

CHAPTER 1

RF/MICROWAVE SYSTEMS

1.1 INTRODUCTION

This book is similar to many well-known texts in the field [1.1–1.13]; they all have a different slant on the same engineering topics, that is, radio-frequency (RF) and microwave circuit design, from an intuitive and engineering point of view. This book was first written in 1982 [1.14] using only linear techniques, which is out of print, and was later augmented in 1990 [1.15] including both linear and nonlinear techniques. The second extensively updated edition published in 2005 [1.16] especially considered particularly the advances in microwave computer-aided design (CAD). Another 16 years later, we present this third edition that reflects the recent advancements in microwave semiconductor technology, CAD capabilities, and microwave systems in general.

The audience is both graduate students of RF/microwave courses and practicing engineers in this industry. We expect you have already mastered the fundamentals (component definitions for amplifiers, oscillators, and mixers; two-port network theory; power gains; Smith chart matching; direct current (DC) biasing; etc.), but these are also included in the text for careful review. Prior to using the CAD tools, the practicing engineer should be able to do general RF/microwave problems with a calculator and Smith chart. Even though circuit design and optimization today requires detailed nonlinear calculations that are in their complexity only feasible to be solved by CAD, we are still far from automated circuit design, layout or floorplanning in the microwave and millimeter wave field. It should also be considered that the simulation accuracy depends on the accuracy of the component models, which is a research topic itself. A microwave component is only useful if either the manufacturer or a third-party modeling company like Modelithics provide a reliable model. Anyway, microwave and millimeter circuit design, especially concerning

integrated circuits, still requires a skilled engineer using the CAD tools. Commonly, it is considered that an monolithic microwave circuit (MMIC) designer requires about ten years in the business to be considered really experienced.

A book of this magnitude must begin with a brief history of the nineteenth and twentieth-century communications achievements, which are tabulated in Table 1.1 [1.17]. While key scientific events occurred over a century ago, the present digital wireless era was demonstrated in 1962 and introduced commercially in 1988. From a solid-state device perspective, the key events were the inventions of the bipolar junction transistor (BJT) and GaAs metal-semiconductor field-effect transistor (MESFET), the predecessors of today's heterojunction bipolar transistors (HBTs) in SiGe, GaAs, or InP and high electron-mobility transistors (HEMTs) in GaAs, InP, or GaN technologies. The Ge BJT was quickly replaced by the Si BJT due to temperature considerations and the discovery of SiO₂ (the planar process). Bell Labs accidentally discovered the Ge BJT while attempting to build a variable resistor, or field-effect transistor (FET).

The first solid-state X-band radar was developed by Texas Instruments during the period 1966 to 1970 under contract to Wright Patterson Air Force Base [1.18]. This contract, which was called the Microwave Electronics Radar Applications (MERA) program, revolutionized microwave engineering, providing new insights into the use of hybrid microwave integrated circuit (MIC) construction using microstrip transmission lines on alumina, after determining silicon would never succeed in this role. This is a phased array antenna which is pointed by the phase shifters preceding the 1-W transmitters, 640

TABLE 1.1 Historical Events in Communications

Event	Names	Year
Maxwell's equations	James Clerk Maxwell	1873
Invention of telephone	Alexander Graham Bell	1876
Validation of Maxwell's theory	Heinrich Hertz	1891
Transatlantic communications	Guglielmo Marconi	1901
Galena (lead sulfide) detector	J. C. Bose	1901 (patent filed)
Superheterodyne receiver	Edwin H. Armstrong	1917
X-band radar	MIT Radiation Labs	1942
Invention of transistor	John Bardeen, William Brittain, and William Shockley, Bell Labs	1947
Digital voice transmission	ATT	1962
Invention of GaAs metal-semiconductor field-effect transistor (MESFET)	C.A. Mead Cal Tech	1965
First solid-state X-band radar	Texas Instruments	1970
First GaAs MESFETs in satellites	SPAR/CRC	1975
Analog cellular radio	ATT/Motorola	1983
Digital cellular radio	ATT	1988
Digital personal communication service (PCS) radio code division multiple access (GSM/CDMA)	Europe/Qualcomm	1993
Wide-band code-division multiple access (WCDMA) 4G CDMA networks	Mobile Internet	2000

of them. This was replaced in the 1990s by GaAs MESFET modules by Raytheon and Texas Instruments for the BMDO (Ballistic Missile Defense Operation for ground-based radar [GBR]) when 60 000 units were shipped about 1996 [1.19].

The first introduction of GaAs MESFETs into space deserves some comments. At this point in time (1973), two major companies were producing devices with about 2 μm gate lengths, Fairchild and Plessey, at a price of about US\$500 each. A satellite was about to be launched in 1975 by SPAR, which hired Communications Research Center CRC (Canada), which selected the new MESFETs from both suppliers, to design the low-noise amplifier (LNA). The first purchase of space-qualified GaAs MESFETs was 23 devices for US\$40 000 from Fairchild. These were all burned out in two months due largely to electrostatic discharge (ESD) problems, so they purchased an entire wafer next. To shorten the story, which is documented in [1.20, 1.21], both suppliers provided transistors for five- or six-stage amplifiers (see Fig. 1.1) with 26 dB gain and 10 dB noise figure at 12 GHz (300 MHz bandwidth [BW]), and two satellites were launched in 1975; the project was a complete success, with a lifetime of about 3 years [1.22] for the Plessey amplifier; the Fairchild amplifier never turned on due to switching problems in the satellite. The circuits were made on 25-mil polished alumina with TiW/Au metal 6 μm thick. The resistance of the TiW was 50 Ω/square . Some photographs of these amplifiers, which were used in the world's first direct broadcast TV satellite, which was launched in Australia in late 1975, are shown in Figure 1.1.

Turning to cellular telephone, analog cellular systems introduced in 1984 are commonly referred to as first-generation systems. The first commercial digital systems such as GSM, personal digital cellular (PDC), CDMAOne (IS-95), and US TDMA (IS-136), were second-generation systems. These systems served both voice communications and introduced other services such as text messaging and access to data networks. Beginning with third-generation systems, the main focus shifted from voice calls to multimedia content. Although these devices seem to be dominated by their computing power and screen size, it were the advancements in wireless data rate in combination with other wireless interfaces like WiFi, Bluetooth, and GPS, that enabled the killer applications that were location services and social media access. WCDMA technology emerged as the most widely adopted third-generation air interface. Its specification has been created in the Third Generation Partnership Project (3GPP). Within 3GPP, WCDMA is called UTRA (Universal Terrestrial Radio Access), FDD (Frequency Division Duplex), and TDD (Time Division Duplex). The differences between WCDMA FDD and WCDMA TDD are explained in [1.23].

The importance of the RF circuit design for the technical advancement of the systems relying on them should be obvious, but if this issue is overlooked, it may end in tremendous cost. The introduction of the 3G cellular service is an example for such a disaster. In Germany, for example, telecom companies bought the right to use 3G frequencies in the year 2000 at a total cost of 98.9 billion Deutschmarks, roughly US\$50 billion. But it took until 2004 for affordable mobile devices to be available on the market. Two consortia did not reach that state: they gave up their licenses in 2003. What happened? When the 2G standard was designed, it was quite a bit of a challenge to build an affordable mobile device capable of digitally transmitting phone calls at frequencies around 1–2 GHz at all. One decided therefore to use narrowband signals, and the Gaussian minimum shift keying (GMSK) digital modulation scheme that transports few bits per second but provides a constant envelope. It was still advanced technology at that time, but designing a high-efficiency linear amplifier is feasible for narrowband signals

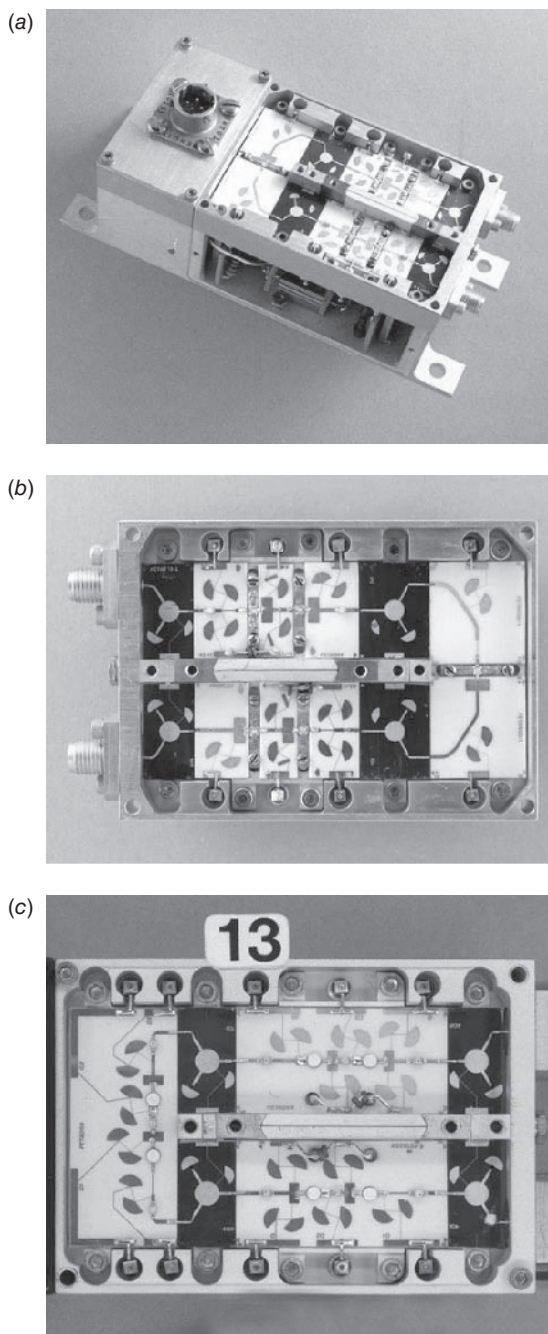


FIGURE 1.1 First GaAs MESFET amplifiers for 12-GHz satellite application for direct broadcast TV. (a) Amplifier module with opened package. (b,c) Close-up of amplifiers with package lid removed. (Courtesy of Rene Douville [1.20–1.22].)

of a constant amplitude. When 3G was designed, a lot more computer power was available in the phones, and also the integration of the RF part was promising. One thereby decided to move to a WCDMA coding scheme that provides a number of advantages like much higher data rate, flexible channel assignment, and tolerance against multipath propagation (echoes). Unfortunately, the RF semiconductors were not yet there, and the 2G amplifier concepts were not suitable for broadband WCDMA signals that show an envelope that resembles a noise signal. Consequently, the modulation scheme chosen for the fourth generation mobile standard, long-term evolution (LTE), again considers the waveform to be amplified in the mobile handset. It allows the handset to use a single-carrier frequency-division multiple access (SC-FDMA) scheme that reduces the complexity on the side of the mobile, which is compensated by higher data processing efforts on the base station side. The higher data rates in 4G systems and beyond also require not only additional bandwidth and better modulation schemes. They also introduce carrier aggregation (parallel use of frequency bands) and on the side of the base station multiple-input multiple-output (MIMO) and spatially selective transmission. The upcoming 5G standard will for the first time also include mm-bands.

Even an inexpensive mobile phone today includes a large number of digital two-way radios, one for each band, and different types for the different transmission standards like 3G, 4G, WiFi, and Bluetooth. But for a long time, the private usage of RF transmission was one-way: radio and TV receivers.

Almost 100 years ago, the simplest radio receiver was the crystal radio receiver shown in Figure 1.2, which uses no battery. The diode (or crystal) demodulates the amplitude-modulated (AM) carrier to excite the headphones into sound. This circuit is also called “the foxhole radio” because of its use during World War II. In this case, the components were the antenna (a length of wire), the LC tank, the detector, which was made from pencil lead touching a Gillette razor blade, and a headphone set. The capacitor of 1000 pF in parallel with the headphones is an RF ground for the carrier frequency. The headphones detect the envelope of the received signal, which is the desired information. This type of receiver is the simplest of all, and there are numerous web sites which can sell you one for your evaluation.

A similar invention which also uses no battery is the telephone [1.24]. The detector is a diaphragm which transmits sound to the human ear drum, which has a threshold sensitivity of one hydrogen atom displacement [1.25], where the frequency is roughly 5 kHz.

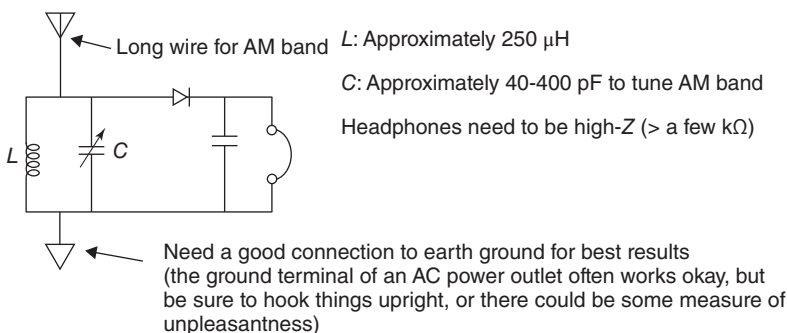


FIGURE 1.2 Crystal radio receiver or foxhole radio. (From Lee [1.13].)

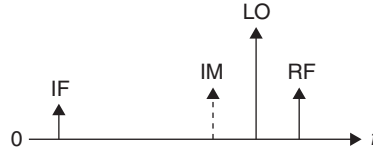


FIGURE 1.3 Frequency spectrum of radio receiver.

The frequency spectrum of a receiver is shown in Figure 1.3, where the image signal may also produce an unwanted intermediate frequency (IF) output, so the image should be filtered. The image signal is the mirror image of the desired RF signal. The radio is a tuned resonant tank circuit at the carrier frequency, which maximizes the input voltage to the heterodyne receiver shown in Figure 1.4. The incoming signal is converted to a lower IF by the local oscillator (LO), where the pertinent mathematics is

$$\begin{aligned}\cos \alpha \cos \beta &= 0.5 \cos(\alpha - \beta) + 0.5 \cos(\alpha + \beta) \\ &= 0.5 \cos(\text{IF}t) + 0.5 \cos(\text{RF}t + \text{LO}t)\end{aligned}\quad (1.1)$$

where α and β are the RF and LO frequencies. The frequency spectrum is shown in Figure 1.3, where the image signal may also produce an unwanted IF output, so the image should be filtered. The image signal is the mirror image of the desired RF signal. The basic heterodyne receiver invented by Armstrong in 1917 is given in Figure 1.4. The modulated carrier is amplified, converted to an IF, demodulated, and amplified at baseband (audio). This gives a single tuning control (the LO) and allows high gain and selectivity at the IF. A superheterodyne receiver has two (or more) mixers, so the frequency is converted once or twice to a lower frequency. Most of the gain is done at the first or second IF, where the cost is generally lower. Most electrical engineers have worked in various aspects of

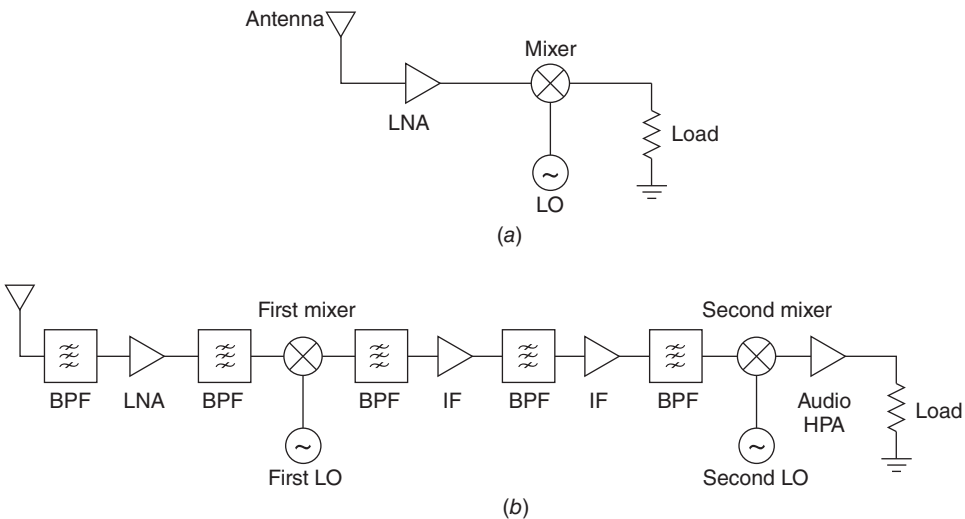


FIGURE 1.4 (a) Heterodyne receiver, single-conversion superheterodyne receiver, (b) double-conversion superheterodyne receiver [1.29].

RF or microwave design. This book is addressed to the designers of these circuits. The circuits are organized into three frequency ranges:

RF	1 MHz (or less) to 1 GHz
Microwave	1–30 GHz
Millimeter wave	30–300 GHz
Terahertz	300 GHz and above

The word wireless was used by Marconi in 1901, and it reoccurred as a replacement for the word radio in about 1991. The design techniques tend to be different for these four groups, but there are many similarities. A typical RF CAD package may be used for all three groups. Another summary of wireless applications is given in Table 1.2 [1.27]. These applications include all frequency groups as well as communications, radar, navigation, remote sensing, RF identification, broadcasting, automobiles and highways, sensors, surveillance, medical, and astronomy and space exploration.

At low frequency, we use lumped components with transistors and diodes as needed, that is, R , L , and C . When the components become about $\lambda/8$ long, about 500 MHz to

TABLE 1.2 Wireless Applications [1.27]

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1. Wireless communications: cellular telephones, space, long-distance, cordless phones, mobile, WiFi, aircraft, marine, citizen's band (CB) radio, vehicle, satellite, global, etc.
 2. Radar (standing for radio detection and ranging): airborne, marine, vehicle, collision avoidance, weather, imaging, air defense, traffic control, police, intrusion detection, weapon guidance, surveillance, etc.
 3. Navigation: microwave landing system (MLS), global positioning system (GPS), beacon, terrain avoidance, imaging radar, collision avoidance, auto-pilot, aircraft, marine, vehicle, etc.
 4. Remote sensing: Earth monitoring, meteorology, pollution monitoring, forest, soil moisture, vegetation, agriculture, fisheries, mining, desert, ocean, land surface, clouds, precipitation, wind, flood, snow, iceberg, urban growth, aviation and marine traffic, surveillance, etc.
 5. RF identification: security, antitheft, access control, product tracking, inventory control, keyless entry, animal tracking, toll collection, automatic checkout, asset management, etc.
 6. Broadcasting: amplitude- and frequency-modulated (AM, FM) radio, TV, direct broadcast satellite (DBS), universal radio system, etc.
 7. Automobiles and highways: collision warning and avoidance, GPS, blind-spot radar, adaptive cruise control, autonavigation, road-to-vehicle communications, automobile communications, near-obstacle detection, radar speed sensors, vehicle RF identification, intelligent vehicle and highway system (IVHS), automated highway, automatic toll collection, traffic control, ground penetration radar, structure inspection, road guidance, range and speed detection, vehicle detection, etc.
 8. Sensors: moisture sensors, temperature sensors, robotics, buried-object detection, traffic monitoring, antitheft, intruder detection, industrial sensors, etc.
 9. Surveillance and electronic warfare: spy satellites, signal or radiation monitoring, troop movement, jamming, antijamming, police radar detectors, intruder detection, etc.
 10. Medical: magnetic resonance imaging, microwave imaging, patient monitoring, etc.
 11. Radio astronomy and space exploration: radio telescopes, deep-space probes, space monitoring, etc.
 12. Wireless power transmission: space-to-space, space-to-ground, ground-to-space, ground-to-ground power transmission.
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1 GHz, we may add transmission line components (usually microstripline) in addition to lumped components. The transition from lumped elements to distributed elements will be covered in Chapter 2. When the free-space wavelength becomes less than 1 mm (millimeter wave), the designers are usually forced to use distributed transmission line elements wherever possible. Different forms of transmission are also used depending on the application [1.28], such as waveguide, slotline, coplanar waveguide, inverted microstripline, and suspended microstripline [1.26]. It is useful to keep in mind that it requires two wires in a homogeneous medium for transversal electromagnetic (TEM) wave propagation. Other types of waveguides allow for transverse-electric (TE) and transverse-magnetic (TM) mode propagation which is dispersive, which means that the line properties change with frequency. However, microstrip and coplanar lines are considered quasi-TEM as the field component in the propagation direction is not dominant for a wide range of frequencies. The transmission line models provided by RF circuit simulators account for the dispersive effects. At least, the simulator will issue a warning when simulating outside the frequency range for which the model is valid.

Transmission lines (see Figure 1.5) may be modeled as in Figure 1.6, which leads to the telegrapher equations, which are a time-domain description of the line:

$$\frac{\partial V(z, t)}{\partial z} = -RI(z, t) - L \frac{\partial I(z, t)}{\partial t} \quad (1.2)$$

$$\frac{\partial I(z, t)}{\partial z} = -GV(z, t) - C \frac{\partial V(z, t)}{\partial t} \quad (1.3)$$

Transmission lines

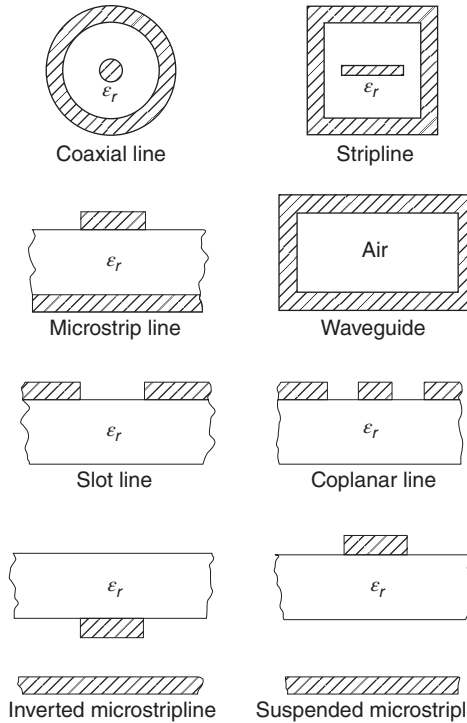


FIGURE 1.5 Geometry for microwave transmission.

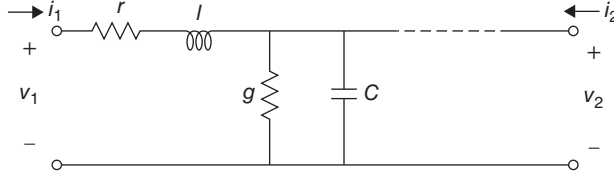


FIGURE 1.6 Lumped-element equivalent circuit of transmission line.

For sinusoidal steady-state conditions, this may be simplified to

$$\frac{dV(z)}{dz} = -(R + j\omega L)I(z) \quad (1.4)$$

$$\frac{dI(z)}{dz} = -(G + j\omega C)V(z) \quad (1.5)$$

which is noted to be very similar to Maxwell's curl equations:

$$\nabla E = -j\omega \times H \quad (1.6)$$

$$\nabla H = j\omega \times E \quad (1.7)$$

Combining the telegrapher equations leads to

$$\frac{d^2 V(z)}{dz^2} = \gamma^2 V(z) \quad (1.8)$$

$$\frac{d^2 I(z)}{dz^2} = \gamma^2 I(z) \quad (1.9)$$

where $\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$ is the complex propagation constant, which is a function of frequency. A traveling-wave solutions can be found as the superposition of waves traveling in forward and reverse direction, indicated by the superscript (+) and (−):

$$V(z) = V_0^+ e^{-\gamma z} + V_0^- e^{\gamma z} \quad (1.10)$$

$$I(z) = I_0^+ e^{-\gamma z} + I_0^- e^{\gamma z} \quad (1.11)$$

With a few more steps, we may obtain the voltage waveform in the time domain as

$$V(z) = |V_0^+| \cos(\omega t - \beta z + \phi^+) e^{-\alpha z} + |V_0^-| \cos(\omega t + \beta z + \phi^-) e^{\alpha z} \quad (1.12)$$

where $\phi^\pm$ is the phase angle of the complex voltage $V_0^\pm$. The wavelength on the line is

$$\lambda = \frac{2\pi}{\beta} \quad (1.13)$$

and the phase velocity is

$$v_{ph} = \frac{\omega}{\beta} \quad (1.14)$$

where the actual time delay must be calculated using the group velocity, defined by

$$v_{\text{gr}} = \frac{d\omega}{d\beta} \quad (1.15)$$

For a TEM wave, these two velocities are the same.

Some useful design aids for the microstripline case are given in Figures 1.7 and 1.8, which are the solutions provided by Wheeler in 1965 for wide and narrow lines [1.29]. These curves allow you to calculate the characteristic impedance and effective dielectric constant for the TEM mode and hence to draw the mask for your design. Many examples of this procedure will be given in this book.

There are many books which treat the solution to transmission line problems, but they all eventually lead to the Smith chart, which is the primary circuit design tool for these problems. The use and applications of the Smith chart will be described in detail in Chapter 5, where we find that graphical solutions provided by Smith charts are much more intuitive and faster than the analytic solutions provided from algebra.

Maxwell's equations become more useful as the frequency range exceeds 30 GHz, and the solutions to three-dimensional (3D) electromagnetic (EM) problems have now

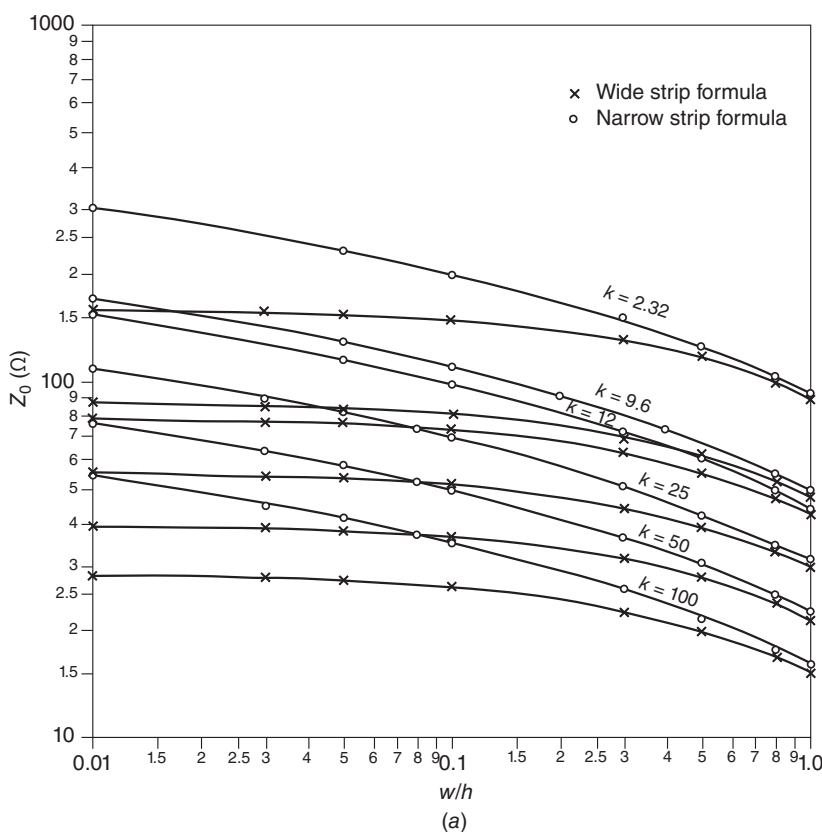


FIGURE 1.7 (a) Characteristic impedance for (i) narrow and (ii) wide microstripline and (b) characteristic impedance for wide microstripline. (Modified from Wheeler [1.30].)

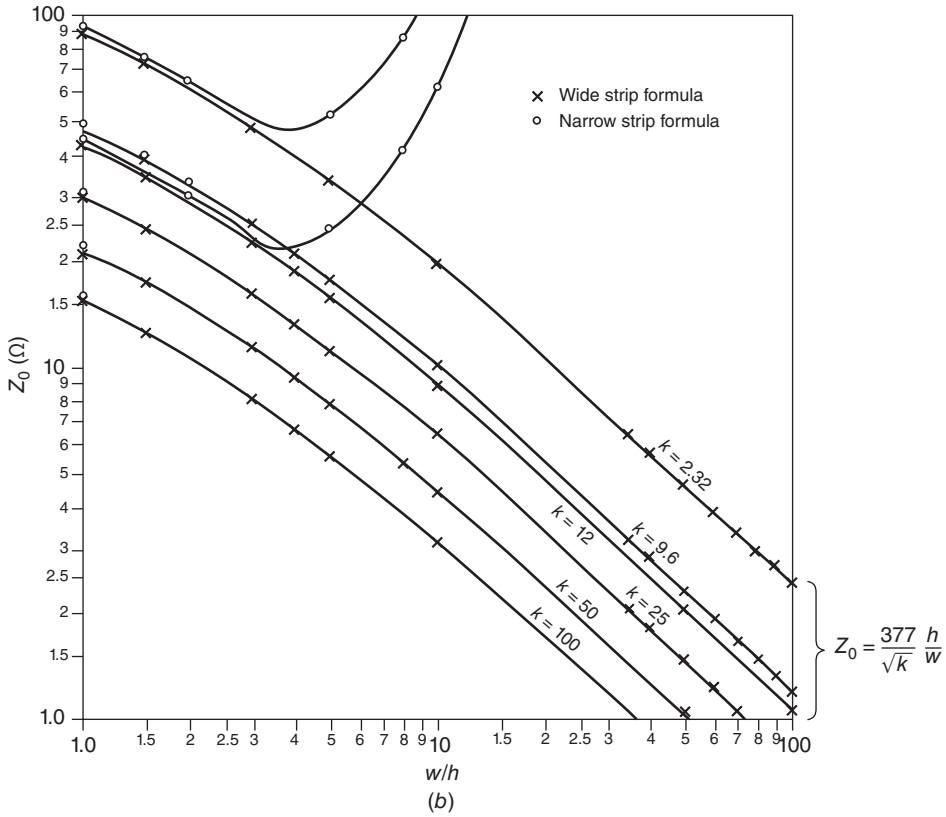


FIGURE 1.7 (Continued)

become a valuable engineering design tool that are commonly even integrated with microwave circuit simulation packages that allow for fast EM simulation of a layout structure. Both $2\frac{1}{2}$ D and 3D solutions exist. $2\frac{1}{2}$ D is faster but 3D is more accurate.

1.2 MAXWELL'S EQUATIONS

All forms of modern communications are based upon Maxwell's equations, which are treated in numerous textbooks [1.26, 1.29–1.31]. These four equations are

$$\nabla \times \vec{E} = -\frac{\partial \vec{B}}{\partial t} \quad (1.16)$$

$$\nabla \times \vec{H} = \frac{\partial \vec{D}}{\partial t} + \vec{J} \quad (1.17)$$

$$\nabla \cdot \vec{D} = \frac{\rho}{\epsilon} \quad (1.18)$$

$$\nabla \cdot \vec{B} = 0 \quad (1.19)$$

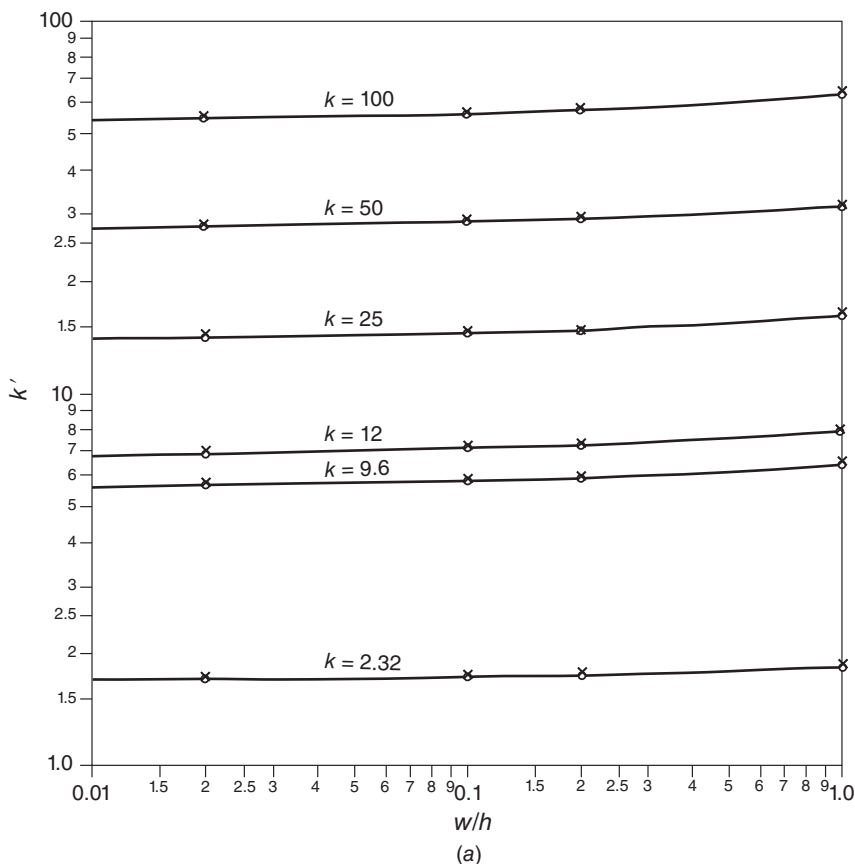


FIGURE 1.8 (a) Effective dielectric constant for (i) narrow and (ii) wide microstripline and (b) effective dielectric constant for a wide microstripline. (Modified from Wheeler [1.29].)

These equations are also known as Faraday's law, Ampere's law, and Gauss's laws.

An interesting interpretation of these laws is shown in Figure 1.9, where physical examples of four antennas are shown to obey Maxwell's equations. The first equation is illustrated in Figure 1.9a, where a circular alternating current (ac) electric field requires an ac magnetic field. This is known as a loop antenna or electric antenna. The dual is shown in Figure 1.9b, which is a circular loop antenna or a magnetic antenna. Again, the circular ac magnetic field requires an ac electric field. The half-wavelength dipole antenna illustrates Gauss's law for electric charge, as given in Figure 1.9c. The electric fields terminate on charges only. The distribution of fields along the length of the dipole give a current, or $\vec{H}$, maximum at the center and a voltage, or $\vec{E}$, maximum at the open-circuited ends of the antenna. The final equation is illustrated in Figure 1.9d by a ferrite rod antenna. An air coil does not disturb the uniform $\vec{H}$ field, but a high micro ferrite inside the coil will attract the $\vec{H}$ fields inside the ferrite but there is no source of magnetic fields. Thus, the four Maxwell equations have been demonstrated by simple antenna examples.

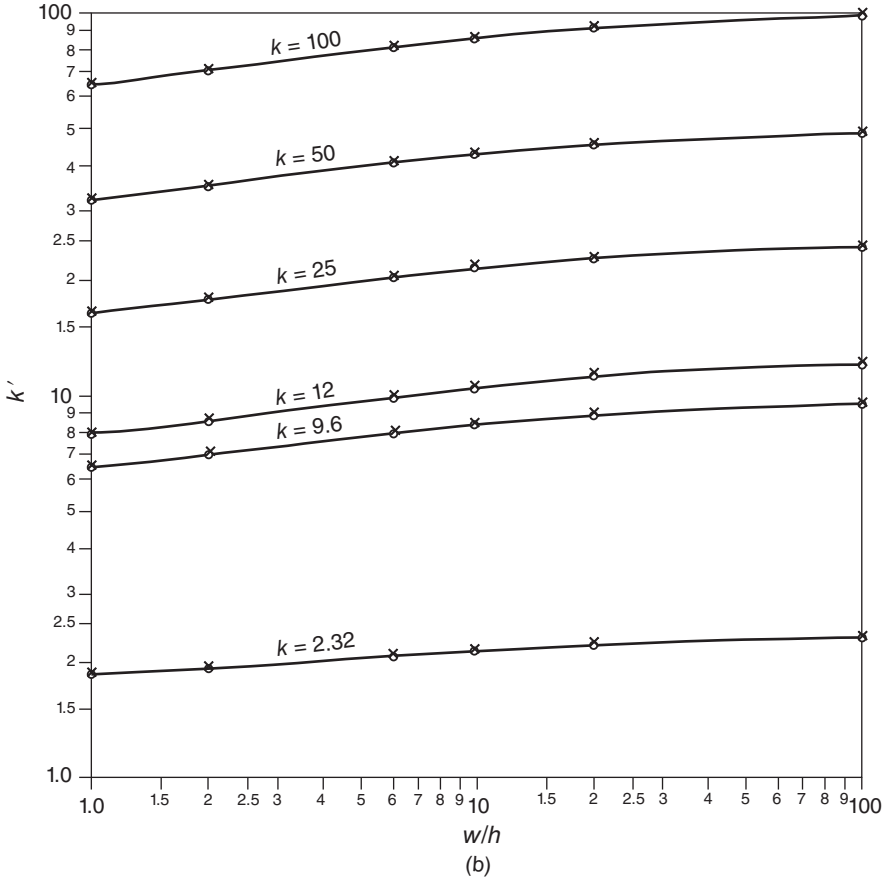


FIGURE 1.8 (Continued)

1.3 FREQUENCY BANDS, MODES, AND WAVEFORMS OF OPERATION

A useful list of the frequency bands is given in Table 1.3. As the frequency moves up, the wavelength of sinusoidal signals reduces and the bandwidth in hertz of the communications systems continues to increase, including optical circuits. The human eye operates at $\lambda = 5000 \text{ \AA}$, which is a frequency of

$$f = \frac{c}{\lambda_0} = \frac{3 \times 10^8 \text{ m/s}}{5000 \times 10^{-8} \text{ m}} = 0.6 \times 10^{15} \text{ s}^{-1} = 600 \text{ 000 GHz} = 600 \text{ THz} \quad (1.20)$$

The carrier or unmodulated sinusoidal signal may propagate in several modes which satisfy Maxwell's equations. The simplest preferred mode is TEM, where there is no electric or magnetic field in the direction of propagation. This is the mode for low-frequency coaxial lines and parallel-plate waveguides. Microstrip lines operate in their useful frequency range in a quasi-TEM regime. When the parallel-plate waveguide is enclosed

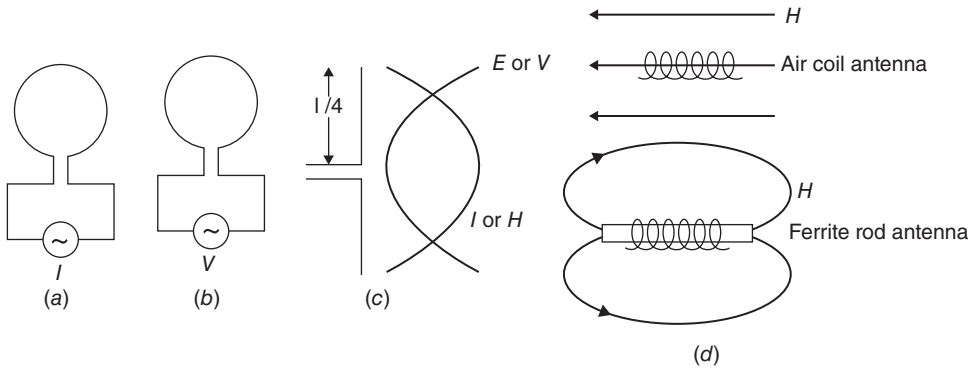


FIGURE 1.9 Four antennas illustrating Maxwell's four equations: (a) electric, (b) magnetic, (c) dipole, and (d) ferrite rod antennas.

TABLE 1.3 Frequency Bands

Band	Frequency
RF	1500 MHz
P	500 MHz–1 GHz
L	12 GHz
S	24 GHz
C	4–8 GHz
X	8–12 GHz
Ku	12–18 GHz
K	18–26 GHz
Ka	26–40 GHz
U	40–60 GHz
E	60–90 GHz
W	75–110 GHz

(see Fig. 1.5), the modes change to TE and TM, and the low-frequency propagation is not possible. The transmission medium has become a single piece of metal versus two pieces of metal for TEM modes. Another possible mode of propagation at microwave and millimeter-wave frequencies is a surface mode, TM or TE [1.26, 1.28]. Few applications have been found, but the very simplicity invites consideration. As the dielectric thickness approaches about $\lambda/4$ at high frequencies, the TE_1 surface mode is propagated with a cutoff frequency of

$$f_{c1} = \frac{c}{4h(k-1)^{1/2}} \quad (1.21)$$

This mode is dispersive, that is, the velocity is increasing with frequency. When the TE mode begins, the velocity of the TM mode is close to the velocity of the microstrip TEM mode and the fields are very similar in nature, that is, well coupled. Does this invite ideas? These concepts are only applicable when the frequency is high (millimeter wave) or the dielectric constant is high (100). For 25-mil alumina,

$$f_{c1} = \frac{3 \times 10^{10}}{4 \cdot 0.025 \cdot 2.54 \cdot 3} = 39.4 \text{ GHz} \quad (1.22)$$

But for 250-mil rutile ($\epsilon_r = 100$)

$$f_{c1} = \frac{3 \times 10^{10}}{4 \cdot 0.025 \cdot 2.54 \cdot 10} = 1.2 \text{ GHz} \quad (1.23)$$

and indeed effects of this were observed in the S band at Texas Instruments in 1967 [1.32].

When considering waveforms, we are thinking or calculating in the time domain instead of the more common frequency domain. The sinusoidal waveform is the basis of analog communications, and the reader should be familiar with AM and FM. The digital waveforms are more difficult to generate and visualize, especially when modulated in multicarrier (orthogonal frequency-division multiple [OFDM]) or spread-spectrum (code-division multiple access [CDMA]) modulation schemes. Fortunately, the actual circuits are essentially the same for both analog and digital, although some of the specifications will change and the methods of testing are obviously different. Modern CAD tools will give us the performance in the frequency and time domains, so the limitations may be studied in detail.

1.4 ANALOG AND DIGITAL SIGNALS

Analog signals travel continuously in real time, so it is very difficult to multiplex signals to increase the number of customers on the frequency band. For digital signals, there are many ways of multiplexing the signals so several customers receive communications simultaneously over the same frequency band. This is one of the foremost advantages of digital communication systems. The process of converting analog signals to digital format uses high-speed sampling analog-to-digital converter (ADC) circuits. Once the signals are in the digital domain, it becomes very easy to multiplex the information. In addition, signal processing can be done using digital signal processing (DSP) circuits to perform filtering, interpolation, and so on. Once the above techniques are performed, high-speed digital-to-analog converter (DAC) circuits may be used to reconstitute the original analog signal. Another feature of digital communications is error correction, which is not possible in analog communications. Analog signals may fade or become lost in the noise intermittently; with digital computer data, the digital format allows the information to be corrected for transmission errors, the accuracy is essentially 100%.

Analog signals are basically AM or FM (wider bandwidth). Digital signals usually carry the information in a mixture of AM and PM, that is best interpreted as discrete points in a plane given in polar coordinates.

The three most important forms of digital multiplexed signals are as follows:

TDMA means each user is sharing the same frequency with his or her own time slot, typically eight users on the same frequency (for GSM).

FDMA means the carrier frequency is hopping in a pattern known by the transmitter and receiver.

CDMA means the entire bandwidth is shared by all users, who have orthogonal signals which do not interfere with each other. The bandwidth for CDMA is 1.25 MHz, and it has increased to 5.0 MHz for WCDMA.

For the designer of the microwave front end, it is important to know the signals that are to be transmitted or received. In the analog domain, these would be defined by the

bandwidth, and in case of AM, also by the statistical peak-to-average power ratio (PAPR). The same holds for digital signals. In the following, we will recall basics of digital modulation schemes with a focus on the wave form they produce. Remember that the circuits that we have to design need to process these signals in a highly linear analog way.

As said before, the digital signal codes its information basically in the amplitude A and phase ϕ relative to a reference amplitude and phase. Let the reference be unity for the amplitude and zero for the phase for the moment. At the decision time, the digital signal therefore reads:

$$S = A \cos(\omega t + \phi) = I \cos(\omega t) + Q \sin(\omega t) \quad (1.24)$$

Amplitude A and phase ϕ define one point in polar coordinates on a plane. The signal can also be interpreted in Cartesian coordinates with $I = A \cos(\phi)$ and $Q = A \sin(\phi)$. In fact, IQ modulators are used to generate and demodulate these digital signals. First of all, single-carrier modulation schemes differ by the number of states they define. Figure 1.10 shows two examples of quadrature amplitude modulation (QAM) and 64QAM. Increasing the number of symbols increases the number of bits that a symbol represents. In QAM, for example, the $4 = 2^2$ symbols represent 2 bits each, while in 64QAM, each of the $64 = 2^6$ symbol transmits 6 bits. There are many possible combinations. For example, the symbols are not necessarily defined on a Cartesian grid.

Figure 1.11 shows a generic system for the realization of such a digital modulation scheme. The incoming bit stream is first mapped to the symbols, i.e. real part I and imaginary part Q of the complex signal amplitude are calculated. The next step, pre-coding, is to define the time-dependent shape of the symbols, which determines the required bandwidth and the resulting analog envelope signal. The signal is then converted to analog and upconverted by an $I - Q$ modulator.

Increasing the order of the digital modulation increases the data rate. But it comes at a price. In order to save energy, the signal amplitude (defining the difference between the symbols) would be chosen as low as possible where the channel noise level and the tolerable symbol error rate define the minimum transmit power. Higher order modulation schemes require therefore either better channel quality or higher transmit powers.

This book is on the design of microwave circuits, and considering the discrete constellation points alone is not sufficient. Each of the symbols is to be transmitted linearly, and

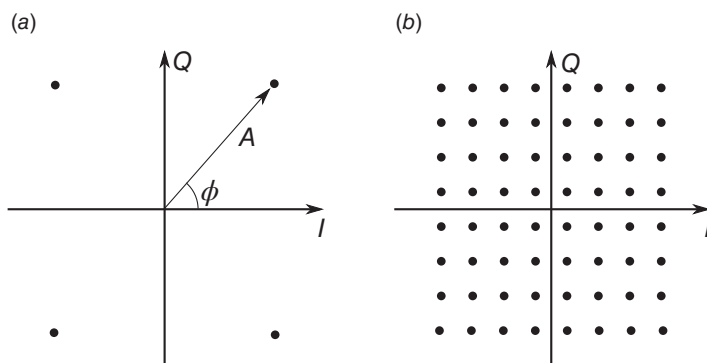


FIGURE 1.10 Constellation diagram for (a) QAM (quadrature phase shift keying [QPSK] and GMSK would look the same) and (b) 64QAM.

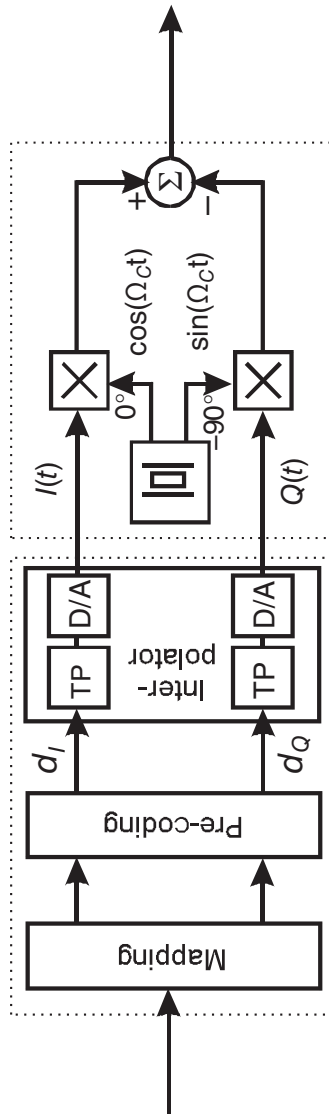


FIGURE 1.11 Block diagram of a digital single-carrier modulation system.

it is useful to consider the digitally modulated bitstream as an analog signal varying phase and amplitude over time. The envelope shape of a digitally modulated signal depends on the shape of the symbols I and Q , that will not be ideal square functions. Instead, the rising and falling edges of the symbols are smoothed by convoluting the square-shaped symbol with a rolloff function like a cosine half-wave or by a root-raised cosine function that is basically a burst of the shape of $\cos^2$. The width of these rolloff functions relative to the symbol length is given by the rolloff factor ρ .

Figure 1.12 shows the simulation of four modulation schemes that would provide identical constellation points at the decision time, but that show quite different continuous-time behavior.

These diagrams were simulated by plotting the resultant vector in the IQ plane, using a random bit series in the I and Q channels and assuming different symbol shapes. Comparing the QPSK signals with cosine and root-raised cosine rolloff, it is observed that the cosine rolloff yields amplitudes that show less variation. Both modulation schemes, however, show zero-crossings of the signal amplitude.

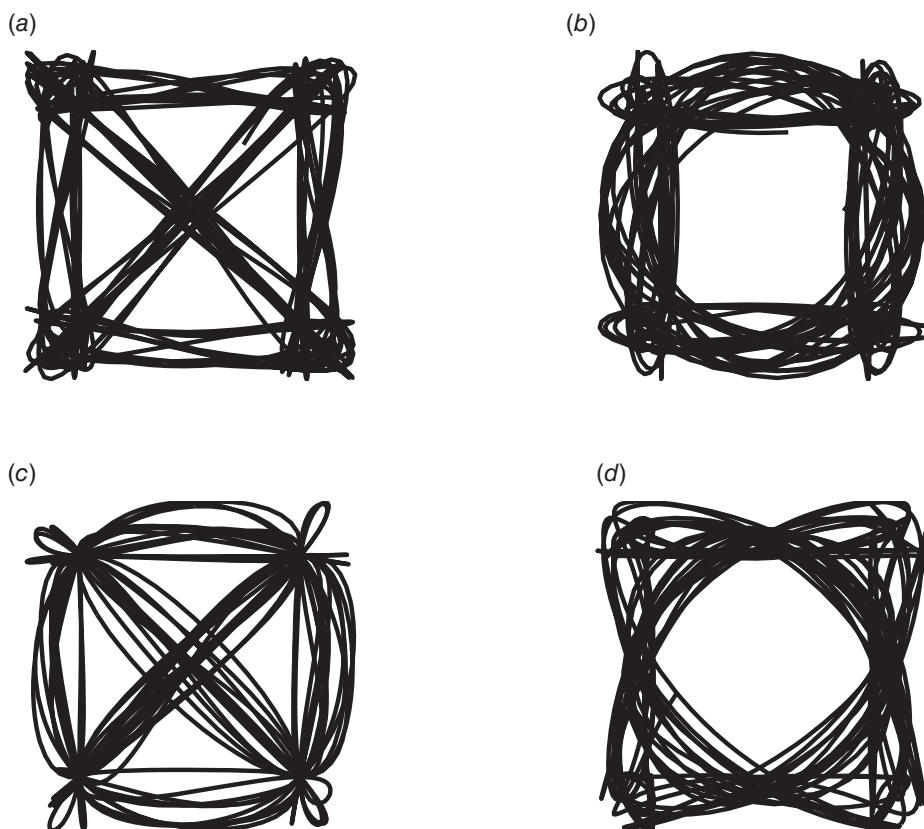


FIGURE 1.12 Constellation diagrams drawn in continuous time for a random bit sequence. QPSK signals with raised cosine and root-raised cosine rolloff, and the corresponding OQPSK signals. (a) QPSK root raised cosine, (b) OQPSK root raised cosine, (c) QPSK raised cosine, and (d) OQPSK raised cosine.

A varying amplitude is disadvantageous for the design of the transmitter. First of all, although the decision points are all at the same power level, we require to process the envelope over all possible amplitudes with high linearity. In the design, a linear operation would be required between zero output power and the power required for the maximum amplitude. In addition, the average power of the transmitted signal will be lower than the maximum. This requires a kind of power back-off in the design, and especially the power amplifier is required to be capable of providing the maximum power while it is running on average at a lower power level. From the system perspective, the zero-crossing adds difficulties. On the receive side, for example, it is required to synchronize in frequency and phase to the transmitted signal. If the signal disappears frequently, this task becomes more challenging. On the transmit side, LO leakage into the transmit channel can become challenging if the signal should be zero.

In the case of QPSK, however, it is well possible to reduce the PAPR (or Crest factor). A zero crossing in the $I + Q$ envelope only appears if I and Q are zero at the same time. Thus, offsetting the symbols of the two branches by half a symbol (offset QPSK, OQPSK) prevents it. The resulting complex envelope is shown in Figure 1.12. The PAPR is significantly reduced and the signal shows no zero-crossing at all.

But why not forcing the continuous signal to a constant envelope? A constant envelope signal can be amplified using a saturated amplifier. Its nonlinearity is not an issue at constant amplitude, and we can profit from a high-efficiency mode of operation. A constant amplitude A restricts I and Q to $I = \cos(A)$ and $Q = \pm \sin(A)$; thus, the shape of the symbols has the form of a cosinus half-wave. The information is coded in the change of the phase from symbol to symbol. This coding scheme is called minimum shift keying (MSK). Figure 1.13 shows the complex envelope of an MSK-coded signal as a function of time. It is clearly seen that only the phase changes, but not the amplitude.

Even though the MSK carries its information in the phase only, it was explained here in terms of a quadrature AM signal with appropriately shaped symbols in the two channels. Since digital modulation schemes are always designed to be bandwidth efficient, the MSK can also be interpreted as a narrowband PM that requires the same bandwidth as the QAM.

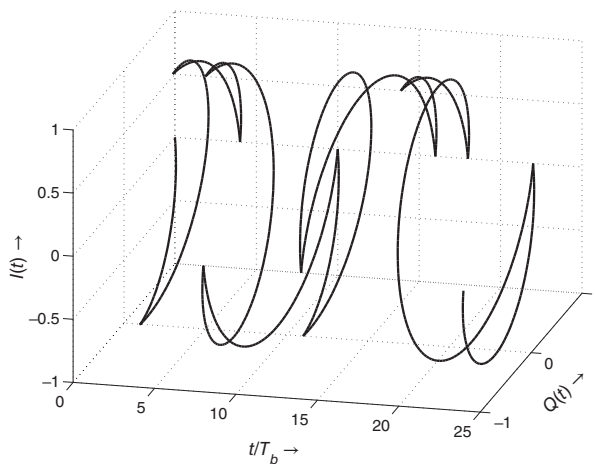


FIGURE 1.13 Complex envelope of an MSK-coded signal as a function of time.

In order to keep a constant envelope but at a significantly reduced bandwidth, the cosinus-shaped symbols of MSK can be replaced by Gaussian-shaped symbols. This coding is called GMSK and was chosen for the GSM standard.

The advantage of GMSK is that it is a spectrally efficient coding scheme that is probably the easiest to be realized in hardware. But it has certain limitations. The first is the limited data rate. One would like to use higher-order modulation schemes in order to fit more bits into each expensive Hz of bandwidth. The GSM standard also used dedicated channels and time slots, and therefore a fixed maximum number of connections. And the channel suffers from fading and echos, which requires adaptive distortion correction.

The UMTS standard therefore relied on a WCDMA-type of modulation. The baseband signal processing scheme is shown in Figure 1.14. A (wideband) code-division multiple access scheme starts with a traditional single-carrier modulated signal, but one performs an XOR of each bit with a predefined known pseudo-random bit stream in the baseband. Each bit results thereby in a long series of short so-called chips. In order to retrieve the original bit, one requires again to perform an XOR using the same bit stream that was used for the encoding. Since one bit gets subdivided in a large number of much shorter chips, the bandwidth increases accordingly. This type of modulation is called direct-sequence

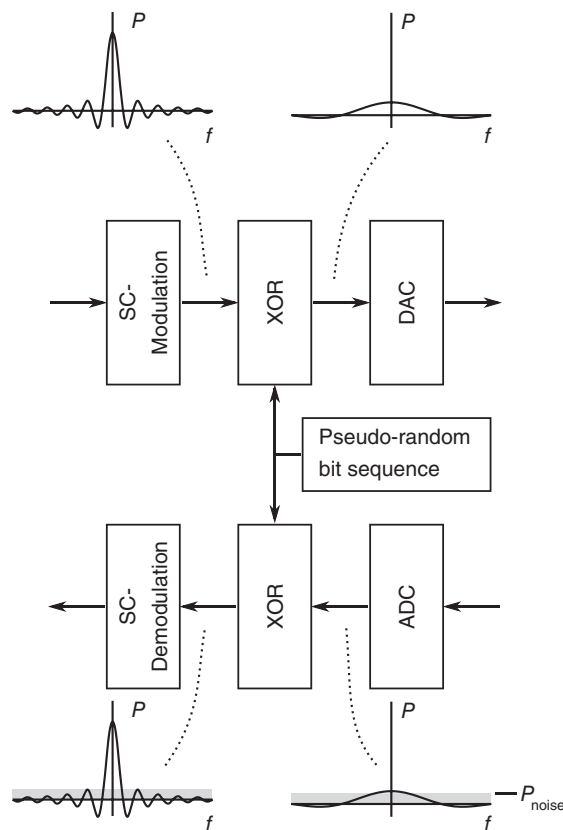


FIGURE 1.14 Block diagram of a baseband CDMA modulation and demodulation system, indicating the bandwidth required (for the theoretical case of rectangular symbols).

spread-spectrum technique. Unlike FM where the increasing bandwidth corresponds to an improvement in signal-to-noise ratio (SNR), the SNR of CDMA is identical to AM, since spreading and de-spreading would not affect white noise. So what is the benefit of such a technique?

Spread-spectrum has its roots in military communications. Unless you know the spreading code, you cannot decode the signal. It therefore provided channel encryption in times where digital encryption was not available. The signal is also hard to be detected, it looks like broadband noise to an observer, and it is hard to be jammed since a narrowband jamming signal gets spread in frequency at the receiver.

The usage in mobile communications has other reasons. Today, encryption of the transmission is done digitally, while the spreading codes are standardized. The first benefit of this approach is that gold codes are used for spreading and despreading. These codes are pseudo-random, but also pseudo-orthogonal. Using orthogonal codes allows to use the same bandwidth for multiple transmissions, since spreading with one code and despreading with another code will not reconstruct the bit, but yield a random function that gives no signal when integrating over the length of the bit. Figure 1.15 shows the effect. If the same code was used on spreading and despreading (Fig. 1.15a), a clear signal is obtained, while for different codes (Fig. 1.15b), a low residual error signal remains. The spreading code is used to separate the channels that share the same frequency at the same time: it is CDMA.

UMTS now relies on pseudo-orthogonal codes, with means that the different parallel transmissions do not cancel out completely. Other subscribers effectively add to the channel noise, that is why the data rate in UMTS is variable, it needs to be adjusted to the channel quality. On the positive side, there is no hard limit to the number of parallel communication channels.

The other benefit is that the spreading code is orthogonal to a time-shifted replica of itself. The receiver requires to be perfectly synchronized in order to receive a signal. Figure 1.15 shows the autocorrelation function of a gold code and the cross-correlation of two different codes. Regarding the autocorrelation, the signal is fully restored at $\tau = 0$, i.e. in case of a synchronized receiver. Any delayed signal $\tau \neq 0$ yields no signal at the receiver. The cross-correlation is observed not to be zero, but to yield something close to an increase of the noise floor. Any echo or multipath propagation is canceled out, or we might even use parallel receivers that synchronize also to the time-delayed echos, which reduces the bit-error rate. If the channel is no longer impeded by multipath-propagation problems thanks to CDMA, higher-order modulation schemes can be applied, and data rate can be increased significantly.

Unfortunately, the designer of the front-end now faces the issue that it is required to work with a signal that has a considerably high PAPR, and also wide bandwidth. CDMA is commonly combined with a single-carrier modulation up to 64QAM in high speed downlink packet access (HSDPA), and therefore comes with the typical PAPR for such a modulation scheme. However, the transmitter of the mobile base station is required to work with the superposition of a high number of channels, resulting in a signal that more or less looks like noise.

An alternative approach to mitigate the problem of multi-path transmission is to use symbols that are significantly longer than the time difference for the different propagation paths. Long symbols mean a slow data rate, but also a very narrow bandwidth. If you take, say, 256 bits, and transmit them in parallel: you could take 256 times the time per bit, using 256 parallel subchannels that require in total exactly the same bandwidth. This is

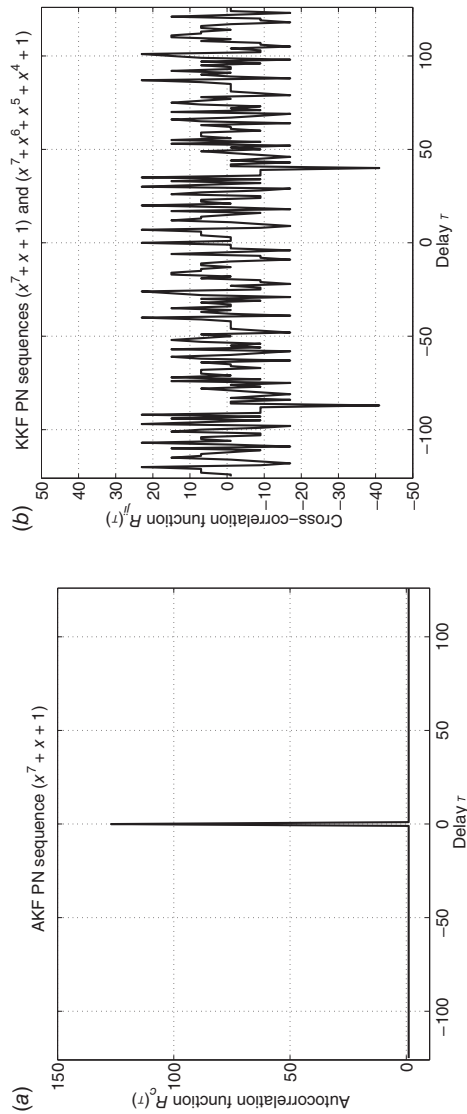


FIGURE 1.15 *a* Auto- and *b* crosscorrelation functions of pseudo-random signals generated through seven-bit feedback shift registers.

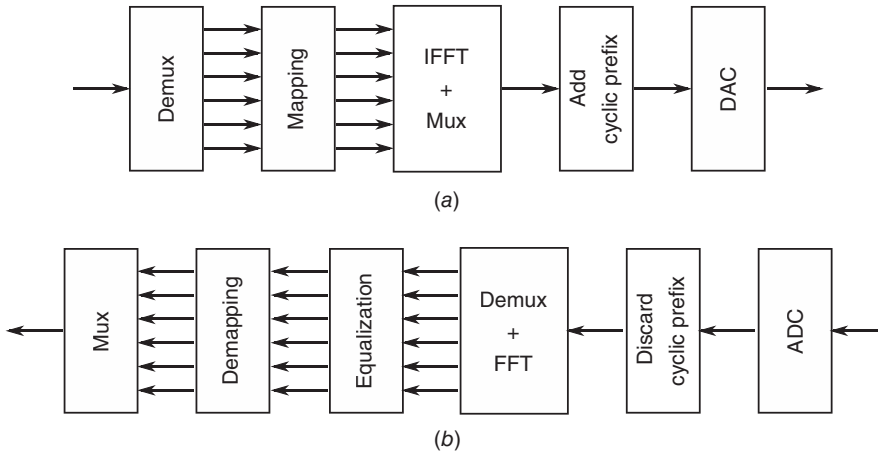


FIGURE 1.16 Block diagram of a baseband OFDM modulation (a) and demodulation (b) system.

the basic idea behind the OFDM access modulation that is used in many applications such as WiFi or DVB-T.

The OFDM transmitter works as follows, see Figure 1.16. The sequential bitstream de-multiplexed into a number of lower-datarate parallel streams. These parallel streams determine the sub-channels, thus data in the parallel streams can be understood to be Fourier components defining DC and a number of harmonics. What needs to be done now is to transform this mathematical description of the frequency components into a time-domain signal that can be sent out. This is done by an inverse fast Fourier transform (IFFT) followed by multiplexing the output, as the IFFT algorithm returns a vector of time samples. Between Demux and IFFT, it can be advantageous to rearrange the sub-channel (mapping), which will be discussed later. As a last step on the transmitter side a guard interval, the so-called cyclic prefix, is added between two symbols, in order to improve the signal quality. The guard interval should cover the time difference of the different transmission channels, but it should be short compared to the bit length. This guard interval is filled with meaningful contents, as it is filled with a replica of the signal from the end of the payload signal. A typical simulated baseband OFDM signal is shown in Figure 1.17 for 280 subchannels coded in the 64QAM modulation scheme. The contents of the guard interval is obtained by copying the time series from the tail of the symbols before the start of the signal.

On the receiver side, first of all the guard interval is discarded. The time-domain signal is then converted into the frequency domain through a fast Fourier transform (FFT), followed by an equalization step. The FFT requires a vector of time steps as an input, therefore a demultiplexer precedes it. Equalization is required since the superposition of signals in a multipath transmission will either be constructive or destructive, strongly depending on wavelength and frequency. The radio channel becomes thereby dispersive, probably also with some sub-channels being completely lost for the transmission. Through constant adaptive equalization for each of the subchannels and using error-correction codes, a secure data transmission is still possible. The combination of OFDM with digital error correction is known as coded orthogonal

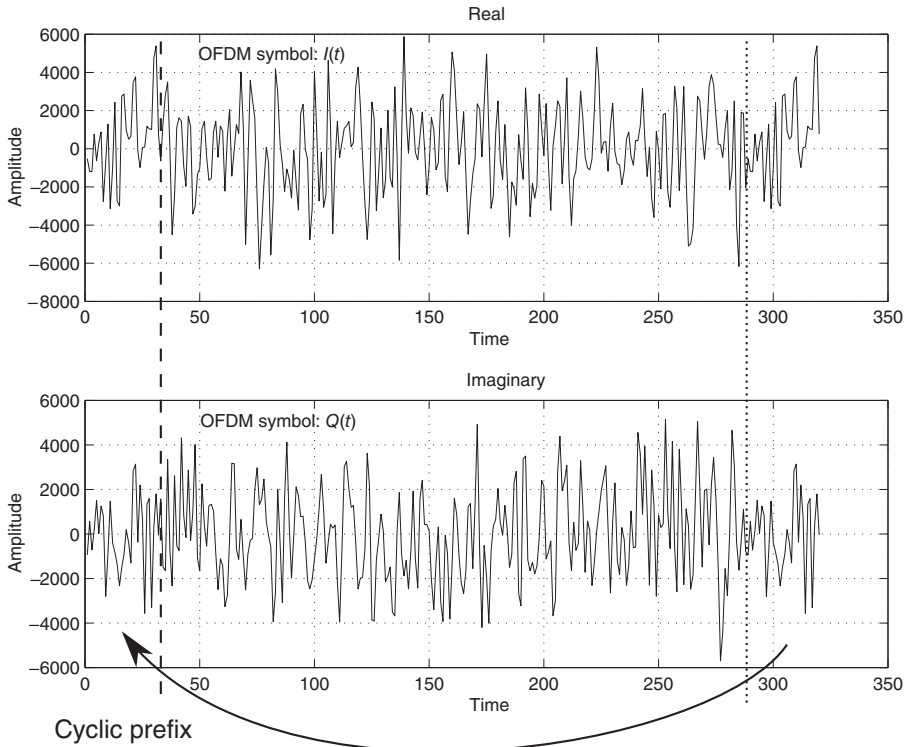


FIGURE 1.17 Time-domain OFDM signal.

frequency-division multiplexing (COFDM). When the equalization reconstructed the data in the parallel sub-channels, a multiplexer fits them back into a serial time-domain bit-stream.

The advantage of OFDM is clearly the better performance in correcting the channel distortion due to multipath propagation as compared to a single-carrier modulation. Even highly distorted channels rather lose a few subchannels, while for the single-carrier transmission, one would be required to restore a frequency-independent response of the channel.

The downside of OFDM is the fact, that we have to deal with the superposition of a high number of subchannels that are often coded with some higher-order QAM scheme, resulting in a signal that looks more or less like noise, with a high PAPR. This can be seen in Figure 1.17 that shows the complex envelope of an OFDM signal. Several digital approaches to reduce PAPR in order to relax the requirement to the transceiver frontend include appropriate mapping of the subchannels, limiting of peak signals, or adding channels that are filled with waveforms that compensate extreme signal amplitudes. Figure 1.18 shows the impact of gain compression resulting in the clipping of high amplitudes. The original broadband signal is well confined within the bandwidth of 20 MHz in this case. If the transmitter limits the signal to 3 dB below the maximum amplitude, significant spectral regrowth and out-of-band transmission is observed. 3 dB is a lot, but considering the fact that the amplitude density distribution is approximately Gauss-shaped as for a noise signal, clipping occurs only in a very small fraction of time.

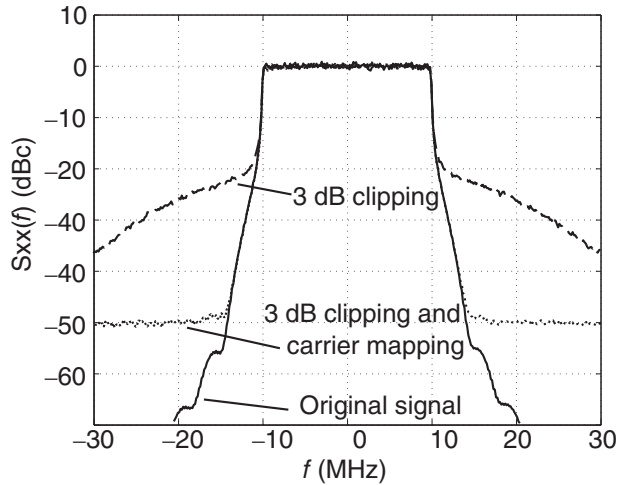


FIGURE 1.18 Frequency-domain OFDM signal, showing spectral regrowth due to clipping of signal peaks.

The figure also shows that by adding a few auxiliary channels that can be filled with random data, it is possible to reduce the peak-to-average ratio significantly and the impact of the signal clipping becomes much less pronounced.

Equalizing in the frequency domain is, however, intriguing. In a perfect world, we would like to combine the benefits of OFDM in terms of channel equalization together with the reduced PAPR of single-carrier modulation schemes. The method to achieve this is called SC-FDE, and developed further it is known as SC-FDMA used in the upstream (mobile to base station) in LTE.

The starting point for a modulation system as shown in Figure 1.19 is the use of a single-carrier modulation scheme in order to keep some control over the PAPR. When

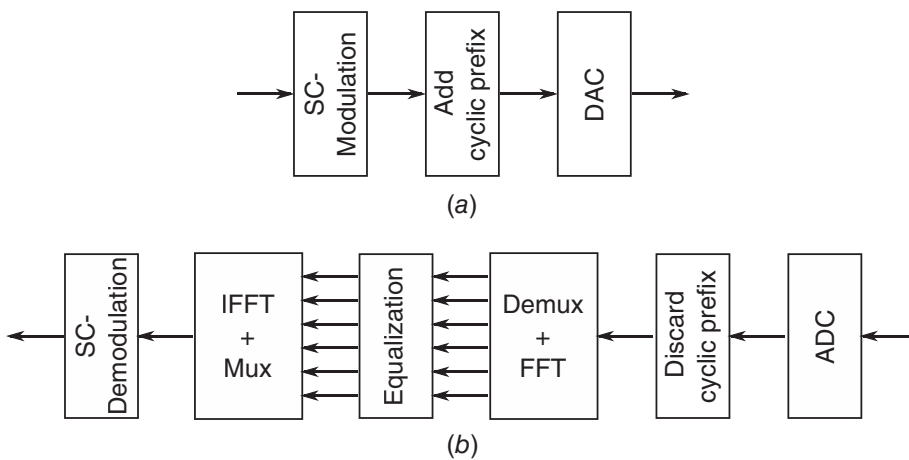


FIGURE 1.19 Block diagram of a baseband single-carrier frequency-domain-equalization (SC-FDE) modulation (a) and demodulation (b) system.

we receive this signal, we will observe echoes and frequency-dependent fading due to the multipath-propagation. How to work with these issues we learned from OFDM: Equalizing the channel in the frequency domain. Therefore, the received signal is transformed into the frequency domain, the fading is numerically corrected for where possible, and it is transformed back to the time domain for demodulation. By doing so, we treat the fast bit sequence that is transmitted as if it were some random low-bitrate OFDM signal in the receiver anyways, so we can easily add a cyclic prefix to the data we send out. An alternative way to explain it: The transmitter-channel-receiver chain is linear to the signal. Thus, it is possible to move building blocks without any impact on the output.

Such a coding takes the load from the mobile device, not only in terms of PAPR but also concerning computing power that is completely required on the receiver (base-station) side, where power consumption is less critical.

In LTE, the SC-FDE is modified to enable frequency-division multiple access, called SC-FDMA. In a mobile network, it cannot always be guaranteed that the full required bandwidth is available. Instead, the basestation assigns multiple sub-bands to the mobile, and the mobile splits the frequency band. This distribution of information to subbands is again performed in the digital domain through a FFT algorithm, rearrangement of the Fourier components to the available frequencies, and IFFT to obtain the corresponding time-domain signal.

1.5 ELEMENTARY FUNCTIONS

Before concluding this chapter, a few elementary definitions are needed, including noise figure, minimum detectable signal (MDS), dynamic range (DR), spurious-free dynamic range (SFDR), P_{IdBc} , intermodulation distortion (IMD), third-order intermodulation (TOI), and so on. These definitions are used throughout the book.

The noise contributed by the receiver may be calculated from the noise figure, which is the ratio of S/N (signal-to-noise power) input to output:

$$\frac{(S/N)_{\text{in}}}{(S/N)_{\text{out}}} = F \geq 1 \quad (1.25)$$

or equivalently the noise temperature (in kelvin),

$$T_e = (F - 1)kTB \quad (1.26)$$

where k is Boltzmann's constant equal to $1.381 \times 10^{-23} \text{ J/K}$, B denotes the bandwidth in Hz, and T is the ambient temperature (290 K is the IEEE standard for room temperature).

Since the receiver always adds additional noise while amplifying the input signal and noise, the input S/N is always greater than the output S/N . When we cascade components in a receiver, the total noise figure is calculated from Friis's noise figure equation:

$$T_{\text{tot}} = F_1 + \frac{F_2 - 1}{G_{A1}} + \frac{F_3 - 1}{G_{A1}G_{A2}} + \dots \quad (1.27)$$

where F_N is the noise figure of the N th component and G_{AN} is the available gain of that component, in ratio form (not decibels).

Before beginning the next section, one must realize the noise level of any resistor is kTB :

$$kTB = 1.38 \cdot 10^{-23} \text{ J/K} \cdot 290 \text{ K} \cdot 1 \text{ Hz} = -204 \text{ dBW/Hz} = -174 \text{ dBm/Hz} \quad (1.28)$$

where k = Boltzmann's constant $T = 290\text{K}$ by IEEE definition $B = 1\text{Hz}$.

This number is used constantly throughout engineering, and you must understand the meaning of this very important fundamental noise limit.

Another very important concept is the Friis transmission equation, which discusses the range of the communication system [1.33]. Consider the simplest communications system shown in Figure 1.20. The transmitter with a power of P_t is fed into a transmitting antenna with a gain of G_t . The received power is P_r at a distance of R . The received power density can be calculated assuming no atmospheric losses, mismatch losses, and so on, as

$$S_D = \frac{P_t}{4\pi R^2} G_t \text{ (W/m}^2\text{)} \quad (1.29)$$

The received power is the power density multiplied by the effective area of the receiving antenna (which is related to the antenna gain)

$$P_r = \frac{P_t G_t}{4\pi R^2} A_{\text{er}} \text{ (W)} \quad (1.30)$$

$$A_{\text{er}} = \frac{G_r \lambda^2}{4\pi} \text{ (m}^2\text{)} \quad (1.31)$$

Substituting gives the Friis power transmission equation:

$$P_r = P_t G_t G_r \left(\frac{\lambda}{4\pi R} \right)^2 \text{ (W)} \quad (1.32)$$

which can also be put in the form [1.34]

$$\frac{P_L}{P_T} = \frac{k A_{\text{et}} A_{\text{er}}}{\lambda^2 R^2} \quad (1.33)$$

where k is an efficiency factor, generally 0.4–0.7, which accounts for factors such as misalignment, polarization mismatch, impedance mismatch, and atmospheric losses, and it is also called $1/L_{\text{sys}}$.

If $P_r = S_{i,\text{min}}$, the MDS required for the system, we have the maximum range

$$R_{\text{max}} = \left[\frac{P_t G_t G_r}{S_{i,\text{min}} L_{\text{sys}}} \left(\frac{\lambda}{4\pi} \right)^2 \right]^{1/2} \quad (1.34)$$

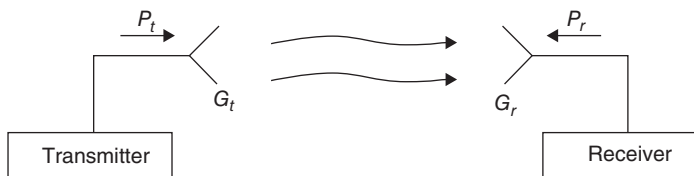


FIGURE 1.20 Simplified wireless communication system. (From Chang [1.33].)

This is a very important result for understanding the many components which constitute the communication system or wireless system. An example will illustrate these concepts. Consider a transmitter at 2 GHz,

$$\lambda_0 = 15 \text{ cm} \quad G_t = 10 \text{ m} \quad P_t = 1 \text{ W} \quad (1.35)$$

sending to a receiver with

$$G_r = 10 \text{ dB} \quad S_{i,\min} = -90 \text{ dBm} = -120 \text{ dBW} = 10^{-12} \text{ W} \quad L_{\text{sys}} = 1 \quad (1.36)$$

The maximum range is

$$R_{\max} = \left[\frac{1 \cdot 10 \cdot 10}{10^{-12} \cdot 1} \left(\frac{0.15}{4\pi} \right)^2 \right]^{1/2} \text{ m} = (1.42 \cdot 10^{10})^{1/2} \text{ m} = 119 \cdot 10^3 \text{ m} \quad (1.37)$$

which is about 75 miles. The range increases as the square root of the transmitted power increases. The frequency dependence indicates that the maximum range decreases with increasing frequency, although any frequency-dependent losses on the transmission path are not considered yet. The reason is that the size of an antenna providing a certain gain scales with the wavelength, and as a result, the effective area as defined in (1.31) shrinks.

Another important observation is the difference between coaxial transmission and antenna-to-antenna transmission, shown in Figure 1.21 [1.34]. As the distance increases beyond 3000 m (1.86 miles), the coaxial system is impractical compared to antennas. Also, the losses reduce at higher frequencies for the antenna system, but the opposite is true for coaxial systems.

The dynamic range of the system is illustrated in Figure 1.22, which is a plot of P_{out} versus P_{in} at the carrier frequency, where power is always given in dBm, dB to 1 mW (e.g. 1 W is 30 dBm, 10 W is 40 dBm). Because of nonlinearities in every component,

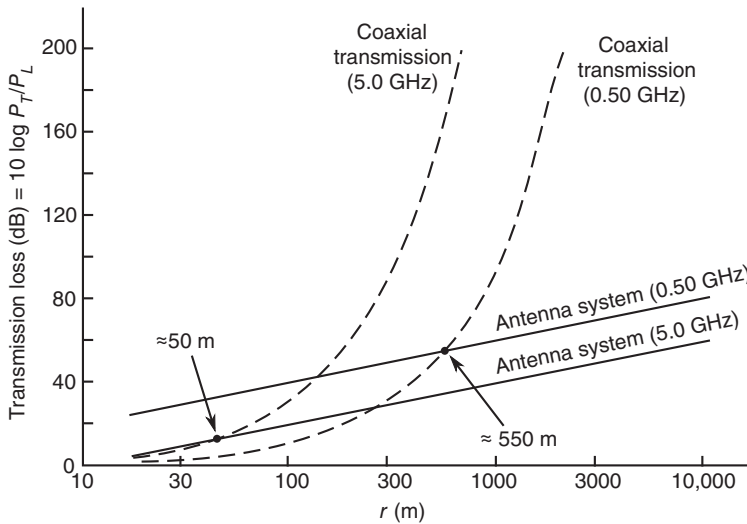


FIGURE 1.21 Attenuation for coaxial and antenna system communications. (Rizzi [1.34].)

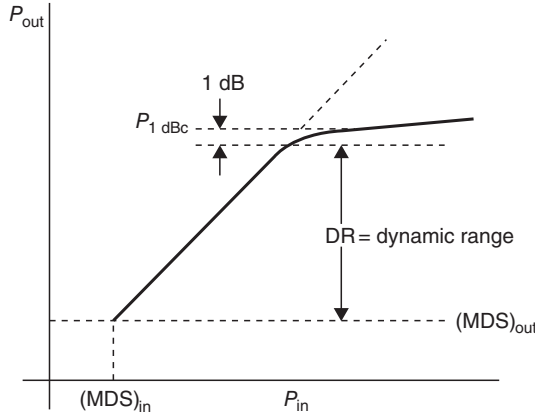


FIGURE 1.22 Dynamic range.

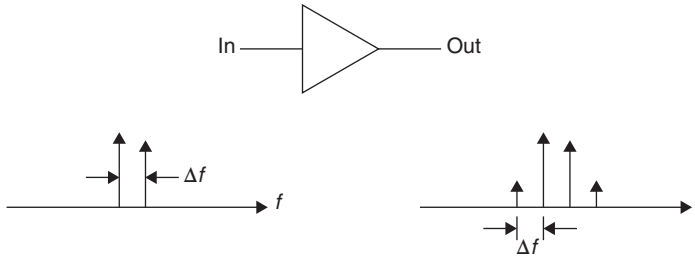


FIGURE 1.23 Output power spectrum for two input signals.

if two signals very close in frequency are introduced at the input, the output spectrum contains TOI products which are impossible to filter at the output; the resulting power spectrum at the load is shown in Figure 1.23. The intercept of the extended linear gain and the extended TOI products is a useful figure of merit called the TOI intercept point, or IP3, which is usually 10 dB or more above P_{1dBc} , the 1-dB compression point where the linear gain has been reduced by 1 dB. The SFDR refers to the output power range where no third-order products are observed (Fig. 1.24). An example will illustrate these various definitions. Consider an amplifier with a bandwidth of 30 MHz, transducer gain of 30 dB, and noise figure of 6 dB, the MDS is

$$MDS_{in} = -114 \text{ dBm} + 15 \text{ dB} + 6 \text{ dB} + 3 \text{ dB} = -90 \text{ dBm} \quad (1.38)$$

If the P_{1dBc} is at +15 dBm, the linear dynamic range is

$$DR = P_{1dBc} - MDS_{out} = 15 \text{ dBm} + 60 \text{ dBm} = 75 \text{ dB} \quad (1.39)$$

And the SFDR is

$$SFDR = \frac{2}{3}(TOI - G - MDS_{in}) = \frac{2}{3}[25 - 30 - (-90)] = \frac{2}{3}[85] = 57 \text{ dB} \quad (1.40)$$

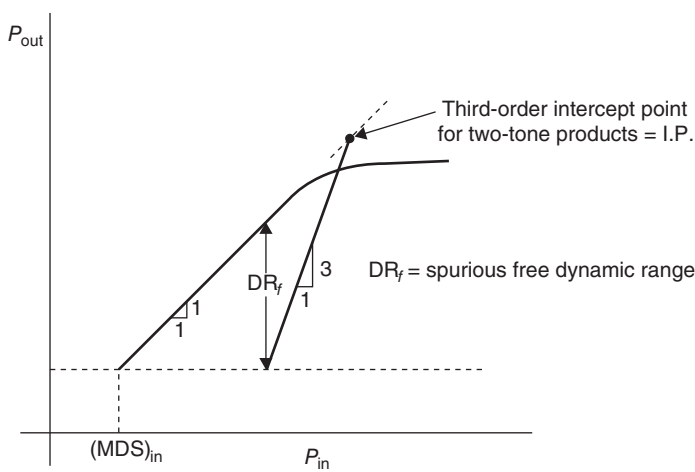


FIGURE 1.24 Spurious-free dynamic range.

Which is 18 dB less than the linear dynamic range. All of these calculations pertain to analog signals.

For digital transmission, the equivalent way of expressing P_{ldBc} is the adjacent channel power ratio (ACPR), which expresses the amount of signal “spilled over” to the next channel at high output powers. Typical requirements are -55 dB (minimum) down for the measurement channel carrier power. A good amplifier is given in Figure 1.25 using the MWT17 MESFET, which shows ACPR of 75 dBc at a low bias point, with the

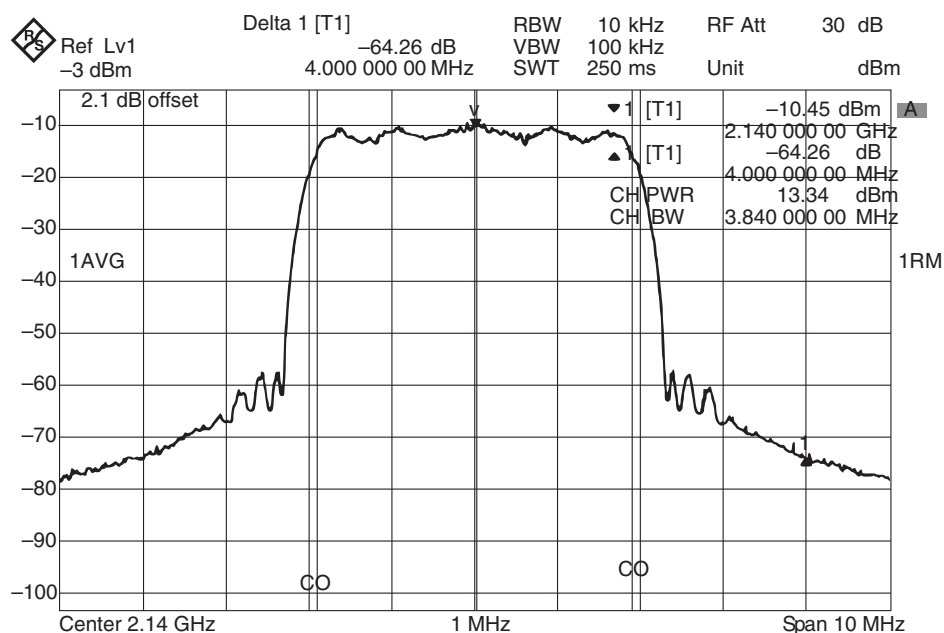


FIGURE 1.25 ACPR for a MWT-213011-82 amplifier. (Courtesy of Microwave Technology, Greg Zhou.)

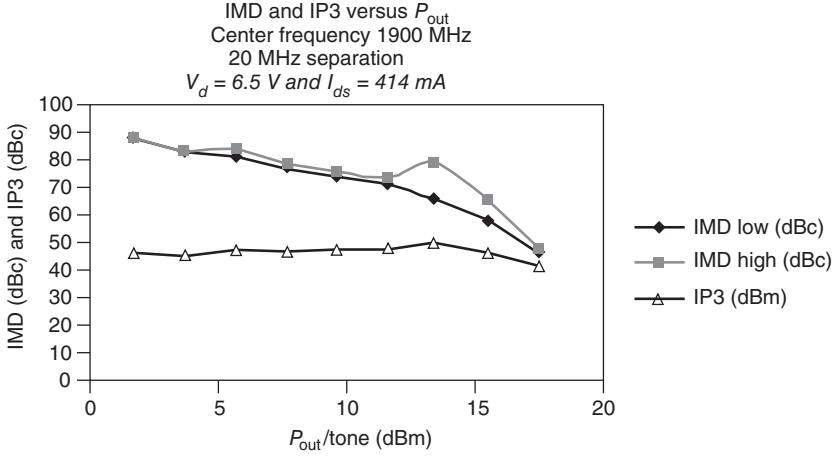


FIGURE 1.26 IP3 or TOI of MWT-213011-82 amplifier versus P_{out} . (Courtesy of Microwave Technology, Greg Zhou.)

output power backed off from 1 W to 13 dBm. The equivalent TOI measurements for this amplifier are given in Figure 1.26 versus output power for comparison.

A new look at this subject has been published by Rohde and Rudolph [1.6]. He proposes the dynamic measure (DM), a new figure of merit for receivers. The equation for the DM is

$$DM = [(IP3_{in} + Att) - (NF_r + Att + NF_{ant})] \frac{NF_r}{NF_r + Att} \quad (1.41)$$

where DM is a dimensionless number, $IP3_{in}$ is the receiver's third-order input intercept in dB μ V (instead of dBm) without any added attenuation; NF_r is the receiver noise figure in dB without any added attenuation; NF_{ant} is the antenna-system noise figure in dB; and Att is the decibel value of any added attenuation, which is sometimes used to shift the upper and lower limits of a system's dynamic range to higher absolute signal levels. The Att will increase NF_r and increase $IP3_{in}$, but the DM will decrease, which is probably undesirable. The usefulness of the DM figure of merit is illustrated further in Ref. [1.6].

One final topic on definitions is the calculation of load power. Circuit simulators usually offer a function that returns the power provided to a certain node. Commonly, it is assumed that the load at this node equals 50 Ω . However, if the load changes to 25 Ω or 100 Ω due to load-pulling effects in the real world, the computer will not understand this change during harmonic balance calculations, so it incorrectly calculates the power from the node voltage only, assuming $R_L = 50$ Ω :

$$P_{L50} = \frac{V_L^2}{2R_L} = \frac{V_L^2}{100} \quad (1.42)$$

which is of no physical meaning if $R_L \neq 50$ Ω . For correct results, one requires either to correct this value through

$$P_L = P_{L50} + 10 \log \left(\frac{50}{R_L} \right) \text{ (dBm)} \quad (1.43)$$

or to calculate the power directly from current and voltage by placing the two respective sensors in the circuit schematic. Of course, the value of V_L is always the peak value unless otherwise stated.

1.6 BASIC RF TRANSMITTERS AND RECEIVERS

A typical RF receiver (based upon Armstrong's superheterodyne receiver; Fig. 1.4) is given in Figure 1.27 [1.35]. This could be an AM radio receiver for the commercial AM broadcast band. The signal is twice down-converted to the low-cost 10.7-MHz IF band, amplified, detected, and amplified again in the audio band (0–8 kHz).

The basic transmitter is given in Figure 1.28 [1.35]. The characteristics of interest include power output and operating frequency, efficiency, power output variation, frequency tuning range, stability, oscillator quality factor (Q_u), noise (AM, FM, and phase noise), spurious signals, frequency variations due to frequency jumping, frequency pulling (load variations), frequency pushing (DC supply variations), and post tuning drift (frequency and power variations due to heating of a solid-state device). The oscillator noise is discussed in great detail throughout the book and is defined by

$$\mathcal{L}(f_m) = \frac{\text{noise power in 1-Hz bandwidth at } f_m \text{ offset from carrier}}{\text{carrier signal power}} = \frac{N}{C} \quad (1.44)$$

$$= 10 \log \left(\frac{N}{C} \right) \text{ dBc/Hz}$$

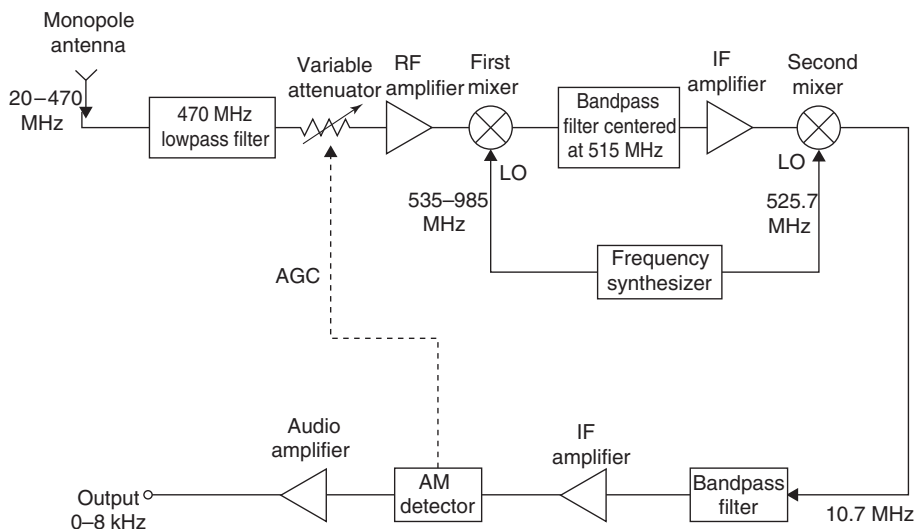


FIGURE 1.27 Typical radio receiver. (From Chang [1.38].)

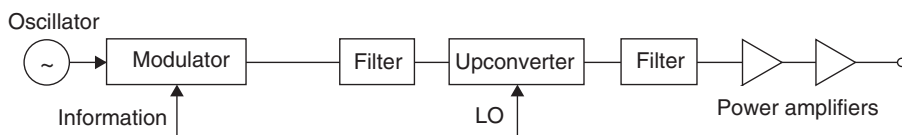


FIGURE 1.28 Transmitter system. (From Chang [1.38].)

where the negative sign in the answer is usually omitted. For example, if the carrier is 1 mW (0 dBm) and the noise is -80 dBm at $f_m = 100$ kHz, the $\mathcal{L}(100 \text{ kHz})$ is 80 dBc/Hz.

The cascading of circuit components produces an increase in noise figure and a reduction of TOI (IP3). The total noise figure is given by

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots \quad (1.45)$$

and the total IP3 is given by

$$\frac{1}{\text{IP3}_{\text{tot}}} = \frac{1}{\text{IP3}_n} + \frac{1}{G_n \text{IP3}_{n-1}} + \dots + \frac{1}{(G_n \dots G_2) \text{IP3}_1} \quad (1.46)$$

Unless otherwise stated the TOI or IP3 usually refers to the output of the component, but it could also refer to the input. We also use the notation OIP3 for output and IIP3 for input to avoid confusion. Another way to express the total IP3 at the input is to refer all of the IIP3 components to the input of the circuit, giving

$$\text{IIP3}_{\text{tot}} = 10 \log \left(\frac{1}{\text{IIP3}_1} + \frac{1}{\text{IIP3}_2} + \dots + \frac{1}{\text{IIP3}_n} \right)^{-1} \quad (1.47)$$

An example of the total IP3 is calculated from Figure 1.29 [1.35]. For this circuit, the total gain is 14 dB and the TOI is calculated for the input as

$$\begin{aligned} \text{IIP3}_{\text{tot}} &= 10 \log \left(\frac{1}{10} + \frac{1}{15.85} + \dots + \frac{1}{4} \right)^{-1} \text{ dBm} \\ &= 10 \log (0.10 + 0.0631 + 0.25)^{-1} \text{ dBm} \\ &= 10 \log (0.4131)^{-1} 10 \log (2.4208) \text{ dBm} = 3.84 \text{ dBm} \end{aligned} \quad (1.48)$$

So the output TOI is

$$\text{OIP3}_{\text{tot}} = \text{IIP3}_{\text{tot}} + G_{\text{tot}} = 17.84 \text{ dBm} \quad (1.49)$$

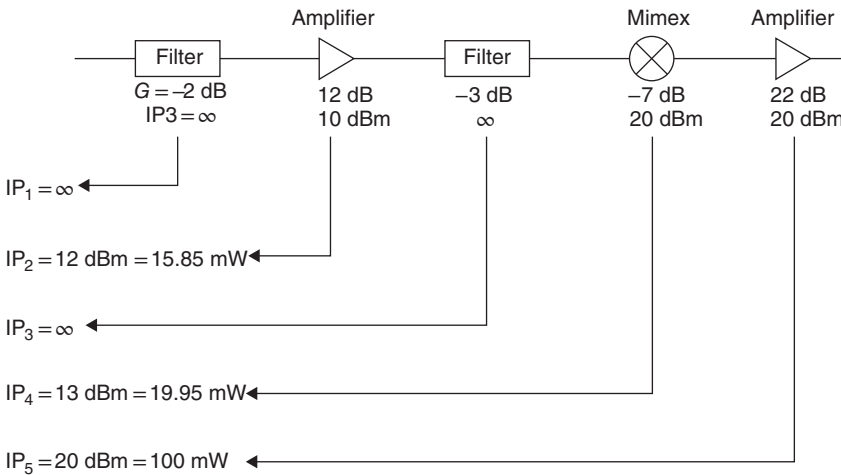


FIGURE 1.29 Receiver and its input intercept point.

1.7 RF WIRELESS/MICROWAVE/MILLIMETER WAVE APPLICATIONS

The primary applications of the three frequency groups are essentially the same: communications receivers and transmitters (or transceivers). The simplest example to envision is your cellular telephone operating between 800 MHz and 5 GHz. The essential components are the amplifiers, oscillators, and mixers, which will be covered in detail in this book. The complete derivations may be found in the references, but the necessary formulas are given in each chapter when appropriate.

The amplifier, oscillator, and mixer functions will use the lowest cost transistors which satisfy the specifications: including RF and SOI CMOS, GaN HEMTs, SiGe HBTs, InGaP/GaAs HBTs, InP HBTs, AlGaAs pHEMTs, InGaP pHEMTs, Si LDMOS transistors, and the list continues to expand.

The oscillators use the same low-cost transistors with the additional requirement of low phase noise. The material properties and manufacturing methods presently favor silicon-based devices due to lower $1/f$ flicker noise, but this could change quickly. Designers of amplifiers and oscillators are usually the same engineers using the same software, transistors, and circuit technology; however, oscillator designers also need a high- Q resonator. These oscillator designs may also be approached from a linear S -parameter viewpoint; however, a more complete design requires a nonlinear CAD solution. This will be shown in Chapter 10.

A basic cellular RF wireless transceiver is shown in Figure 1.30 [1.6]. The circuits are basically the same for both analog and digital modulation, it is also the same in all frequency bands, even at 100 GHz [1.36]. The chipset shown is not the most recent generation, as it provides only penta-band (four GSM and one UMTS bands) coverage and no LTE transceiver. But it shows nicely how close digital data processing already came to the RF circuit section.

Starting from the antenna, there is a switch that selects either GSM or UMTS operation. Next comes a duplexer in the UMTS branch, and a switchplexer in the GSM branch. While UMTS runs in full duplex mode (transmit and receive at the same time), GSM either transmits or receives bursts of data. Therefore, the GSM duplexer consists of a switch. A band-selection filter for each of the supported bands follows that are required to keep out-of-band signals, especially also from the own transmitter, from driving the LNAs into saturation. The next stage is a downconverting stage, consisting of an IQ mixer.

The IQ mixer consists of two AM mixers, one driven in phase with the transmit reference signal, and one 90° out of phase. This receiver configuration is able to detect mixed phase and amplitude modulated signals. An information coded in the phase ϕ relative to the reference phase results in a amplitude variation in the I and Q channels forwarding the angle ϕ to the demodulated signal, if it is interpreted as polar coordinates in the I - Q plane. This mixer topology is therefore well suited to determine the GMSK coded GSM signal, that provides a constant amplitude and contains the data in a relative phase shift, and the WCDMA-coded UMTS signal using phase and amplitude to transmit data. Since digital modulation schemes using frequency or phase modulation techniques always are narrow-band PM or FM schemes, those can as well be understood in terms of IQ -modulated AM signals.

After downconversion to baseband, the I and Q channels are low-pass filtered to prevent the alias effect, and then directly sampled and further processed digitally.

A modern cellular handset uses a direct conversion topology instead of the heterodyne concept (cf. Fig. 1.4). The advantages are obvious: the functionalities that had to be



realized at the IF by analog building blocks are much cheaper realized in the digital domain. The signal is now digitally filtered, amplified, and decoded.

In the analog world, the heterodyne concept provides the additional advantage of suppressing second-order intermodulation products. Assuming that the LNA and mixer are only weakly nonlinear, one can expand the output power as a function of input power into a Taylor series. The magnitude of the Taylor coefficients can be expected to decrease with increasing order, so that it can be sufficient to focus on the linear, quadratic, and cubic terms. The quadratic nonlinearity (second order) results in a mixing of any signal to DC and to its double frequency. Especially, a portion of the own transmit signal that gets unintentionally coupled into the receive part will therefore result in an arbitrary time-varying DC shift. Analog filtering or adjusting the DC in a direct conversion receiver is not possible, but the heterodyne concept cancels the issue by first mixing to IF. The digital direct conversion receiver solves this issue digitally by recalibrating the origin of the IQ plane.

The receive path of UMTS and GSM share basically the same architecture, but the transmitted signals differ significantly. GSM relies on a constant envelope signal that can be provided by an offset phase-locked loop (PLL), while UMTS requires a highly linear transmitter architecture. Thus, a heterodyne transmitter is used.

The baseband I and Q signals are first upconverted to an IF that is used as control signal of a PLL. The PLL consists of a voltage-controlled oscillator and a feedback branch. The feedback signal is downconverted and compared to the IF signal in a phase detector. The output baseband signal provides the control voltage for the voltage controlled oscillator (VCO). This architecture exceeds the performance of heterodyne transmitter approaches since it inherently eliminates LO feedthrough and spurious sideband conversion while providing image rejection, which reduces the requirements concerning filtering at the output [1.37]. The diagram in the figure shows that two of these transmitters are implemented, each of them covering two GSM bands (850 and 900 MHz and 1800 and 1900 MHz) (Table 1.4).

TABLE 1.4 RF/Wireless Applications

RF/Wireless	Microwave	Millimeter Wave
AM radio, 1 MHz	Ultrahigh frequency (UHF) TV, 300 MHz–1 GHz	Local multiple-distribution systems (LMDSs), 30 GHz
FM radio, 100 MHz	Microwave oven, 2.45 GHz	Communication at maximum attenuation, 60 GHz
Very high frequency (VHF) TV, 45–300 MHz	Cellular telephone, 850 MHz and 1.85 GHz	Communication at minimum attenuation, 10 and 94 GHz
Cordless telephone, 50 MHz	Bluetooth, 2.45 GHz	Optical communication at 40 GHz and above
Garage/auto openers, 1–5 MHz	Wireless video, 2.45 GHz coded orthogonal frequency-division multiplexing (COFDM), 5.25 GHz Ultra wideband, 3–10 GHz	Laser frequency, typically 300 000 GHz Ultra wideband, 60–66 GHz

The PLL approach can only modulate the phase of the signal. WCDMA modulation schemes require also high linearity amplitude modulation. Therefore, the I and Q signals are upconverted by an IQ upconverter that linearly shifts the phase and amplitude information to an intermediate frequency signal. The signal is then amplified and upconverted to the UMTS transmit frequency. A reason for the usage of the heterodyne architecture is a requirement to vary the power level of the transmitted signal in a wide range, depending on the wireless channel quality, which requires multiple adjustable gain stages [1.37].

A useful way to categorize the applications is to list some everyday products which the consumer uses (Table 1.3).

But consumer applications are no longer restricted to the lower GHz range. At the time of writing, automotive radar is transforming from a fancy extra in luxury cars to a standard security device. While autonomous driving is being explored by car makers and IT companies, we will probably first benefit from multiple radar systems that help mitigating car accidents and assist drivers. Current systems work at 77 GHz, and future system are assumed to operate at 94 and 24 GHz depending on the task.

Also in the millimeter wave range are short-range communication systems at 60 GHz that are about to be deployed at the time of writing. The task of these systems is to provide high data rates over short distances. The idea is to get rid of the Gbit LAN and TV cable in our livingroom. An example is the WiGig standard, which became IEEE 802.11ad in 2013. It operates in four bands of a bandwidth of 2.16 GHz each in the frequency range between 58.32 and 64.8 GHz, providing a maximum data rate up to 7 Gbps. Similar to Bluetooth, the interface can provide different functionalities. Two examples are the replacement of the video cable in order to wirelessly stream uncompressed high definition video, the other is the replacement of a docking station for notebook computers that allows to connect LAN, monitor, and all other peripherals by placing the laptop in the receiving range of the WiGig-enabled monitor.

Leaving the area of consumer devices, radar applications are not only entering the automobiles. Also in industrial applications, for the location or analysis of assets, many opportunities exist to expand and provide new functionalities. One is radar imaging, that currently is mainly known in its security application as a body scanners.

1.8 MODERN CAD FOR NONLINEAR CIRCUIT ANALYSIS

In way of introductory material, the present-day nonlinear CAD for any circuit is available in at least four forms:

1. Small-signal (S -parameter)
2. Transient time-domain form
3. Harmonic balance form
4. Envelope form (including modulation)

For circuits with no steady-state solution, transient time-domain solutions are found using some form of SPICE (which means Simulated Program with IC Emphasis). This can be the slowest form of the solution, but if needed, it is essential. SPICE was developed by the University of California at Berkeley about 1965, and it was made available to the public at no charge. This sparked commercial development with improved graphics and frequency-domain solutions using the FFT, but these solutions take great care since the

minimum time step can change the answer dramatically. Many companies thrived on SPICE development, including the free software package LTSPICE provided by Linear Technology.

If there is a steady-state solution (e.g. an oscillator), the computation can be reduced dramatically by using harmonic balance techniques. This has become available since about 1980, and it is widely used. The number of harmonics is usually chosen between three and seven and the output power spectrum may be displayed at the load. In addition, the dynamic load line of the transistor may be displayed. The harmonic balance method is very valuable to the design of mixer circuits, where there are many frequencies to find the respective power levels.

The third form of calculation is Envelope, which is most suitable for carriers with modulation. This CAD can dramatically decrease the time of the calculations of complicated digital circuits with all kinds of phase modulation. The output baseband information may be displayed in many ways, including constellation diagrams and eye diagrams. This tool became available about 1995.

1.9 DYNAMIC LOAD LINE

Basic electronics teaches us that when two sinusoidal signals are applied to the vertical and horizontal inputs of an oscilloscope circles will occur known as Lissajous patterns when the load is a pure capacitor or inductor. This is important information related to the nature of the dynamic load line of a transistor amplifier. It can easily be shown that a capacitor produces a clockwise (CW) pattern while an inductor produces a counterclockwise (CCW) pattern [1.38].

For a well-designed amplifier, both types of circles are frequently produced, that is, a figure-8 pattern, where part of the load line is capacitive (CW) and the other part is inductive (CCW). For most high-power amplifiers, the load line is an elliptical RC load (CW), where the closer this approaches a straight line, the higher the output power becomes (see Fig. 1.31).

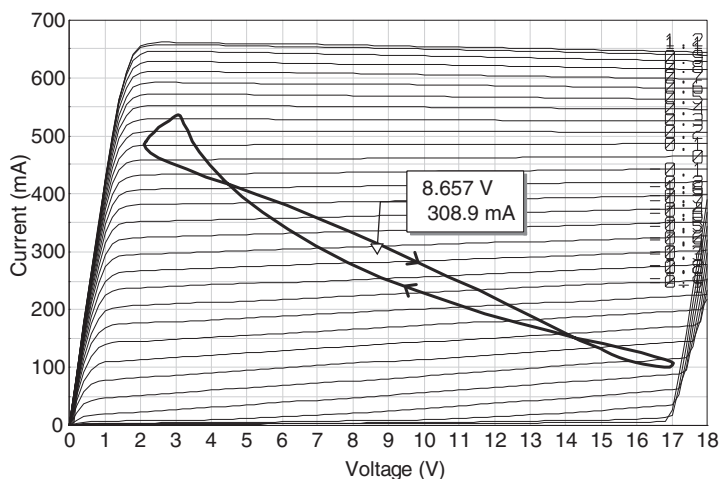


FIGURE 1.31 Dynamic load line of 2.4-GHz LP1500 pHEMT amplifier with $P_{\text{LOAD}} = 1\text{W}$.

A simple test for understanding the Lissajous pattern concept is to plot the time dependence of the input voltage and input current to a L or C element at a single frequency, then plot the input current versus the input voltage, also in the time domain. This gives a circle (or ellipse) which is CW for a capacitor and CCW for an inductor. This simple test will give an intuitive understanding of the nonlinear reactive impedance of the transistor in both amplifiers and oscillators. These techniques are especially applicable to high-power amplifiers and oscillators, where nonlinearities produce saturation effects of important significance. Several examples of this concept will occur throughout this book.

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PROBLEMS

- 1.1** Given a superheterodyne AM receiver for high fidelity (Fig. P1.1), design the gain of each stage for linear operation (multiple solutions are possible):

$$P_{\text{in}} = -90 \text{ dBm} \quad f_{\text{in}} = 1 \text{ MHz (FM radio)} \quad P_{\text{out}} = 10 \text{ W} \quad (1.50)$$

The gain of each bandpass filter is -1 dB and of each mixer is -6 dB .

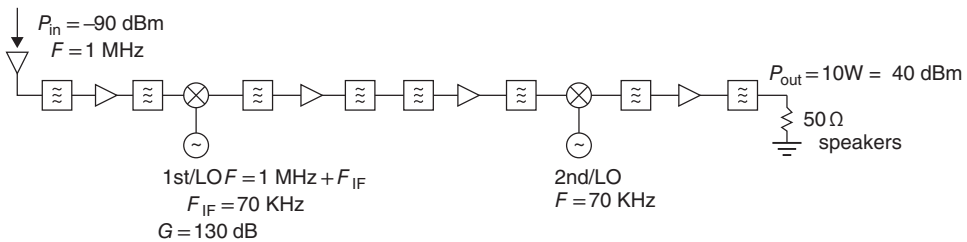


FIGURE P1.1 Superheterodyne receiver design.

- 1.2** For a high-power amplifier, the dynamic range is given below. Find the SFDR and the DR:

$$\text{MDS}_{\text{out}} = -60 \text{ dBm} \quad (1.51)$$

$$\text{BW} = 10 \text{ MHz} \quad (1.52)$$

$$\text{Transducer gain} = 30 \text{ dB} \quad (1.53)$$

$$P_{\text{1dBc}} = 40 \text{ dBm} \quad (1.54)$$

$$\text{TOI} = 50 \text{ dBm} \quad (1.55)$$

Refer to Figures 1.22 and 1.24.

- 1.3 Consider a substrate with $h = 10$ mils and $\epsilon_r = 10$ (alumina). Find the f_{TE1} cutoff frequency and explain its significance. What thickness would you need for operation at 94 GHz with no effects from surface modes?
- 1.4 Explain why digital communications has finally overtaken analog communications in the modern world.
- 1.5 If a crystal radio receiver is designed for 1 MHz, what are some values for L and C (see Fig. 1.2)? Take L to be about 10 μH .
- 1.6 Given a three-stage LNA (low-noise amplifier) with

$$\begin{aligned} F_1 &= 2.0 \text{ dB} & G_{A1} &= 10 \text{ dB} \\ F_2 &= 4.0 \text{ dB} & G_{A2} &= 12 \text{ dB} \\ F_3 &= 6.0 \text{ dB} & G_{A3} &= 14 \text{ dB} \end{aligned}$$

Find the total noise figure in dB and the T_e in kelvin.

- 1.7 Derive the SFDR from Figure 1.22 using simple geometry (similar triangles).
- 1.8 Given the modern 6-GHz heterodyne receiver shown in Figure P1.2, find the total TOI and the total NF of this receiver.

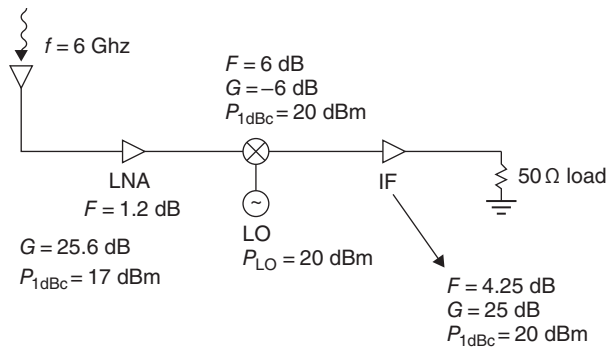


FIGURE P1.2 Modern 6-GHz heterodyne receiver.