power control requests the highest power level. The highest DMO power class is class 2, which is equivalent to 40 dBm or 10W. Further classes were created by reducing the peak power successively by 2.5 dB down to 22.5 dBm or 0.18W. These classes are denoted 2L, 3, and so on up to 5L. Note that the classes 1 (45 dBm or 30W) and 1L (42.5 dBm or 17.5W) are only defined in TMO but not in DMO.

The bearer services of TETRA DMO are the same as those of TMO, as shown earlier in Table 6.8, with the important difference being that only one time slot is available, which means that all multislot bearer services are excluded from DMO. Hence, the main teleservices and facilities of DMO are similar to those for TMO, shown earlier in Table 6.9, but of course only those services are available that do not rely on specific network resources.

The restrictions mentioned also prohibit supplementary services that are not feasible or are meaningless in DMO. Of course, this does not apply to individual and group calls for voice transmission, SDS, status messages, or encryption. Short number addressing, preemptive priority calls, transmitting party identification, and OTAR are also available. The latter allows the mobile to generate and distribute static cipher keys to every other mobile that has received a sealing key from an authorization center. Finally, late entry is also available in DMO, for example, for joining a group during an ongoing call, just after switching on the radio, after leaving another communication channel, after switching from TMO to DMO, or after changing the coverage area. A number of telephone-type supplementary services were originally planned for proprietary implementation, but due to a lack of capacity in the signaling format these services were skipped.

Note that in DMO a third type of service called *intrinsic service* was specified. An intrinsic service is inherently available within a normal bearer or teleservice. It is an integral part of the DMO air interface U_d and the signaling associated with the basic services but it differs from the services in TMO because it requires no explicit invocation. The intrinsic services are offered in association with all the voice and data services described earlier [12].

6.7 TETRA PDO

The standardization of TETRA PDO was started in parallel to that of TETRA V+D. However, TETRA V+D is much more flexible compared to TETRA PDO because it allows voice *and* packet data transmission in the circuit mode *or* connectionless mode. Moreover, the maximum data rate of TETRA PDO is only a little bit higher than that of TETRA V+D. For these reasons interest in TETRA PDO began to fade away and, consequently, in 1995 to 1998 the standardization work on different parts of the TETRA PDO standard was put on hold by ETSI. TETRA PDO is now obsolete because it was superseded by new initiatives that are aiming for higher bit rates. In the end the work on TETRA PDO was replaced by the promotion of DAWS [35].

Nevertheless, a short description of TETRA PDO is useful. It is a specific TETRA variation dedicated solely to packet data transmission. It uses the same 25-kHz carrier separation and $\pi/4$-DQPSK modulation as TETRA V+D. Other basic parts of the air interface and protocol are also similar. However, its primary use is intended in the 900-MHz region using several carriers in separate uplink and downlink bands. The main difference from TETRA V+D is that TETRA PDO always employs packet data transmission over the air interface. A major issue is a short transit time of less than 100 ms, while TETRA V+D only guarantees less than 500 ms for its comparable CONS [18].

Unlike the frame and slot structure used in TETRA V+D, TETRA PDO uses a burst transmission structure without time slots or frames. Instead a statistical access method called *statistical multiplexing/multiple access* (STM/STMA) is employed and data packets are transmitted in variable length bursts (single or multiple packets). Due to the reduced overhead, the user gross bit rate is increased from 28.8 kbit/s (four concatenated time slots in TETRA V+D) to 32 kbit/s. Each burst is transmitted individually using the packet data transmission principle in connection oriented or connectionless mode. This transmission mode is particularly well suited to those types of data transfer that have a bursty nature such as Internet access. If TETRA PDO is used in separate service areas, then these may be connected to each other and to other networks via X.25 links. Note that TETRA PDO also offers unrestricted mobility with full roaming and handover capability.

6.8 TETRA Networks

Because TETRA provides handover and roaming facilities, large multisite networks with local, regional, or national coverage can be built. Moreover, even transborder networks covering regions of two or more countries can be created. As shown in Figure 6.1, the switches can control the traffic of several BS controllers that are connected to a number of base stations. If a TETRA network employs only one gateway, then the whole network will show a star configuration. If one switch is too small for all of the network traffic, then several switches are necessary and in this case these can be connected in a star configuration or a mesh can be built, as Figure 6.21 illustrates.

In reality, very often the best solution is to have a mixture of a star and a meshed network. The service area size, the traffic distribution, the main network applications, and the capabilities of the switches, which are different from manufacturer to manufacturer, will determine which is the best compromise. Issues of costs, future expandability, and various other aspects will also influence the network topology. Hence, all of the different networks will be tailored to the needs of their users and the plans of their operators.

The whole topic of network design cannot be standardized. What can be standardized and have been are the main elements and their standard interfaces. Their optimum combination and finding the best network configuration is a difficult design problem to be solved on a case-by-case basis.

6.9 DAWS, MESA, and Future TETRA Development

Years ago TETRA and the American APCO 25 standard were seen from many sides as competitors and there was a hot race between them. Later, cooperation between TIA, where the APCO standards are written, and ETSI was established, applying in particular to Phase II of the APCO 25 project. To APCO's original plan to introduce a narrowband system with a carrier separation of 6.25 kHz a proposal from the TETRA MoU has been added: 4TDMA. This is equivalent to four FDMA channels with a carrier separation of 6.25 kHz for use in urban areas.

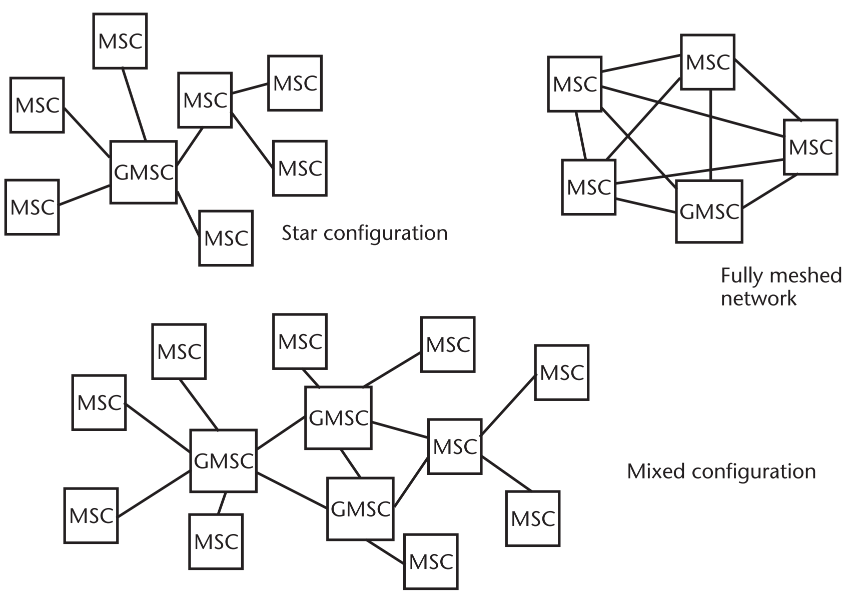


Figure 6.21 TETRA network configurations.

However, the IMBE codec already introduced for APCO 25 Phase I and DES encryption have been proposed instead of the TETRA ACELP codec and the TETRA encryption algorithm as a concession to APCO. However, the plan to develop an ANSI/TIA standard by its working group TR-8.14 [31, 36] was skipped in the end (see Chapter 7).

The basic properties of TETRA are also well suited to higher transmission rates. Therefore, *Digital Advanced Wireless Service* (DAWS) was derived from TETRA PDO as an *enhanced PDO* (E-PDO) version. It was intended as a new means of high-speed mobile radio access to the Internet based on TETRA principles but providing much higher data rates. DAWS employs $\pi/4$-DQPSK modulation similar to TETRA with the uplink and downlink on separate carriers in the 5-GHz band but also capable of TDD. It was intended for mobile access to high speed ATM networks. These work at 155 Mbit/s, whereas UMTS is limited to 2 Mbit/s. The network access is based on IPv6 and is therefore capable of supporting multimedia applications. A transmission rate of 25 Mbit/s in a 25-MHz bandwidth has already been demonstrated and this has shown that the concept is viable although the maximum speed of 155 Mbit/s has not been demonstrated. Full mobility and roaming capability are offered but the high carrier frequency and the large bandwidth may impose some restrictions [37–39].

The standardization of DAWS started in 1997 and should have been completed in 2000. However, the work was superseded by new initiatives such as new RLAN systems within the already mentioned project BRAN (see Chapter 2). As another spin-off of the DAWS project, ETSI and TIA have agreed to work together on the requirements and specifications for wireless broadband technologies in a new project called *Mobility for Emergency and Safety Applications* (MESA), which supersedes the previous work on DAWS and TIA Project 34. MESA is a *public safety partnership project* (PSPP) of ETSI and TIA aimed at creating a set of internationally accepted standards for globally applicable implementations. MESA comprises representatives of industry and user organizations with participation from Europe, the United States, Africa, and Asia and also observers such as the *Telecommunications Technology Association of Korea* (TTA).

MESA has identified a growing demand for mobile broadband services for fire fighting, disaster recovery, civil defense, suppression of organized crime and terrorism, telemedicine, mobile robotics, and emergency operation on international waters. Therefore, cooperation between different public services within a country and also between different countries while maintaining service and national integrity and information security is necessary. The key characteristics of MESA are mobility up to aeronautical speeds, advanced broadband services even throughout remote areas, satellite links, self-healing service cells, and interlinking to fixed networks via master nodes—all of this with core networks based on IPv6 and ATM access at very high data rates in excess of 2 Mbit/s. In fact, the MESA specifications are beyond those of third-generation systems [40–42].

TETRA standardization has been completed with the exception of small maintenance and upgrading tasks, especially for DMO. However, the third generation of public cellular systems (3G, meaning UMTS, IMT-2000, and so on) will provide much higher bit rates than did the second-generation systems. Hence, the demand for higher transmission rates has also arisen in the TETRA context. TETRA

Release 2 will therefore be created to meet this demand by offering multimedia services, high-speed data rates, and improved interoperability and roaming with GSM, UMTS, and other 3G systems providing IP compatibility. In short, TETRA is envisaged to become a member of the third generation of mobile systems. TETRA Release 2 was planned as a 3-year standardization project started in late 2000 with a set of ambitious goals, some of which seem difficult to achieve in combination [43–47]:

- High-speed packet data in excess of 300 kbit/s for multimedia transmission and other high-speed data;
- Range extension up to 120–200 km and beyond;
- Improved spectrum efficiency;
- Reduced size, weight, and cost of terminals with longer battery duty cycle;
- Improved network capacity, system performance, and QoS;
- Improved performance-to-cost ratio for rural areas with low traffic and subscriber density;
- Improved interworking with public networks such as GSM/GPRS, UMTS, and other 3G/IP networks;
- Additional speech codec(s) with enhanced quality for interconnection to 3G networks without transcoding;
- Evolution of the TETRA SIM to a *universal SIM* (USIM) for interworking and roaming with GSM and 3G networks as well as electronic banking and electronic cash;
- Full backward compatibility with TETRA Release 1.

By the way, this catalog does not meet all PSPP and MESA requirements, but nevertheless these help to push for TETRA Release 2 by stressing the general need for improved characteristics. The introduction of TETRA Release 2 products is envisaged from 2005 onward.

The route to full TETRA Release 2 will comprise two important steps. Phase I is *TETRA Advanced Packet Service* (TAPS), which is planned as an overlay to PAMR networks for rapid market deployment. For TAPS new infrastructure and terminals are needed, based on GPRS and EDGE principles at up to 384 kbit/s in a channel raster of 200 kHz.

Phase II is *TETRA Enhanced Data Service* (TEDS) providing data rates between 28.8 and 384 kbit/s. It will based on a new modulation scheme using several 25-kHz carriers in parallel, similar to OFDM. At the end of 2001 five different candidate technologies were under consideration for TEDS. The one eventually selected should ensure backward compatibility to TETRA Release 1 and easy upgradability of mobiles as two of the main goal for TEDS.

As is well known, the *electromagnetic compatibility* (EMC) debate has substantially heated up in the last years. Besides possible health hazards, which we will discuss in Chapter 9, electromagnetic interference by radio transmitters under certain circumstances may create problems with other electronic devices. In particular, in hospitals the use of mobile phones on public systems such as GSM is prohibited. In TETRA TMO the mobiles are always under control of the base station so there is the possibility of inhibiting transmission of TETRA terminals in such critical areas by command from a dispatch center as long as their location is known with sufficient accuracy. The inhibited terminals can receive but are not allowed to transmit except when making emergency calls. This transmit-inhibit function was standardized in 2001 [48].

We can expect TETRA, with its existing variations, to be able to cover significant parts of the PMR markets and to meet the requirements of a large percentage of PMR users. Not the least reason is that TETRA is supported by many manufacturers and that interworking of products from different suppliers is ensured. To remove even the slightest doubt about official TETRA *interoperability* (IOP) testing and certification service was inaugurated in early 2002 [49, 50]. Nevertheless new requirements will arise and TETRA has to continue to develop so as not to become obsolete in the near future.

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CHAPTER 7

European and American Standardized Digital PMR Systems

In addition to TETRA, we should take a look at other standardized digital PMR systems. In particular, these are DSRR, DIIS, narrowband PMR, and APCO 25. DSRR was the first attempt to introduce a standardized digital trunked PMR system in Europe. It was initiated within ETSI in the early 1990s for use in the 900-MHz band for the low-cost end of the market, but was never launched and the standard was withdrawn in December 1995. Nevertheless, a short description is given due to its historical importance. DIIS is a new digital FDMA system currently going through the ETSI standardization process. It is a simple but flexible system for those market segments that cannot be economically served by TETRA. Narrowband PMR is intended for specific purposes. It provides a transparent channel for speech and data. APCO 25 is a digital FDMA system that has been standardized in the United States, mainly for public safety tasks.

7.1 DSRR: The First Attempt at European Digital PMR Standardization

Digital Short Range Radio (DSRR) was initially planned as a digital trunking system with 80 channels in the 900-MHz band and an address space of 10^7. It was intended to be an improved alternative to analog CB radios for the lower market segments, but upgraded with digital speech and capable of transmitting data. However, the idea was to address small business enterprises and not to play a major role in the CB and leisure market.

DSRR was mainly intended for traffic in simplex mode between mobile stations between 933 and 935 MHz (called *high band,* HB). Two-frequency semiduplex mode was specified and could have been implemented between 888 and 890 MHz (*low band,* LB) in those countries where the necessary frequency allocations were available. To set up a call, a free traffic channel had to be found by hunting before signaling could take place on one of two control channels [1, 2].

GMSK with BT = 0.3 was chosen and the gross bit rate was set at 16 kbit/s. This permitted the use of the GSM speech codec working at 13 kbit/s with a small amount of error protection. This was done simply for economic and timescale reasons because it would not have been possible to develop a dedicated speech coder for DSRR. It was suggested at that time that if a second-generation DSRR were to be developed, then a more efficient codec would be implemented so that simultaneous voice and data could be carried.

For the control channels only a quarter of the modulation bit rate was selected to ensure a robust signaling scheme: GMSK with BT = 0.5 at 4 kbit/s. Moreover, the adjacent channel power could be reduced by this low bit rate, and initially the two border channels of the DSRR band were allocated for control purposes. This would have guaranteed operation adjacent to other systems with similar properties without the need for guard channels. However, this argument did not really count since the frequency allocation for the GSM core band had been made directly above the DSSR band. Eventually, the control channels were moved into the DSRR band to channels 26 and 52 to ensure undisturbed operation. Interference from adjacent GSM base station transmitter sideband noise would then only spoil some border channels of DSRR in areas to which they were allocated and would have little effect on overall DSRR system performance. In effect, the interference would locally reduce the number of channels from which a traffic channel could be selected. Bit synchronization was achieved with a 256-bit preamble, and for frame synchronization 16 bits were assigned. The code word length was 88 bits including 16 bits of CRC. No format was fixed for data transmission, but the block length was limited to 10 seconds.

The main characteristics of DSRR as specified in I-ETS 300 168 are shown later with the properties of the other standardized systems (see Sections 7.5 and 7.6). Compared to a conventional system, some of the limits were relaxed because the equipment costs had to be matched to the intended market segment. Most of the limits, however, meet the requirements of EN 300 113.

There were several reasons why DSRR did not become generally accepted. At first there were no firm frequency allocations and when these were made the window for the DSRR market was beginning to close. Moreover, it seemed to be unfavorable to introduce a completely new digital PMR technology with some relaxed limits that were not applicable to other bands and for other PMR applications. At the same time, proponents of GSM were looking to extend their band and had allocated channel numbers across the proposed DSRR assignment. Eventually, the standard was withdrawn and the frequency allocations for DSRR were canceled. Subsequently, the bands from 880 to 890 MHz and 925 to 935 MHz were allocated as an extension band for GSM so a standardized digital PMR solution for the low-end market was never implemented. This is one of the reasons why DIIS was created later.

7.2 DIIS: The New Digital FDMA Standard for PMR

ETSI has already created the TETRA digital PMR standard, so why is another in preparation? The reason is because TETRA cannot provide an economic solution for every PMR application, and it seems to be economically unreasonable to create one single technology to cover the many, partly contradictory PMR requirements. TDMA systems such as TETRA are well suited to networks with high user densities and large coverage areas in which a comprehensive set of features including handover, roaming, and encryption is needed. TDMA is also beneficial if a number of separate user groups with similar operational requirements and a common limited service area, such as a city, can be placed on a community repeater. FDMA, and in particular DIIS, is appropriate for small nontrunked systems without handover, roaming, and encryption, provided that it meets the user requirements and is cheaper. Thus, both systems clearly address different PMR market segments. Moreover, sometimes specific marketing and technical reasons have to be taken into account, for example, the nonavailability of TETRA in the 2-m band.

The idea of DIIS first arose in May 1986 when ECTEL began to draft the later BIIS 1200. At that time, it was thought possible that a system could be much faster than 1.2 kbit/s if direct modulation were used. However, this concept was set aside to concentrate on BIIS 1200, but in the mid-1990s it was taken up again by ETSI RES 02 (later renamed ERM RP 02 with an Ad-Hoc Group for DIIS that now works under TG 32). At first, the intention was merely to accelerate the system to a higher bit rate, for example, 9.6 kbit/s. Very soon more thorough considerations led to an improved concept with digital speech, extended data transmission capabilities, and other PMR features called *Digital Interchange of Information and Signaling* (DIIS). The original goal was a European nontrunked digital PMR standard based on FDMA and well matched to small systems and the lower market segments. Initially, only scrambling at the air interface was planned (nevertheless there is always the possibility to use proprietary end-to-end encryption). Encryption and trunking were not included in the scope to avoid direct competition with TETRA, but eventually they were introduced into the considerations.

The main feature of DIIS is that it provides a modulation rate of 12 kbit/s within a channel separation of only 12.5 kHz, which results in very good frequency economy even if it is unable to match TETRA. Moreover, DIIS employs a constant envelope modulation scheme that provides good transmitter efficiency and lower broadband noise properties than linear modulation, as has been considered in more detail in Chapter 5. Another important point is that no coexistence problems with systems in adjacent channels or with distant cochannel cells are to be expected because DIIS has been designed to meet the limits of EN 300 113 [3, 4]. Table 7.1 lists the main technical characteristics of DIIS.

DIIS can be used in all PMR bands from 30 to 500 MHz. Primarily it has been developed for a channel separation of 12.5 kHz, but adaptations for 20 and 25 kHz are also possible if the market dictates. However, it is much more likely that sub-bands with 12.5 kHz of separation will be introduced by refarming bands with existing 20- and 25-kHz channel allocations. To start, spectrum for DIIS around 446 MHz has been requested.

Because the sensitivity of DIIS receivers is similar to analog receivers, a change from analog PMR to DIIS will not reduce the coverage as long as similar frequencies and transmitting power levels are used (see Chapter 3). Hence, the old BS sites can be used without having to install additional infill equipment, and DIIS does not need new frequency planning methods. For DIIS systems without a fixed infrastructure and for bands with unregulated spectrum, *dynamic channel allocation* (DCA) would be a suitable access method, possibly together with *dynamic number assignment* (DNA). If there is a fixed infrastructure, then PSTN and Internet access will be possible with DIIS, which will also support IPv6.

Table 7.1 General Characteristics of DIIS

Property	*Remark*
Frequency range	30–500 MHz
Channel separation	12.5 kHz
Channel access	FDMA
Mode of operation	Simplex, semiduplex, duplex, TDD[1] and VTDD,[1] circuit switched and packet data
Repeater types	Transparent, reclocking, error correcting (delay ≤ 7 ms)
Transmitter power	BS: 40–50 dBm, MS: ≤ 40 dBm, all according to national regulations and licensing conditions
Frequency tolerance	≤3 kHz between unsynchronized mobiles
Modulation rate	12 kbit/s gross, 4GFSK with BT = 0.28 and h = 1/3
Error protection	Reed-Solomon codes with r_C = 1/4 to 1/2 plus CRC[2]
Data rates	9.6-kbit/s transparent and unprotected; 7.2-kbit/s light protection at good propagation conditions; 3.6-kbit/s strong protection at bad propagation conditions
Speech coding	Blocks of 20, 22.5, and 30 ms feasible, ≈2.4–4.8 kbit/s depending on propagation conditions and mode of operation; speech encoding algorithm yet undecided
Encryption	Scrambling only (end-to-end encryption applicable)
Symbol duration	166.7 μs, symbol rate 6 ksymbols/s
Time slot duration	20 ms
Frame structure	18 time slots per frame, 2 of them for signaling
Superframe length	1 superframe = 4 frames of 360 ms each = 1,440 ms
Synchronization	30 symbols followed by 7 symbols for phase correction
Tx soft keying	6 symbols with 3 symbols or 500-μs overlap for delay compensation
Rx settling time	1 symbol
Coverage distance	Maximum 75 km without repeater, 500-μs permissible delay[3]
Rx sensitivity	≤–107 dBm with Rayleigh fading at BER = 1% for the transparent channel without error protection[4]
Cochannel rejection	≤–17 to –19 dB with Rayleigh fading at different speeds
Adjacent channel rejection	≈50 dB with Rayleigh fading
Doppler spread	≤140 (230) Hz for 150 (250) km/hr at 500 MHz
Delay spread	≤50 μs; equalizer optional for high-end equipment
Mobile speed	≤150 km/hr, ≤1 dB of degradation at 250 km/hr
Services	All important PMR features including TI[5] and LE[6]
Other limits	As EN 300 113
Coexistence	With PMR in accordance to EN 300 086, 113, 219, and so on

[1]*Time division duplex* and *variable* TDD on one single carrier.
[2]Not yet finally agreed.
[3]Two-way distance 150 km.
[4]For digital speech with FEC up to 5 dB better.
[5]Transmitter interrupt.
[6]Late entry.

Despite the narrow channel separation, the relatively high gross bit rate of 12 kbit/s has been achieved by the introduction of 4GFSK with appropriate modulation index and filtering, that is, h = 1/3 and BT = 0.28. This type of modulation may be processed in the receiver by a coherent demodulator, but simple noncoherent demodulation for cheap products will also be feasible. Due to the long symbol duration and the robust FSK scheme, DIIS is not very vulnerable to Doppler shift, Doppler spread, and delay spread. The permissible maximum mobile speed is around 150 km/hr, but even at 250 km/hr only a small degradation will be experienced.

Figure 7.1 shows the basic burst structure of DIIS. After six symbols of soft *carrier ramping up* (CRU) and one subsequent symbol for receiver *settling time* (ST), seven pilot symbols (*channel tracking sequence,* CTS) are provided for fine-tuning of the receiver. The first symbol in a time slot is always the first of these pilot symbols. For self-contained messages or at the beginning of a transmission, a full synchronization sequence (*master synchronization sequence,* MSS) lasting 5 ms and consisting of 30 symbols is inserted between ST and CTS in the preceding time slot. This is advisable because in the case of self-contained messages or before the start of a payload transmission it is usually unused. Furthermore, this has the benefit that the core of the burst always has the same format. The remainder of the burst carries user information and the error correction overhead. Each burst is terminated by six symbols for soft *carrier ramping down* (CRD). The soft ramps are needed to reduce the switching bandwidth to a negligible extent as has been explained in Chapter 5. Moreover, there is an *overlap* (OVL) period of three symbols for the ramping times of consecutive bursts from different sources. Hence,

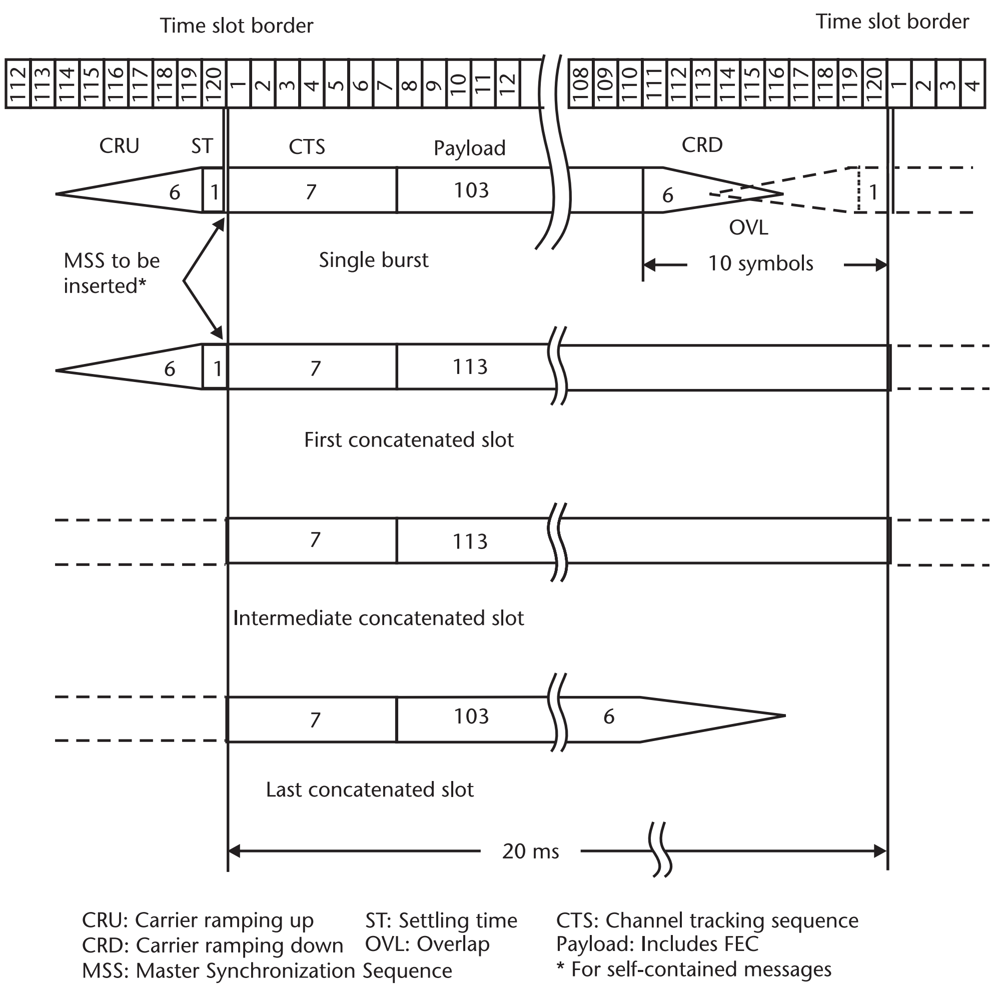


Figure 7.1 The burst structure of DIIS.

a shift of three symbols equivalent to 500 μs is permissible without interference of the burst content. This is of specific interest for sales in countries with low population densities because it permits different mobiles to have large distance differences from base stations of up to 75 km (equivalent to 150-km two-way delay). At these distances there is usually no coverage, but in specific cases and, in particular, at low frequencies such distances can be catered to without the need for additional guard time.

With repeaters, up to twice this distance can be achieved. Three different types of repeaters have been defined. Simple transparent repeaters retransmit incoming bursts without further processing, but there is always the danger that unwanted or erroneous signals could also be retransmitted. In contrast, error-correcting repeaters remove all errors from the first transmission path before retransmission of the burst. Hence, retransmission will usually take place in the next but one time slot. Specific timing problems are associated with reclocking repeaters, which remove channel impairments from every symbol without performing error correction or further processing of the message contents, but they have the advantage of introducing a transmission delay of only 7 ms or less.

The time slot duration of 20 ms is equivalent to 120 symbols. Due to ramping up, pilot information, ramping down, and overlap 103 symbols or 206 bits remain available for useful information (message or payload content) and redundancy. In the case of concatenated bursts, 113 symbols are available for information in all bursts except the last. The additional 10 symbols may be used for a slow associated signaling channel so that the payload contents for every burst is 103 symbols. The code rates are between approximately r_C = 1/2 and 3/4, which gives good error correction capability. For short self-contained messages that include the MSS inside the burst the useful information content including redundancy is reduced to 73 symbols or 146 bits.

One DIIS frame comprises 18 time slots each of 20-ms duration and, therefore, lasts 360 ms. This has been set to be a multiple of 9 time slots to get a good match with speech coders that block a duration of 20, 22.5, or 30 ms. The first 2 time slots of a frame are used for signaling, for example, for the transmission of *channel status* (CS) information, *forward signaling* (FS), and *transmitter interrupt* (TI) commands. The latter are needed to interrupt an ongoing transmission, to establish an emergency call, to join an ongoing group communication by *late entry* (LE), to switch off a faulty transmitter, or for control of TDD with varying transmit-to-receive ratios. The duration of signaling messages is only equivalent to 1 time slot, but the processing time needed by reclocking repeaters means that 2 time slots are required. The remaining 16 time slots of a frame are available for messages and other payloads. Four of these frames comprise a superframe with a duration of 1,440 ms as shown in Figure 7.2.

A single time slot as well as concatenated messages with a maximum duration of 16 time slots are supported within the frame structure. Circuit mode links such as speech or real-time data transmission for slow scan video and so on will be transmitted in this concatenated burst format.

As with most other PMR systems DIIS will offer simplex, two-frequency semi-duplex and full-duplex operation. It has not yet been decided which speech codec should be specified for DIIS. However, the use of a variable-rate codec with two

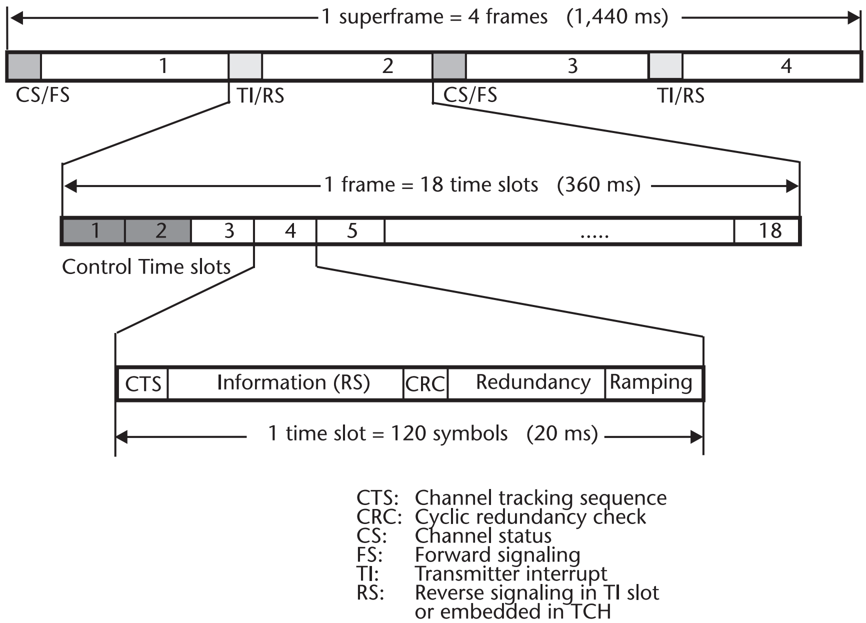


Figure 7.2 The frame structure of DIIS.

transmission rates as a minimum is envisaged. These should be on the order of 2 and 4 kbit/s, which would offer very interesting additional system features. Normal operation would use the higher rate whose error protection would provide sufficient speech quality under normal propagation conditions. Under severe conditions due to very weak signal reception or strong multipath distortions the speech coder would be switched to about 2 kbit/s with slightly reduced speech quality, but considerably stronger FEC and would therefore allow much better intelligibility. Additionally this low bit rate can be used for single-frequency TDD or for simultaneous transmission of speech and data.

Usually in duplex mode only one direction at a time carries a signal, at least if speech is transmitted. This has led to the idea of *variable time division duplex* (VTDD) with a variable transmit-to-receiving ratio. If 90% of the transmission capacity is devoted to the direction of main information flow, both users will have the impression that they have full capacity available because there is always the possibility to reverse the direction by speaking. However, if the direction of information flow is reversed then the capacity assignment for both directions also has to be reversed. Without any preference for any direction 50% of the transmission capacity is assigned to each.

With nothing happening the DIIS channel is in the idle state. If any activity is envisaged, the channel turns into the reserved state for this specific activity. In this state the available capacity may, under some circumstances, be used for short messages and commands. At the time the expected activity really starts the channel enters the transmission or payload state. If the activity is terminated, the channel is released. This may be a transition back to the idle state or to the reserved state if another activity has been prepared in the meantime. The DIIS protocol stack has a layered structure similar to that of TETRA. DIIS layers 1 and 2 are common for the control and user planes, but the control plane of layer 3 in particular contains additional functions.

DIIS offers four main types of services: *digital voice* transmission, *short data* transmission of various kinds, *general data* transmission including file transfer, and a *transparent digital channel*. The maximum unprotected user bit rate is 9.6 kbit/s. With light error correction 7.2 kbit/s is available, and with strong error correction 4.8 kbit/s is provided. Similar to BIIS 1,200 there will be the possibility to introduce user-specific applications without affecting the basic functionality. Application-specific speech coders may also be used in addition to the default speech coder. However, communication with the default coder must always be possible.

As with GSM and TETRA, a MoU has been set up for DIIS to promote the standard and to ensure that manufacturers' products comply with the standard. This MoU communicates the properties of DIIS equipment and systems to the outside world and will introduce a suitable brand name for DIIS, maybe DISCUS. One of the most important goals is that all DIIS products carrying the final brand name will be interoperable over the air interface.

As mentioned in Chapter 2, PMR 446 is expected to be very successful in Europe and therefore the combination with the digital transmission principle of DIIS is an interesting perspective. It could lead to a digital version of PMR 446 that might be called DPMR 446. If this happens, it is most likely that a chip set will be created to enable large production numbers at low cost. These chips could also be used for normal DIIS radios and cut down their costs considerably.

7.3 Narrowband PMR

For a long time, the conventional method for improving frequency efficiency has been by channel splitting. However, we showed in Chapter 3 that this is not necessarily the best approach. If the transmission bandwidth could be reduced without any drawback such as reduced transmission capacity and increased vulnerability to noise and interference then this would be the superior strategy and a reduction of the carrier separation would result automatically. Hence, for at least the last two decades studies and field trials using SSB techniques have been carried out, mainly in the United Kingdom and United States. In the United Kingdom, specific PMR frequency allocations were made for SSB applications in Band III around 215 MHz and in the United States at 225 MHz.

This concept has two drawbacks. First, SSB has comparable or greater vulnerability to single-frequency interference than frequency or phase modulation and, second, the modulation process needs increased expenditure because a linear transmitter power amplifier is needed (see Chapter 5). Nevertheless SSB makes sense in PMR. The receiver noise bandwidth is small and all of the transmitted signal power is carrying useful information while in AM, ASK (without suppressed carrier), FM, PM, FSK, and PSK a large amount of power is wasted in the unmodulated carrier. Hence, SSB offers a favorable link budget. If a reduced transmitting power suffices, then the requirements for the linear transmitter power amplifier can be met relatively easily and the poor efficiency of a highly linear power amplifier has less impact on the total power consumption. SSB performance may be further improved by the use of level compression, which increases coverage by reducing the dynamic AF signal range in the transmitter. The level compression is compensated by matched level expansion in the receiver to maintain acceptable audio quality.

By these means a SSB system can outperform conventional PMR systems, but at the expense of more complicated software and hardware. Today at least some of these difficulties can be solved by mass-produced integrated circuits, particularly *digital signal processors* (DSPs), so SSB has become an important alternative even if it is not yet supported broadly. However, to be fair, it should be added that if the compander principle were to be applied to 12.5-kHz angle modulation systems, these would perform even better, at least as far as interference suppression is concerned.

This is the background that obliged ETSI to commence standardization of a narrowband system for PMR applications. EN 301 166 is applicable for all channel separations below 10 kHz in the frequency range from 30 to 3,000 MHz, but it does not specify the type of modulation or the format for data transmission. It specifies a transparent channel usable for analog transmission. Digital transmission may be performed if a radio modem is additionally applied. Hence, this standard is another coexistence standard similar to EN 300 086 and 113, but not a system standard [5].

Depending on the type of code, the code rate, and the user bit rate, different limits, in particular for CCR, have been defined as shown in Table 7.2. The limits for receiver sensitivity and adjacent channel rejection are less stringent and therefore no distinct limits for different bit rates have been specified.

Table 7.2 Main Properties of the Narrowband Standard EN 301 166

Property	*Remark*
Frequency range	30 MHz to 3 GHz
Channel separation	≤10 kHz
Channel access	FDMA
Mode of operation	Simplex, semiduplex, duplex, repeaters
Transmitter power	PEP[1] according to national regulations
Frequency tolerance	BS: ≤0.3 kHz below 300 MHz and ≤0.5 kHz above; MS and HP: ≤10% of the channel separation
Modulation	Not specified, constant envelope not required
Speech coding	Format not specified
Data transmission	Format not specified
Signaling	Format not specified
Error protection	Code and properties not specified
Rx sensitivity	≤–107 dBm static for 20-dB SINAD ≤–110 dBm static for BER = 1%
Cochannel rejection	≥–15 dB static for 14-dB SINAD ≥–12 dB static for ≤2.4 kbit/s ≥–15 dB static for ≤4.8 kbit/s ≥–18 dB static for ≤9.6 kbit/s ≥–24 dB static for ≥9.6 kbit/s
Adjacent channel rejection	HP: 50 dB, BS and MS: 60 dB, all static for 14-dB SINAD and BER = 1%
Identification	Mandatory
Other limits	As EN 300 086 and 113

[1]Peak envelope power.

So far the only implementation of interest of this new standard has been made by Securicor Wireless using the *reference vector equalization* (RVE) method to improve robustness against transmission impairments. However, the Securicor implementation of RVE in narrowband channels is not a standardized system, therefore it is described in Chapter 8.

7.4 APCO 25: The North American PMR Standard

In 1989 the *Association of Public Safety Communication Officials* (APCO) supported by federal, state, and local governments set up Project 25 for the definition of a digital radio communication system for public safety tasks. This system, usually called APCO 25, has been standardized by TIA primarily for the use in the United States and in Canada with strong support of Motorola. It is an open standard with proprietary elements covered by mutual licensing agreements with the involved manufacturers [6]. The mobiles usually operate digitally at a 12.5-kHz channel separation, but they are backward compatible to traditional analog 12.5-kHz FM (or more accurately PM) systems. It is mandatory that any mobile be capable of selecting an analog channel with CTCSS and analog voice transmission if needed.

The heart of APCO 25 is a *common air interface* (CAI) for 12.5-kHz channels with 4FSK modulation that ensures interoperability of radios from different manufacturers. It is intended for use in several frequency bands and can be run on single frequencies or in trunked mode and with conventional or trunked repeaters. In the latter case mobiles may roam actively or passively, meaning that they are not registered, but just monitor the network. During wide-area roaming the mobiles are identified by unique electronic serial numbers, network IDs, and user group IDs. Roaming between radio subsystems is controlled through an *inter-RF subsystem interface* (ISSI). User groups can operate as closed groups and may create their own group hierarchy [7–11]. Table 7.3 shows the main properties of APCO 25.

APCO 25 was specified for three different bands with high transmit power well suited to large distance coverage. Base stations may operate at up to 350W or even 500W, while mobile RF power is usually between 10 and 110W. With hand portables such high power levels are not feasible—they provide only 1–5W carrier power. For improved hand portable duty cycles it is of importance that power control and a sleep mode be available (duty cycle of 8 hr with 90–5–5% standby–receive–transmit operation).

The modulation used is a four-level FSK type with constant envelope called *continuous four-level FM* (C4FM) [11]. In the transmitter the original binary digital signal is converted into a four-level baseband signal, which is then passed through a Nyquist RC filter and a shaping filter before being fed into a linear FM modulator. For the transmitter a nonlinear class C power amplifier may be used. The relationship between the dibits (four-level symbols) and the frequency deviation is 1.8 kHz for "01," 0.6 kHz for "00," –0.6 kHz for "10," and –1.8 kHz for "11." Compared to binary coding, the first two states have been exchanged such that, similar to Gray coding, there is never more than a 1-bit difference between adjacent states, which remains true even when they are mapped onto a QPSK scheme as it was intended for Phase II of APCO 25.

Table 7.3 Main Properties of the APCO 25 CAI

Property	*Remark*
Frequency bands	138–174, 406–512 (later extended to 380–512), and 764–869 MHz
Channel separation	12.5 kHz, 6.25 kHz[1]
Channel access	FDMA
Mode of operation	Simplex, semiduplex, duplex, circuit switched and packet data
Transmitter power	BS: ≤55 dBm, MS: 40–50 dBm, HP: 30–37 dBm
Frequency tolerance	According to FCC
Modulation	4FSK[2] with $h = 0.25$ and $r_O = 0.2$, CQPSK[1,3]
Modulation rate	9.6 kbit/s, including 2.4-kbit/s signaling
User bit rate	7.2-kbit/s transparent without error protection; 4.8 kbit/s with FEC
Speech coding	IMBE[4] at 4.4 kbit/s plus 2.8 kbit/s for FEC
Data transmission	Messages and packet data
Throughput delay	250 ms for DMO; 350 ms via single conventional repeater; 500 ms within an APCO 25 subsystem
Rx sensitivity	≤−105 dBm with Rayleigh fading
Cochannel rejection	≥−16.5 dB with Rayleigh fading
Adjacent channel rejection	≥60 dB static
Services	All relevant PMR services for public safety tasks
Backward compatibility	Analog voice and CTCSS on operational demand

[1]Initially envisaged for Phase II.
[2]Constant envelope, also known as C4FM.
[3]Also QPSK-C or compatible QPSK, for backward compatibility to Phase I.
[4]Improved multiband excitation.

A bit rate of 9.6 kbit/s means a symbol rate of 4.8 ksymbols/sec for C4FM. According to (5.57) we find a modulation index of $h = 0.25$. The filter response $H_N(f)$ of the RC Nyquist filter is given by

$$H_N(f) = \begin{cases} 1 & \text{for } f < 1{,}920 \text{ Hz} \\ \dfrac{1}{2} + \dfrac{1}{2} \cdot \cos\left[\pi\left(\dfrac{f - 1{,}920}{960}\right)\right] & \text{for } 1{,}920 \text{ Hz} < f < 2{,}880 \text{ Hz} \\ 0 & \text{for } f > 2{,}880 \text{ Hz} \end{cases} \tag{7.1}$$

Because the symbol rate is $R_S = 4.8$ ksymbols/sec, the 6-dB cutoff frequency of the filter is 2.4 kHz and (7.1) suggests that the RC filter slope runs from 1,920 to 2,880 Hz, from which a roll-off factor $r_O = 0.2$ results. The group delay of this filter is flat for $f \leq 2{,}880$ Hz. The subsequent C4FM shaping filter has the following characteristic $H_S(f)$:

$$H_S(f) = si\left(\frac{\pi f}{4{,}800}\right) \quad \text{for } f < 2{,}880 \text{ Hz} \tag{7.2}$$

The demodulator consists of a linear FM detector, an integrate and dump filter, and a stochastic clock recovery device. The filter is the same as the shaping filter in the transmitter. The maximum usable bit rate of an APCO 25 channel is the same as TETRA provides per time slot, namely, 7.2 kbit/s without error protection.

With FEC and interleaving, 4.8 kbit/s can be provided with a BER $\leq 10^{-6}$ and data transmission works up to BER = 7% measured on the unprotected channel.

Voice encoding is performed with an *improved multiband excitation* (IMBE) speech coder running at 4.4 kbit/s plus 2.8 kbit/s for FEC, which sums up with the signaling of 2.4 kbit/s to the modulation rate of 9.6 kbit/s. This coder delivers very good speech quality despite the relatively low bit rate. It is one of the latest coders, having been introduced only recently. Due to its good quality and low bit rate it was also selected by INMARSAT for digital voice transmission over maritime satellite channels [9].

A message service complements speech, circuit switched data, and packet data transmission facilities. For the various tasks different coding schemes are employed, for example, BCH codes for channel coding and network IDs, trellis coding for data, and Hamming codes for embedded signals.

For encryption of data and speech, the *Data Encryption Standard* (DES) is used. It was introduced about 25 years ago and is still strong enough to withstand all attacks made with reasonable computational power even today [12–14]. The encryption keys can be changed dynamically by OTAR. Hence, handling is more flexible and makes more frequent key changes possible to obtain better protection against misuse and fraud. APCO 25 employs four levels of encryption, but since systems and devices employing strong encryption are prevented from export by law only type 4 encryption can be used for export purposes:

- Type 1 Classified national government communications;
- Type 2 Unclassified security-related communications;
- Type 3 Unclassified sensitive government communications such as for public safety applications;
- Type 4 Commercial or other communications.

The frame structure of APCO 25 is relatively simple. A superframe lasts 360 ms and comprises 3,456 bits. It consists of two frames called *logical data units* (LDUs) with a duration of 180 ms or 1,728 bits each. For transmission of digitized voice nine voice coder blocks of 20 ms each carrying 144 useful bits (88 bits for IMBE and 56 for FEC) result in 1,296 information bits per frame. These 20-ms blocks are interleaved for better error resistance of the digitized speech. Each voice transmission starts with a header carrying all of the necessary synchronization, address, and encryption information. The encryption process does not change during a transmission and all control information is repeated every superframe to enable late entry [8, 15].

The *digitized voice header* consists of 48 synchronization bits and 64 bits for network ID plus 120 information bits protected by 528 bits for error correction. The 120 information bits comprise 72 bits of an encryption initialization vector, 8 bits of a manufacturer ID, 8 bits of an encryption algorithm identification, 16 bits of an encryption key identifier in systems with multiple encryption keys, and 16 bits of a talk group address. Additionally, 22 bits of channel status information for collision resolving are sent.

In the first LDU of a superframe, 72 information bits and 168 FEC bits for link control are contained. The link control information for *talk groups* comprises

8 bits of a talk group type of transmission, 8 bits of a manufacturer ID, 1 bit of an emergency flag, 15 bits reserved for future use, 16 bits of a talk group address, and 24 bits of a transmitting radio ID, which altogether is 72 bits. The caller ID is continuously transmitted during a communication for identification at any time. The address of an *individually called radio* not belonging to a talk group is composed of the last 56 of 72 bits. In detail these are 8 reserved bits, 24 bits for destination radio ID, and 24 bits for transmitting radio ID, all of which are protected by 168 bits for FEC. The second LDU of a superframe repeats 96 bits of encryption information protected by 144 FEC bits. Additionally, 16 information bits and 16 FEC bits for embedded *low-speed data* (LSD) are set in each LDU resulting in 88.9 bit/s for supervisory and monitoring tasks, for example, channel quality and location updating information.

APCO 25 also provides circuit mode and packet data transmission capabilities. Data messages are always divided into fragments of 4,096 bits maximum (due to past mobile memory restrictions) and each packet has a header comprising 196 bits. Fragments are divided into blocks of $N \times 8$ bits where $N = 12$ for unconfirmed messages and $N = 16$ for confirmed messages (except header blocks, which are 96 bits in length). A fragment, preceded by a header, is called a *data packet*. The data packet header is preceded by 48 synchronization bits plus 64 bits of network ID and contains 80 bits of address and control information plus 16 bits for error detection. Confirmed data packet blocks contain a 7-bit serial number for ARQ, 9 bits of error detection, and 112 data bits. Unconfirmed data blocks contain 96 bits plus 32-bit error detection code. Hence, the packet length is always 128 bits where the error protection is appended to the end of data packets. The header blocks use 1/2 rate trellis code for error correction and unconfirmed data packets also use 1/2 rate trellis code, whereas confirmed data packets usually use a 3/4 rate trellis code. Additionally, interleaving over the data blocks is applied.

Voice and data transmission can be integrated on the same channel because data messages usually last 100–300 ms while the duration of average voice calls is about 5 seconds. Telephone calls to the fixed network usually have a duration of less than 1 minute and are relatively seldom because they are only allowed for privileged officers. Voice and data integration is only feasible on lightly loaded channels of about 20 users maximum per channel for average traffic profiles. Besides normal voice calls in APCO 25, additional specific voice call types are available:

- *Talk group call:* All group members can be addressed simultaneously and latecomers can join a group communication by late entry.
- *Alert call:* This call enables the dispatcher to leave a message in an unattended mobile like a voice mail in a telephone system.
- *Emergency alarm:* This alarm is triggered by an emergency button and sent to the dispatcher and all group members. It initiates a visible and audible alert serving as a need-for-help indication. However, this alarm can also be silent to avoid local detection.
- *Digital priority scan:* This feature protects high-priority calls against interruption or termination by calls of lower priority.

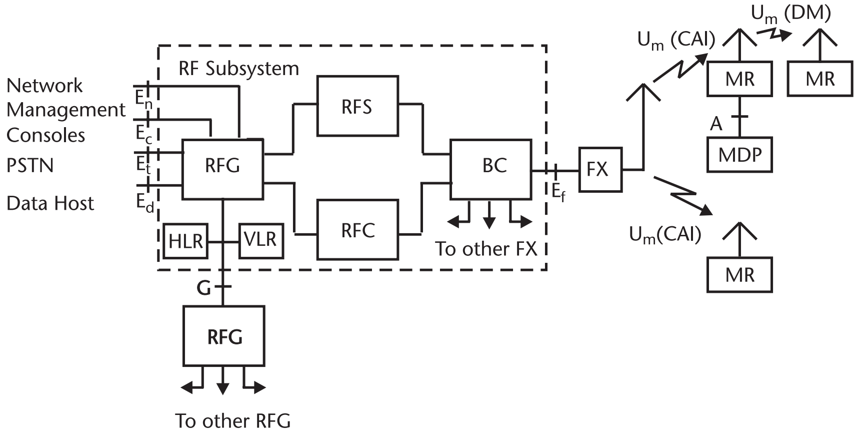
- *Selective radio inhibit:* This command allows the dispatcher to disable a mobile over the air to prevent misuse if it is lost or stolen.
- *Telephone interconnect:* Calls to the PSTN or other fixed networks can be made.
- *Private call:* Privacy is guaranteed, nobody can listen in.
- *Radio check:* This command serves to check whether a radio is operational.

All of the services needed for public safety tasks are available, for example, late entry, discreet listening, silent emergency listening (similar to ambiance listening in TETRA), priority calls and preemptive priority (for emergency calls), call restriction, and call alert. Groups can be created and dynamically reorganized. *Over-the-air programming* (OTAP) of the mobiles can be performed or the user can, in the field, change system group and user ID numbers or alter the transmission power level, encryption capability, priority level, assignment of talk group numbers, and analog channel selection. APCO 25 offers many trunking-like features, but is less expensive and complex than trunked systems. In fact, APCO 25 systems may be operated in a trunked or nontrunked mode and the organization is managed over a control channel. Let's look briefly at some of the most important trunking functions and features:

- *Emergency access:* This highest priority level overrides all others.
- *Queuing:* Call attempts are put in a queue if all channels are busy.
- *True priority control:* The queue is sorted according to the priority of the call requests.
- *Talk prohibit tone:* If the PTT is pressed while the radio is in the queue, the user gets a busy tone until he releases the PTT.
- *Call back:* An indication is given that the call in the queue has received a channel assignment and is ready for communication.
- *Automatic retry:* Automatic PTT repetition if the first trial did not succeed.
- *Continuous assignment updating:* Regular channel assignment messages are sent to provide late entry capability.
- *Out-of-range tone:* If connection to the system is lost due to a weak signal the user is informed by a specific tone.

In the infrastructure usually the *Simple Network Management Protocol* (SNMP) is employed, which is a widely used standard protocol for effective management of data networks. Network hosts provide X.25 and TCP/IP facilities and therefore connection to the Internet poses no problem. Apart from the air interface (U_m), seven additional interfaces as shown in Figure 7.3 have been standardized for compatibility of system elements from different manufacturers, as well as for easy system upgrading and enlarging. These are the mobile data port (A), the console interface (E_c), the data host interface (E_d), the base station interface (E_f), the network management interface (E_n), the telephone interconnect to the PSTN (E_t), and the intersystem interface for gateways to other networks (G).

For mobility management the user data are kept in *home location* and *visitor location registers* (HLR and VLR, respectively). Hence, user authentication is also possible. Besides normal frequency reuse in multisite systems for improved coverage,



Network Elements

BC	Base Control
FX	Fixed Station
HLR	Home Location Register
MDP	Mobile Data Peripheral
MR	Mobile Radio
RFC	RF Subsystem Control
RFG	RF Subsystem Gateway
RFS	RF Subsystem Switch
PSTN	Public Switched Telephone Network
VLR	Visitor Location Register

Standardized Interfaces

A	Mobile Data Port
E_f	Fixed Station Interface
E_t	PSTN Interface
E_d	Data Host Interface
E_n	Network Management Interface
G	RFC or Gateway Interface
U_m	Air Interface:
	CAI: Common Air Interface
	DM: Direct Mode

Figure 7.3 The standardized interfaces of APCO 25.

simulcast operation can also be implemented. Very often the systems are controlled by a dispatcher, which is the usual mode of operation for public safety. APCO 25 permits full-duplex operation with the base stations, but also direct communication between mobiles.

Figure 7.3 also shows the general structure of an APCO 25 system. Note that the names of many network elements are different from those of other systems even if they have the same task. The acronym FX represents the fixed station and is equivalent to BS or BTS and the mobiles radios are called MRs instead of MSs or MTs. However, the tasks to be performed are in some cases differently grouped compared to other systems; for instance, RFC is the RF subsystem control, RFS is the RF subsystem switch, and RFG is the RF subsystem gateway [7, 16]. The ISSI supports unit registration and tracking, group affiliation and tracking, unit and group migration, and individual and group calls by several ISSI service classes, which may be assigned to single users or to groups, and combinations of different service classes are possible, for example, the combination of classes 1, 4, and 5:

- *Service class 1:* Roaming with voice and status message service;
- *Service class 2:* Circuit and packet switched data service;
- *Service class 3:* Telephone interconnect service;
- *Service class 4:* Encryption service;
- *Service class 5:* OTAR service.

APCO 25 equipment is supplied by several manufacturers, for example, Motorola delivers ASTRO 25 and RACAL 25 equipment comes from Thales Communications, a result of the merger between Racal and Thomson CSF.

The goal for the development of APCO 25 Phase I was to double spectrum efficiency and to employ digital transmission including digitized voice as well as circuit mode and packet data transmission. For Phase II, which was considered for introduction around 2005, a further improved spectrum efficiency with an effective channel separation of 6.25 kHz is the goal. The system specification is laid down in the TIA/EIA-102 series of standards written by the TIA Committee on Mobile and Personal Private Radio Standards (TR-8) and comprises 33 documents for APCO 25 Phase I and an additional 6 documents for Phase II.

Additionally, Phase II radios should be backward compatible to Phase I radios for easy upgrading. This would have been possible without too much expenditure by replacing C4FM by a quarternary PSK scheme called *compatible QPSK* (CQPSK or QPSK-C). This occupies only half of the modulation bandwidth at an unchanged bit rate fitting in the 6.25-kHz channels. In the modulator a look-up table would have been used for the creation of the I and Q components of the RF signal. Then both, one shifted by 90°, would have been passed through Nyquist RC lowpass filters and added, generating a nonconstant envelope QPSK signal with *continuous phase* (CP) transitions, but requiring a linear class AB transmitter power amplifier. This modulator can also be used for C4FM and in turn the C4FM demodulator described above is also well suited for CQPSK because the two modulation schemes belong to the same family. Finally, the linear transmitter needed for CQPSK can also process C4FM and, therefore, a common transceiver design for Phase I and Phase II radios to ensure backward compatibility would have been possible with this design strategy [15]. However, Phase II characteristics have been subsequently changed.

Future development in the United States will be strongly influenced by Project 34, more precisely called Project 25/34, which is Phase II of Project 25 [8, 17–19]. The initial idea of CQPSK on 6.25-kHz channels has been abandoned and cooperation between TIA and the ETSI TETRA project has been started instead. Interestingly during the debate about how to proceed with APCO 25 Phase II, Ericsson took up the old idea of a two-slot TETRA system with a channel separation of 12.5 kHz as an alternative to the two other possibilities under consideration, which were 6.25-kHz FDMA narrowband and 4TDMA TETRA in 25-kHz channels [20, 21]. All three proposals result in the same spectral efficiency. However, later they were all suspended and in early 2002 a new 2TDMA proposal from *EADS Defence & Security Networks* (EDSN) and Nortel Networks was discussed. Based partly on TETRAPOL technology, it would offer a good match to the geographical requirements of the U.S. public safety agencies and provide backward compatibility with Project 25 Phase I conventional and trunked systems.

To achieve reliable high data transmission speeds over the mobile radio channel for purposes such as the transmission of video, photos, maps and so on, the future main goals of Project 34 have been defined as follows:

- Maximum radio spectrum efficiency;
- Transport network for multiple standardized protocols;

- Two-way gross transmission rates up to 1.544 Mbit/s and eventually 155 Mbit/s, initially 384 kbit/s in 125-kHz channels;
- Seamless handover;
- Multiple levels of security and network integrity;
- Short transmission delay with minimum errors even under harsh propagation conditions.

In the EU countries similar goals are envisaged as already mentioned in conjunction with TETRA Release 2. Hence, Project 34 decided to cooperate via TIA with ETSI, initially specifically on DAWS, which as mentioned earlier has been superseded by the MESA project.

7.5 Comparison of Standardized PMR Systems

PMR system engineers are very often required to compare the main properties of different radio systems. This is usually the starting point of frantic action as they try to collect the relevant data. From my own experience, I know this situation very well and it has always been a difficult task to retrieve the needed data quickly and completely. To aid the reader who may face similar tasks, Table 7.4 has been compiled to aid in comparing the most important characteristics of standardized digital PMR systems including TETRA V+D.

Comparing the properties of the different systems leads to the conclusion that in most cases the characteristics are similar—This is true for very simple reasons. The channel separations of all of the systems are not dissimilar, the physical laws governing the systems are the same, and expenditures must be limited to a reasonable extent for economic reasons. The main parameters of a system can be traded against each other, however; if one is improved another one necessarily gets worse. Moreover, all systems should be suited to very different applications and therefore a reasonable trade-off between all the relevant parameters is necessary. Therefore, good engineering practice always leads to similar results even if the approach is different.

7.6 Perspectives of Standardized PMR Systems

At present it is uncertain which PMR systems will get the biggest market share worldwide. However, many market analysts believe that the chances for standardized systems in the long term are considerably better than for nonstandardized systems. Hence, APCO 25 and TETRA will have best chances and DIIS, after completion and launch, may attract users of small systems and will provide solutions in frequency allocations where the "big" systems are not available.

However, the PMR market is currently changing dramatically and moving faster than ever. Hence, a well-polished crystal ball would be necessary to make sound predictions. This risk is left for market analysts to take. Another problem is that PMR experts do not always have the right feeling for influences from other areas, while on the other hand most communications market analysts are not real PMR experts. Hence, all PMR market reports and predictions have to be read very carefully.

Table 7.4 Most Important Properties of Standardized Digital PMR Systems

Property	*TETRA V+D*	*DSRR*	*DIIS*	*EN 301 166*	*APCO 25*
Frequency bands	380–520, 806–933[1] MHz	888–890,[2] 933–935 MHz	30–500 MHz	30–3,000 MHz	138–174, 380–520, 800–900 MHz
Channel separation	25 kHz	25 kHz	12.5 kHz	≤10 kHz	12.5 kHz
Channel access	4TDMA	FDMA	FDMA	FDMA	FDMA
Mode of operation	S, SD, D, packet data	S, SD[2], D[2]	S, SD, D, packet data	S, SD, D	S, SD, D, packet data
Tx power BS:	28–46 dBm	36 dBm	≤50 dBm	According to national regulations	≤55 dBm
MS:	24–45 dBm	36 dBm	≤40 dBm		≤50 dBm
HP:	15–35 dBm	≤36 dBm	≤35 dBm[3]		≤37 dBm
Unsynchronized frequency error	—[4]	≤2.5 kHz	≤3 kHz	≤0.3/0.5 kHz, ≤10%[5]	According to FCC
Modulation	$\pi/4$-DQPSK $r_O = 0.35$	GMSK $h = 0.5$ BT = 0.3/0.5	4GFSK $h = 1/3$ BT = 0.28	—[4]	C4FM $h = 1/2$ $r_O = 0.2$
Gross bit rate	36 kbit/s	4/16 kbit/s	12 kbit/s	—[4]	9.6 kbit/s
Maximum user bit rate	28.8 kbit/s transparent	—[4]	9.6 kbit/s transparent	≥9.6 kbit/s protected	7.2 kbit/s transparent
Speech coding	ACELP, 4.6 kbit/s	RPE-LTP[6], 13 kbit/s	Undecided, 2.4/4.8 kbit/s	—[4]	IMBE, 4.4 kbit/s
Sensitivity BS:	≤–106 dBm	≤–107 dB	≤–107 dB	≤–110 dBm	≤–105 dBm
MS:	≤–105 dBm	≤–107 dB	≤–106 dB[9]	≤–110 dBm	≤–104 dBm[9]
HP:	≤–103 dBm all dynamic[7]	≤–107 dB all static[8]	≤–104 dB[9] all dynamic[7]	≤–110 dBm all static[8,10]	≤–103 dBm[9] all dynamic[7]
Cochannel rejection	≥–19 dB dynamic[7]	≥–18 dB static[8]	≥–19 dB dynamic[7]	≥–12–24 dB static[8]	≥–16.5 dB dynamic[7]
Adjacent channel rejection	≥57 dB dynamic[7]	≥50 dB static[8]	≥50[2] dB dynamic[7]	≥50/60 dB static[8,11]	≥60 dB static[8]
MS speed	200 km/hr	—[4]	≤250 km/hr	—[4]	—[4]
Delay spread	≤100 μs	—[4]	≤50 μs	—[4]	—[4]
Backward compatibility	—	—	Analog PM[12]	—	Analog PM, CTCSS

[1]Various European and overseas bands.
[2]Only some European countries.
[3]Expected.
[4]Not specified.
[5]Absolute for BS for below/above 300 MHz, percent of channel separation for MS.
[6]Same codec as GSM.
[7]Rayleigh fading only.
[8]Dynamic values; see Chapter 9.
[9]Estimated.
[10]Analog 3 dB worse.
[11]Lower limit HP only.
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CHAPTER 8

Proprietary Digital PMR Systems

In the past, several digital trunked PMR systems have been developed by different manufacturers for public and nonpublic applications. In the United States these systems are known as digital *specialized mobile radio* (SMR), whereas in Europe this service is called *public access mobile radio* (PAMR). The first attempts to develop digital systems of this type date back to the second half of the 1980s. In the meantime some manufacturers have left the PMR business or have merged with others. Consequently, some digital PMR systems have disappeared, so a reduced number of proprietary systems are currently available or have market relevance. In particular, in Europe most of the proprietary systems no longer play a significant role because with one exception their manufacturers are strongly promoting TETRA.

Currently, in Europe only EDACS and TETRAPOL have any significance but analysts predict that the market share of EDACS will decline. However, worldwide the situation is different and iDEN is growing fast and holds a significant market share in the United States and many other countries. Specific data transmission systems have been developed and were launched many years ago but did not succeed as expected because they show a lack of flexibility compared to those that are capable of transmitting speech and data.

8.1 ASTRO

ASTRO is an FDMA system developed by Motorola that has been available for a number of years in the 2-m band and also at higher frequencies. It is available in four standard channel separations and can be configured from a single site with one channel up to multichannel, multisite systems. ASTRO systems can be operated in a conventional or trunked mode and simulcast networks can also be built. To balance the link budget, ASTRO is capable of receiver voting. Despite not being standardized, ASTRO is close to a standard because the APCO 25 system is broadly based on ASTRO. Furthermore, ASTRO is backward compatible with Motorola's SmartNet single-site systems and SmartZone multisite networks that operate with 3.6 kbit/s signaling. ASTRO employs C4FM modulation method with a gross bit rate of 9.6 kbit/s. For error correction BCH, RS, and Golay codes are used. For the user a maximum of 7.2 kbit/s is available in an unprotected format. The available MS power is 10W to 100W in the VHF/UHF bands and a maximum of 35W from 806 MHz to 870 MHz [1, 2]. The main characteristics of ASTRO are given in Table 8.1. For further details the APCO 25 description in Chapter 7 may be helpful.

For speech coding the *vector sum excited linear predictive coding* (VSELP) algorithm is used. Speech is sampled in 30-ms slices called *voice frames*. The VSELP runs at 4.8 kbit/s and is protected with 2.1-kbit/s Golay-coded FEC. A further 2.7 kbit/s are available for embedded signaling, which gives a total of 9.6 kbit/s. If the FEC capability is exceeded, the speech signal is extrapolated to replace faulty voice frames. In the case of voice frames that are too severely damaged for correction, progressive muting is used. Instead of the VSELP coder a *continuously variable slope delta* (CVSD) modulation speech codec is available as an alternative on demand, and in ASTRO 25 equipment for APCO 25 the IMBE coder is mandatory.

ASTRO provides an integrated voice and data transmission capability and employs the *radio data link access procedure* (RD-LAP) for data transmission. Voice transmission can also be performed using *Voice over Internet Protocol* (VoIP), which means that it is based on packet data transmission. AVL using GPS data transmission can be applied for fleet management.

The superframe duration is 360 ms and the structure is similar to the APCO 25 CAI. Channel access is performed using slotted *carrier sense multiple access* (CSMA), and call setup requires 500 ms in the worst case. In conventional mode 10,000,000 user IDs are available, whereas in trunked mode this is limited to 48,000 user IDs, together with 4,096 (2^{12}) additional network access codes and 4,096 user group IDs. For comparison, APCO 25 provides 65,000 user group IDs (2^{16}) in conventional mode and 4,094 in trunked mode.

ASTRO also provides encryption and OTAR. Two encryption algorithms and 16 different keys are available per radio. The centralized system management is

Table 8.1 Main Properties of ASTRO

Property	*Remark*
Frequency bands[1]	136–174-, 380–520-, and various 700/800/900-MHz bands
Channel separation	12.5, 20, 25, and 30 kHz
Channel access	FDMA
Transmitter power[2]	BS: 40–50 dBm, MS: 40–50/≤45 dBm for VHF and UHF/800 MHz HP: ≤37/36/35 dBm for VHF/UHF/800 MHz
Modulation[3]	C4FM of QPSK-C family
Modulation rate	9.6 kbit/s
User bit rate	7.2 kbit/s unprotected
Speech coding	VSELP[4] at 4.8 kbit/s plus 2.1 kbit/s for FEC
Receiver sensitivity[2,3,5]	MS: ≤−117 dBm, HP: ≤−115 dBm, all static for 12.5 kHz at BER = 1%
Cochannel rejection[2,3,5]	−16.5 dB for 12.5 kHz with Rayleigh fading
Adjacent channel rejection[2,3,5]	MS: ≥70[6]/80 dB for 12.5/20, 25, and 30 kHz HP: ≥≈65/75 dB for 12.5/20, 25, and 30 kHz, all static

[1]Main frequency bands, each type of equipment covers different sub-bands.
[2]Properties taken from various data sheets.
[3]See also APCO 25.
[4]Vector sum excited linear predictive coding.
[5]Dynamic values; see Chapter 9.
[6]≥65 dB at 800 MHz.

based on PCs running Windows and for network upgrading software downloads can be made. A wide range of multichannel mobiles, handhelds, and BS equipment is available for ASTRO networks. The latest digital ASTRO mobiles are fully digital from baseband to IF and can be equipped with either an RS-232 or USB serial data interface.

ASTRO does not play a significant role in Europe due to the reasons already mentioned. However, in other important worldwide PMR markets ASTRO has a healthy market share.

8.2 EDACS

The *Enhanced Digital Access Communication System* (EDACS) was originally developed by General Electric and then taken over by Ericsson, which was pursuing a straight migration strategy from an analog to a digital FDMA system oriented toward the migration path from analog APCO 16 to digital APCO 25. EDACS is an FDMA system available in the major PMR bands and also in the 800- and 900-MHz region, but only for channel separations of 12.5 and 25 kHz. GFSK modulation provides a modulation bit rate of 9.6 kbit/s and delivers good cochannel rejection. Delay spread is no problem since multipath propagation delay differences of more than 50 μs are permissible. The main technical properties of EDACS are listed in Table 8.2.

Conventional and trunked EDACS networks are well suited to the needs of public safety, large utilities, and commercial users. EDACS provides fast channel access, good voice clarity, security against fraud, and ease of maintenance. It also supports analog and digital voice and data communication. Networks with coverage from small towns to multisite configurations including repeaters serving large regions or countries can be built, and most of the important PMR services including late entry and direct mode are provided [1, 3–6].

Table 8.2 Main Properties of EDACS

Property	*Remark*
Frequency bands	160, 450, 700, 806–870, and 896–941 MHz
Channel separation	12.5 and 25 kHz
Duplex separation	4 to 45 MHz according to the frequency band
Channel access	FDMA
Transmitter power	BS: 56 dBm, MS: 40–50 dBm,[1] HP: 38 dBm
Modulation	GFSK
Modulation rate	9.6 kbit/s
User bit rate	9.1 kbit/s unprotected
	7.77 kbit/s lightly protected
Time slot duration	21 ms
Speech coding	AME or IMBE[2], ≈4.4 kbit/s; block length 21 ms
Sensitivity	–115/–116 dBm static[3] for 12.5/25 kHz
Cochannel rejection	12.5 kHz: –7/16 dB static/dynamic
	25.0 kHz: –5/14 dB static/dynamic
Delay spread	≤52 μs

[1]Several types of mobiles are available.
[2]Advanced multiband excitation, improved multiband excitation.
[3]Dynamic values; see Chapter 9.

EDACS offers encrypted digital voice and integrated voice and data transmission capability with one standard user bit rate. For AVL purposes the transmission of GPS data can be used. EDACS provides three types of channels: *control* or *signaling channels, traffic channels for voice,* and *data transmission channels.* Dispatcher control of large fleets is possible and up to 13,684 individual users and 2,048 call groups as a maximum can be served by a single EDACS system. For priority calls eight different levels including emergency are available and a mobile in emergency transmits its ID automatically to all recipients. Large EDACS networks consist of elements belonging to three different hierarchical levels called *tiers:*

- *Tier 1:* At the system tier all basic configurations from single-site to multicast systems are based on a specific *trunked failsoft architecture.*
- *Tier 2:* At the multisite tier two or more multisite systems constitute a fault-tolerant wide-area configuration controlled by an *integrated multisite controller* (IMC).
- *Tier 3:* At the extended network tier two or more IMCs connected in a star configuration provide the basis of a fault-tolerant distributed network.

In Tier 1 four different call types are available: *individual calls, group calls, group emergency calls,* and *broadcast,* or *all-calls.* Working groups can be established on four levels, namely, subfleet, fleet, agency, and system. EDACS is also capable of providing CUGs. Groups can be dynamically reorganized and late entry is possible. The gross transmission rate of 9.6 kbit/s allows high data throughput, fast late entry and reentry, high call capacity, and fast access times of less than 250 ms. To save transmission capacity, transmission trunking is preferred but message trunking is offered on demand. Roaming users are tracked by the IMCs and stolen or faulty mobiles can be remotely disabled. The most important specific EDACS features are listed next:

- *Dispatch operation:* Status messages, request-to-talk messages, paging, supervising functions, and so on can be handled by a dispatcher via a console.
- *Caller ID display:* In individual and group calls the caller ID is displayed.
- *Busy condition:* Call requests are put into a queue according to their priority level. If all channels are busy, eight priority levels are available.
- *Transmit busy lock-out:* If the call originating MS keys its transmitter all others are prohibited from transmitting. In any receiving mobile the user needs to push a button to indicate an emergency and to immediately send an emergency message.
- *Trunking priority group scan:* Any MS can monitor the calls of unselected groups while it remains able to receive a priority call.
- *Telephone interconnect:* Authorized users can make and receive telephone calls without dispatcher assistance. Calls from a telephone network may address individuals or groups as a whole.
- *Pro-scan:* In multisite networks roaming is based on decisions made by the MS according to signal strength from different BSs.
- *Pro-file:* This specific feature allows the system administrator to perform PC-based reprogramming of radio personalities over the air.

Digital voice coding can be achieved with the *advanced multiband excitation* (AME) algorithm, the *ProVoice* digital voice coding based on DSP technology, or the IMBE coder, which exhibits superior quality compared to older products. Bit interleaving and FEC make it less vulnerable to transmission errors and multipath distortion. ProVoice is backward compatible to ordinary EDACS, and ProVoice radios can operate in analog as well as in clear and encrypted digital mode based on DES encryption.

The high fault tolerance of EDACS networks is achieved by distributed processing and nonstop control channel operation. All protocol layers monitor each other and if one fails the others provide alternative modes of operation. The loss of a traffic channel has only a minor effect on a multichannel system. If, however, a control channel fails, then another channel takes over since any channel can be used as a control channel. If the site controller fails, then the system goes to *trunked failsoft mode* retaining emergency access priority. If the link from the network to a BS fails, then this BS remains operating autonomously. The modular design of EDACS networks offers easy expandability. Coverage problems and link budget imbalances can be solved with receiver voting diversity at the base stations. For the future development of EDACS a 3TDMA version is envisaged and also a migration path to TETRA has been considered.

In Europe EDACS is not very well positioned in the PMR markets. Worldwide it looks better because it is in use in the Americas, Asia, and Eastern Europe but currently EDACS exhibits only a relatively small growth rate.

8.3 iDEN

iDEN means *integrated Digital Enhanced Network* formerly called *Motorola Integrated Radio System* (MIRS). It was developed by Motorola but in contrast to ASTRO it is based on TDMA rather than FDMA. It is the only DPMR system employing a modulation scheme in excess of four levels. The 16QAM approach with *trellis coded modulation* (TCM) delivers 64 kbit/s for six TDMA channels, that is, a gross bit rate of 10.67 kbit/s per channel. Large multipath propagation delay differences do not present problems to iDEN because it has a relatively low symbol rate and a proprietary four subcarrier data recovery system. iDEN is one of the rare systems designed for frequencies beyond 1 GHz. It employs power control, and the transmitting power of mobiles can be controlled in the range from –6 to 27 dBm for frequencies below 870 MHz and from 0 to 40 dBm above 890 MHz. Speech coding is achieved using the VSELP method at 4.8 kbit/s. This codec suppresses background noise well and delivers an end-to-end speech delay of 160 ms. Channel access is performed with a slotted ALOHA procedure in about 210 ms [2–4, 7–9]. The main characteristics of iDEN are shown Table 8.3.

iDEN is a trunking system for PAMR applications based on an integrated voice and data design that offers a variety of PMR features and is well suited to the needs of many user groups. Remarkably one of the most popular iDEN services is mobile telephony. With the bundling of two slots, near toll voice quality can be provided with minimal speech delay, and interconnection to the PSTN is also available. Other very much appreciated features are that messages with up to 140

Table 8.3 Main Properties of iDEN

Property	*Remark*
Frequency bands	806–870, 890–960, and around 1,500 MHz
Channel separation	25 kHz
Duplex separation	45 MHz
Channel access	6TDMA
Transmitter power[1]	BS: 51 dBm, MS: 27–40 dBm, HP: 22–35 dBm
Modulation	16QAM, TCM
Modulation rate	64 kbit/s
User bit rate	6×7.2 kbit/s lightly protected
Time slot duration	40 ms
Speech coding	VSELP[2] at 4.8 kbit/s plus 2.4 kbit/s for FEC
Data transmission	Messages and packet data
Receiver sensitivity	≤−105.5 dBm with Rayleigh fading at BER = 1%
Cochannel rejection	≥−10 dB static, ≥−19.5 dB dynamic
Adjacent channel rejection[3]	≥70 dB static
Delay spread	≤40/66 μs with class A/B equalizer

[1]Burst peak power.
[2]Vector sum excited linear prediction.
[3]Dynamic values; see Chapter 9.

characters can be sent and voice mails can be stored if the called subscriber is absent. In dangerous situations emergency dialing is possible without the use of a PIN or SIM card, but this is network dependent.

In iDEN data transmission is generally based on packet switching, and an *interworking function* (IWF) for the interfaces to other networks is also provided. If more PMR-like kind of operation is needed, then iDEN also supports dispatcher operation to control a fleet of mobiles, but this is only possible across a single switch while interswitch calls are carried on duplex links via a GSM-style interconnection network. In addition to the already mentioned features, iDEN provides a comprehensive list of services:

- *Call handling:* Call forwarding is possible in unconditional and conditional modes; call waiting, call hold, and call transfer are also available.
- *Call barring:* Barring is possible for incoming and outgoing calls and the barring status is displayed.
- *Alternate phone lines:* Simultaneous access to two phone lines.
- *Three-way calling:* Conference calls for three subscribers can be made.
- *Phone only:* Private and group services can be switched off.
- *Private call:* A phone can be identified by a name or a private ID for a one-to-one conversation to ensure privacy.
- *Call alert:* This feature allows the paging of an individual by sending an audible tone to be responded to when it is convenient.
- *Call alert queuing:* Up to eight call alerts can be queued.
- *Group calls:* Calls can be made to a group of subscribers in a specific service area.
- *Area selection:* Can select a specific service area in group mode.
- *Auto call-back:* Responds to a message containing a call-back number.

iDEN has many user friendly features for easy handling, for example, automatic redial, on-hook dialing, and short number dialing. If during the operation of an iDEN phone a failure occurs, the user will be informed by an error status message stating the reason. The comprehensive list of status messages comprises: *system busy try later, please try again, user busy in private call, user not available, user not authorized, please try later, number out of service, service restricted, service not available, service conflict, self-check error,* and *self-check fail*. Software patches for mobile stations can be downloaded over the air interface, and to improve battery duty cycle a sleep mode has been implemented.

iDEN mobiles are equipped with RS-232 interfaces to provide wireless laptop Internet access. Multimode hand portables are available that can also be used with AMPS, *U.S. digital cellular* (USDC), and GSM. Hence, roaming with GSM and other networks is already possible or belongs to the group of envisaged future enhancements. Increased data rates of more than 100 kbit/s are also under consideration for a second generation iDEN system. A compact *mobile switching office* (MSO) for dispatch and telephone interconnect, remote access, and network and alarm management is available for small iDEN systems: The Harmony system was made to cater to up to 5,000 SMR, commercial, and industrial users in the 806- to 825-MHz band.

However, one major drawback is that iDEN does not offer a direct mode, which seems to be inevitable for a number of PMR applications, such as public safety. However, iDEN was never intended for this purpose and Motorola recommends ASTRO for such applications.

In the European markets iDEN is virtually unknown but it has acquired a big market share in the United States, where the largest iDEN network is operated by Nextel, and this system is also proceeding well in other overseas markets, such as Japan.

8.4 Enhanced Specialized Mobile Radio

SMR was originated by the FCC in the United States as early as 1974 for voice dispatch, data broadcast, and mobile telephony over mainly private analog PMR networks. These operated in the 800- and 900-MHz bands. *Enhanced specialized mobile radio* (ESMR) is the digital evolution of SMR. In 1993 the first iDEN-based ESMR network was opened by Nextel. In the United States today Nextel runs the world's largest ESMR network and offers a cellular-like service with nationwide coverage that competes with public mobile telephone networks and PCN/PCS. However, there are no roaming agreements with other wireless providers in the United States [1, 3].

All of the main iDEN features are available in ESMR. For PMR-style applications additionally a PTT service is available that provides direct contact with individual and group calls in half-duplex mode. Currently, this service is only available in a user's home area but can be extended for predefined groups to sites in other service areas. There are three versions of direct links: *local-area calls* in one service area formed by a combination of several continuous coverage areas, *selected service area calls* to every group member roaming in the selected service area, and *wide-area calls* to group members roaming in any part of the Nextel network.

The future development of ESMR is closely linked to the general progress in mobile telephony and will presumably follow the same trend as all 3G networks.

8.5 TETRAPOL

At present TETRAPOL is the main competitor to TETRA in Europe and it is spreading in other important worldwide PMR markets [10]. It is an FDMA trunking system that was developed by EADS Telecom (formerly Matra Communications and later Matra Nortel Communications) for the French public safety forces. However, many other civil applications are well served by TETRAPOL, for example, public transportation systems like buses, trams, and subways and coverage of airports, harbors, manufacturing plants, and various other industrial sites.

TETRAPOL is available for channel separations of 10 and 12.5 kHz but only in the 80- and 450-MHz bands; however, a 900-MHz version needed for specific regions of the world has been announced. The smaller channel separation is used in large networks with a very high traffic capacity that require a large number of carrier frequencies, for example, in the heavily loaded Paris conurbation. The Matra MC 9600 equipment, for instance, covers the UHF range from 380 to 512 MHz.

TETRAPOL employs GMSK with BT = 0.25 and a somewhat limited modulation rate of 8 kbit/s, but this on the other hand is why the TETRAPOL air interface is very robust against multipath propagation impairments. For a channel separation of 12.5 kHz, the spectral efficiency is only 0.64 bit/s/Hz, but due to the good dynamic cochannel rejection of –15 dB the frequency economy is 2.2 times better than that of analog 12.5-kHz systems (see Chapter 3). Additional improvements to the frequency economy can be achieved by trunking gain and in particular by message and quasi-transmission trunking as offered by TETRAPOL. The small bandwidth and robust modulation offer good receiver sensitivity and a favorable link budget, which permits large-area coverage. To improve the link budget further, antenna diversity may be used and simulcast transmission is offered where needed. Being an FDMA system TETRAPOL provides *direct mode* (DM) and open channel operation without difficulties when there is no support from an infrastructure. If necessary, *repeaters* (RPs) can be established in a TETRAPOL system to increase coverage distance for direct mode. In DM the mobiles can also monitor infrastructure activity by use of a dual-watch function [1, 11–18].

TETRAPOL meets the radio and EMC limits of EN 300 113 and EN 301 489-1 and -5. Therefore, there are no coexistence problems with analog or other digital systems operating in an adjacent channel or at a distance on the same frequency. The technical characteristics of TETRAPOL have been collected in Table 8.4.

TETRAPOL voice coding requires a DSP capable of 20 MIPS and uses the *regular pulse code excited linear prediction* (RPCELP) algorithm with 20-ms voice frames at 6 kbit/s. This bit rate is higher than in other systems but the RPCELP suppresses severe background noise very well and its coding scheme is robust against interference. Two hundred of these voice frames, carrying 120 information

Table 8.4 Main Properties of TETRAPOL

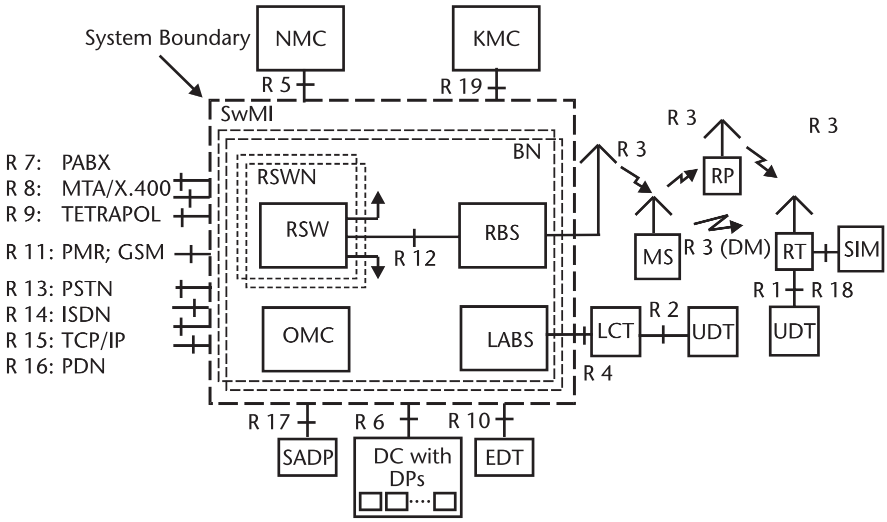
Property	*Remark*
Frequency bands	68–88, 380–512, and 830–930 MHz
Channel separation	10 and 12.5 kHz
Channel access	FDMA
Duplex separation	10 MHz in the UHF bands
Mode of operation	Simplex, duplex
Transmitter power	BS: 30, 34, 38, 42, and 44 dBm; 5 power classes MS: 30, 33, and 40 dBm; 3 power classes, including HP with 30 or 33 dBm
Tx power control range[1]	MS: 30 dB, HP: 20 dB
Modulation	GMSK with BT = 0.25
Modulation rate	8.0 kbit/s
User bit rate	≤7.4 kbit/s transparent[2] ≤4.6 kbit/s with weak protection[3] ≤3.3 kbit/s with strong protection[4] ≤3.2 kbit/s protected packet data[5]
Time slot duration	20 ms, 160 bits/slot, superframe length 4 seconds
Speech coding	RPCELP[6,7] at 6.0 kbit/s
Data transmission	Messages, circuit mode, and packet data
Sensitivity	≤−121/−119 dBm for BS/MS static, ≤−113/−111 dBm for BS/MS with Rayleigh fading
Cochannel rejection	≥−7 dB static, ≥−15 dB dynamic (≥−17 dB for unprotected transparent channel)
Services	All relevant PMR services for public safety tasks
Coexistence	EN 300 113 is fulfilled

[1]Dependent on the type of equipment.
[2]148 data bits/slot.
[3]92 data bits/slot.
[4]62 data bits/slot.
[5]64 data bits/slot.
[6]Regular pulse code excited linear prediction.
[7]Twenty protected plus 100 unprotected speech bits per slot.

bits plus flags and error protection each, build a superframe lasting 4 seconds. For error correction convolutional coding together with interleaving and scrambling has been implemented. Protected data transmission is performed with 92 or 66 data bits per 20-ms time slot giving an upper bound of 4.6 or 3.3 kbit/s, respectively, and uses BCH codes. Packet data transmission runs at 3.2 kbit/s (64 data bits/slot). Error protection coding is modified with the frequency band: At UHF additional differential precoding is applied. Efficient compression techniques allow the transfer of images, maps, slow-moving pictures, and the like even at the relatively low transmission rate. Security mechanisms ensure privacy.

TETRAPOL is a trunking system that enables the user to build large cellular-like networks with gateways to the PSTN and other terrestrial or mobile radio networks as Figure 8.1 shows. Automatic registration in HLRs and VLRs is a prerequisite for easy roaming in large multisite systems. However, handover was introduced as late as 2001 and is available only for group calls. The reason is that because of the wide-area cell coverage, the fact that calls on average are short, and the use of dispatcher-controlled groups, handover is not urgently needed for many applications.

TETRAPOL is a very flexible system offering a variety of services that meet most of the PMR user requirements. A common digital radio platform allows the



DC, DP	Dispatch Center, Dispatch Positions
EDT, UDT	External Data Terminal, User Data Terminal
LABS, LCT	Line Access Base Station, Line Connected Terminal
MS, MT, RT	Mobile Station, Mobile Terminal, Radio Terminal
RBS, RP	Radio Base Station, (independent) Repeater
RSW	Radio Switch
SADP	Standalone Dispatch Position
SIM	Subscriber Identity Module
KMC	Key Management Center
NMC	Network Management Center
OMC	Operation and Maintenance Center
SwMI	Switching and Management Infrastructure
ISDN	Integrated Services Digital Network
MTA	Message Transfer Agent (X.400)
PDN	Public Data Network
PSTN	Public Switched Telephone Network
TCP/IP	Transmission Control Protocol/Internet Protocol

Figure 8.1 Block diagram of a TETRAPOL system.

smooth integration of voice and data transmission based on a layered protocol architecture with three basic transmission modes: *network mode*, *direct mode*, and *repeater mode*. TETRAPOL is based on three main types of logical channels: *control channels*, *traffic channels*, and *data channels*. The *voice teleservices* called core services in TETRAPOL, which are provided in network mode, are *individual call*, *group call* (where one single user may be member of several groups), *multiparty call*, *multisite open channel*, *multisite trunked open channel*, *emergency call*, *emergency open channel*, and *broadcast open channel call*.

In network mode TETRAPOL offers three different types of data bearer services called *support services*, namely, *circuit data service*, *connected packet data service* (X.25), and *connectionless packet data service*. A variety of teleservices for data are offered: *broadcast without acknowledgment*, *short data messaging*, *status transmission* (also used for AVL with GPS), and *Internet access* (TCP/IP). Initially short preset messages carried 127 characters but an enhancement to 255 characters is envisaged [10]. The most important data applications are *external application messaging, fast local messaging, interpersonal messaging* (X.400), and *paging*. TETRAPOL offers about 30 supplementary services and many of them are similar to those in TETRA even if they have different names:

- *Access priority:* This means priority access on the uplink.
- *Adaptive area selection:* Definition of a particular mobile's service area.
- *Ambiance listening:* Also called *remote monitoring* meaning that a MS can be remotely activated to listen in discreetly, for example, in emergency situations. This is only possible if the mobile is not engaged in another call.
- *Area selection:* This is a service area definition on a call-by-call basis.
- *Automatic call-back:* A call that was blocked due to a busy system will be automatically processed when a free channel becomes available.
- *Call completion to busy subscriber:* Automatic redialing to a called busy party after it has become free.
- *Call barring:* Different types of calls can be barred by this service.
- *Dispatcher-authorized call:* A caller needs authorization by the dispatcher for making a call to another party.
- *Call forwarding:* A call can be diverted to another subscriber in the absence of the called party, on no reply, or if the called party is busy or unconditionally.
- *Calling/called party identification:* At call setup the subscriber IDs can be stored and displayed.
- *Call-me-back:* During an unsuccessful call attempt the caller ID can be stored in the called terminal for later manual calling back.
- *Call transfer:* The called party, which can also be the dispatcher, may transfer a call to another.
- *Call waiting:* An engaged user will be notified of a second incoming call.
- *Dual watch:* During network mode DM can be monitored to receive incoming call requests.
- *Discreet listening:* The dispatcher may listen to a voice call between two other parties.
- *DTMF dialing:* This allows DTMF dialing for fixed network access, for example, to PABX, PSTN, or ISDN.
- *Dynamic group number assignment:* Allows group merging and changing of group size, structure, and calling numbers over the air.
- *Intrusion:* An authorized user may intervene in an ongoing call.
- *Include call:* One or more users can be added to an ongoing group communication.
- *Interconnect access:* Interconnection to other networks, for example, PABX, PSTN, or ISDN.
- *Late entry:* Entrance into an existing group communication at any time.
- *Listening restriction:* Mobiles are prevented from direct communication in a group call except for the dispatcher.
- *List search call:* A predefined list of mobiles can be called one by one until one of the called party replies.

- *Priority call:* Multilevel priority with override of busy channels.
- *Priority scanning:* Every user can be a member of several groups. In the case of competing calls, the one with the highest priority will be accepted.
- *Preemptive priority call:* Immediate channel access by frame stealing, in particular, in emergency situations.
- *Short number addressing:* Faster channel access by abbreviated numbers.
- *Shortened numbering:* Only the last digits have to be entered if the first digits of the called party are identical with those of the caller.
- *Stroke signal:* This signal initiates an alert tone to the whole group.
- *Talking party identification:* All parties engaged in a group call can get an indication of the talking party's ID.

To meet all operational requirements, 15 additional network procedures are available, some of which can only be controlled by dispatchers and network managers: *attach-detach, call duration limitation, call reestablishment, call recording, call retention, dynamic regrouping, group merging, migration, presence check, power saving mode, PTT priority, roaming, terminal location registration, transmitter power control,* and *user profile management.* Note that the services available in repeater and direct mode are similar to those in network mode as long as they are feasible.

In addition to most of the usual PMR features, TETRAPOL offers a large variety of security services. Ten of them are available for the protection of the control channel against interception, jamming, and misuse, and 12 others ensure proper system function. Moreover, different and easy-to-use security services are available to protect communication. These include end-to-end encryption and encryption over the air interface, which is of vital importance for many organizations devoted to public safety. End-to-end encryption guards against attack at the air interface or repeaters and also avoids difficult encryption procedures at the repeaters. The distribution of security keys via the air interface allows fast and frequent encryption key changing and avoids the usual inconvenient manual key handling, which is also a weak point at which system security may be attacked. Additionally, to avoid fraud and misuse of stolen equipment, a sophisticated mutual authentication procedure is available. Mobiles may be disabled remotely if necessary. System control and administration can be performed with Windows-based PCs, at least for small systems.

TETRAPOL makes wide use of existing standards; for example, QSIG for the *intersystem interfaces* (ISIs) and CMIP and SNMP for network management. Therefore, it demonstrates high flexibility for integration into large communication systems supported by a large number of reference points as shown in Figure 8.1 and Table 8.5.

Physically some of the reference points shown in Table 8.5 are identical and are standardized interfaces. Most of the reference points denote interfaces in the *base network* (BN), which is the smallest complete basic system able to operate in normal network-connected mode providing all services and features. It comprises one *radio switch network* (RSWN) consisting of several radio switches in which the interface between them, called *inter-radio switch interface* (IRI), is proprietary

Table 8.5 TETRAPOL Interfaces

Reference Point	*Connected System Units*	*Purpose*
R 1	RT-UDT	Radio Terminal[1]–User Data Terminal
R 2	LCT-UDT	Line Connected Terminal–User Data Terminal
R 3	RT-RBS	Radio Terminal–Radio Base Station
R 4	LABS-LCT	Line Access Base Station[2]–Line Connected Terminal
R 5	BN-NMC[3]	Base Network–Network Management Center
R 6	BN-DC	Base Network–Dispatch Center
R 7	BN-PABX	Base Network–Private Automatic Branch Exchange
R 8	BN-X.400 MTA[3]	Base Network–X.400 Message Transfer Agent
R 9	BN-BN	Base Network–Base Network
R 10	BN-EDT[3]	Base Network–External Data Terminal
R 11	BN-PMR	Interface to other PMR systems or GSM
R 12	RBS-RSWN	Radio Base Station–Radio Switch Network
R 13	BN-PSTN	Base Network–Public Switched Telephone Network
R 14	BN-ISDN	Base Network–Integrated Services Digital Network
R 15	BN-TCP/IP	Base Network–Internet Access
R 16	BN-PDN	Base Network–Public Data Network
R 17	BN-SADP	Base Network–Stand Alone Dispatch Position
R 18	RT-SIM	Radio Terminal–Subscriber Identity Module
R 19	BN-KMC	Base Network–Key Management Center

[1]A MS comprises RT and UDT or *terminal equipment* (TE). The MS is also called the *mobile termination unit* (MTU).
[2]Also *line connection interface unit* (LCIU).
[3]Via *Private Data Network* (X.25) or *Public Data Network* (PDN).

and one or more RBSs. The *switch at management infrastructure* (SwMI) comprises one or more BN which also includes the *operation and maintenance center* (OMC). Part of the RSWN is the *radio switch* (RSW), which may have one or more hierarchical levels as a manufacturer's option.

The structure of a TETRAPOL network can be flexibly matched to the user's need and the geographic characteristics of a planned service. It may have a star, ring, or mesh configuration or any combination of them, and separate networks can be merged later. TETRAPOL networks also have ISIs to other PMR networks and to nonpublic and public mobile and fixed networks such as GSM, PSTN, ISDN, and PABX. Of course, there is also a specific interface to other TETRAPOL networks.

ETSI refused to accept the TETRAPOL documentation for standardization as a *publicly available specification* (PAS) for the reason that it would not standardize two different systems for the same purpose and ETSI had already standardized TETRA [19, 20]. Hence, the TETRAPOL air interface is proprietary and very distinct from that of TETRA. Therefore, there is no direct interoperability at the air interface between the two and interoperation is currently possible only via wireline interfaces. The basis for future interworking could be a transponder, dual-mode equipment, or 30-km mutual coverage overlap at country borders. Direct interoperability over the air could only be achieved with dual-standard terminals, and their development is currently under discussion.

The question of interoperability is of paramount importance in Europe because transborder public safety operation between the various countries is highly desirable. However, despite the Schengen Treaty, the NATO requirements, and the CEPT recommendations, some countries have opted for TETRAPOL and others for TETRA, and so the chance to get a single system for public safety all over Europe between 380 and 400 MHz has been squandered. Eventually the TETRA MoU Association and the TETRAPOL User Group decided to cooperate to develop a common interface to make transborder services for the police easier [21].

The TETRAPOL name is the result of a clever marketing move. One of the brand names of installed TETRAPOL systems is ACROPOL, in which the last three characters point to public safety and, in particular, the police. On the other hand, the name TETRA was never protected as an ETSI brand name. Therefore, the name TETRAPOL implies that this system is useful for purposes similar to those of TETRA and that it is particularly well suited to police requirements.

For next-generation TETRAPOL, the implementation of a codec with a lower bit rate has already been envisaged. At first it was felt that the gross bit rate should be increased to 9.6 kbit/s but now doubling it to 16 kbit/s and also improving the spectrum efficiency are within the scope of possibilities (P. Mège, personal communication, 2002). This will enable TETRAPOL networks to combine contradicting requirements such as wide-area coverage, high traffic density, and increased transmission rates in one single network. It would also allow the implementation of PAMR and PMR networks based on an identical technology for divergent applications with very different traffic density and service area sizes in rural and urban areas [10]. For narrowband applications, a variation of TETRAPOL with a channel separation of 6.25 kHz is under consideration [14]. TCP/IP access and interworking with 3G systems are already available and are also necessary features for future TETRAPOL applications.

With the TETRAPOL Forum the TETRAPOL community has formed its own very active platform for the exchange of information, to support TETRAPOL users and to promote TETRAPOL all over the world, obviously with nonnegligible success against strong competition from TETRA and iDEN.

8.6 Narrowband Systems and RVE

The main intention for the use of narrowband FDMA systems is to increase the number of available frequencies; therefore, specific bands may be allocated outside the usual PMR allocations which can them employ a channel separation below 10 kHz, in particular, 5.0 or 6.25 kHz. However, there can be other reasons for the introduction of such systems as the following example shows. In the middle of the 1990s, the Hungarian frequency managers did not have full information about frequency assignments because in the past this topic had been classified as top secret and controlled by the armed forces. Therefore, narrowband channels were inserted between existing 25-kHz channels carrying unknown services to overcome this problem [22].

Narrowband SSB would be useful only for slow data transmission and speech transmission with very low bit rates. For some years codecs with bit rates as low as 2 kbit/s have been feasible and therefore such a system with a modulation bit rate around 4–5 kbit/s to include error correction could be made. However, for all kinds of transmissions, the initial frequency error could be no more than 100–500 Hz. This would require expensive reference crystals for channel frequency generation in the higher frequency bands and the Doppler shift at high mobile speed would also be critical. Note that in some studies channel separations as low as 2.5 kHz have been investigated. However, for analog transmission the baseband would be even narrower than in a 12.5-kHz PM system and the resultant speech quality would be too poor.

At the moment, the only implementation of the more general narrowband coexistence standard EN 301 166 already mentioned in Chapter 7 which has wider interest is a SSB system developed by Securicor Wireless (formerly known as Securicor and later Linear Modulation Technology). It is based on *reference vector equalization* (RVE), originally called *transparent tone in band* (TTIB). In the United Kingdom several channels with 5-kHz separation have been allocated in Band III around 225 MHz for this system.

The Securicor system uses a transparent analog SSB channel well suited to channel separations of 5 and 6.25 kHz. All modes of operation are feasible: Simplex, semiduplex, duplex, and repeaters can also be used. A pilot tone in the middle of the passband is used to control distortion caused by multipath propagation and interference. The proper reconstruction of the pilot signal in the receiver removes most of the channel impairments because, with extremely few exceptions, only flat fading is experienced due to the very small bandwidth. However, this method fails if a single frequency interfering tone occurs within the modulation bandwidth. To prepare the channel for insertion of the pilot signal, the passband is split and both parts are moved apart by a small frequency difference. After reception the signal is reconstructed by removal of the pilot and moving of both parts of the passband back to the original position, which allows the message content to be extracted. Figure 8.2 shows the principle of RVE operation.

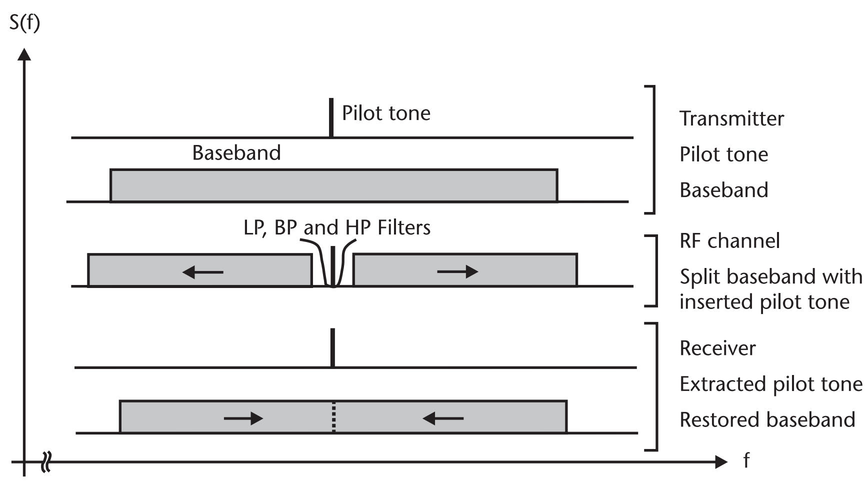


Figure 8.2 The RVE transmission method.

Because RVE primarily provides "only" a transparent analog channel different types of user- and application-specific speech coding and signaling and various data transmission formats can be applied. The transparent channel can be used, together with a compression scheme, to provide analog voice transmission even more robust than in conventional 12.5-kHz analog channels. It can also be used to insert digital information via a radio modem to provide gross bit rates up to 14.4 kbit/s or even higher with mQAM modems. Depending on the type of code, the code rate, and the user bit rate, the residual bit error rate may vary widely and appropriate error protection should be provided. The main technical characteristics of the RVE system should be considered together with Table 7.2 and the narrowband PMR standard EN 301 166 whose limits are fully met.

We should mention that PTTs and manufacturers, in countries other than the United Kingdom and United States, do not seem to be very interested in the RVE system and therefore it is difficult to predict its further market evolution. However, Motorola in the United States also offers SSB systems, and the FCC policy for improved spectrum utilization may be a strong promoter for enlarged future narrowband applications [23]. For this purpose RVE is an interesting candidate technology for cases not well served by TDMA systems.

8.7 PMR Systems for Data Transmission: ARDIS, MOBITEX, and MODACOM

The systems to be discussed now are solely intended for data transmission. A decade or so ago many PMR experts believed that dedicated mobile data transmission systems would be needed for many applications and that the use of speech transmission would decrease rapidly and be replaced by data transmission. However, pure data transmission systems have developed considerably worse than DPMR capable of carrying both voice and data. In the long term it is expected that packet data will dominate, in particular if VoIP asserts itself against current digital voice transmission principles [24].

The first separate wireless data networks were launched in the early 1980s. One of these early data networks was developed in 1983 by Motorola in cooperation with IBM for IBM's service engineers, and in 1990 the network was commercialized as the *Advanced Radio Data Information Service* (ARDIS). In 1998 ARDIS was acquired by the *American Mobile Satellite Corporation* (AMSC), which is the current operator. ARDIS is a proprietary system widely used in the United States where 90% of the population is covered and up to 10 frequencies in the 800-MHz band are available specifically for ARDIS in every city [1, 25, 26].

ARDIS is an FDMA system employing 4FSK modulation with SRRC premodulation filtering at $r_O = 0.2$. The initial gross bit rate of 4.8 kbit/s (GFSK with the proprietary MDC-4800 protocol) was later increased to 9.6 kbit/s in 12.5-kHz channels and 19.2 kbit/s in 25-kHz channels in the 800-MHz band. The static receiver sensitivity is better than –110 dBm and the dynamic cochannel rejection is about –20 dB. A BS transmit power of 46 dBm allows a cell radius of around 20 km and the mobiles transmit at 36 dBm. Channel access is performed with slotted *digital sense multiple access* (DSMA). The transmission is controlled by *radio data-link access procedure* (RD-LAP) employing block coding, interleaving, and ARQ. Besides CRC for error detection, several strategies are pursued for error correction. Random error protection is achieved by convolutional coding with $r_C = 0.5$ and $k = 7$ or trellis coded modulation with a coding rate of $r_C = 0.75$. For packets at 4.8 kbit/s, 16-bit interleaving protects against burst errors up to 3.3 ms, and at 19.2 kbit/s 32-bit interleaving protects against 1.7-ms fade duration. The packet length is 256 bytes; therefore, long messages are subdivided into 240 character packets and reassembled at the recipient. ARDIS is hierarchically structured: The mobiles are on the lowest level, followed by BS, *radio-frequency network control processors* (RF/NCP), and four *network control centers* (NCC) at the highest level. The backbone of the network uses leased telephone lines and packet data transmission (for example, with X.25). ARDIS employs a cellular design ($N_C = 7$) with overlap to improve coverage and transmission quality.

The second U.S. system is based on MOBITEX, which has a long history. Ericsson developed it the early 1980s in collaboration with Televerket, the Swedish PTT, as a portable alarm system for isolated workers. In 1986 the world's first public MOBITEX network was opened by Televerket and found numerous applications in Scandinavia. This system operated at 1.2 kbit/s with BCH (15,10) coding and ARQ. Later Ericsson developed MOBITEX II, also known as MOBITEX 8K. In 1989 Cantel, a Canadian operator, started the first MOBITEX network in North America, which was a major milestone. RAM Mobile Data, now Transcom UK, was founded in 1990 (around 1992–1993 Bell South bought in but left later) and established MOBITEX networks in the United States and United Kingdom. It is now in widespread use in the United States and operated by Bell South Wireless Data. In the United States the BS transmit band is 935–940 MHz, and 896–901 MHz is used for the mobiles. In the meantime MOBITEX systems are to be found all over the world [1, 24, 27–30].

MOBITEX II is an FDMA system available in the 450- and 900-MHz bands that runs with a gross bit rate of 8 kbit/s on 12.5-kHz channels in duplex or semiduplex mode employing GMSK modulation with a BT = 0.3. Additional filtering external to the modulator ensures compliance with adjacent channel power limits. The dynamic receiver sensitivity is –104 dBm and the dynamic cochannel rejection about –20 dB. The transmitter power levels are up to 46 dBm for the base station, 40 dBm for mobiles, and 33 dBm for hand portables. Dynamic power control covers the range from 20 dBm up to the maximum transmitting power of the equipment. The protected data rate is 4.6 kbit/s based on packet data transmission with a maximum size of 512 bytes. Longer messages have to be subdivided and reassembled at the receiving party. Channel access is performed with a modified *carrier sense multiple access* (CSMA) procedure. For large service areas cellular frequency reuse is feasible. MOBITEX also provides *automatic vehicle location* (AVL) and has been upgraded and now also offers IP connectivity.

Each message starts with a frame header consisting of 56 bits, bit and frame synchronization (16 bits each), BS and country ID (12 + 4 bits), and CRC (8 bits) followed by data blocks comprising 160 bits, including 16 bits for CRC, which are coded with a shortened (12,8) Hamming code giving a total block length of 240 bits. The first block carries additional address and control information. The code can correct at least one error per block and additional interleaving with a depth of 20 bits is used. Hence, burst errors up to a length of 20 bits equivalent to fades up to 2.5 ms can be corrected. FEC and ARQ ensure a given BER in user data. MOBITEX networks provide a store-and-forward capability for undeliverable packets until the recipient MS reregisters.

MOBITEX employs a proprietary protocol called MPAK (MOBITEX Packet) that consists of several layers as shown in Figure 8.3. Its first layers are specifically suited to the mobile environment, and the upper layers provide all other necessary means to complete the MOBITEX communication protocol. The different parts of MPAK are as follows:

- MCP/1: MOBITEX Compression Protocol for optional data compression;
- MTP/1: MOBITEX Transport Protocol for packet transfer;
- MASC: MOBITEX Asynchronous Protocol, provides machine interface and allows packets to be transferred over the data link layer where the next packet will only be sent after acknowledgment of the preceding;
- ROSI: Radio Open System Interface protocol, which breaks layer 3 into a series of 144 bits blocks in an M-frame format. Each block is appended with a 16 bits CRC to protect against errors not corrected by the physical FEC. Lost blocks are fetched by selective ARQ.

MOBITEX is based on an open architecture with a hierarchically structured infrastructure comprising base stations, local switches, and regional switches. All cells of one local switch form a service area or subnet and in each of them 10–30 frequency pairs or channels are available. Usually each BS carries one to four channels. HDLC or X.25 links are used from the BSs to the local switches and further to the regional switches. The top of the hierarchy is the main exchange, which interconnects the MOBITEX network with other networks.

The *network control center* (NCC) supports network-wide management and supervisory functions. Message switching always occurs at the lowest level, to ensure quick response times and reduced backbone traffic; for example, two MSs in the same cell involve only the cell's BS. If the link between a BS and the

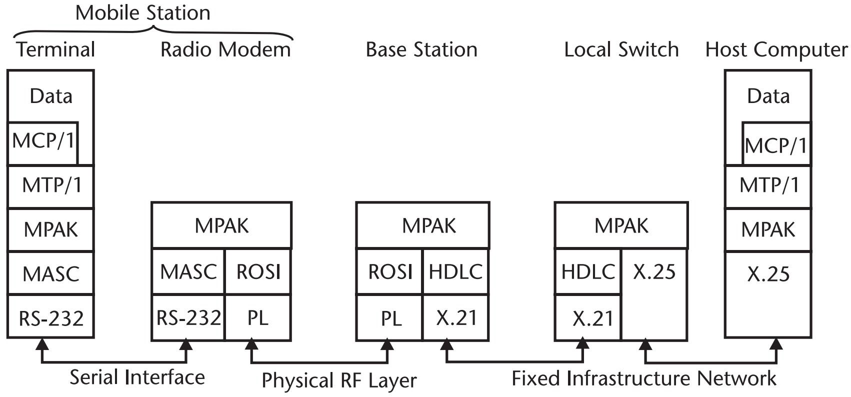


Figure 8.3 The MOBITEX protocol stack.

local switch fails, the BS can operate autonomously for intracell communication. Roaming and authentication are also available and *electronic serial numbers* (ESN) are used for authentication.

In principle, in MOBITEX II networks speech transmission is also possible. However, the transmission protocol is not well suited to speech transmission and excessive delays of up to 5 seconds may occur. It has turned out that there is no simple way to reduce speech delay in heavily loaded networks because not only would the air interface have to be modified, but also the routing of packets through the network to ensure the correct order of their arrival. Nevertheless, MOBITEX has won the biggest market share by far of dedicated mobile data systems and it has spread to many countries, including Belgium, Chile, France, Germany, Korea, Poland, Singapore, the Netherlands, Scandinavia, and the United Kingdom.

The *Motorola Mobile Data Communications System* (MODACOM) is based on the RD-LAP developed jointly by Motorola and IBM. Originally it was launched as ARDIS in the United States at only 4.8 kbit/s. Later 9.6 kbit/s in 12.5-kHz channels and 19.2 kbit/s in 25-kHz channels were achieved. For the German market, where the system is run by T-Mobile, a carrier separation of 12.5 kHz in the 410- to 430-MHz band is used. This system has also been implemented in Switzerland. Compared to MOBITEX II it exhibits a somewhat higher gross bit rate of 9.6 bit/s due to its 4FSK modulation. The effective data rate is about 6.33 kbit/s. The BS transmitting power level is 38–44 dBm while the mobiles operate at 38 dBm. Premodulation filtering is performed with a SRRC filter and a roll-off factor of $r_O = 0.2$. The static receiver sensitivity is –110 dBm or better for BER = 1% without FEC and the dynamic cochannel rejection is –20 dB or better. MODACOM uses trellis coding with additional CRC and ARQ delivering a typical BER $\leq 10^{-8}$. Channel access is performed using the slotted *digital sense multiple access* (DSMA) technique, which checks the channel occupancy before attempting unsolicited access in defined time slots [27, 31].

The transmission scheme is based on packet data with a maximum packet size of 4,096 bits (512 bytes) and acknowledgment. Status transmission as well as file transfer are available. Moreover, group calls, CUGs, roaming, and a mailbox service are provided. The system also offers security features such as encryption, authentication, and passwords. A number of BSs is linked to an *area communication controller* (ACC), which comprises a *radio network controller* (RNC) and a *radio network gateway* (RNG). The latter is connected to the fixed network via X.25 links. The ACC has access to HLRs and VLRs for mobility management. The whole network is controlled by a *network management controller* (NMC).

There are some other mobile data transmission systems of minor importance such as PAKNET, a joint venture of Racal, Chubb, and Mercury. PAKNET operates at around 160 MHz on 12.5-kHz channels. It uses GFSK at 8 kbit/s and transmits packets that are 1,024 bits in length (128 bytes). Error protection is achieved by use of CRC and a Golay code able to correct three errors per block of 24 bits. Another is the COGNITO system operating in the United Kingdom in Band III also on 12.5-kHz channels with 3FSK modulation and a 6.144-kbit/s gross bit rate. It employs CRC and an interleaved BCH (63,51) code for error correction. ARQ is used to retransmit uncorrectable packets. The mobile transmitting power is 37 dBm [26].

In the United States, *radio data networks* (RDN) are operating in the 800- and 900-MHz bands at a 4.8-kbit/s gross bit rate resulting in protected data transfer at 2 kbit/s. Currently the trend for trunked data transmission systems in the United States is to go from 4.8 to 19.2 kbit/s. Hence, for the future a user bit rate of 10.2 kbit/s is envisaged. RDN interfaces are proprietary but a number of widespread network protocols are supported to interconnect network control centers with host computers [30].

Regarding data transmission in dedicated networks, note that the U.S. cellular industry introduced *cellular digital packet data* (CDPD) at 19.2 kbit/s with DES encryption for 1G and 2G U.S. cellular networks. This used a separate overlay network and utilized free traffic channels when they were available. GSM started its *circuit switched data* (CSD) service with a user bit rate of 9.6 kbit/s at the beginning of the 1990s. With a modified coding scheme 14.4 kbit/s has been achieved per time slot and so the recently introduced HSCSD can provide bit rates up to a maximum of 57.6 kbit/s by aggregating two to four time slots. However, this data rate is rarely achieved on normally loaded networks. Much faster is the GPRS, in theory, with bit rates up to 171.2 kbit/s, which in practice operates also with significantly reduced speed. With an improved modulation called EDGE modulation rates up to around 384 kbit/s are expected [32].

We can state that the market for mobile data transmission systems will begin to decline since most modern applications need both voice and data transmission. TETRA and most proprietary digital PMR systems are providing both. Additionally, a certain percentage of data users will make the transition to the (under favorable propagation conditions) much faster 3G mobile networks.

8.8 Other Digital PMR Systems

As already mentioned, the worldwide spread of MPT 1327 means that it should also be seen as a digital system despite its analog speech transmission. However, its low gross bit rate of only 1.2 kbit/s restricts reasonable data transmission to short messages.

Much more interesting is the digital FDMA system SR 440 developed by Ascom and later also offered by Bosch under the brand name DISCO. It shows that a different trade-off of radio parameters allows an improvement of the RF properties if some transmission speed is sacrificed. Hence, this system shows excellent sensitivity and cochannel rejection. Due to the slow binary modulation, Doppler spread and delay spread impose no problems in the mobile radio environment [33].

Moreover, the good LO noise and IF filter properties permit an extraordinarily good adjacent channel interference rejection. [Radio experts know that adjacent channel selectivity, receiver intermodulation, and adjacent channel power depend not only on the filters but also on the noise characteristics of the receiver and transmitter. Hence, in modern PMR low-noise *voltage-controlled oscillators* (VCOs) with good tunability are required to produce cheap but good PLL synthesizers. This is as easy as squaring a circle!] SR 440 will not be reviewed in depth because only a few networks are in the field and no new ones will be produced because the PMR division of Ascom was bought by Bosch and a short time later the whole of Bosch's PMR business was sold to Motorola. The main properties of SR 440 are listed in Table 8.6.

In the marketplace various additional brand names appear. However, in most cases these are derivatives or renamed versions of the major systems already discussed.

8.9 Comparison of Selected Nonstandardized Digital PMR Systems

Again it is of interest to compare the different proprietary digital PMR systems but it may also be useful to make comparisons with a standardized system such as TETRA. In Table 8.7 the properties of the most important proprietary digital PMR systems are listed for easy comparison. Due to the relatively low importance of the other digital PMR systems their characteristics have not been incorporated.

8.10 Perspectives of Proprietary Digital PMR Systems

Predictions of market shares for proprietary digital PMR systems expect some growth for TETRAPOL and much better prospects for iDEN. Some analysts believe that iDEN will be the strongest competitor to TETRA, or the other way round.

In the United States the FCC frequency management policy will have a strong impact on future market development. Due to the rarity of frequencies suited to mobile services, the FCC has decided that from 2007 onward all new systems going into service must use 6.25-kHz equipment. The real goal, however, is to ensure that one voice channel per 6.25-kHz channel is provided [23]. That can be achieved, for example, with 6.25-kHz FDMA narrowband systems as being initially offered

Table 8.6 Main Properties of SR 440

Property	*Remark*
Frequency bands	80, 160, and 450 MHz
Channel separation	12.5 and 25 kHz
Channel access	FDMA
Duplex separation	1–20 MHz
Mode of operation	Simplex, semiduplex, duplex
Transmitter power	BS: 33–44 dBm, MS: 33–44 dBm, HP: 20–37 dBm
Modulation	CP-BFSK
Modulation rate	4.8 kbit/s
User bit rate	4.7 kbit/s unprotected, 2.7 kbit/s with strong protection
Time slot duration	20 ms, frame length 100 ms
Speech coding	IMBE at 4.0 kbit/s
Data transmission	Messages and packet data
Rx sensitivity	≤–118/120 dBm static and ≤–108/110 dB with Rayleigh fading for 12.5/25 kHz
Cochannel rejection	≥–5/3 dB static and ≥–15/13 dB with Rayleigh fading for 12.5/25 kHz
Adjacent channel rejection	≥65/75 dB static and ≥55/65 dB with Rayleigh fading for 12.5/25 kHz
Delay spread	40 μs with reduced data rate
Coexistence	EN 300 113 is fulfilled

Table 8.7 Main Properties of the Most Important Nonstandardized Digital PMR Systems

Property	*ASTRO*	*EDACS*	*iDEN*	*TETRAPOL*	*TETRA V+D*
Current allocated frequency bands[1]	136–174, 380–520, 700/800/900 MHz	160/450/70, 806–870, 896–941 MHz	806–870, 890–960, 1,500 MHz	68–88, 380–512, 830–930 MHz	380–520, 806–825, 851–870, 870–888, 915–933 MHz
Duplex separation[2]	10 and 45 MHz	4–45 MHz	45 MHz	10 MHz	10 and 45 MHz
Channel separation	12.5, 20, 25, and 30 kHz	12.5 and 25 kHz	25 kHz	10 and 12.5 kHz	25 kHz
Channel access	FDMA	FDMA	6TDMA	FDMA	4TDMA
Mode of operation[3]	S, SD, D[4], voice & data	S, SD, D[4], voice & data	S[4], SD, D, voice & data	S, SD[4], D, voice & data	S, SD, D, voice & data
Tx power					
BS	40–50 dBm	56 dBm	51 dBm	≤44 dBm	28–46 dBm
MS	40–50 dBm	40–50 dBm	27–40 dBm	≤40 dBm	24–45 dBm
HP	35–37 dBm	38 dBm	22–35 dBm	≤33 dBm	15–35 dBm
Modulation	C4FM (QPSK-C)	GFSK	16QAM, TCM	GMSK, BT = 0.25	π/4-DQPSK, $r_0 = 0.35$
Gross bit rate	9.6 kbit/s	9.6 kbit/s	64 kbit/s	8.0 kbit/s	36 kbit/s
Maximum user bit rate[5]	7.2 kbit/s[6]	9.1 kbit/s[6], 7.77 kbit/s[7]	6 × 9.6 kbit/s[6], 6 × 7.2 kbit/s[7]	7.4 kbit/s[6], 4.6 kbit/s[7], 3.4 kbit/s[8]	4 × 7.2 kbit/s[6], 4 × 4.8 kbit/s[7], 4 × 2.4 kbit/s[8]
Speech coding	VSELP 4.8 kbit/s	AME or IMBE ≈ 4.4 kbit/s[4]	VSELP 4.8 kbit/s	RPCELP 6.0 kbit/s	ACELP 4.6 kbit/s
Rx sensitivity	–117/–115 dBm static[9] for MS/HP	–115/116 dBm static[9] for 12.5/25 kHz	–105.5 dBm dynamic[10]	–113/–111 dBm dynamic[10] for BS/MS	–106/–105/–103 dBm dynamic[10] for BS/MS/HP
Cochannel rejection	–16.5 dB dynamic[10]	–16/–14 dB dynamic[10] for 12.5/25 kHz	–19.5 dB dynamic[9]	–15 dB dynamic[10]	–19 dB dynamic[10]
Delay spread	—[11]	≤52 μs	≤40/66 μs[12]	—[11]	≤100 μs

[1]New bands may be added according to international demand.
[2]According to national and international regulations.
[3]Simplex, semiduplex, duplex.
[4]No precise information available.
[5]Aggregation of time slots possible.
[6]Transparent.
[7]With weak FEC.
[8]With strong FEC.
[9]Dynamic values; see Chapter 9.
[10]Rayleigh fading only.
[11]Not specified.
[12]Equalizer dependent.

by APCO 25 phase II and ASTRO narrowband equipment, but also with TDMA technology as being offered by TETRA or iDEN.

In Europe it seems that in the long term, proprietary digital PMR systems will struggle against standardized systems such as TETRA and APCO 25. It is too early to estimate the acceptance of DIIS. Anyway the hottest competitors to all digital PMR systems are the second and increasingly the third generation public cellular systems.

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CHAPTER 9

Specific PMR Topics

This chapter is devoted to a number of issues that are of specific interest for PMR, including methods of measurement, comparison of analog and digital system characteristics, evaluation of the differences between FDMA and TDMA, radio interference, migration from old to new systems, and other specific topics that do not fit well into the previous chapters.

9.1 Methods of Measurement for PMR Equipment

For a long time a proper comparison of radio equipment characteristics from different manufacturers has been close to a secret science, in particular, in those cases where equipment has come from different manufacturers located in different parts of the world. This had a simple reason: Different methods of measurement had been used, the measuring criteria were different, and last but not least the limits set by national authorities were also different. In the past these have been big obstacles for free competition in many markets and even now not all of the hurdles have been removed.

Since the end of the 1960s the IEC has tried to collect and standardize the methods of measurement for PMR equipment in IEC Publication 60489 (formerly Publication 489). Simultaneously the CEPT tried to do the same with their Recommendation T/R 24-01 [1, 2]. However, their methods were never aligned with those of the IEC. In the second half of the 1980s, the work originally started by the CEPT was transferred to ETSI and shortly afterward the PMR coexistence standards ETS 300 086, ETS 300 113, I-ETS 300 219, and the report on methods of measurement, ETR 027, were published. The other PMR coexistence standards ETS 300 296, 341, and 390 and another report, ETR 028, followed. Since then some of them have been revised and updated and others are being revised. Recently most of them have been upgraded to European Norms (EN) [3–14].

The same radio characteristic can be measured in different ways. As an example, with analog transmission intermodulation distortion can be measured by a two-generator method or alternatively a three-generator method can be used. For analog measurements different criteria are in use, for example, 20- or 12-dB SINAD. Finally, various administrations have set different limits in their countries. These are the reasons why it is so difficult to compare just one single radio property if the methods of measurement, the measuring criteria, and the limits are different.

The same problems apply to digital radio equipment. But there is an additional trap: The methods of measurement for analog equipment are mostly based on static methods (AWGN channel), whereas the methods of measurement for modern digital equipment are mostly based on dynamic measurements. Usually Rayleigh fading is assumed but in many cases different propagation conditions are taken as a reference, for example, TU 50 or HT 200. The result is that the receiver sensitivity of a multichannel set is represented by a multidimensional table where one coordinate is the channel frequency and another is the propagation profile. Moreover, the different types of signaling and traffic channels may exhibit different sensitivities. Unfortunately, the measured properties may additionally vary with ambient temperature and battery voltage.

It would be an easy exercise to fill a separate book with descriptions of these methods of measurement and to comment on their strengths and weaknesses, but that is not the goal being pursued here. In this book we merely intend to warn the reader that there might be traps if different equipment from different manufacturers and different parts of the world or if analog and digital equipment are to be compared.

9.2 Properties of Digital PMR Systems in Comparison to Analog PMR

Prior to comparing static and dynamic parameters, it is necessary to evaluate their relationship. For analog PMR systems the receiver parameters have always been evaluated by static methods of measurement, assuming an AWGN channel. This is not true just for conventional PMR systems tested in accordance with EN 300 086, but also for digital systems that meet EN 300 113. Even some standards for systems employing digital transmission such as EN 300 168 (DSRR) and EN 301 166 (narrowband PMR) assume measurements under static conditions [5, 15]. All modern system standards such as GSM, TETRA, and DIIS, however, are required to be evaluated under dynamic conditions, and often Rayleigh fading but sometimes more specific propagation models, but never shadowing, is taken into account. For some parameters static values are also often given [16].

For transmitters there is no significant difference between analog and digital because there is no feedback from the channel properties as long as specific functions such as power control and frequency synchronization are disregarded. The only point is that the old PMR systems are based on constant envelope modulation, while new ones such as narrowband PMR do not necessarily make that assumption or are based on linear modulation with nonconstant envelope like TETRA.

In Chapter 3, we showed that for Rayleigh and Ricean fading an average standard deviation of $\sigma_F = 6$ dB or somewhat less can be assumed while for the combination of fading *and* shadowing $\sigma_P \approx 9$ dB is appropriate. It has also been shown that the resulting standard deviation σ_I can be calculated from the ratio of the power P_I of the interferer to the power of the carrier P_C if two signals are involved that exhibit independent fading characteristics. If the desired signal is much stronger than the interferer, then there is only a small difference between σ_P and σ_I.

The measurement of dynamic receiver properties such as sensitivity and cochannel rejection was considered in Chapters 4 and 5. When conducting fading measurements, the method is similar and other properties such as adjacent channel rejection or spurious response rejection can also be handled in the same way. Care should be taken in cases for which three instead of two generators are involved, as in the case for intermodulation, because in the dynamic case both interferers fade independently. In the formula P_I has to be replaced by the sum of the power for the two interferers:

$$P_I = P_{I1} + P_{I2} \tag{9.1}$$

Usually, as the worst case, we assume that $P_{I1} = P_{I2}$; hence, we have $P_I = P_{I1} + 3$ dB. If merely the receiver properties are to be compared, then only fading should be assumed. On the other hand, if coverage distance calculations or frequency economy considerations are required, then fading *and* shadowing should be taken into account. This means that static analog characteristics have to be converted with the help of σ_P and also faded dynamic values should be replaced by those based on a combination of fading and shadowing.

There are two reasons why the transformation from static to dynamic signals is useful. First, we can use it to make possible a direct comparison of the properties of old and new systems and, second, the coverage distances can be calculated more easily from the dynamic link budgets. To do a very precise conversion is often difficult and in the case of voice quality measurements comprehensive statistical investigations based on a large number of subjective evaluations are necessary. However, in most cases an accuracy of a fraction of a decibel is neither necessary nor reasonable and, therefore, a good estimate suffices.

For analog voice transmission the receiver sensitivity is defined by the level of input power that produces a certain minimum signal quality at the output of the receiver, for example, 20-dB SINAD psophometrically weighted or 12-dB SINAD unweighted or a similar measure. Note that the second method of measurement gives 3 dB better results [17]. In the case of degradation, for example, by a cochannel interferer, a certain impairment of the quality must be defined, for example, a reduction from 20- to 14-dB SINAD. In the fading case it must be ensured that the voice quality is better than this limit with at least a 90–95% probability both in time and location, and this is the case for $k \approx 1.5$. Under this assumption the dynamic $(P_C/P_I)_{AD}$ ratio for analog voice transmission can be estimated based on the known static value of $(P_C/P_I)_{AS}$ by simply adding 9 or 13 dB. (The latter should be 13.5 dB but this would be overly accurate because $\sigma_F = 6$ dB is already a worst case assumption.)

$$\begin{aligned} 10\log_{10}\left(\frac{P_C}{P_I}\right)_{AD} &\geq 10\log_{10}\left(\frac{P_C}{P_I}\right)_{AS} + 1.5\cdot\sigma_{F,P} \\ &\approx 10\log_{10}\left(\frac{P_C}{P_I}\right)_{AS} + \begin{cases} 9 \text{ dB for fading only} \\ 13 \text{ dB for fading and shadowing} \end{cases} \end{aligned} \tag{9.2}$$

This result corresponds well with the value of 18 dB already mentioned in Chapter 3 after (3.89) which is the reuse margin used in U.S. 30-kHz systems which exhibit a static cochannel rejection of −6 to −7 dB. For cases other than cochannel rejection, a similar approach is feasible, for example, for adjacent channel rejection and spurious response. The conversion from a dynamic value based on pure fading to one dependent on fading and shadowing means the replacement of $\sigma_F = 6$ dB by $\sigma_P = 9$ dB or simply adding 4 dB in general situations, according to (9.2). However, if propagation conditions or the applied quality criteria are different, then the given estimates have to be modified accordingly.

For many cases the cochannel rejection is measured as the ratio P_C/P_I at a certain criterion, that is, a particular value of SINAD. However, note that in the ETSI methods of measurements defined in EN 300 086 this is different. The desired signal is set to the limit of sensitivity +6 dB μVEMF (a radio must not be worse than this) and the interferer is increased until the psophometrically weighted SINAD reduces from 20 dB or more to 14 dB. As long as the actual receiver sensitivity is at the limit, the result corresponds with the direct ratio at this given criterion. If receiver has 10 dB better sensitivity than the limit, it exhibits a SINAD around 30 dB (only achievable with very low AF distortion) and this may allow the interferer to be a bit stronger before the 14-dB reference is met. This means that relative to other recognized measurement methods a somewhat better result will be seen in this case. Once again this example shows that care is needed if different methods of measurement and their results are being compared.

For digital transmission the unprotected channel (more or less equivalent to the transparent channel offered by systems such as TETRA) is usually what is measured. It is widely accepted that a BER of 1% is an appropriate criterion (see Chapter 3). For the maximum sensitivity measurement according to EN 300 113, the desired signal is adjusted to that level where the bit error rate BER = 1% or the message error rate MER = 20% is achieved. (This corresponds to 22-bit messages since $0.99^{22} = 0.8$.) For interference measurements the desired signal is set 3 dB above the sensitivity limit and the interferer(s) is (are) adjusted until the sensitivity criterion of BER or MER is reached [3, 4]. If the ETSI methods are not used for analog sensitivity, often 12-dB SINAD unweighted or BER = 3% or 5% for transparent digital channels is taken as the measurement criterion. Note that this is daily practice in many parts of the world outside Europe [18].

As has already been shown in Chapter 3, to find an estimate for the conversion of static to dynamic cochannel rejection outside a fading null, a BER << 1% is assumed while inside the null it is 50%. Thus, the signal level must be such that the probability in time and location to be in a fading null is 2%. Hence $k \approx 2$ is necessary, and 12 dB has to be added for fading, or in a general propagation situation 18 dB to include fading and shadowing, for the approximation of the dynamic digital $(P_C/P_I)_{\mathrm{DD}}$ ratio from the static ratio $(P_C/P_I)_{\mathrm{DS}}$:

$$
\begin{aligned}
10\log_{10}\left(\frac{P_C}{P_I}\right)_{\mathrm{DD}} &\geq 10\log_{10}\left(\frac{P_C}{P_I}\right)_{\mathrm{DS}} + 2\cdot\sigma_{F,P} \\
&\approx 10\log_{10}\left(\frac{P_C}{P_I}\right)_{\mathrm{DS}} + \begin{cases} 12 \text{ dB for fading only} \\ 18 \text{ dB for fading and shadowing} \end{cases}
\end{aligned}
\tag{9.3}
$$

However, remember from Chapter 3 that in practical systems the difference very often is much smaller due to coding and FEC. There are two reasons for this. First, the E_B/N_0 or the P_C/P_N ratio is often smaller in digital transmission than for analog transmission and, second, the FEC introduces an improvement. Hence, with the exception of unprotected transparent channels $(P_C/P_I)_{DD}$ it is usually much better than shown above, even $(P_C/P_I)_{DD} \approx (P_C/P_I)_{DS}$ is possible if a strong error protection scheme is applied as GSM has demonstrated [19]. Then corrections with $k = 1.5$ similar to the analog case are appropriate [see (9.2)]. If only fading is assumed, then for many cases the difference can be estimated to be 9 dB as it has been found in TETRA where the limits of some parameters are given under dynamic and static conditions [16]. This is also true for other systems because most practical systems employ similar code rates and also show similar FEC properties. If in such cases shadowing has to be taken into account as well, then 13 dB will be a reasonable estimate.

Now we know all we need to compare the main characteristics of analog and digital PMR systems. To avoid overcomplication, only analog PMR is compared with TETRA, which may serve as a good reference model because the differences are in many cases not very large. If needed, the readers can carry out the comparison for other systems themselves according to the methods given.

Usually the properties of radio equipment are given relative to a defined level. That is the case for most of the receiver limits in the ETSI PMR standards. With the exception of cochannel rejection, all of these values have been converted to absolute power levels because this eases the comparison of the limits in different standards and, in particular, it simplifies the calculation of coverage and maximum interference distances. The dynamic receiver properties given in Table 9.1 are based on fading without any shadowing [20].

The sensitivity limits for conventional analog and digital PMR systems are not very stringent and, hence, typical values may be up to 10 dB better. In particular, analog sensitivity can be made much better. For modern digital systems there is only a small margin for improved sensitivity. From the finding that the real receiver sensitivity of TETRA and analog PMR systems is similar, an additional important conclusion can be drawn: Since a TETRA channel is carried by only one of four time slots the *average* receiver power at sensitivity level is 6 dB smaller, which is a significant improvement compared to analog PMR [17].

Evaluating all of the methods of measuring different receiver capabilities to reject the various types of interferers, such as cochannel, adjacent channel, and spurious response signals, it turns out that the differences between the results of static and dynamic measurements are similar to those of static and dynamic sensitivity. Therefore, the same difference can be used as an approximation in all these cases.

Intermodulation as measured according to the three-generator method is a combination of two different effects. The first one is the nonlinearity of the receiver front-end amplifier and/or the mixer and all IF stages prior to the channel filter that produces an undesired signal at a certain level. This intermodulation signal interferes with the desired signal, causing the cochannel rejection of the receiver to come into play. Hence, for two receivers with identical front ends and IF circuitry design, the measured intermodulation rejection is dependent on channel separation simply because desired signals with a higher deviation exhibit better cochannel rejection. This is not the only example of a combination of effects influencing the result of a measurement. Looking more deeply into the intermodulation mechanism

Table 9.1 Comparison of Analog Receiver Properties with Those of TETRA

Characteristic		*PM 12.5*	*PM 20*	*PM 25*	*TETRA V+D*
Sensitivity [dBm]	static	≤–114[1]	≤–117[1,2]	≤–119[1]	≤–112[3]
	dynamic[4]	≤–105	≤–108[2]	≤–110	≤–103[3]
Cochannel rejection [dB]	static	≥–12	≥–8	≥–8[5]	—[6]
	dynamic[4]	≥–21	≥–17	≥–17[5]	≥–19[5]
Adjacent channel rejection [dBm]	static	≥–47	≥–37	≥–37	—[6]
	dynamic[4]	≥–56[7]	≥–46[7]	≥–46[7]	≥–55
Spurious response rejection [dBm]	static	≥–37	≥–37	≥–37	≥–45
	dynamic[4]	≥–46[7]	≥–46[7]	≥–46[7]	≥–54[8]
Intermodulation rejection [dBm]	static	≥–42[5]	≥–42[5]	≥–42[5]	≥–47
	dynamic	≥–51[7,8,9]	≥–51[7,8,9]	≥–51[7,8,9]	≥–56[8]
Blocking [dBm]	static	≥–23[10]	≥–23[10]	≥–23[10]	≥–25[11]
	dynamic[4]	≥–32[7,10]	≥–32[7,10]	≥–32[7,10]	≥–34[8,11]
Maximum receiver input level [dBm]	static	≥–7	≥–7	≥–7	≥–20
	dynamic[4]	≥–16[7]	≥–16[7]	≥–16[7]	≥–29[8]

[1]Typical limit: –107/–110 dBm for EN 300 086/113. Note: –107 dBm is equivalent to 1 μVcc at 50Ω.
[2]FM 20: 5 dB worse.
[3]BS: 3 dB better, MS typically 1–2 dB better.
[4]Rayleigh fading only.
[5]Typically at least 1 dB better.
[6]Not specified.
[7]Approximated with $1.5\sigma_F = 9$ dB.
[8]Approximated with $1.5\sigma_F = 9$ dB due to the specified difference between static and dynamic sensitivity.
[9]BS: 5 dB better.
[10]For frequency differences from 1 to 10 MHz.
[11]For frequency differences >500 kHz.

reveals that there is also a dependency on the *local oscillator* (LO) noise performance. This is why intermodulation is also dependent on the frequency difference between the interferer and the nominal channel frequency even if the front-end filters are flat in the passband.

Sometimes additional unexpected problems arise. Let us look at an example. Third-order receiver intermodulation is produced by two interfering signals where the second harmonic of one mixes with the fundamental of the other to produce an in-band signal. A problem can occur in the region where the front-end filters are flat and the LO noise is minimal even if the necessary frequency relationship is not fulfilled. The interferers may produce an interfering signal at, for example, half the first IF. If its frequency is doubled in the mixer or in the first IF stage before the IF filter, then an interfering cochannel signal is the result. This is one example that shows why receiver design is tricky and why new requirements such as broadband front ends may introduce new problems not present in conventional receivers.

Going through all of the details shows that there are only small differences between different systems if dynamic properties are compared. This is particularly true for sensitivity and cochannel rejection, which respectively determine coverage and cluster size for frequency reuse. Some of the limits for 20- and 25-kHz systems are more stringent than for 12.5-kHz systems. However, in the last two decades much experience has been gained from 12.5-kHz systems that has shown that the limits are sufficient. In particular, adjacent channel power and adjacent channel rejection are more relaxed. This has led to the TETRA limits being relaxed to these values because tighter limits would cause unnecessary expenditure without real benefit. Overall, results have shown that there is no significant difference in performance but considerable cost benefits for practical systems.

To get a more detailed view, the transmitter properties and link budgets of different systems have been compared [20] and the results are given in Table 9.2. The values and results presented in this table need some explanations beyond the footnotes. The transmit power range of conventional equipment varies considerably dependent on the application and national regulations. However, with few exceptions these are in the ranges given in the table. Disregarding the last decibel, transmitter power levels are very similar for different systems serving similar applications. For link budget calculations it is the transmitter burst power and not the average power that should be taken into account for TDMA systems such as TETRA.

Table 9.2 Comparison of Analog Transmitter Properties with Those of TETRA

Characteristic		*PM 12.5*	*PM 20*	*PM 25*	*TETRA V+D*
Transmit power [dBm]	HP	≈30–38[1]	≈30–38[1]	≈30–38[1]	15–35[2,3]
	MS	≈37–40[1]	≈37–40[1]	≈37–40[1]	15–45[2,3]
	BS	≈40–44[1]	≈40–44[1]	≈40–44[1]	28–46[3,4]
Adjacent Tx channel power	BW[5] [kHz]	8.5	14	16	18
	Limit [dBc]	≤–60[6]	≤–70[6]	≤–70[6]	≤60/–70[7]
Spurious Tx emissions	Discrete [dBm]	≤–36	≤–36	≤–36	≤–36
	Broadband [dBc]	—[8]	—[8]	—[8]	≤–75–90/–100[6,9]
Transmitter intermodulation	Rejection [dBc]	–40/–70[10]	–40/–70[10]	–40/–70[10]	–60/–40/–60[6,11]
	Conversion loss[12] [dB]	10/40	10/40	10/40	10/10/—[6,8,11]
Dynamic link budget[13] [dB]	BS ⇒ MS	149	152	154	151[15]
	BS ⇒ HP[14]	146	149	151	147[15]
	MS ⇒ BS	145	148	150	153[15]
	MS ⇒ MS	145	148	150	150[15]
	MS ⇒ HP[14]	142	145	147	146[15]
	HP ⇒ BS[14]	140	143	145	140[15]
	HP ⇒ MS[14]	140	143	145	137[15]
	HP ⇒ HP[14]	137	140	142	143[15]

[1]Typical range for conventional PMR. Note: –30 dBm is equivalent to 1 μW.
[2]Power classes in 5-dB steps.
[3]Average burst peak power.
[4]Power classes in 2-dB steps.
[5]Measuring receiver bandwidth.
[6]No need for <–36 dBm.
[7]For 25-, 50-, and 75-kHz frequency difference.
[8]Not specified, but not critical for constant envelope.
[9]For frequency range of 100 kHz to Rx band edge, limits dependent on MS/BS and Tx power; no need for <–70 dBm.
[10]BS only, stronger limit only under specific licensing conditions or in case of interference.
[11]MS/single BS/intra-BS only; note 6 refers to BS only; note 8 refers to intra-BS only.
[12]For MS/BS, conversion loss is better suited to interference calculations.
[13]Fading only.
[14]For HP 3-dB body loss has been estimated, depends on frequency and how the HP is held.
[15]For TETRA for MS and BS, class 1 transmitters have been assumed, and for HP mobile class 3 transmitters; the sensitivity has been assumed for HP 1 dB better and for MS 2 dB better than the limit, whereas for the BS 3 dB better than the MS limit has been assumed; it has been indicated whether the links are symmetric.

For HPs the transmitting power is simply limited by battery capacity and the maximum operating time. For mobile stations more transmitter output power is feasible, but the car battery does not have unlimited capacity. As for the base station, too high a power level makes no sense because the power budget for uplink and downlink must be balanced and the limits are usually determined by the uplink. Finally, for handheld radio sets health considerations have to be observed and this also limits the maximum transmitting power.

For conventional equipment only limits for discrete unwanted emissions are specified because there has been no need to stipulate broadband noise limits. For constant envelope modulation systems, if the PLL synthesizer VCO meets the noise requirements needed for adjacent channel power and adjacent channel rejection, then there is usually no problem with transmitter broadband noise. However, for linear modulation there is a nonnegligible danger of transmitter broadband noise.

Transmitter intermodulation occurs if large external signals, entering the transmitter power amplifier via the antenna, mix with the desired transmitter signal. Over a wide range of levels this happens with a nearly constant conversion loss. The ETSI method of measurement specifies levels relative to the desired carrier from which the conversion loss has been derived. This makes it easier to calculate interference distances (that is, coupling) between transmitters.

To get an impression of what all this means in practice, the dynamic link budgets and the coverage distances have been calculated for some examples. For the purpose of comparison, we have assumed that a 2-m band TETRA system is available, which currently is not the case. For propagation between the base station and mobiles or hand portables, the Hata model has been used, whereas for DMO distances between mobiles and hand portables the dual-slope model has been taken. To obtain realistic results, fading and shadowing have been taken into account. All other assumptions have been made as already mentioned, in particular, that the environment is a built-up city center. The results are given in Table 9.3.

It turns out that the coverage distances for TETRA and conventional analog PMR systems are very similar if the carrier frequency is similar. The main differences are introduced by the different frequencies and not by the equipment parameters. For analog 25-kHz systems with equal transmitting power, the link budget is 5 dB better than for 12.5-kHz systems (see Table 9.2). This means that the maximum coverage distance, which for 12.5 kHz is somewhat worse than for TETRA, would be slightly superior for 25-kHz systems. Note that a link budget improvement of 5 dB is equivalent to a coverage distance improvement of 33% for $\alpha = 4$.

Hence, there is no reason for uncertainty when replacing analog with digital PMR systems if the frequency is comparable. The reason is that the main parameters and limits for all PMR systems are similar or equal because this has been one main precondition for standardization. However, if the frequency is changed, for example, by going from one band to another, then a considerable coverage difference may occur. This is not a new insight because this has always been valid for conventional systems. One critical case is the replacement of an analog PMR system in either the 4-m or 2-m band by a digital PMR system such as TETRA in the UHF band between 380 and 470 MHz. Then about 10–16 dB of the link budget is missing and this requires a recalculation of system coverage. Improved antenna gain in combination with antenna diversity may help, but in some cases only a

Table 9.3 Examples of Coverage Distances

Property	*Band*	*System*	$BS \Rightarrow MS$[2]	$BS \Rightarrow HP$[2]	$MS \Leftrightarrow MS$[3]	$MS \Rightarrow HP$[3]	$HP \Leftrightarrow HP$[3]
Link budget a_L [dB][1]	All bands	PM 12.5	145	142	141	138	133
		TETRA	147	143	146	142	139
Maximum coverage distance d_{max} [km][2,3]	150	PM 12.5	12.7	10.5	2.7	2.2	1.7
	150	TETRA[4]	14.5	11.2	3.6	2.8	2.4
	450	PM 12.5	5.7	4.7	2.4	2.1	1.5
	450	TETRA	6.5	5.0	3.3	2.6	2.2
	900	PM 12.5	3.4	2.8	3.0	2.5	1.9
	900	TETRA	3.8	3.0	4.0	3.2	2.7

[1]Dynamic link budget according to Table 9.2 but with fading and shadowing, hence 4 dB additional loss.
[2]Hata model with $h_B = 30$m and $h_M = 1.5$m; distance according to (3.37) with $a_T = 0$ dB.
[3]Dual-slope model according to Table 3.7 with $\alpha = 4$ and $d_S = 30$m; hence, $d_0 = 4/10/30$m for 150/450/900 MHz; distance according to (3.39).
[4]Fictitious.

larger number of base stations will solve the problem and this can be a costly exercise.

9.3 Interference Scenarios and the Near-Far Problem

Propagation of desired signals has been discussed in depth and undesired signals obey the same propagation laws—thereupon the evaluation of interference scenarios can be based [21–23]. Here only the principles of how to deal with them and some specific hints can be given.

In the first step the power level of the interfering signal level p_I at a certain frequency has to be determined from the transmitter power level p_T, the out-of-band rejection a_T, and the propagation loss a_d for a certain distance d:

$$p_I = p_T - a_T - a_d = p_U - a_d \tag{9.4}$$

Here p_U is the unwanted transmitting power of the emission at a different frequency where a filter loss a_T that has already been taken into account has to be added. In many cases this can be directly replaced by the effective level of the unwanted emission if it is known. [Recall that the power level p in dBm can always be calculated from the power P by $p = \log_{10} P(P/\mathrm{mW})$ where 1 mW is the reference level.] The undesired signal experiences a propagation loss a_d when traveling from its source to the receiver that is suffering the interference. In the second step the conditions for good reception without interference can be set up:

$$p_W \geq p_I - a_R \quad \text{while} \quad p_W \geq p_{\mathrm{Rx\,min}} \quad \text{and} \quad a_R = 10 \cdot \log_{10}(P_I/P_C) \tag{9.5}$$

This simply means that at the receiver the level of the desired signal p_W must exceed the level of the interfering signal p_I minus the receiver rejection a_R of the undesired signal. In the case of a spurious response or an adjacent channel signal, a_R is usually on the order of 60–70 dB or more. The exact value has to be taken from the relevant standard or the equipment specification. In the case of cochannel interference, the rejection is small and negative and therefore the undesired signal must be smaller than the desired signal at least by the absolute value of the cochannel rejection.

If the condition for undisturbed reception is not fulfilled at zero distance, then the missing attenuation can only be provided by propagation loss a_d. From this the necessary separation distance d_I can be calculated. If this is very small compared to the coverage distance d_C, then the overall interference probability is low. If in the simplest case a system with many equally loaded radio cells and a uniform distribution of users is assumed, then the probability is simply equal to $(d_I/d_C)^2$. If the coverage distance is, say, 10 km and the maximum interference distance is 100m, then the overall probability is 0.01%. That is very small compared to the outage probability due to fading and shadowing, which is usually on the order of at least 2–5%. In practice, the overall interference probability in a cell is even smaller because in the inner parts of the cell the desired signal is much greater than at the cell border and therefore the interference coverage distance is considerably reduced or even zero. Hence, it can be concluded that an additional outage due to interference compared to a lack of coverage is mostly indistinguishable to the user, and thus the outage due to interference usually has little effect on the probability of good reception. However, sometimes specific traffic scenarios may strongly promote certain interference mechanisms and then additional care has to be taken.

All of these calculations can be made for different interference scenarios, for example, adjacent channel power and adjacent channel rejection, spurious responses and spurious emissions, and receiver and transmitter intermodulation rejection. Because all of these parameters are similar or equal for the different analog and digital PMR systems, there is no significant difference between the different systems in this respect.

However, two specific issues should be looked at more closely. The first one is the near-far problem, which can become very serious under specific conditions. It is based on the fact that signals received from different distances have different strengths due to the different propagation losses even if the transmitting power is the same. As long as any interference is much weaker than the desired signal, there is no impairment. However, if the interferer gets too strong, then the weaker received signal can be corrupted [24, 25]. Figure 9.1 allows a closer look into the near-far problem.

In the case in Figure 9.1 we consider a BS or MS 3 nearby that receives a strong signal from MS 1, which is at a short distance. That presents no problem because any signal from the remote MS 2 is too weak to disturb the desired signal. The second case, however, is the critical one where the BS or MS 3 receives a very weak signal from the distant MS 2, but the reception of this desired signal might become impossible due to intermodulation or unwanted emissions from MS 1, which is nearby. However, in the case of discrete spurious frequencies only a few combinations produce real interference and therefore the probability is low in most cases. As an example of this particular case, we look at the spurious emissions of transmitter of MS 1 as shown in Figure 9.2.

First, there are some discrete emissions that meet the requirements and limits of a particular standard; these are typically –70 dB or less relative to the carrier power. Second, the transmitter may also produce broadband noise. The crucial point is that the latter can be critical even if it is within the specified limits. In linear modulation systems, the transmitter broadband noise floor usually is only

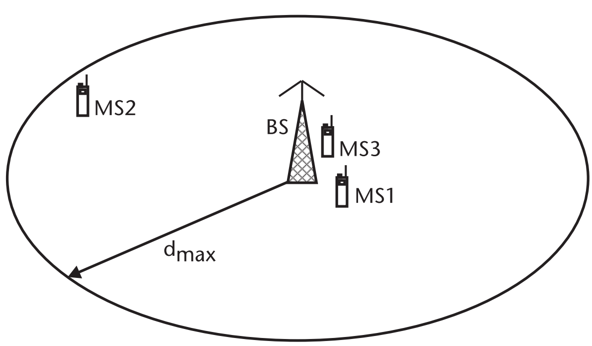

Figure 9.1 The near-far-problem.

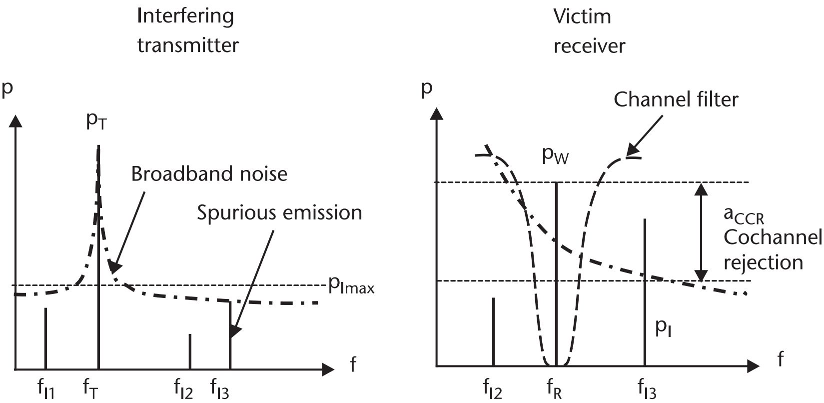

Figure 9.2 Discrete and broadband interference.

about 65–70 dB down, and extends, with relatively little change, much further from the carrier in comparison to constant envelope modulation. Looking at the receiver, it is obvious that only those emissions that fall into the receiving channel produce interference at a P_I/P_C ratio smaller than the cochannel rejection.

It all then depends on the numeric relationships of the various frequencies. If the critical emission does not equal the receiving channel frequency, then there is no problem. The same is true for interference caused by an intermodulation product. However, in the case of broadband noise the numerical relation does not count since the whole band around the receiver might be spoiled and this may result in local blocking of terminals. Only the P_I/P_C ratio matters.

Hence, to avoid interference caused by transmitter broadband noise, it is necessary for the limits for broadband noise to be well below those for discrete emissions. For conventional FM and PM radios, this is usually no problem. However, for systems with linear modulation such as SSB, TETRA, or iDEN, the necessary transmitter linearization may produce a higher broadband noise level than in the case of FM or PM. The reason is the limited dynamic range of the transmitter linearization feedback loop. In the last decade considerable progress has been made and, in principle, this problem has been reduced to a manageable level. Varying antenna mismatch can cause severe problems to linearization but this can be avoided by the insertion of circulators or isolators. Unfortunately, these have a limited bandwidth for the most part and this is a drawback for broadband design. Moreover, the dimensions of such devices are proportional to the wavelength and this can be a disadvantage at low frequencies, particularly in hand portables where the mismatch problem is most acute.

Note that receiver blocking is also an effect strongly influenced by broadband noise, even including LO broadband noise, which can produce a similar effect by reverse mixing with a strong clean interfering carrier. Hence, as long as transmitter and LO broadband noise are similar, there is a balance. Whichever is worse will limit the system properties. For this reason the TETRA transmitter broadband noise specification has been improved significantly since the first drafts of the standard.

If new frequency allocations are to be made, frequency managers must ensure that there is no unacceptable mutual interference between existing and new systems and that all interference inside a new system does not limit its use. For this purpose a comprehensive analysis has to be conducted. Today this is more demanding than in the past because the traffic density in all systems is steadily increasing. This also means that such an analysis has to be based not only on the current situation, but has to take into account the future development of the services under consideration. For old and new analog and digital PMR, it has been shown that no really new problems have occurred nor that specific guard bands between the different systems are necessary. This has been shown for TETRA and TETRAPOL and for general DPMR systems and is a consequence of the fact that they meet the requirements of EN 300 086 and 113 [26–28].

The evaluation of interference situations by taking worst-case scenarios is only useful in the case where this can prove that there is no problem at all. However, very often the quick and easy-to-handle worst-case method leads to unacceptable results and therefore statistical methods and complex models have to be chosen to get a realistic answer. This means that the distribution of interfering transmitters and victim receivers must be modeled properly in the coverage area under consideration. Additionally, reasonable assumptions on frequency distribution, power levels, propagation conditions, and so on have to be made. Then an evaluation can be carried out which reveals that under given circumstances a certain interference probability can be expected. All this can be done with Monte Carlo simulations, and the CEPT has been using such a tool for several years [22, 23].

9.4 Comparison of FDMA with TDMA

Fair comparison of FDMA and TDMA systems is a difficult problem. If real FDMA and TDMA systems are compared, then their benefits and drawbacks would not be easy to assess properly. The reason is that a trade-off is always possible between different parameters that leads to a similar result concerning the overall system performance, but which has nothing to do with the method of channel access. Therefore, a hypothetical system called PMR 6 is introduced. This is a fictitious TETRA variant with a channel separation of 6.25 kHz that indeed was discussed in the middle of the 1990s within ETSI but has never been standardized. For the comparison of FDMA and TDMA, the two systems PMR 6 and TETRA are ideal candidates because all properties not related to the channel access are identical [20, 29–31]. Before making the comparison, the main characteristics of PMR 6 are as follows:

- Carrier separation 6.25 kHz;
- Compatible with channel rasters of 12.5, 20, and 25 kHz; some loss with two channels in 15 kHz; five channels in 30 kHz requires slightly reduced frequency tolerance or somewhat smaller roll-off factor for premodulation filtering;
- Quasi-FDMA system with TDMA properties (1TDMA) such as backward signaling and transmitter interrupt;

- Modulation, data rate per traffic channel, speech coder, protocols, and services are for the most part equal or similar to TETRA.

At first the question of maximum coverage distance has to be evaluated carefully so as not to draw the wrong conclusions. For the quality of a digital transmission only the received E_B/N_0 ratio is of importance. The received useful signal energy, however, is the product of signal power density, bandwidth, and symbol duration T_S. The same is true for the received noise energy, which is the product of noise power density bandwidth and symbol duration [see also (5.112)]. If the symbol energies for TETRA and PMR 6 are equal for both receiver and transmitter, then the maximum coverage distance must be equal as Figure 9.3 demonstrates.

What does that mean in detail? We assume that the received signal and noise power densities are identical for both systems. If the bandwidth for TETRA is four times that of PMR 6, then the noise power as well as the signal power will be also four times higher. Hence, equal coverage will be achieved. However, the symbol duration of TETRA is only a quarter of that of PMR 6 and therefore the average power is identical! The only case in which TETRA exhibits less coverage distance compared to PMR 6 is when in both cases the peak power of the carrier is limited at the transmitter to the same level [see also (5.112)]. In the past some misunderstandings have occurred in public discussion on this topic when different digital PMR systems have been compared.

But there are other differences to be evaluated. Due to the smaller modulation bandwidth PMR 6 exhibits steeper sidelobes compared to TETRA. The main point is that the region below –60 dBc exhibits broader shoulders in the case of TETRA. This is of interest at the borders of a band because the region where the adjacent channel power may become critical is four times wider for TETRA than for PMR 6 as Figure 9.4 suggests. In real life, however, this is of small practical importance.

As a last specific problem intermodulation should be addressed. In practice, third-order products are the major intermodulation problem in receivers. Assuming that there are n carriers, then the number of possible interference products can be calculated:

$$n_I = \binom{n}{2} = \frac{n!}{2! \cdot (n-2)!} = \frac{n^2 - n}{2} \rightarrow \frac{n^2}{2} \quad \text{for} \quad n >> 1 \qquad (9.6)$$

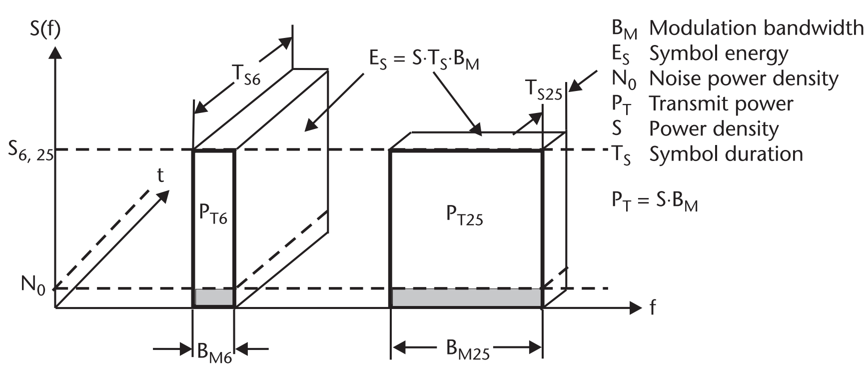

Figure 9.3 Symbol energy comparison for PMR 6 and TETRA.

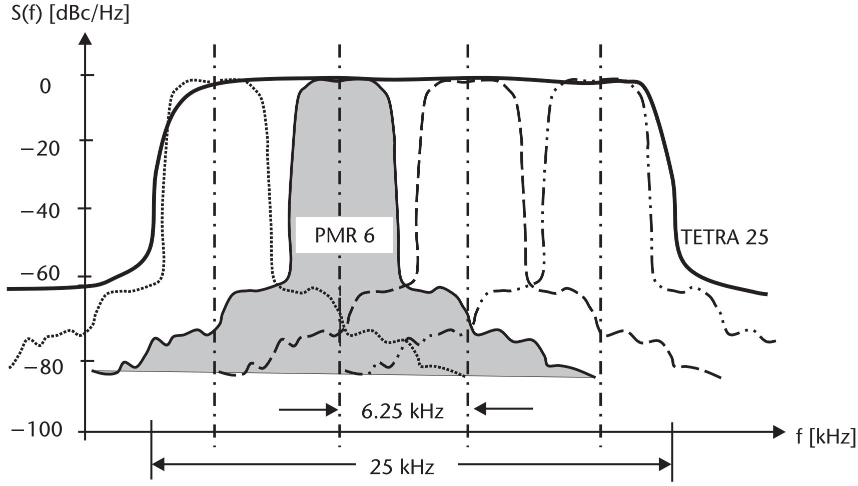

Figure 9.4 Modulation bandwidths of PMR 6 and TETRA.

Assuming equal interferer power P_I in PMR 6 for all carriers, the resulting total power P_{I6} can be determined:

$$P_{I6} = \frac{n^2}{2} \cdot P_I \tag{9.7}$$

For TETRA the number of carrier frequencies in a given bandwidth is only a quarter of those of PMR 6. However, for equal coverage four times the transmitter peak power is needed. Unfortunately, the level of an intermodulation product increases at a third-order rate as the fundamental increases. Thus, for TETRA the total intermodulation power is four times higher than that for PMR 6:

$$P_{I25} = \frac{n^2}{2} \cdot \frac{4^3}{4^2} \cdot P_I = 4 \cdot P_{I6} \tag{9.8}$$

Hence, the ratio of the transmitting power to the intermodulation power is identical in both systems. Consequently, there is no difference *inside* the band. However, TETRA also produces four times higher intermodulation interference power in the *adjacent* bands, but the practical consequences are mostly negligible.

Some general arguments are related to the hardware differences of FDMA and TDMA systems and their consequences. Table 9.4 compares the main issues of both systems for the mobile station.

Comparing the hardware for mobile stations, it is obvious that there are several differences. The much higher transmitter peak power needed for an equal link budget reduces the efficiency and needs more battery capacity for the same duration of operation due to one more power amplifier stage in the transmitter chain. FDMA, however, needs a duplex filter if real duplex operation is needed, except for TDD. This filter has a slightly higher insertion loss than the antenna switch that suffices for TDMA, but this is only a small additional burden on the FDMA transmitting

Table 9.4 Hardware Expenditure for FDMA and TDMA Mobiles

MS Characteristics for FDMA	*MS Characteristics for TDMA*
Duplex filter needed for duplex operation[1]	Antenna switch only[2]
Small transmitter peak power[3]	High transmitter peak power[3]
Equalizer advisable only for simulcast operation[4]	Equalizer beneficial
Small battery capacity	Increased battery capacity[5]
Transmission capacity limited	Transmission capacity enlargement by time slot aggregation
BS scanning only if DTX and DRX are applied	BS scanning during unused time slots

[1]TDD excluded.
[2]Duplex filter necessary if more than half of the time slots per direction are needed.
[3]Equal maximum coverage distance assumed.
[4]If any.
[5]Due to additional RF power amplifier stage.

efficiency. Finally TDMA provides a higher flexibility and additional features like dual watch, which is not possible with FDMA except when discontinuous transmission and reception (DTX and DRX) are employed. TDMA also allows several time slots to be assigned to one user while FDMA has a fixed transmission capacity. However, time slot aggregation requires more battery capacity, and a duplex filter if in the TDMA system more than half of the available time slots per direction are assigned to one user.

An equalizer is not necessarily needed in either case because primarily flat fading is experienced for the bandwidths under consideration. FDMA simulcast systems may lead to large propagation time differences but usually the signal strength is nearly always so different that rarely one of the severely delayed signals disturbs the much stronger one from the nearest base station. Conversely, signals of similar strength most likely exhibit only small delay differences. The main general differences between FDMA and TDMA base stations are listed in Table 9.5.

The antenna coupler needed to combine all of the transmitters onto one antenna is less complex in TDMA, and all channels of one TDMA carrier can be processed in the RF, IF, and baseband sections simultaneously. However, the footnotes in the table show that alternatives exist; for example, the antenna coupler problem disappears if linearized high-power RF amplifiers are available. In principle, these are not very different from the normal linearized RF power amplifiers needed, for

Table 9.5 Hardware Expenditure for FDMA and TDMA Base Stations

BS Characteristics for FDMA	*BS Characteristics for TDMA*
Complex antenna coupler[1]	Simplified antenna coupling
Each traffic channel has to be processed separately[2]	Simultaneous processing of several traffic channels
For analog-to-digital transition old BS sites can be used	High-gain antennas and/or diversity or additional base stations necessary[3]

[1]Alternatively, highly linearized (but expensive) RF power amplifiers can amplify several channels simultaneously.
[2]Fast ADCs and powerful DSPs allow several traffic channels to be processed simultaneously.
[3]Only valid if *peak* power is equal or if the FDMA system operates on a *lower* frequency.

example, for TETRA mobiles or base stations, but they have to meet somewhat higher linearity requirements at considerably higher RF output power and will therefore be expensive. We should not overlook, however, the fact that complex conventional antenna couplers are also not very cheap. The (actively linearized) linear amplifier is better suited to handle several channels over a small band in contrast to the conventional coupler, which is better suited to larger frequency differences.

However, we must stress that a complete comparison of real TDMA and FDMA systems is less easy because other system-specific factors must be taken into account, mainly modulation and coding together with error correction, the different speech coders, and a different combination of available services.

In the end, whether FDMA or TDMA is the better solution for a particular application depends on various parameters, for example, the whole system structure, the number of subscribers, and the carried traffic. FDMA and TDMA both have benefits and drawbacks and only a thorough analysis can reveal which is more appropriate in the case considered.

9.5 Shared Use of Channels

Where *exclusive use* is assigned in PMR systems, whether with one or many channels, then frequency allocations in the band present no problems with the differences between different systems. However, in many cases PMR frequencies are *not* assigned exclusively. This means that a certain number of user groups share one channel. If the systems used employ different technologies, then interference, message collisions, and other problems should be avoided or at least minimized. This means that there must be rules to ensure the proper coexistence of different systems in one channel without these negative effects.

For this purpose ETSI has created EN 300 471, which offers an access protocol and occupation rules for the transmission of data on shared channels. The mix of data and speech transmission is also addressed. No requirements defining the mode of operation, the bit rate, or the type of modulation have been fixed. Within the framework given by this standard, different user groups may use their own protocols [32]. However, this standard is not well suited to fast digital PMR systems such as DIIS because its mechanisms are too slow and thus too much transmission capacity is wasted. Hence, an extension or a revision of this standard will become necessary or another one suited to fast systems has to be created in the future. At present there is no related work item under consideration within ETSI.

The U.K. Radio Agency started its own work item under this topic and the result is known as Interface Requirement 2008 (IR 2008) for *Private Business Radio* (PBR). It defines two TDMA channel access procedures for data only on shared wide-area PBR channels. One for time slots of 250-ms duration and the other one for 500 ms. In both cases there are eight time slots per frame, allowing eight users to share one frequency. Thus, the frame length is 2 or 4 seconds, and corresponding downlink and uplink slots are shifted by four time slots. Any modulation scheme, bit rate, or data protocol is permitted but no digital voice. All synchronization and timing is derived from the atomic-controlled *coordinated universal time* (UTC) with fixed rules and an accuracy of ±1 ms. Moreover, every BS transmission must contain an 8-bit fleet code [33].

9.6 Frequency Band Refarming: The European Frequency Allocation Table

In Europe the use of PMR bands and the technical parameters are very inconsistent; therefore, several CEPT reports and the European Table of Frequency Allocations, which is under preparation within the CEPT ECC (formerly ERC) and CEPT ERO, are trying to achieve unified use of the PMR bands and consistent technical parameters [34–36]. However, such a refarming of the PMR bands requires time and money because existing licensed users cannot be thrown out of a band at short notice, and the replacement of obsolete technology with new equipment and infrastructure is costly. Nevertheless, the European Table of Frequency Allocations is envisaged to be in force beyond 2008.

In Phase II of the CEPT *Detailed Spectrum Investigation* (DSI), the future need for PMR frequencies in Europe was estimated in the late 1990s. The result was that, on average, 115 MHz of spectrum is needed for smaller countries and at least 155 MHz for larger ones. The predicted increase in the number of users was about 50% within 5 years and 100% after 10 years. Traffic per user is expected to triple from 8.2 to 25 mErl caused by more intensive use of radio communications and, in particular, by increased data transmission. Hence, up to six times the transmission capacity is predicted to be needed [37–40]. Improved frequency economy aided by advanced modulation, coding techniques, and trunking gain will provide improvements by up to a factor of 3. However, the replacement of existing analog technology is likely to need more than 10 years in total, maybe 15 to 20, and trunking is not well suited to all PMR applications. Therefore, at least doubling of the total bandwidth required for PMR was calculated for the 10 years following the DSI. Admittedly the timetable estimated appears to be too short, given current knowledge, but there are few doubts that it is only a question of time until a huge tranche of additional frequencies will be needed. And, finally, we should keep in mind a statement made by Joseph Gelas, the chairman of the ECTEL EFSG, around 1995. He said succinctly and to the point: "No frequency—no market!"

However, the lack of frequencies is not just a European problem. Also around 1995 groups in the United States conducted several studies concerning new spectrum requirements for U.S. land mobile service. It was estimated that more than 200 MHz would be needed within the next 10 years and nobody knew where it would come from [41]. Since then additional allocations have been made in the United States.

In the past to obtain additional channels, channel splitting was performed in several steps by taking 100-kHz channels to 50, 25, and eventually to 12.5 kHz, whereas in the United States the last step was the split from 30-kHz into 15-kHz channels. In Germany and some of the surrounding countries, a direct jump from 100- to 20-kHz channels was performed. The split is usually made by inserting new carrier frequencies between the existing ones and cutting the modulation bandwidth in half. However, the drawback is that at the borders of countries with different channel separations or in the case of distant reuse of frequencies with different channel separations, one of the old broader channels has to be coordinated with up to three of the new narrower ones. This coordination problem becomes increasingly serious due to the increasing congestion of PMR bands.

A much cleverer channel-splitting strategy that permits more simplified frequency coordination has been discussed within the CEPT. It is much better not to match the center frequencies but the borders between channels. Then one broad channel needs only be coordinated with two channels of half the bandwidth. The necessary frequency offset imposes no problem for equipment design today. Bundling of channels to achieve higher bandwidth and transmission rates is possible by using the same principle in reverse. For example, it would make it easier to introduce TETRA or other future PMR broadband systems into a band where only 12.5-kHz channels are currently available [42].

9.7 Migration from Analog to Digital PMR Technology

The migration question is very important for all of those user groups that already have an analog PMR system and want to replace it with a digital one because they need fast data transmission, encryption, or one of the other new services offered. For them the possibilities of soft migration and of saving at least part of their investment is of major interest because completely replacing their existing system would be a very expensive exercise. Therefore, they would like to keep some parts of their existing equipment and merely upgrade it from, say, MPT 1327 to TETRA. Some manufacturers, for example, Motorola and R&S BICK Mobilfunk, offer this feature [43].

Drawing up a block diagram of a digital mobile radio unit as shown in Figure 9.5 reveals that all of the different digital radio transmission systems need the same basic functional blocks. The system-specific components are mainly the RF circuitry including the duplex filter or antenna switch, channel filter in the IF chain, synthesizer (mainly the VCO), and the (maybe linear) transmitter power amplifier. These depend on the frequency bands and the different channel and duplex separations [20, 44].

In modern equipment many parts of the circuitry are realized in broadband design to reduce the number of variations for different applications. Highly sophisticated and flexible ICs can be adapted to many functions that are different in the different systems. This can be done by adjusting some of the hardware or software parameters or by changing software modules.

The channel filter, which is one of the most critical components, can be implemented by a filtering algorithm running on a DSP provided that a fast enough ADC, with low power consumption and sufficient resolution on the order 14–16 bits, is available at reasonable cost. This has been a serious problem in the past. If the sampling frequency is high enough, a single IF superhet with a high first IF, which simplifies the image frequency rejection problem, can be built. Alternatively, an imageless direct conversion receiver can be made, but these need to be designed very carefully to avoid other drawbacks [45–47]. The necessary computational

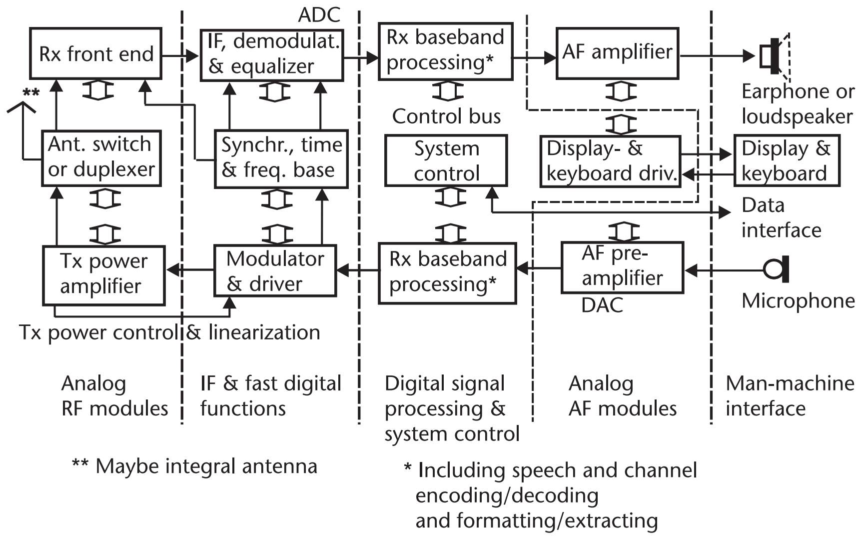

Figure 9.5 Block diagram of a digital mobile radio set.

power for current DPMR systems is on the order of 20–80 MIPS similar to the latest GSM designs, whereas for 3G systems about 1,000 MIPS digital IF and 2,000 MIPS digital baseband computational power will be needed [48–52].

In short, modern components and software algorithms allow very flexible radio structures with very differing properties at an acceptable cost. Highly integrated circuits offer the required functionality at low cost coupled with low power consumption even if the necessary computational power is high. The complexity of such ICs can now exceed what is needed in most of the systems. Hence, multistandard radios will become commonplace and will offer the user system integration inside his radio terminal even if system access has to be made via different air interfaces [44]. Such a flexible radio structure can be used for different digital mobile radio systems and its properties can be determined completely by software. In particular, a software radio can also be run in an analog mode. This is of interest for migration because such radios can be used in an existing system for analog transmission, and after upgrading the system to digital transmission the mobiles can simply be switched over. Technology will give us more and more freedom but the market will decide what is really needed.

The migration from analog to digital systems has several different technical aspects. For the terminal unit it is very simple: Existing analog PMR terminals cannot be upgraded to TETRA or another DPMR system. However, new digital radio terminals can be designed in such a way that they are capable of being switched to analog mode and used in old networks until the network is upgraded. Then these radio terminals are switched to digital operation.

Similarly analog RF hardware in base stations cannot be used for new DPMR systems. Here the same approach as for the mobiles could be taken. New digital modules could in principle also be used to support analog transmission. However, it is doubtful if this is economically feasible. For the switching and transmission tasks in the infrastructure, there is little dependence on the different air interfaces.

Hence, in this area many parts of the equipment may also be used after transition to a digital system. If the design has been made flexible enough, only new software has to be downloaded to the different network elements. Finally, some components are completely independent from the RF structure, for example, the power supplies.

Existing FDMA network configurations may be mapped on to DPMR networks, but with TDMA different network configurations are also possible [20]. The simplest case is a group of mobiles communicating in direct mode without support from any infrastructure. However, for some applications, in particular, TETRA terminals, TDMA may be too expensive and therefore cheaper solutions such as PMR 446 may be preferred. For highly professional applications and if a direct mode besides a complex functionality is needed, only a few alternatives are available.

Large organization like those devoted to public safety need multisite systems with regional or even nationwide coverage. Handover and roaming capabilities are usually required along with access to the terrestrial telephone network. The features of such systems may exceed those of basic PMR and may also employ many mobile telephone functions. In the past large organizations like the police and fire brigades had their own separate mobile radio networks. In future system configurations, this may change completely. A system operator may run a very large trunked TDMA network that provides mobile communication to different CUGs, which remain functionally isolated. Mutual interference is avoided and the different groups cannot block each other. All utilities (electricity, gas, water, oil supply, and so on) have very similar tasks and they also maintain large networks and have common coverage areas. Thus, they may form another group being served by a common digital trunked system, and there are many other examples. Figure 9.6 shows the past and the possible future of TETRA in large combined networks.

TETRA systems with single base stations could, in principle, replace conventional single base station systems directly. However, the minimum TETRA transmission capacity comprises four channels. For small systems with few users this capacity is very often much more than needed and hence part of the system capacity could be rented to other users. However, due to the regulations in many countries,

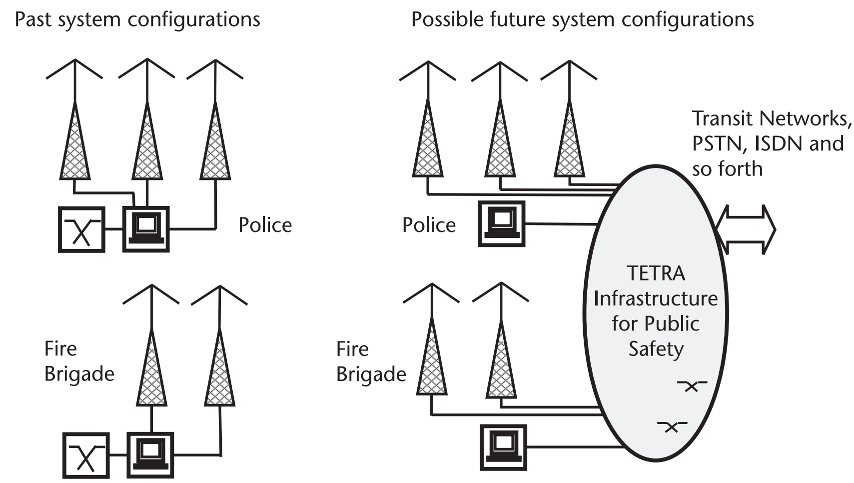

Figure 9.6 Network configurations with TETRA.

the system owner then becomes a *system operator* and needs an *operator's license,* which he will often not get due to existing legislation. Such users are then obliged to join networks operated by third parties, but they will no longer have full control of system availability. Moreover, the third party intends to earn money and make a profit by leasing transmission capacity. In the long term, this means that a privately owned system should be less costly despite the high initial outlay needed to install such a system.

Nevertheless, the introduction of large nationwide digital trunked networks based on digital technology will force many users of conventional PMR networks to change over to the new technology and to outsource their communication tasks to an independent network operator. Moreover, most DPMR networks and, in particular, TETRA, offer full PMR functionality and also mobile telephony access with many advanced features. This makes them attractive to those user groups that need PMR-style operation such as group call and status messaging but also need access to public mobile phone systems.

9.8 EMC Issues and Health Risks

During the last few decades electronic and radio equipment has proliferated. Unfortunately, the different types of equipment have exhibited unwanted interactions. Radio transmitters, in particular, may disturb the proper functioning of other sensitive electronic equipment. The interference on a TV set or a terrestrial telephone caused by a mobile telephone operated nearby are only two, but very annoying, examples. Malfunction of equipment in a hospital, however, must be avoided under all circumstances because the life of a patient can be endangered; consequently, the use of mobile phones is often forbidden in such locations. However, hardening of electronic equipment against strong RF signals presents no great technical difficulty, but it might require some extra cost.

A much more serious technical problem is that of unwanted emissions that fall into highly sensitive radio or TV receivers. One example is that of the harmonic of the sampling frequency of CD players, which falls directly into the flight radio communication and navigation band. This is one of the reasons why CD players are not permitted to be used in airplanes. Of course, there are many more interference scenarios and this is the reason why *electromagnetic compatibility* (EMC) has become an important issue in the last decade.

Two main strategies are used to minimize EMC problems. First, frequency managers have to be careful with new frequency allocations. Proper frequency management can prevent many EMC difficulties. The other point is that for unwanted emissions as well as for immunity against intentionally transmitted radio signals, reasonable limits have to be set, which means that we have to find the best compromise between accepting a certain amount of residual interference and the making the necessary expenditures to reduce it. For PMR equipment operating in the frequency range from 30 MHz to 1 GHz, ETSI has fixed the limits for emissions and immunity in EN 301 489-1 and EN 301 489-5 [10, 11].

There is another problem related to RF equipment and, in particular, to radio transmitters. These are the possible health hazards they present. If a piece of equipment meets the EMC standards mentioned above or those specified in other parts of the world, this does not guarantee safety against harm from radio energy. This is an issue that has been discussed widely by the public for at least two decades. As far as RF energy in mobile services operating in the frequency range from 30 MHz to 3 GHz is concerned, two different kinds of effects should be considered [53–56].

First, there are *thermal effects*. In the frequency range used by mobile services, say, from 30 MHz to 3 GHz, it is very unlikely that RF energy can harm parts of the human body without first being noticed because the body has very sensitive heat receptors and the power emitted by mobile equipment is too low. If, for example, 5W is being transmitted nearby, only a small portion of it can be coupled into the body, let us assume 50% or 2.5W (even this is a very high proportion). Many investigations have found that in reality the absorbed energy can be much lower to produce an effect, but these studies have also pointed out that such an effect does not necessarily cause any harm. In contrast, sunbathing on the beach results in the body receiving energy at a rate of 50–100 mW/cm^2 depending on the location and the time of year. A medium-sized person will exhibit a surface of about 0.5 m^2 to the sun, which is equivalent to the reception of 250–500W, resulting in roughly 5W/kg! This amount of energy poses no problem for the human body provided that sunburn is avoided.

Measurements carried out with handheld equipment at a transmitting power of 6.4W on a simulated man in one of the PMR bands resulted in a *specific absorption rate* (SAR) that scarcely exceeded 0.5 mW/g = 0.5 W/kg for the most exposed parts of the body under normal operating conditions. Note that peak values may be considerably higher, but this is not very likely. Finally, the total absorbed RF power can never exceed the total transmitted power and cannot exceed a few watts in contrast to the sunbathing scenario. Moreover, in PMR the exposure time is much less than 6 minutes, which is the averaging time fixed in the standards, so the total RF energy absorption is well below the limits of, for example, the U.S. standards [57, 58].

Another contentious issue is the alleged health risks posed by high-power base stations, but members of the general public are unlikely to climb up the antenna masts. If they are some distance away, even close to the base of the tower, it is easy to calculate that the power density is much too small (see Chapter 3) to do any harm and is considerably less than the exposure to RF energy from a handheld mobile [55].

Second, the suspicion has arisen that dangerous *nonthermal effects* might exist that could promote, for example, cancer. However, for nonionizing radiation such as that used in the mobile services and for power densities below the level where thermal effects can be noticed, there are no known mechanisms for such effects. The numerous studies carried out since around 1980 and, in particular, in the last few years have not been able to exclude with 100% certainty that such effects do exist, but they have clearly shown that the probability *must be* extremely low. This means that in everyday life there is a greater likelihood of being killed by a flash of lightning in a thunderstorm or of being hurt by a car when crossing the street in front of one's house [53–56].

Being fair, one also has to look at how many lives have been saved by the ability to call a doctor or ambulance from a mobile radio; those lives might not have been saved if those calls had not allowed medical personnel to arrive as quickly at they did at the scenes of accidents. Moreover, the beneficial effects of RF are well known. Doctors use heating of human tissue by RF energy at a level on the order of many dozens of watts to heal specific diseases. Moreover, recall that many of us have been living under showers of RF energy for at least half a century. A large number of broadcast and TV transmitters constantly expose most of us to RF energy in the same frequency range as mobiles and with similar modulation. Even the low modulation frequencies under suspicion in TDMA systems are nothing new. TV transmitters produce signals with similar frequency content. But the distance from the transmitters is usually high enough to ensure that the energy density is too small to be dangerous. Finally, PMR users have operated their radio sets for decades in different frequency bands and with transmitting powers of several watts without any health problems being reported due to exposure to RF energy.

To be clear, the possible dangers from exposure to RF fields should not be neglected, but the discussion must be balanced and put into context with the other dangers we have to live with on a daily basis in order to see the real meaning and relative importance of the problem. Compared to other risks encountered daily, there is no evidence of a significant problem related to the exposure to the RF power distributed by today's mobile radio communication systems, broadcast, and TV transmitters.

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CHAPTER 10

Outlook: PMR in the Decade Until 2010

New developments in electronics depend heavily on semiconductor technology and its progress. This is one of the most important factors determining the further evolution of mobile communication including PMR. Knowing the driving forces and the trends in technical developments assists us with analyzing the PMR market and the prediction of future trends in PMR development.

10.1 The Driving Forces

The major driving forces for the future development of PMR are the speed of semiconductor integration, advances in system and coding theory and digital signal processing, and the evolution of mass markets for mobile phones having exceeded 1 billion subscribers in 2002. Together with extended user needs, these factors are going to revolutionize PMR [1–3].

Current semiconductor technology is based on three important cornerstones. The first was the invention of the transistor in 1947 by Shockley, Bardeen, and Brattein. The subsequent fast replacement of vacuum tubes by transistors changed communication and digital data processing completely. The next big step was the independent invention of the integrated semiconductor circuit by Kilby and Noice in 1958–1959. Despite the fact that tubes with multiple built-in systems had been created much earlier, the transfer of this idea to semiconductors opened a wide new field of applications because semiconductor integration is much easier than the production of multiple-system tubes. The decisive difference is that it is possible to shrink transistors to extremely small sizes and to put very large numbers of them on a single chip. The next big step was made by Ted Hoff who in 1971 developed the first microprocessor, the 4004, at a small and new enterprise called Intel. The project for which the 4004 was made failed, and the processor was offered in the open market with low expectations by the manufacturer, but it had an unexpectedly favorable response:

> The transistor was invented little more than half a century ago. Integrated circuits have been commercially available for four decades and the first microprocessor was created only 30 years ago. Nevertheless, these events have changed our world considerably and modern communication would not be possible without them.

Astonishingly, the progress in semiconductor integration technology has followed a clear trend during the last four decades that allows precise prediction at least for the near future. Starting around 1960 at about 100 μm, the shrinking process of transistors had by the end of 2002 reached the 0.1-μm region. The idealized diagram shown in Figure 10.1 illustrates the so-called "Moore's law," which assumes that about every 18 months the number of transistor functions per chip doubles. On a logarithmic/linear scale Moore's prediction is a straight line running without interruption from the early 1960s to the present day. Currently, the shrinking process seems to have accelerated as the production of devices with 0.09-μm structures was started in early 2003. Clock frequency increased faster than expected when Pentium IV processors became available with clock frequencies above 2.5 GHz by the end of 2002. Already in 1999, *dynamic random access memories* (DRAMs) with 4–16 Gb and a cell size of less than 0.5 μm^2 were to be found in research labs and 100-Mb SRAMs equivalent to 10^9 transistors have also been made [4–9].

The straight line in Figure 10.1 represents *memory chips,* in particular *read-only memories* (ROMs) and DRAMs, which can be packed extremely densely due to their simple and regular structure. For *processors* only about *10%* of the transistors are feasible on the same chip size due to their more irregular structure. However, this does not change the main relationships and conclusions that can be drawn.

A chip representing the latest state of the art is a very expensive item, but 10 years later it is old and well-proven technology available for just a few dollars, as the example of the 68000 processor has shown. The price of such a circuit can fall during its lifetime to 1% of its original value. With shrinking structures and increases in the maximum clock frequency, the computational power of chips is constantly rising. For technological reasons, the maximum chip size remains more or less constant around 100 mm^2, which also limits its power dissipation.

In the past it has been assumed several times that this steady growth must come to an end for technological or physical reasons, but until now all of the supposed barriers have been broken by clever new technological methods. For the

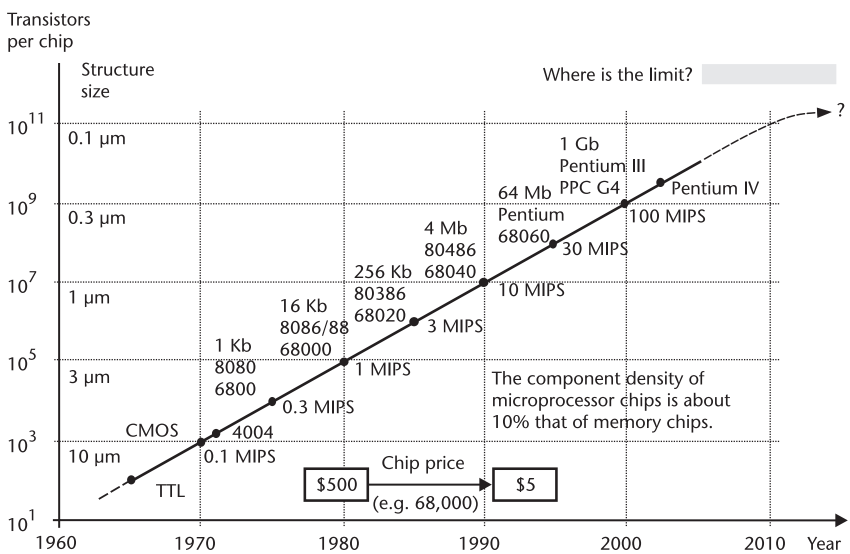

Figure 10.1 Microminiaturization trend of semiconductors.

future this is only feasible with improved and new semiconductor integration technologies, such as advanced CMOS and BiCMOS, strained silicon, SiGe, and *silicon on insulator* (SOI) and their combinations, which are being enabled by considerable progress in lithographic techniques. New materials and technologies are constantly being investigated. Advanced CMOS is going to be used for linear and digital functions and even for RF power devices, thus superseding expensive GaAs technology. For chips with high power density and temperatures up to 450°C, new materials such as SiC are under consideration. Transistors with 20-nm gate lengths have already been made in laboratories, promising that 20-GHz speed CPUs will become available around 2007. Later new solutions will be needed, perhaps double-gate FETs. Furthermore, molecular computing based on nanoelectronics promises extreme shrinking capabilities. A 64-bit memory on a 1-μm^2 chip surface and a molecular half adder needing only 100 nm^2 was demonstrated in 1999. Thus, it can be expected that the trend currently carried by planar integration technology will continue for at least another decade [10–12].

At the beginning of the 1990s, many chips were needed to build, for example, a GSM phone, but now in principle only one chip suffices. For technical reasons RF and analog functions, as of now, are separated from the digital functions, but new technologies are evolving, for example, fast new CMOS processes, that allow integration of active devices for very different tasks on the same chip. Now the chips have reached a complexity that can accommodate more than one access system. Therefore, multistandard radios will soon become widespread and allow access to different systems including PMR. Integrated tri-band GSM transceivers with direct conversion have already become commonplace. Single chips for DECT and a GSM transceiver chipset with only three chips are already available. The first TETRA baseband chip has recently been offered in the open market. GSM/DECT combinations have been made and the combination of GSM with a PDA has been available for several years. One of the latest is the Nokia 9210 communicator, which paves the way for wireless multimedia. Similarly, the latest GSM phones offer integrated camera and transmission of still pictures [13–27].

The integration of GPS and traffic information, baseband and PDA functions, and similar topics are under investigation. The introduction of advanced chipsets has permitted the reduction of the number of discrete components in a mobile phone to 100 or fewer. The number of additional passive components has been reduced together with the replacement of traditional PC boards by new technologies such as *low-temperature cofired ceramic* (LTCC). This allows the integration of high-*Q* RF devices permitting smaller mobile telephones than ever before at reduced costs. Another advanced approach to an old problem is the invention of tiny but efficient ceramic antennas (metal with ceramic coating or ceramic body with a metallized structure printed on) and slot antennas as used at microwave frequencies. As already mentioned, the mixing of baseband analog and digital functions and the integration of IF and RF functions is difficult and is at present nearly impossible or at least still too expensive. However, the trend toward the single-chip radio is steadily progressing because in the end it will allow for smaller designs and will be cheaper. Fast ADCs with high resolution, 14 bits and more that are capable of 80 Msamples/sec and beyond are already available, and they are another precondition that is just becoming available at acceptable cost such that the single-chip radio and the software radio will soon converge. Remember that the first and relatively simple GSM phones needed about 30 MIPS. Hence, Figure 10.1 indicates that a one-chip GSM phone could never have been reasonably expected before the second half of the 1990s. Improved speech coding such as in the GSM half-rate codec requires a significant increase in processing over the much simpler full-rate codec, but it has now become feasible [28–30].

Many techniques and methods have for a long time been known in theory, but it has been impossible to utilize them because the cost, volume, and weight of the hardware would have been much too great. Information and system theory have offered new methods of signal processing that only became feasible with the advent of highly integrated and powerful digital signal processors. Complicated mathematical engines can be embedded in silicon today at very reasonable cost, large memories are now available, and very complex logic and control functions can be carried out by a single fast microprocessor. These are the preconditions for powerful channel equalizing and voice coding, both of which need an incredible amount of processing power. Voice coding and channel equalizing in GSM would never have been commercially feasible without such chips.

Unfortunately, 3G handsets must be multistandard, multiband devices because there will be no single 3G standard but a family of standards with differing frequency allocations all over the world, including different modulation and access methods such as *wideband CDMA* (WCDMA) and *time division-CDMA* (TD-CDMA). The 3G will also comprise *mobile satellite systems* (MSSs) and, moreover, backward compatibility to 2.5G systems is required. The necessary computational complexity exceeds current GSM phones by about 100 times; hence, 3G handsets are a real new technological challenge needing more computational power and, as a consequence, much improved batteries. In addition, multimedia applications require high-resolution color displays with low power consumption. All of these are reasons why technical difficulties were experienced with the first UMTS radios and why their prices were expected to be high. Nevertheless, the first 3G terminals went into series production at the end of 2002 [31].

10.2 Future Development of Radio Communication Technology

A number of additional trends have had a strong impact on public mobile radio communication and have led to steadily improved products at constantly decreasing costs and to more powerful and easier to use communication systems. Moreover, many products are surpassing their old limits and acquiring new functions from other products and fields. However, everything is constantly and speedily moving, meaning that new ideas continually arise as others vanish soundlessly [23, 32, 33].

Radio LANS or *wireless LANs* (RLANs, WLANs) are supported by many manufacturers and offer access to mobile radios, mobile computers, printers, and the like in small wireless local networks at low cost. This applies in particular to Bluetooth employing FH/TDD technology providing data rates of somewhat less than 1 Mbit/s at 2.5 GHz. Single chips for integration into mobile radios, laptops, printers, and so on are envisaged for around $5. Bluetooth provides very limited coverage—only about 10m for indoor applications—but up to 100m is possible with an additional 20-dB gain power amplifier. Therefore, within a short time it is expected to be widespread in the marketplace. More powerful is the admittedly more expensive WLAN family of IEEE 802.11 standards. This operates in the 2.5- and 5-GHz bands with several types of modulation, spread spectrum, and ODFM access schemes and offers data rates from 1 to 54 Mbit/s and a 10-fold coverage increase compared to Bluetooth. Both compete with DECT with respect to fixed network and mobile internet access [34–40].

For some time the *Wireless Application Protocol* (WAP) has been under discussion. It is based on existing Internet standards and procedures (XML or WML, HTTP, SSL, and TCP) and allows Web sites to be shown on the small displays of mobile phones. However, it has had difficulties penetrating mobile markets because it is too expensive to use and has a poor MMI. The Japanese i-Mode is a similar packet and IP-based service that uses simplified *Hypertext Markup Language* (HTML) instead of *Wireless Markup Language* (WML) and permits easy mobile Internet access without a PC or a PDA. Due to its overwhelming success i-Mode was introduced in Europe at the end of 2002 by several cellular operators [41–43].

If an innovation like i-Mode appears at the right time, meets market demands, and is easy to handle, then it has a good chance of success. Other examples are *enhanced messaging service* (EMS), which allows the transmission of pictures and tones from one mobile phone to another. MMS is a further enhancement of SMS and EMS to exchange video clips and music files between mobile phones. The introduction of MMS together with new handsets was planned for the end of 2003. A further step will be unified messaging, which is a common format for all kinds of messages whether they are text, pictures, video clips, or tunes.

Cordless phones are trying to invade the PMR and GSM markets at least for specific segments and tasks. Public access points with improved coverage and a limited mobility management permit the use of cordless telephones in urban areas when away from home. The DECT/GSM dual-standard handset also permits connection to GSM if DECT is not accessible. If additionally an MSS part is built-in and if there is no coverage by terrestrial GSM, then mobile satellite systems could be automatically addressed. The additional expenditure for MSS access is low if Iridium is used since this is based mainly on GSM technology [23, 32].

Unfortunately, MSS has not evolved as predicted; consider, in particular, the example of Iridium, which floundered mainly because GSM coverage has been provided faster than expected for large parts of the world. However, Iridium is back again and may have a bright comeback in the end, and other MSSs are evolving steadily. INMARSAT is targeting new subscribers by offering tariffs that are less than cellular roaming and will introduce an IP-based *Regional Broadband Global Area Network* (RBGAN) at 4 Mbit/s in 2003–2004. TELEDESIC is not yet mature but aims to become—within a decade—the Internet in the sky; however, currently its appearance is uncertain [44–46].

GSM cordless telephone service (GSM-CTS) is a new feature of the GSM standard that competes with DECT and needs only a GSM software upgrade. In the home environment, communication runs via a CTS *home base station* (HBS). Such GSM "home zones" cover distances up to 200m. Leaving the home zone, the handset registers automatically on the subscriber's GSM network. This new

GSM service, by the way, has made DECT/GSM dual-standard sets questionable [33].

The fixed networks are providing steadily increasing bandwidth and are encouraging users to use broadband services for multimedia and information retrieval. This is also one of the driving forces for the further development of GSM and UMTS, and PMR users will also wish to have higher data rates. The promises of UMTS and other 3G systems are manifold, in particular, speedy data transmission for mobile fast Internet access and multimedia applications. Improved connectivity to other networks and access anywhere at any time are among their core ideas. On the other hand, many predictions have shown a more pessimistic attitude for the near future as to the necessity for these most advanced new features such as global roaming or fast broadband services. Moreover, competing ideas such as fast data transmission using OFDM similar to DAB are trying to outdo UMTS. As fixed and mobile networks converge, there is a strong demand to extend number portability so that a single personalized number may be used in any communication network independent of technology. There are differing approaches to come when migrating from GSM to UMTS without disturbing business more than necessary. The most likely is to use advanced 2.5G networks, with packet data and IP capability, as UMTS core networks so that only an additional RF interface is to be added [1, 38, 47–62].

Finally, the reader should keep in mind that with the *advanced speech call items* (ASCI) GSM systems are going to provide some of the most important PMR functions. Very soon they will also offer enhanced bit rates up to about 500 kbit/s and UMTS will provide even higher bit rates up to 2 Mbit/s, but only under very good propagation conditions and over very small distances from the base station and at limited mobile speed.

For certain, more powerful integrated circuits, new components, high-energy density batteries allowing longer battery operational times, improved digital signal processing, and colored displays with better resolution and less power consumption will come, and a mobile phone will become even more of a consumer device. Market forces are surely the primary determining factors for further development of a technology. Competition and large volumes drive prices down, which in turn attracts more users longing for more products. GSM has demonstrated this very impressively during the last decade.

10.3 Long-Term Prospects of Digital PMR

For users, manufacturers, and administrations involved in PMR, it is of great interest to know what will happen with PMR up to the end of the first decade of the new century. Of course, nobody is able to make precise predictions and even market analysts can only take into account what is already known. However, a review of all relevant facts and trends mentioned earlier can at least give us an idea about what might happen. For sure, PMR will not show explosive growth but a steady increase of subscriber numbers as it has in the past.

Components that were originally developed for mobile computers and public mobile radio communication systems will also be available for PMR. New features from public mobile phone systems will also invade PMR. All of this is going to revolutionize the technical basis of PMR. An improved MMI may also have a significant positive impact. A powerful voice recognition capability that allows workers to user their hands for separate actions while controlling their radio communications is only one example [33].

The limitations of conventional analog transmission can be overcome by the introduction of digital PMR, which retains all existing PMR features and adds some new ones based on higher data transmission speeds. Hence, at first many analog PMR systems will be replaced by digital standardized and proprietary PMR systems such as APCO 25, iDEN, TETRA, TETRAPOL, and DIIS. Fast data transmission will become available over PMR; for example, mobile Internet access, e- and m-commerce applications, and the like will appear together with new functions based on Internet protocols. Voice transmission will presumably also be based increasingly on VoIP procedures.

New trends in the public mobile networks will most likely also have a nonnegligible impact on PMR. Hence, we can expect EMS, MMS, unified messaging, WAP, and i-Mode to be in demand in the PMR domain. Most likely this will be the case only for those services that suit the demands of professional users. The general influence of UMTS on PMR is very uncertain and will depend on how far the UMTS functionality can go to meet specific PMR requirements. Based on more advanced technology than is available today, we might eventually see different public and nonpublic mobile radio services being merged in order to fulfill user needs even better [63].

Another impact may come from the nationwide PAMR networks with handover and even international roaming that are based on the TETRA technology. Such systems have been introduced in several European countries, but in the meantime doubts have arisen concerning the viability of their business model because the main player in Europe has failed. However, in late 2002 Dolphin Telecom started a comeback in several countries, for example, the United Kingdom and France. At the same time a number of separate public TETRA systems have begun operating due to the undisputed benefits of this standard. In general, TETRA markets are developing well, and modern meshed full IP-based TETRA network architectures have much better resilience against failure than those employing traditional circuit switching. Moreover, with TETRA Release 2, higher data rates, improved coverage, and a wider scale of traffic densities that can be covered economically are envisaged, and this will allow TETRA to compete in its specific domains with UMTS [64–68].

One other impact on the PMR market could be the growing penetration of new low-cost and leisure applications such as *Family Radio Service* (FRS) in the United States. Some years ago SRBR exhibited unexpected growth, and at the turn of the century its successor PMR 446 was introduced throughout Europe. National administrations are looking to deregulate portions of the spectrum in an attempt to reduce their workload and to stimulate growth in the market. The user groups and their needs are also changing rapidly today. Some of them will leave PMR and change over to other mobile services; others will remain but will have changing needs.

As has already been shown in Chapter 2, public systems can only partly meet PMR operational requirements. On the other hand, there are new features in public networks that might become of interest to PMR users. The consequence is that future digital PMR systems will have to be much more versatile than current ones.

Other influences are also of note, such as the new European regulations including a new type of approval scheme based on the R&TTE Directive, changing licensing conditions and modified fees, and the envisaged refarming of the PMR frequency bands. Changing and extended frequency allocations are also an important topic for U.S. regulations. We must stress again that the further development of the PMR markets and the widespread introduction of DPMR depend heavily on the availability of frequencies [69–71].

PMR belongs to the group of radio communication systems with the best frequency economy. Thus, increased promotion of modern PMR technology provides a very good opportunity to make the best use of the frequency spectrum, which is a very limited resource. Frequency management will have a very strong impact. As discussed in Chapter 9, currently CEPT ECC is putting together a new European Table of Frequency Allocations that should be ready around 2008. PMR technical parameters in Europe are expected to become unified and the PMR bands will have to be refarmed. In the United States, the FCC has already made new PMR frequency allocations and is keeping an eye on further demand.

Finally, we should mention that many PMR applications do not need digital speech transmission or the transmission of high data volumes, and these systems can be operated with data transmission based on MPT 1327 and similar systems. In such cases the users are not obliged to migrate from conventional analog to modern digital PMR systems. As mentioned in Chapter 2, surprisingly in Asia a new boom in trunked radio has been experienced for applications with basic requirements due to its low price and ease of availability. Hence, analog PMR will not disappear overnight. It will stay for another 10–15 years, however, it will experience steadily declining subscriber numbers [70, 71].

How might the transition from analog PMR to digital come about? This transition has already taken place in public systems—it occurred roughly a decade earlier and very quickly. The PMR market only started to accept new digital PMR systems after it was demonstrated in the public domain that they really offered new and improved features that were not available in the former analog systems. One reason might be that professional users didn't risk replacing a smoothly running system with a new immature one. They also have to look very carefully at what they should spend limited budget money on and when they spend it. Hence, in addition to the delay in implementation of digital PMR relative to cellular, the duration of the transition will be considerably longer.

Taking all of this into account it is obvious that PMR has now entered a phase in which it has begun to change rapidly and radically. Hence it is not possible to predict in detail what will happen, however, some general statements can be made [72–76]:

- From a functional point of view, PMR can only partly be replaced by other radio services; only PMR systems grant the user full system control.
- PMR systems are at least as economical as the use of public radio communication systems.

- With PMR on one side and GSM, DECT, and UMTS on the other, although they complement each other, they also partly try to replace each other.
- Standardized digital PMR systems will play an increasing role, while proprietary PMR systems have to fight to maintain their market shares.
- FDMA-based DIIS is not a competitor to trunked or nontrunked TDMA systems, but an important opportunity for niches not viable for "big" digital PMR systems.
- Although in some countries the PMR market size is decreasing, it is increasing in others.
- Expanded, new, and changed frequency allocations will occur.
- Type approval procedures, licensing conditions, and tariffs will change or have already changed.

PMR is inevitable, at least for some specific user groups, because they will stick with PMR for a number of operational and economic reasons. In particular, for public safety applications and others where availability and reliability are crucial, public systems are clearly the second choice. Only two reasons need be given. In the case of a catastrophe, public systems soon get blocked by other subscribers as well as the first responders due to a lack of easily accessible and sufficient priority mechanisms. In contrast, modern digital PMR systems provide multiple priority levels. If there is a prolonged power supply failure due to a major defect at a power plant or in the energy transmission network, the public cellular network breaks down because its emergency power supplies are not capable of sustained failure. Many cell sites even do not have backup supplies because their area can be covered from adjacent cells.

In cases where the economic comparison between PMR and public systems is nearly balanced, the wider range of features may push the decision in the direction of PMR. Without a doubt, some functions can be better served by public mobile radio communication systems. The conclusion to be drawn from all of the considerations mentioned here is that public systems and PMR are not alternatives—instead, they complement each other. In short, PMR is not obsolete but more powerful than ever before. TETRA, iDEN, and other DPMR systems demonstrate this very clearly and their future role cannot be overestimated.

Currently, we see many new wireless initiatives and standards evolving, not all of them successfully. Bluetooth and IEEE 802.11, APCO Project 25/34, BRAN, DAWS, and MESA might be recalled as examples. In PMR it is TETRA Release 2 that will have a new impact around 2005. But what will happen after that? The only fixed point is that the preparation of 4G standards will start around 2004, thus targeting 4G introduction around 2010. This has been announced publicly by the Japanese government. However, nobody knows yet what this system will look like, on which technology it will be based, what its main characteristics and features will be, and if it will be a single worldwide common standard. Currently, some 4G researchers are concentrating on improved WCDMA, while others are concentrating on OFDM. Will it finally take most of the PMR features on board?

In Europe PMR will be dominated in the immediate future by TETRA. Other PMR systems will play a minor role. Many proprietary systems will become increasingly isolated and will most likely suffer from decreasing market shares.

However, well-specified digital PMR systems such as iDEN and TETRAPOL are expected to perform much better in international markets.

Eight assertions on the future development of PMR are offered below. These are the consequences derived from the technological trends, the influence of other radio communication systems, and the changing regulatory environment:

1. PMR users will partially migrate to GSM and 3G systems such as UMTS.
2. PMR will be partially replaced by DECT, RLANs, WLANs, and MSSs.
3. A migration from analog to digital PMR systems will take place.
4. Existing analog trunked radio systems will largely move to TETRA.
5. New PMR markets will appear due to high volumes of equipment with low prices and simplified licensing. These markets will include PMR 446 and FRS or new applications (e- and m-commerce, mobile Internet, VoIP) and new standards (DIIS).
6. Digitalization will also happen in the low-end PMR market segments.
7. In the future digital PMR systems will offer improved coverage and increased data rates (APCO Project 25/34, MESA, TETRA Release 2).
8. Analog PMR (for example, MPT 1327) will survive for some time for applications not requiring comprehensive data or digital voice transmission.

Without a doubt, the *main* competitors of PMR are the public cellular systems. This is also true for 3G, but today its impact on PMR is difficult to predict. Many users and applications will benefit from the use of PMR *and* public cellular systems. Hence, there should be a buoyant market for multistandard, multiband equipment. Consequently, system integration may well happen not in the networks but in the hands of the users who in the future will have access to different public cellular and PMR systems via such a radio terminal. To what extent such versatile equipment will in the end provide PMR services will depend on the market demand. At any rate, PMR and cellular systems will coexist like cars and motorbikes.

In the past the steam engine was replaced by the electric motor. Similarly, PMR may become obsolete somewhere in the future but not before a superior and more flexible and comprehensive mobile radio communication system becomes available. Such a system has not yet appeared on the horizon. Therefore, it is permissible to make the following prediction:

> Due to its outstanding properties and features, the existence of PMR will be justified until new mobile communication systems or a combination of different means will cover, without any gaps, all of the requirements that modern digital PMR systems fulfill.

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Acronyms and Abbreviations

π/4-DQPSK	π/4-Shift Differential Quaternary or Quadrature Phase Shift Keying (TETRA)
π/8-DQPSK	π/8-Shift Differential Quaternary or Quadrature Phase Shift Keying (GSM-EDGE)
16QAM	QAM system with 16 modulation states (iDEN)
1TDMA	TDMA system with 1 time slot per frame, identical to a slotted FDMA system such as DIIS
2.5G	2.5 generation mobile, also known as 2+
2G	Second generation mobile
3FSK	Three-level frequency shift keying
3G	Third generation mobile
3TDMA	TDMA system with three time slots per frame
4FSK	Four-level frequency shift keying
4G	Fourth generation mobile
4GFSK	Gaussian-filtered four-level frequency shift keying
4TDMA	TDMA system with four time slots per frame, for example, TETRA V+D
6TDMA	TDMA system with six time slots per frame, for example, iDEN
8-PSK	Eight-level phase shift keying
8TDMA	TDMA system with eight time slots per frame, for example, GSM
A	Mobile data port (APCO 25), BSC-GMSC interface (GSM)
AACH	Access assignment channel (TETRA)
Abis	BSC-BTS interface (GSM)
AC	Authentication center (GSM), alternate current
ACELP	Algebraic code excited linear prediction (TETRA)
ACK	Acknowledgment
ACP	Adjacent channel power
ACR	Adjacent channel rejection
ACROPOL	TETRAPOL system for Public Safety
ACTS	Advanced Communication Techniques and Services (EU research program)

ADC, A/D	Analog to digital (converter)
ADPCM	Adaptive differential pulse code modulation (DECT)
AF	Audiofrequency
AFC	Automatic frequency control
AGC	Automatic gain control
AI	Air interface
ALOHA	Channel access procedure
AM	Amplitude modulation
AMBE, AME	Advanced multiband excitation
AMPS	Advanced Mobile Phone Service (United States, 800-MHz band)
AMR	Adaptive multirate speech coder
AMSC	American Mobile Satellite Corporation
ANSI	American National Standards Institute (United States)
APCO 16	Association of Public Safety Communication Officials, Project 16
APCO 25	Association of Public Safety Communication Officials, Project 25
ARDIS	Advanced Radio Data Information Service (forerunner of MODACOM)
ARQ	Automatic repeat request, for example, for retransmission of lost messages
ASCI	Advanced speech call items (GSM not ASCII!)
ASCII	American Standard Code for Information Interchange
ASK	Amplitude shift keying
ASTRO	Proprietary digital PMR system from Motorola
ASTRO 25	ASTRO equipment for APCO 25
ATM	Asynchronous transfer mode
AVL	Automatic vehicle location
AWGN	Additive white Gaussian noise (used in propagation models for a static radio channel)
B	GMSC-VLR interface (GSM)
BAK	Betriebsartenkennzeichen (mode of operation identifier, ZVEI)
BBK	Broadcast block (TETRA)
BC	Base control (APCO 25)
BCCH	Broadcast control channel, a logical channel carrying system information broadcast from the system control to all other members of the system (TETRA)
BCH	Bose-Chaudhuri-Hocquenghem Code
BER	Bit error rate, ratio of received erroneous bits to the total number of received bits

BFSK	Binary frequency shift keying
BiCMOS	Bipolar CMOS (semiconductor integration technology)
BIIS 1200	Binary Interchange of Information and Signaling at 1,200 bit/s specified in ETS 300 230
BKN	Block number (TETRA)
BLCH	BS linearization channel (TETRA)
BLER	Block error rate
Bluetooth	Indoor mobile RLAN at 2.5 GHz
BN	Base network (TETRAPOL)
BNCH	Broadcast network channel (TETRA)
BOS	German authorities and organizations responsible for public safety (Behörden und Organisationen mit Sicherheitsaufgaben)
BPSK	Binary phase shift keying
BRAN	Broadband Radio Access Network (ETSI Project)
BS	Base station, a base station can also be a repeater or a fixed terminal with dispatcher
BSC	Base station controller (GSM)
BSCH	Broadcast synchronization channel, broadcast signaling channel (TETRA)
BSS	Base station subsystem (GSM)
BT	Bandwidth-symbol time product
BTS	Base transceiver station (GSM)
BU 50	Bad urban area at 50 km/hr (propagation scenario)
BW	Bandwidth
C 450	Net C in 450-MHz band (first German cellular system; C 900 has been planned but never been realized)
C	GMSC-HLR interface (GSM), control plane (TETRA)
C4FM	Compatible four-level FM (APCO 25)
CAI	Common air interface (APCO 25)
CB	Correction uplink burst, citizen band (27 MHz)
CBS	Common base station
CC	Convolutional code, country code, call control (TETRA)
CCCH	Common control channel (TETRA)
CCH	Control channel (TETRA)
CCIR	Comité Cosultatif Internationale des Radiocommunications (now ITU-R)
CCITT	Comité Cosultatif Internationale Téléphonique et Télégraphique (now ITU-T)
CCR	Cochannel rejection
CD	Compact disc
CDMA	Code division multiple access

CDPD	Cellular Digital Packet Data (United States)
CEC	Commission of the European Community (now EC)
CELP	Code excited linear prediction
CEPT	Conférence Européenne des Postes et des Télécommunications
CEPT ECC	CEPT Electronic Communications Committee
CEPT ERC	CEPT European Radio Committee (now ECC)
CEPT ERO	CEPT European Radiocommunications Office (now merged with ETO)
CEPT ETO	CEPT European Telecommunications Office
CEPT SE	CEPT Spectrum Engineering (working group)
CF	Continuous frequency
CFM	Continuous frequency modulation
CHEKKER	German MPT 1327 version
cHTML	Simplified HTML version
CLCH	Common linearization channel (TETRA)
CLNS	Connectionless network service (TETRA)
CMCE	Circuit mode control entity (TETRA)
CMIP	Common Management Information Protocol
CMOS	Complementary metal-oxide semiconductor
COGNITO	Mobile data transmission system in the United Kingdom
CONS	Connection oriented network service (TETRA)
CP	Control physical channel, continuous phase
CP-2GFSK	Continuous phase binary Gaussian frequency shift keying (also known as CP-BGFSK)
CP-4GFSK	Continuous phase quaternary Gaussian frequency shift keying (also known as CP-QGFSK)
CP-BFSK	Continuous phase binary frequency shift keying
CP-GFSK	Family of continuous phase Gaussian frequency shift keying modulation schemes with different numbers of modulation levels (and BT products)
CPM	Continuous phase modulation
CP-mFSK(nRC)	Family of continuous phase frequency shift keying modulation schemes with different numbers m of modulation levels and differing pulse base line lengths (nT_S) and raised cosine shaping (RC)
CPU	Central processing unit
CQPSK, QPSK-C	Compatible QPSK (APCO 25)
CRC	Cyclic redundancy check
CRD	Carrier ramping down (DIIS)
CRU	Carrier ramping up (DIIS)
CS	Channel status, channel status telegram (DIIS)

CSD	Circuit switched data (GSM)
CSMA	Carrier sense multiple access
CSM/CA	Carrier sense multiple access with collision avoidance
CSM/CDA	Carrier sense multiple access with collision detection
CT	Cordless telephone
CTCSS	Continuous tone controlled squelch system (subaudio signal)
CTS	Cordless telephone system, channel tracking sequence (DIIS, 7 pilot symbols)
CUG	Closed user group
CVSD	Continuously variable slope delta modulation
D	Duplex mode operation, HLR-VLR interface (GSM)
DAB	Digital audio broadcasting
DAC, D/A	Digital-to-analog converter
D-AMPS	Digital Advanced Mobile Phone Service (United States, IS-54, IS-136, and IS-136+)
DAWS	Digital Advanced Wireless Services
dBμV	Decibels referred to 1 μV = 0 dBμV
dBc	Decibels referred to carrier power
DBCPM	Differential binary continuous phase modulation
dBi	Decibels referred to the isotropic radiator
dBm	Decibels referred to 1 mW = 0 dBm
DC	Dispatch center (TETRAPOL)
dc	Direct current
DCA	Dynamic channel allocation
DCS	Digital controlled squelch, Digital Communication System (Europe, identical with GSM-1800)
DCS 1800	Digital Communication System for 1,800 MHz (now GSM-1800)
DECT	Digital enhanced cordless telephone
DES	Data encryption standard
DIIS	Digital Interchange of Information and Signaling
DISCO	DPMR system from Ascom/Bosch (SR 440)
DISCUS	Proposed brand name for DIIS products
DM	Delta modulation, direct mode
DMO	Direct mode operation (TETRA)
DNA	Dynamic number assignment
DP	Dispatch position (TETRAPOL)
DPCM	Differential pulse code modulation
DPMR	Digital professional mobile radio
DPMR 25	Fictitious DPMR system for 25-kHz channel separation
DPMR 446	Proposed digital PMR 446 version

DPSK	Differential phase shift keying
DQPSK	Differential quaternary or quadrature phase shift keying
DRAM	Dynamic random access memory
DRX	Discontinuous reception (GSM)
DS-CDMA	Direct sequence CDMA
DSI	Detailed Spectrum Investigation (CEPT)
DSMA	Digital sense multiple access
DSP	Digital signal processor
DSRR	Digital Short Range Radio
DTE	Data terminal equipment
DTI	Department of Trade and Industry (United Kingdom)
DTMF	Dual-tone multiple frequency (tone dialing)
DTX	Discontinuous transmission (GSM)
DW	Dual watch (TETRA)
E	GMSC-MSC interface (GSM)
EADS	EADS Telecom (formerly Matra Nortel Communications)
E_c	Console interface (APCO 25)
EC	European Commission (formerly CEC)
ECC	See CEPT ERC
ECCH	Extended control channel (TETRA)
ECSD	Enhanced circuit switched data (GSM)
ECTEL	Association of the European Telecommunications and Professional Electronics Industry (now EICTA)
ECTEL EFSG	ECTEL Frequency specialist group
ECTEL MRSG	ECTEL Mobile Radio Specialist Group
ECTEL PMRWG	ECTEL Professional Mobile Radio Working Group
ECTEL TMS	ECTEL Task Group for Mobile Services
E_d	Data host interface (APCO 25)
EDACS	Enhanced Digital Access Communications System (General Electric/Ericsson)
EDGE	Enhanced Data Rates for GSM Evolution
EDSN	EADS Defence & Security Networks
EDT	External data terminal (TETRAPOL)
EEA	Electronic Engineering Association (United Kingdom)
E_f	Base station interface (APCO 25)
EGPRS	Enhanced General Packet Radio Service (GSM)
EIA	Electronic Industries Association (United States)
EIR	Equipment identity register
EIRENE	European Integrated Railway Radio Enhanced Network
EMC	Electromagnetic compatibility
EMC D	EMC Directive 1989/336/EEC

eMLPP	Enhanced multilevel precedence and preemption (GSM, ASCI)
EMS	Enhanced Messaging Service (GSM)
EN	European norm
E_n	Network management interface (APCO 25)
EP	ETSI Project
E-PDO	Enhanced PDO (TETRA)
EQ 200	Equalizer test signal with six equidistant paths at 200 km/hr
ERC	European Radiocommunications Committee (see CEPT ERC)
Erl	Erlang, traffic measure for telephone links
ERM	See ETSI ERM
ERMES	Enhanced (formerly European) Radio Messaging System
ERO	European Radiocommunications Office
ESMR	Enhanced Specialized Mobile Radio (United States)
ESN	Electronic serial number
E_t	PSTN interface (APCO 25)
ETHERNET	LAN standard
ETR	ETSI technical report
ETS	ETSI Telecommunication Standard (now mostly transformed into ENs)
ETSI	European Telecommunications Standards Institute
ETSI ERM RP 02	ETSI Technical Committee ERM Working Group Radio Projects 02
ETSI STC RES 02	ETSI Sub-Technical Committee Radio and Equipment and Systems Working Group 02, later RP 02, now Task Group 32 (TG 32, PMR)
ETSI TC ERM	ETSI Technical Committee Electromagnetic Compatibility and Radio Spectrum Matters
ETSI TC RES	ETSI Technical Committee Radio Equipment and Systems
EU	European Union
EU 15	The current 15 EU member states
F	GMSC-EIR interface (GSM)
FACCH	Fast associated control channel (TETRA)
FCC	Federal Communications Commission (United States)
FCCH	Forward control channel (TETRA)
FDD	Frequency division duplex
FDMA	Frequency division multiple access
FEC	Forward error correction
FER	Frame erasure rate
FET	Field effect transistor
FFSK	Fast frequency shift keying

FH	Frequency hopping
FH-CDMA	Frequency hopping CDMA
FM	Frequency modulation
FM 20	Analog PMR system with 20-kHz channel separation and FM
FMS	Funkmeldesystem, early short message system of the German police (BOS)
FN	Flag for numbering scheme modification
FPLMTS	Future Public Land Mobile Telecommunications System
FR	Full rate (D-AMPS, GSM)
FRS	Family Radio Service (United States)
FS	Forward signaling, forward signaling telegram (DIIS)
FSCH	Forward signaling (logical) channel
FSK	Frequency shift keying
FSS	Full synchronization sequence (DIIS, 30 symbols)
FTP	File Transfer Protocol
FX	Fixed station (APCO 25)
G	Gateway interface (APCO 25)
GaAs	Gallium arsenide (semiconductor)
GDP	Gross domestic product
GERAN	GSM/EDGE Radio Access Network
GFSK	Gaussian-filtered FSK
GMM	Global mobile multimedia
GMSC	Gateway Mobile Switching Center (GSM, TETRA)
GMSK	Gaussian-filtered minimum shift keying
GOS, GoS	Grade of service
GPRS	General Packet Radio Service (GSM)
GPS	Global Positioning System
GSM Pro	Professional GSM equipment (Ericsson)
GSM	Global System for Mobile Communications
GSM-400	Global System for Mobile Communications at 400–500 MHz
GSM-900	Global System for Mobile Communications at 900 MHz
GSM-1800	Global System for Mobile Communications at 1,800 MHz (Europe, also DCS 1800)
GSM-1900	Global System for Mobile Communications at 1,900 MHz (United States, also PCS 1900)
GSM-R	GSM for Railways
GTFM	Generalized tamed frequency modulation
HB	High band (DSRR)
HBS	Home base station (GSM Cordless Telephone Systems)
HDB	Home database (TETRA)

HDLC	High-level data link control
HiFi	High fidelity
HIPERACCESS	High-Performance Radio Access
HIPERLAN	High-Performance Radio Local-Area Network
HIPERMAN	High-Performance Radio Metropolitan-Area Networks
HLR	Home location register
HP	Hand portable
HR	Half rate (D-AMPS, GSM)
HSCSD	High-Speed Circuit Switched Data (GSM)
HT 50, HT 250	Hilly terrain at 50 or 250 km/hr (propagation model)
HTML	Hypertext Markup Language
HTTP	Hypertext Transfer Protocol
I	In-phase signal component, interferer
In	Standardized TETRA interface number $n = 1$ to 6
IBM	International Business Machines
IC	Integrated circuit
ID	Identity of a user, a station or a system
iDEN	Integrated Digital Enhanced Network (formerly MIRS, Motorola Integrated Radio System)
IDN	Integrated data network
IEC	International Electrotechnical Commission
IEEE	Institute of Electrical and Electronics Engineers
IEEE 802.11	IEEE standard for high-speed wireless access
IEEE 1394	IEEE standard for high-speed serial data interface
IEEE SA	IEEE standardization area
IF	Intermediate frequency
IM	Intermodulation
IMBE	Improved multiband excitation
IMC	Integrated multisite controller (EDACS)
IMT-2000	International Mobile Telecommunication System 2000
IN	Intelligent network
INMARSAT	International Maritime Satellite Organization
IOP	Interoperability (TETRA)
IP	Internet Protocol
IPR	Intellectual property right
IPv4	Internet Protocol, version 4
IPv6	Internet Protocol, version 6 (enlarged address space)
IR	Infrared
IRI	Inter-radio switch interface (TETRAPOL)
IRIDIUM	Mobile satellite telephone system (originally 77 LEOs)
IS	Interim standard

IS-54	Digital cellular standard (TDMA, United States)
IS-136	Digital cellular standard (TDMA, United States)
IS-136+	Digital cellular standard (TDMA, United States)
ISDN	Integrated Services Digital Network
ISI	Intersymbol interference, intersystem interface (TETRA)
ISM	Unlicensed industrial, scientific and medical band(s)
ISO	International Standardization Organization
ISSI	Inter-RF subsystem interface (APCO 25)
ITU	International Telecommunication Union
ITU-R	ITU Radiocommunications Sector
ITU-T	ITU Telecommunications Sector
IWF	Interworking function
JDC	Japanese digital cellular system
JPEG	Joint Photographic Image Coding Experts Group
KMC	Key management center (TETRAPOL)
L	Phase impulse length (duration)
L1	Layer 1, physical layer (PHY)
L2	Layer 2, data link layer, consisting of MAC and LINK sublayers
L2A	Layer 2A, MAC, media access and control layer
L2B	Layer 2B, LLC, logical link control layer
L3	Layer 3, network layer
L4	Layer 4, transport layer
L5	Layer 5, session layer
L6	Layer 6, presentation layer
L7	Layer 7, application layer
LABS	Line access base station (TETRAPOL)
LAN	Local-area network
LB	Linearization uplink burst (TETRA)
LCH	Linearization channel (TETRA)
LCIU	Line connected interface unit (TETRAPOL)
LCT	Line connected terminal (TETRAPOL)
LDU	Logical data unit (APCO 25)
LE	Late entry
LED	Light-emitting diode
LEO	Low-Earth orbit (as in a satellite)
LET	Link establishment time
LLC	Logical link control
LMR	Land mobile radio
LO	Local oscillator
LOS	Line of sight

LP, LPF	Lowpass filter
LPC	Linear prediction coding
LR	Location register (TETRA)
LS	Line station (TETRA)
LTCC	Low-temperature cofired ceramic
LTR	Logic trunked radio (E. F. Johnson, United States)
MAC	Medium access control
MAP 27	DTE interface for MPT 1327 radio sets
MASC	MOBITEX Asynchronous Protocol
MCC	Mobile country code
MCCH	Main control channel (TETRA)
MCP/1	MOBITEX Compression Protocol
MDC-4800	Motorola Data Communication Protocol for 4,800 bit/s
MDMO	Managed direct mode operation (TETRA)
MDP	Mobile data peripheral
MDTRS	Mobile Digital Trunked Radio System (now TETRA)
MER	Message error rate
MESA	Mobility for Emergency and Safety Applications (joint ETSI/TIA project)
mFSK	m-ary frequency shift keying
MIPS	Million instructions per second
MIRS	Motorola Integrated Radio System (now iDEN)
M-JPEG	Mobile Joint Picture Expert Group, data compression format
MLB, MLE	MS-BS link control entity, also MLE/MLB (TETRA)
MM	Mobility management
MMI	Man-machine interface
MMS	Multimedia messaging service (GSM)
MNC	Mobile network code
MOBITEX	Mobile Text Transmission System
MOBITEX 8K, II	Mobile Text Transmission II at 8 kbit/s
MODACOM	Motorola Data Communications system
MOS	Mean opinion score, measure for the subjectively perceived speech quality
MoU	Memorandum of understanding
MP3	Motion Picture Expert Group (MPEG) Audio Layer III, widely used format for music file compression
MPAK	MOBITEX Packet Transmission Protocol
MPEG-4	Motion Picture Expert Group, data compression format 4
MPT	Ministry of Posts and Telecommunications (United Kingdom)
MPT 1317	MPT specification for selective calling

MPT 1327	MPT specification for trunked radio
MPT 1343	MPT interface specification for trunked radio
MPT 1352	MPT mobile radio specification for MPT 1327/1343 radios
MPT 1383	MPT specification for Short Range Business Radio
MR	Mobile radio (APCO 25)
MS	Mobile station
MSC	Mobile/main switching center
MSK	Minimum shift keying
MSO	Mobile switching office (iDEN)
MSS	Mobile satellite service, master synchronization sequence (DIIS)
MT	Mobile terminal (TETRA)
MT A	Mobile Telephone A (early analog system, Sweden)
MTP/1	MOBITEX Transport Protocol
MTU	Mobile termination unit (TETRAPOL)
NACK	Nonacknowledged
NADC	North American Digital Cellular System
NATO	North Atlantic Treaty Organization
NCC	Network control center (ARDIS)
ND	Noise plus distortion
NDB	Normal downlink burst (TETRA)
Net C	See C-450
NLOS	No line of sight
NMC	Network management center (MODACOM, TETRAPOL)
NMS	Network management system
NMT	Nordic Mobile Telephone System
NMT-450	NMT System in the 450-MHz band
NMT-900	NMT System in the 900-MHz band
NSS	Network and switching subsystem (GSM)
NT	Network termination
NTIA	National Telecommunications and Information Administration (United States)
NUB	Normal uplink burst
OFDM	Orthogonal frequency division multiplexing
OMC	Operation and maintenance center (GSM)
OQPSK	Offset quaternary or quadrature phase shift keying
OSI	Open Systems Interconnection
OSS	Operation subsystem (GSM)
OTAP	Over-the-air programming
OTAR	Over-the-air rekeying
OVL	Overlap (DIIS)

PA	Power amplifier (usually a transmitting power amplifier)
PABX	Private automatic branch exchange
PAKNET	Mobile data transmission system from Racal, Chubb & Mercury
PAMR	Public access mobile radio
PAS	Publicly available specification (ETSI, TETRAPOL)
PBR	Private Business Radio (United Kingdom)
PC	Personal computer
PCM	Pulse code modulation
PCN	Personal communications(s) network
PCS	Personal Communications Service
PCS 1900	Personal Communications Service at 1,900 MHz (United States, identical with GSM-1900)
PD	Packet data
PDA	Personal digital assistant
PDN	Packet data network
PDO	Packet data optimized (TETRA)
PEI	Peripheral equipment interface (TETRA, formerly TEI, terminal equipment interface)
PEP	Peak envelope power
PHY	Physical layer
PIN	Personal identification number
PLL	Phase locked loop
PLMN	Public land mobile network
PM	Phase modulation
PMM	Personal multimedia
PMR	Professional (formerly, private) mobile radio
PMRWG	See ECTEL PMRWG
PMR 12	Analog PMR system with 12.5-kHz channel separation and PM
PMR 20	Analog PMR system with 20-kHz channel separation and PM
PMR 25	Analog PMR system with 25-kHz channel separation and PM
PMR 446	(Low-cost) Professional/Private Mobile Radio at 446 MHz
PMR 6	Fictitious digital PMR system for 6.25 kHz
POCSAG	Post Office Standardization Advisory Group
PPC G4	PowerPC, fourth generation
PPP	Point-to-Point Packet Data Protocol
PSK	Phase shift keying
PSPDN	Public switched packet data network
PSPP	Public Safety Partnership Project

PSRSC	Public Safety Radio Communication Service (United Kingdom)
PSS	Partial synchronization sequence (DIIS, 15 symbols)
PSTN	Public Switched Telephone Network
PT	Project team
PTM	Point to multipoint
PTN	Private telephone network
PTP	Point to point
PTT	Push or press to talk; Postal Telephone and Telegraph Administration
Q	Quadrature phase component or quaternary signal
QOQAM	Quaternary offset quadrature amplitude modulation
QoS	Quality of service
QPSK	Quaternary or quadrature phase shift keying
QPSK-C	Compatible QPSK (APCO 25)
R&TTE D	Radio and Telecommunication Equipment Directive
RA	Radio Agency (United Kingdom)
RA 200	Rural area at 200 km/hr (propagation scenario)
Racal 25	Racal radios for APCO 25
Radix 40	Data compression method specified in ETS 300 230 (BIIS 1200)
RAM	Random access memory
RBGAN	Regional Broadband Global Area Network (INMARSAT)
RBS	Radio base station (TETRAPOL)
RC	Raised cosine
RCPC	Rate-compatible punctured code
RD-LAP	Radio data link access procedure
RDN	Radio Data Network (United States)
RE	Repeater
RES	See ETSI TC RES
RES 02	See ETSI STC RES 02
RF	Radio frequency
RF/NCP	Radio-frequency network control processor (ARDIS)
RFC	RF subsystem control (APCO 25)
RFG	RF subsystem gateway (APCO 25)
RFS	RF subsystem switch (APCO 25)
RLAN	Radio local-area network
RLC/MAC	Radio link control/medium access control
RLP	Radio Link Protocol
RM	Reed-Muller code
Rn	Interface $n = 1$ to 19 (TETRAPOL)

RNC	Radio network controller (MODACOM)
RNG	Radio network gateway (MODACOM)
ROM	Read-only memory
ROSI	Radio open system interface (MOBITEX)
RP	Repeater, radio project
RP 02	See ETSI STC RP 02
RPCELP	Regular pulse code excited linear prediction
RPE-LTP	Regular pulse excitation—long-term prediction (GSM)
RS	Reed-Solomon code, reverse signaling (DIIS)
RS 232	A standard serial data interface (EIA)
RSSI	Received signal strength indication
RSW	Radio switch (TETRAPOL)
RSWN	Radio switch network (TETRAPOL)
RT	Radio terminal (TETRAPOL)
RTC	Radio traffic channel (traffic measure for one-way channels)
RVE	Reference vector equalization (Securicor Wireless)
Rx	Receiver
S	Simplex mode operation
S/N, SNR	Signal-to-noise ratio
SACCH	Slow associated control channel (TETRA)
SADP	Standalone dispatch position (TETRAPOL)
SAR	Specific absorption rate
SB	Synchronization downlink burst (TETRA)
SCCH	Secondary control channel (TETRA)
SCH	Signaling channel (TETRA)
SCH/F	Full-size signaling channel (TETRA)
SCH/H	Half-size signaling channel (TETRA)
SCH/HD	Half-size downlink signaling channel (TETRA)
SCH/HU	Half-size uplink signaling channel (TETRA)
SD	Semiduplex operation
SDCCH	Standalone dedicated control channel (TETRA)
SDS	Short data service (TETRA)
SER	Symbol error rate
SFH	Slow frequency hopping
SiC	Silicon carbide (semiconductor material)
SiGe	Silicon-germanium (semiconductor material)
SIM	Subscriber identity module
SINAD	Signal plus noise plus distortion to noise and distortion ratio (see also SND/ND)
SmarTrunk	Trunked radio system (SmarTrunk Systems, United States)

SMG	Special Mobile Group (formerly GSM, Groupe Speciale Mobile)
SMR	Specialized Mobile Radio (United States)
SMS	Short message service (GSM)
SNAF	Subnetwork access function (TETRA)
SND/ND	Signal and noise and distortion to noise and distortion ratio (see also SINAD)
SNDCP	Sub-Network Dependent Communication Protocol
SNMP	Simple Network Management Protocol
SOI	Silicon on insulator
SoR	Statement of requirements
SR 440	DPMR system from Ascom/Bosch (for the 440-MHz band)
SR	Satellite receiver (remote BS receiver)
SRBR	Short Range Business Radio
SRRC	Square root raised cosine
SS	Supplementary service
SSB	Single sideband
SSI	Side signaling information, short subscriber identity
SSN	Subslot number (TETRA)
ST	Settling time of the receiver (DIIS, 1 symbol)
STC	Sub-TC (ETSI)
STCH	Stealing channel (TETRA)
STM	Statistical multiplexing
STMA	Statistical multiplexing multiple access
STS	Settling time symbol (DIIS)
SwMI	Switching and management infrastructure
TACS	Total Access Communications System (European version of AMPS in the 900-MHz band)
TAPS	TETRA Advanced Packet Service
TBR	Technical basis for regulations
TC	Trellis code, technical committee (ETSI)
TCH	Traffic channel
TCH/D	Traffic channel for data
TCH/F	Traffic channel at full rate (GSM)
TCH/H	Traffic channel at half rate (GSM)
TCH/S	Traffic channel/speech (TETRA)
TCH/2.4	Traffic channel with 2.4 kbit/s (TETRA)
TCH/4.8	Traffic channel with 4.8 kbit/s (TETRA)
TCH/7.2	Traffic channel with 7.2 kbit/s (TETRA)
TCM	Trellis coded modulation
TCP	Transmission Control Protocol

TCP/IP	Transmission Control Protocol/Internet Protocol
TD-CDMA	Time division CDMA (on UMTS simplex frequencies)
TDD	Time division duplex
TDMA	Time division multiple access
TE	Terminal equipment
TEDS	TETRA Enhanced Data Service
TEI	TETRA equipment identity (TETRA, formerly terminal equipment interface)
TELEDESIC	LEO-based broadband satellite data service ("Internet in the sky")
TETRA	Terrestrial (formerly, Trans-European) Trunked Radio
TETRA-IP	TETRA Internet Protocol
TETRA PDO	TETRA packet data optimized
TETRA V+D	TETRA Voice plus Data
TETRAPOL	Proprietary digital PMR system from EADS Telecom (formerly Matra Nortel Communications)
TFM	Tamed frequency modulation
TG	Task group
TI	Transmitter or terminal interrupt, terminal interface
TIA	Telecommunications Industry Association (United States)
TIA TR-8	TIA Working Group Transmission/Reception 8
TM	Trunked mode
TMO	Trunked mode operation (TETRA)
TP	Traffic physical channel
TTA	Telecommunications Technology Association (Korea)
TTIB	Transparent tone in band (see also RVE)
TTL	Transistor-transistor logic
TU 50	Typical urban area at 50 km/hr (one of the GSM propagation models)
TV	Television
Tx	Transmitter
U	User (plane, TETRA)
UADG	User Access Definition Group (has specified MAP 27)
U_d	Air interface (TETRA DMO)
UDT	User data terminal (TETRAPOL)
UHF	Ultra-high frequency (300–3,000 MHz)
UIC	Union International des Chémins des Fer
U_M	Air Interface (APCO 25, GSM, TETRA TMO)
UMTS	Universal Mobile Telecommunications System
UP	Unallocated physical channel
USB	Universal serial bus (external serial PC data bus)

USDC	U.S. Digital Cellular System
USIM	Universal SIM
UTC	Universal Coordinated Time
UTRA	UMTS Terrestrial Radio Access
UTRAN	UMTS Terrestrial Radio Access Network
V.24/V.28	Standard serial computer interface (RS 232)
V.42bis	ITU-T recommendation for a transmission protocol employing data compression
V+D	Voice plus Data (TETRA)
VAD	Voice activity detection
VBR	Variable-bit-rate speech coding
VBS	Voice broadcast service
V_{CC}	Volts closed circuit
VCO	Voltage controlled oscillator
VDB	Visitor database (TETRA)
VDEW	Vereinigte Deutsche Elektrizitätswerke (German association of electricity suppliers)
V_{EMF}	Volts electromotive force
VGCS	Voice group call service (GSM, ASCI)
VHE	Virtual home environment
VHF	Very high frequency (30–300 MHz)
VLR	Visitor location register
VoIP	Voice over Internet Protocol
VPN	Virtual private network
VSELP	Vector sum excited linear prediction
VTDD	Variable time division duplex
WAP	Wireless Access Protocol
WCDMA	Wideband CDMA (on UMTS duplex frequencies)
WG	Working group
WLAN	Wireless wide-area network
WLL	Wireless local loop (also RLL, radio in the local loop)
WML	Wireless Markup Language
WWW	World Wide Web
X.25 and X.28	ITU-T Recommendations for packet data transmission
X.400	ITU-T recommendation for data transmission
XML	Extensible Markup Language
ZVEI	Zentralverband der Elektrotechnik-und Elektronikindustrie (German industrial association for manufacturers of electrical and electronic equipment)

About the Author

Hans-Peter A. Ketterling studied telecommunications at the Technical University (TU) in Berlin from 1961 to 1968. In 1969 he received his degree as diploma engineer (Dipl.-Ing.) from TU and joined the land mobile radio development department of Standard Elektrik Lorenz AG in Berlin, where he was involved in the development of PMR equipment, and later became the head of this department.

In 1983 he joined the land mobile radio development department of Robert Bosch GmbH in Berlin, where he was responsible for public mobile radio telephones and type approvals. Mr. Ketterling also worked on improving methods of measurement for land mobile radio sets and worked on the design of anechoic chambers and the verification of their properties. He then spent several years on the development of infrastructure equipment for the GSM system and became the technical project manager for the DMCS 900 Consortium. In the 1990s, his work was devoted to new radio systems including TETRA and UMTS.

Mr. Ketterling holds more than 30 patents in the field of land mobile radio communication and has been working for almost 25 years in various national and international industrial and standardization organizations such as CCIR, CEPT, DKE, ECREEA, ECTEL, ETSI, ÖVEI, and ZVEI. Until 1997 he chaired the ECTEL Mobile Radio Specialist Group, the ECTEL UMTS Focus Group, and the ZVEI AK Radio. From 1986 to 1997 he also was vice chairman of the ECTEL PMR Working Group. He has been a member of the IEEE since 1996.

After a period at Motorola in 1997, in 1998 Mr. Ketterling became an independent author in the field of mobile radio and founded Mobile Radio Consulting Berlin, which obtained contracts from several industrial enterprises and national and European organizations, for example, the EC, the ERO, and the German Ministry of Economics. He is involved in the standardization of DIIS, gives lectures and training courses on digital PMR, has written numerous articles on mobile radio topics and a textbook on digital PMR, and has made many presentations at international radio conferences. He can be reached at h.-p.ketterling@t-online.de.

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