
INTRODUCTION TO WIRELESS CIRCUIT DESIGN

1.1 INTRODUCTION

Wireless circuits are not that different from commonly known two-way radio, television, and broadcast arrangements. Some of them require high linearity in modulation (TV picture); some work via relay stations (two-way radio). The real differences lie in the fact that the cell sizes are much smaller, and that in most cases we attempt multiple channel use (reuse) using time-division multiplex, spread spectrum, or some other efficient means of reducing the bandwidth required for communication. One can argue that the wireless circuits include simple devices, such as garage-door openers and wireless keys for automobiles (we have seen many cases where strong interfering signals prevented the car owners from reclaiming their cars until the interfering signal disappeared). Another longtime favorite is cordless telephones: initially, 50-MHz models with essentially no privacy protection; later, more sophisticated models that operate at 900 MHz; and now, dual-band designs that use 900 MHz and 2.4 GHz.

The largest wireless growth area is probably the cellular telephones. The two major applications are the handsets, common referred to as cell phones or occasionally as “handies,” and the base stations. The base stations have many more problems with large-signal-handling linearity at high power, although handset users may run into similar problems. An example of this is the waiting area of an airport, where many travelers are trying to conduct last-minute business; in one instance, we concluded that about 30% of all the people present were on the air! It would have been fun to evaluate this receiver-hostile environment with a spectrum analyzer.

From such use comes anxiety factors, the lesser of which is “When will my battery die?”—a spare battery tends to help—and the greater of which the ongoing question, “Will this cell-phone transmitter harm my body?” A brief comment for the self-proclaimed experts in this area: a 50–100-kW TV transmitter, specifically its video or picture portion, connected

to a high-gain antenna, emits levels of energy in line-of-sight paths that by far exceed the pulsed energy from a cell phone. Specifically, the duration of energy is significantly smaller, and the absolute energy is more than a thousandfold higher, than the RF supposedly harming us from the cellular phone. Hand-held two-way radios have been used for the last 30 years or so by police and other security interests, operating in the frequency range from 50 to 900 MHz with antennas close to the users' heads, and there are no known cases of cancer or any other illnesses caused by these handheld radios. Recent studies in England, debatably or not, showed that the reaction-time level of people using cell phones drastically actually increased—but then there are always the skeptics and politically motivated who ignore the facts, try to influence the media, and have their 15 min of fame (as Andy Warhol used to say).

The question if cellphone radiation is harmful for the user is a question of ongoing research. However, the issue is quite complicated, and the results of different studies are even contradictory, or hard to reproduce. Luckily, Professor James C. Lin, from the University of Illinois-Chicago, takes the effort to write review articles on recent studies in his series *Health Effects* in the *IEEE Microwave Magazine* since 2001. This series provides us with the respective information and it is not only comprehensive but also written for engineers. From the vast number of articles, we just cite the first [1] and the one on the multinational study of the possible relation of tumors and cell phone use [2]. The other articles are easily found by a IEEEXplore database search.

Concerning radiation, Figure 1.1 shows the simulated near-field radiation of a Motorola mobile phone. The antenna is hidden inside the phone for optical reasons, and it is most likely a radiating structure that looks quite different from the linear antennas that were used in the past. The field is mainly concentrated at the top of the phone, where it is unlikely that



Figure 1.1 Antenna near-field radiation of a mobile phone (courtesy Prof. D. Manteuffel, Univ. Kiel, Germany).

the user's hand attenuates the transmit and receive power, and it is also directed away from the head. If the user will find a "warm" sensation, it will have more to do with the efficiency of the RF power amplifiers heating up the case than the effect of radiation, especially when many frontends are in use, like GSM, GPS, and WiFi.

With this introduction in place, we will first take a look at a typical UHF/SHF transceiver and explain the path from the microphone to the antenna and back. After this, we will inspect the radio channel and its effect on various methods of digital modulation. Analysis of wireless receivers and transmitters will be next, followed by a look at available building blocks and how they affect the overall system. To validate proper system operation, a fairly large number of measurements and tests must be performed, and conveying their purpose and importance will necessitate the definition of a number of system characteristics and concepts, such as dynamic range. Finally, after this is done, we will look at the issue of wireless system testing. Again, we intend to give guidance applicable to battery-operated, hand-held operation as well as high-powered base stations.

1.2 SYSTEM FUNCTIONS

A cellular telephone is a hybrid between a double-sideband and FM (PM) transceiver. The actual transmission is not continuous, but is pulsed, and because of the pulse spectrum there is a signal bandwidth concern due to keying transients, not unlike intermodulation products of an SSB transceiver cluttering up adjacent channels. The cellular telephone is also a linear transceiver in the sense that its signal-handling circuitry must be sufficiently amplitude- and phase-linear to preserve the modulation characteristics of the AM/PM hybrid emissions it transmits and receives. Containing such an emission's spectral regrowth, which affects operation on adjacent channels, is not unlike the linearity requirements we encounter in single-sideband (SSB) transceivers—requirements so stringent that amplifiers must be run nearly in Class A to meet them. The time division multiple access (TDMA) operating mode, which allows many stations to use the same frequency through the use of short, precisely timed transmissions, requires a system that transmits with a small duty cycle, putting much less thermal stress on a power amplifier than continuous operation. Power management, including a sleep mode, is another important issue in handset design.

Figure 1.2 shows the block diagram of a hand-held transceiver. This example shows an example chipset for a mobile phone that can operate in four 2G GSM (800/900 and 1800/1900 MHz) bands and in one 3G WCDMA (UMTS) band. Many functionalities are integrated into a few chips.

For those not too familiar with transceivers, here is a "walk" through the block diagram. The RF signal is intercepted by the antenna is fed to a switch that selects whether the GSM or WCDMA path is in use.

If we follow the GSM path first, the signal is fed to a diplexer/switch (called switch-plexer in the figure). Since GSM operates transmits in bursts, the RF front end is switched between transmit and receive. Then follows a band-selection filter for each of the four GSM bands, and low-noise amplifiers. The signal is then directly downconverted to baseband by a quadrature demodulator using a local oscillator signal generated on chip. The baseband signal is low-pass filtered, amplified, and converted to the digital domain for further processing.

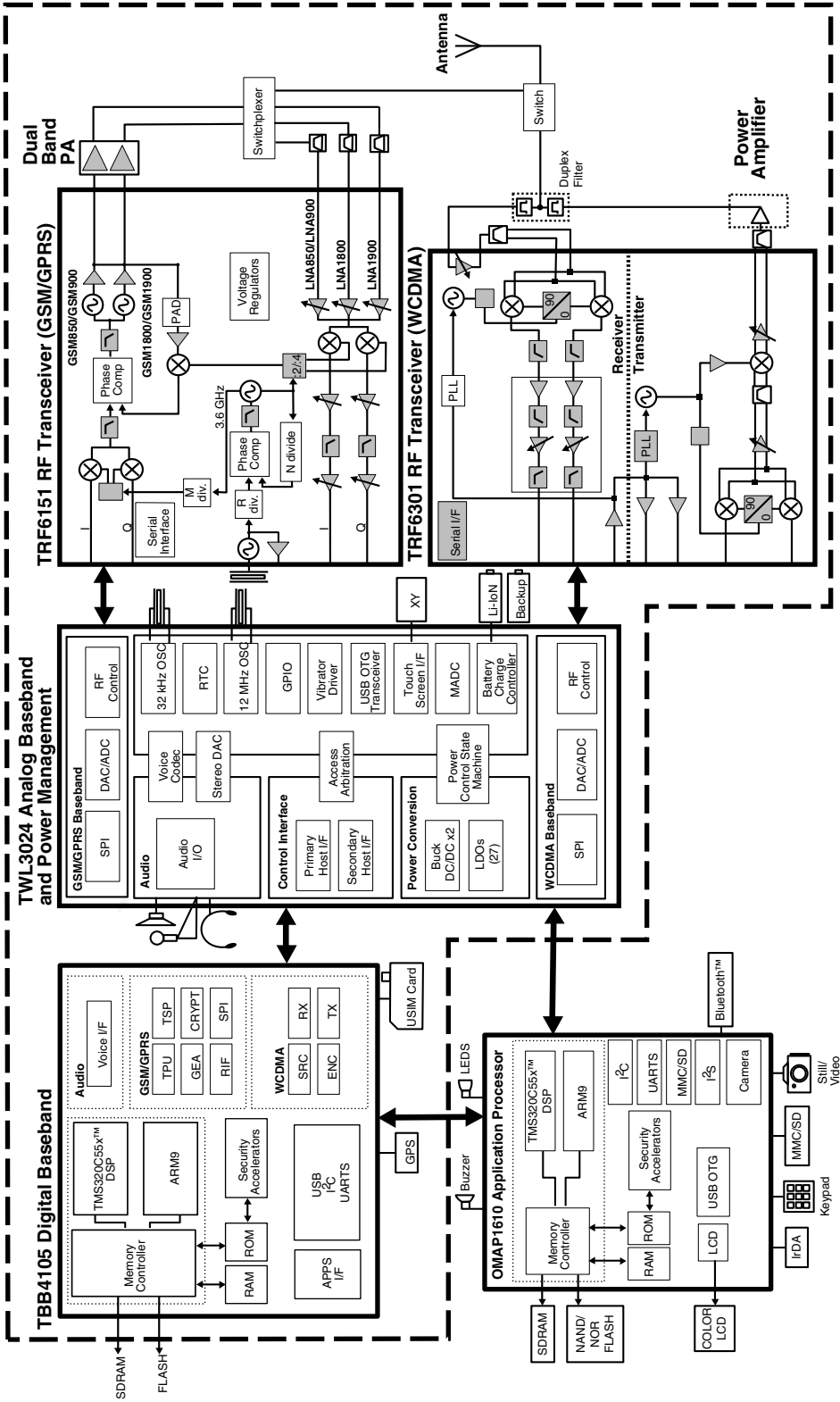


Figure 1.2 Block diagram of a handheld cellular telephone transceiver (courtesy Texas Instruments).

The WCDMA path is different from the GSM path at first sight since it requires a duplex filter instead of the switch/duplexer used in GSM. The reason is that in UMTS, the transmitter is in full-duplex operation and transmits and receives at the same time. The duplex filter therefore has to separate the two paths, especially it has to prevent crosstalk at the transmit frequency that could overload the input or the receiver. Again there is a combination of band-selection filter and low-noise amplifier before the received signal is downconverted into the baseband, and its I and Q channels are converted to the digital domain.

A nice overview about DSP in “readable” form is Ref. [3].

The GSM transmit portion consists of an on-chip synthesizer that is modulated. Both receive and transmit frequencies are controlled by a miniature temperature-compensated crystal oscillator (TCXO). The output of the voltage-controlled oscillator (VCO) is then amplified and fed to the antenna through the same switch/duplexer as the receive portion.

The WCDMA transmit branch relies on the I-Q modulator architecture. In contrast to GSM it uses the GMSK coding scheme with constant envelope, WCDMA signals show high peak-to-average ratios and need good control over amplitude and phase of the RF signal.

Figure 1.2 also reveals which part of the functionality can be integrated in CMOS today. In fact, it is almost everything, so let us talk about what is still off-chip and what will not be integrated in CMOS anytime soon.

- The antenna, for obvious reasons. Unlike early mobile phones, radiating parts are integrated inside the phone that do not even look like classical antennas.
- Band-selection filters need very low insertion loss and high selectivity. The low loss is required to obtain high sensitivity. This type of high- Q filters are commonly realized as surface-acoustic wave (SAW) or bulk acoustic resonance (BAR) filters on piezo crystals.
- The antenna switch must be able to carry the high transmit power, provide low insertion loss, and high isolation. Besides of pin diodes, GaAs HEMT devices are often used.
- The power amplifiers need to provide high powers, high linearity, and a maximum of power-added efficiency. The technology of choice to date is InGaP/GaAs HBTs.

A mobile phone transmitter is more involved than, say, a WLAN or DECT transmitter, as it requires to operate at multiple frequencies according to different standards. And since it needs to operate over much wider distances and at higher powers. Less critical transceivers might be fully integrated in CMOS or in BiCMOS. However, an in-depth discussion of different transceiver architectures is beyond the scope of this book, and we stop at this point. A nice overview of different architectures is found in Ref. [4].

1.3 THE RADIO CHANNEL AND MODULATION REQUIREMENTS

1.3.1 Introduction

The transmission of information from a fixed station to a mobile is considerably influenced by the characteristics of the radio channel. The RF signal arrives at the receiving antenna not only on the direct path but is normally reflected by natural and artificial obstacles in its way. Consequently, the signal arrives at the receiver several times in the form of echoes that are

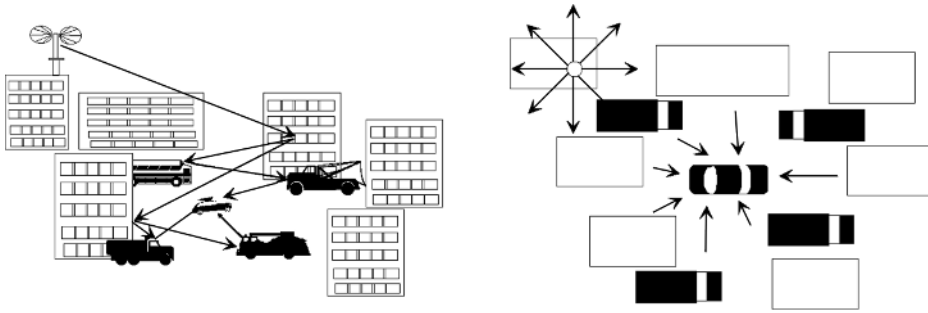


Figure 1.3 Mobile receiver affected by fading.

superimposed on the direct signal (Figure 1.3). This superposition may be an advantage as the energy received in this case is greater than in single-path reception. This feature is made use of in the DAB single-frequency network. However, this characteristic may be a disadvantage when the different waves cancel each other under unfavorable phase conditions. In conventional car radio reception, this effect is known as fading. It is particularly annoying when the vehicle stops in an area where the field strength is reduced because of fading (e.g., at traffic lights). Additional difficulties arise when digital signals are transmitted. If strong echo signals (compared to the directly received signal) arrive at the receiver with a delay in the order of a symbol period or more, time-adjacent symbols interfere with each other. In addition, the receive frequency may be falsified at high vehicle speeds because of the Doppler effect so that the receiver may have problems to estimate the instantaneous phase in the case of angle-modulated carriers. Both effects lead to a high symbol error rate even if the field strength is sufficiently high. Radio broadcasting systems using conventional frequency modulation are hardly affected by these interfering effects. If an analog system is replaced by a digital one that is expected to offer advantages over the previous system, it has to be ensured that these advantages—for example, better AF S/N and the possibility to offer supplementary services to the subscriber—are not at the expense of reception in hilly terrain or at high vehicle speeds because of extreme fading.

For this reason, a modulation method combined with suitable error protection has to be found for mobile reception in a typical radio channel, which is immune to fading, echo, and Doppler effects.

With a view to this, more detailed information on the radio channel is required. The channel can be described by means of a model. In the worst case, which may be the case for reception in built-up areas, it can be assumed that the mobile receives the signal on several indirect paths but not on a direct one. The signals are reflected, for example, by large buildings; the resulting signal delays are relatively long. In the vicinity of the receiver, these paths are split up into a great number of subpaths; the delays of these signals are relatively short. These signals may again be reflected by buildings but also by other vehicles or natural obstacles like trees. Assuming the subpaths being statistically independent of each other, the superimposed signals at the antenna input cause considerable time- and position-dependent field-strength variations with an amplitude obeying the Rayleigh distribution (Figures 1.4 and 1.5).

If a direct path is received in addition, the distribution changes to the Rice distribution and finally, when the direct path becomes dominant, the distribution follows the Gaussian distribution with the field strength of the direct path being used as the center value.

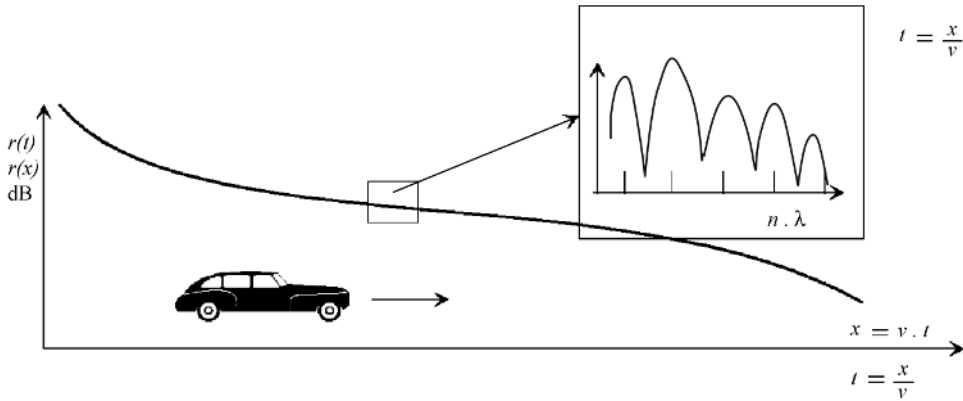


Figure 1.4 Receive signal as a function of time or position.

In a Rayleigh channel, the bit error rate increases dramatically compared to the BER in an additive white Gaussian noise (AWGN) channel produces (Figure 1.6).

1.3.2 Channel Impulse Response

This scenario can be demonstrated by means of the channel impulse response. Let us assume that a very short pulse of extremely high amplitude [in the ideal case a Dirac pulse $\delta(t)$] is sent by the transmitting antenna at a time $t_0 = 0$. This pulse arrives at the receiving antenna direct and in the form of reflections with different delays τ_i and different amplitudes because of path losses. The impulse response of the radio channel is the sum of all received pulses

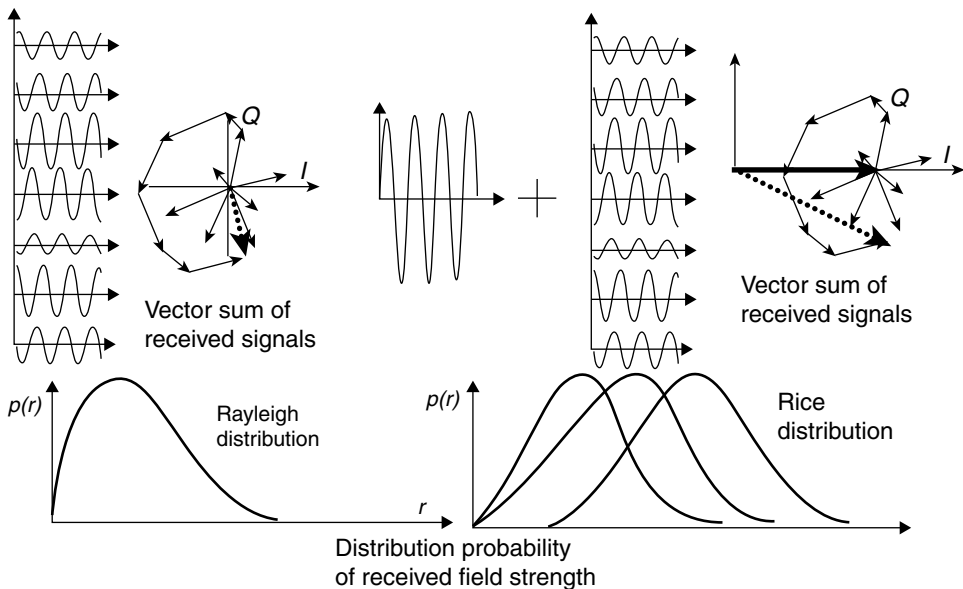


Figure 1.5 Rayleigh and Rice distribution.

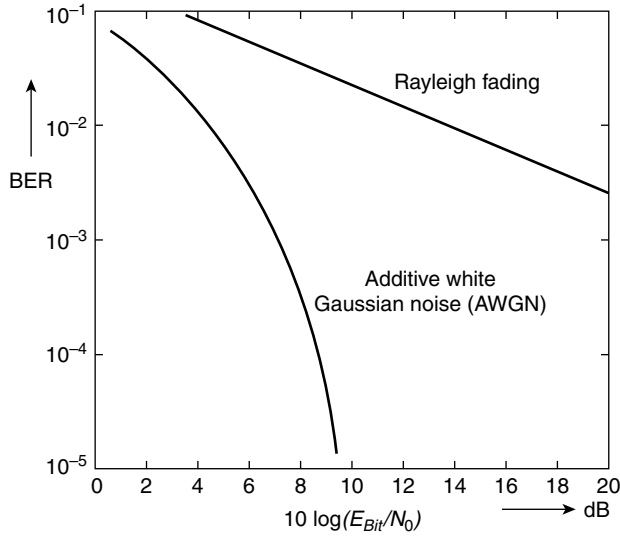


Figure 1.6 BER in a Rayleigh channel.

(Figure 1.7). Since the mobile receiver and also some of the reflecting objects are moving, the channel impulse response is a function of time and of delays τ_i , that is, it corresponds to

$$h(t, \tau) = \sum_N a_i \delta(t - \tau_i) \quad (1.1)$$

This shows that delta functions sent at different times t cause different reactions in the radio channel.

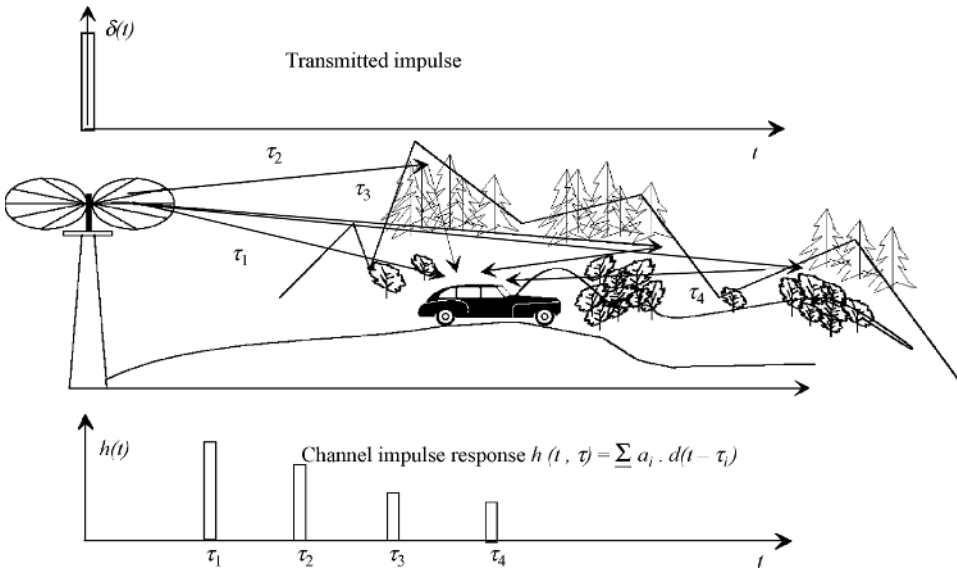


Figure 1.7 Channel impulse response.

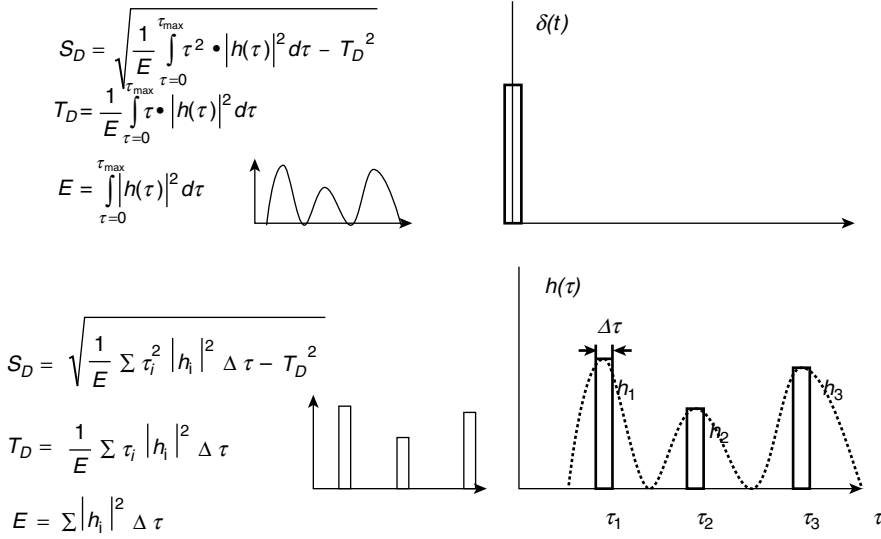


Figure 1.8 Calculation of delay spread.

In many experimental investigations, different landscape models with typical echo profiles were created.

The most important are

- rural area (RA),
- typical urban area (TU),
- bad urban area (BA), and
- hilly terrain (HT).

The channel impulse response informs on how the received power is distributed to the individual echoes. A parameter, the “delay spread” can be calculated from the channel impulse response, permitting an approximate description of typical landscape models (Figure 1.8).

The delay spread also roughly informs on the modulation parameters carrier frequency, symbol period, and duration of guard interval, which have to be selected in relation to each other. If the receiver is located in an area with a high delay spread (e.g., in hilly terrain), echoes of the symbols sent at different times are superimposed when broadband modulation methods with a short symbol period are used. In the case of DAB, this problem is aggravated by the use of single-frequency networks. An adjacent transmitter emitting the same information on the same frequency has the effect of an artificial echo (Figure 1.9).

A constructive superposition of echoes is only possible if the symbol period is much greater than the delay spread. The following holds:

$$T_s > 10T_d \quad (1.2)$$

This has the consequence that relatively narrowband modulation methods have to be used. If this is not possible, channel equalizing is required.

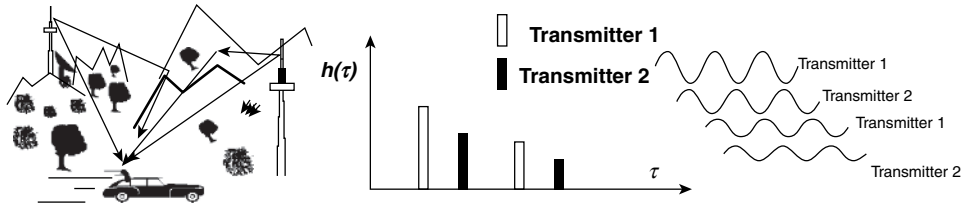


Figure 1.9 Artificial and natural echoes in the single-frequency network.

For the channel equalizing, a continuous estimation of the radio channel is necessary. The estimation is performed with the aid of a periodic transmission of data known to the receiver. In networks according to the GSA standards, a midamble consisting of 26 bits—the training sequence—is transmitted with every burst. The training sequence corresponds to a characteristic pattern of I/Q signals that is kept in a memory in the receiver. The baseband signals of every received training sequence are correlated with the stored ones. From this correlation, the channel can be estimated, the properties of the estimated channel will then be fed to the equalizer (Figure 1.10).

The equalizer uses the Viterbi algorithm (maximum sequence likelihood estimation) for the estimation of the phases that most likely have been sent at the sampling times. From these phases, the information bits are calculated (Figure 1.11). A well-designed equalizer then will superimpose the energies of the single echoes constructively, so that the result in an area, where the echoes are not too much delayed, delay times up to $16 \mu\text{s}$ have to be tolerated by a receiver, are better than in an area with no significant echoes (Figure 1.12).

Remaining bit errors are eliminated using another Viterbi decoder at the transmitter convolutionally encoded data sequences.

The ability of a mobile receiver to work in an hostile environment such as the radio channel with echoes must be proven. The test is performed with the aid of a fading simulator.

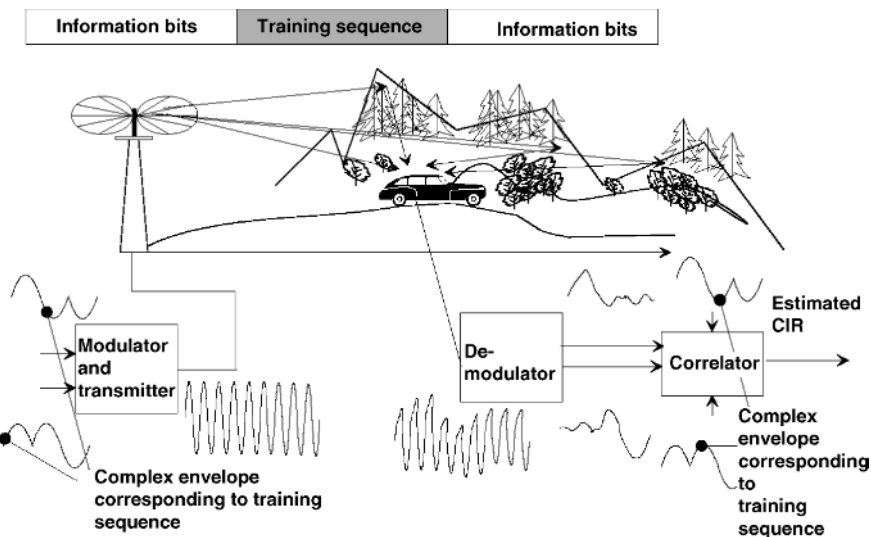


Figure 1.10 Channel estimation.

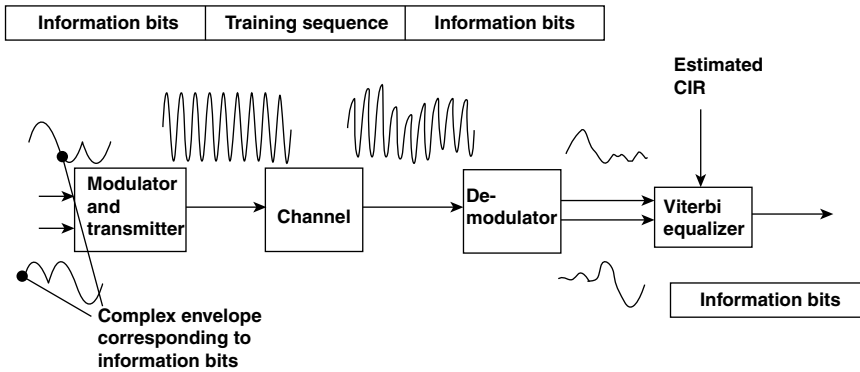


Figure 1.11 Channel equalization.

The fading simulator simulates different scenarios with different delay times and different Doppler profiles. A signal generator generates undistorted I/Q modulated RF signals that are downconverted into the baseband. Here, the I/Q signals are digitized and split into different channels where they are delayed and attenuated and where Doppler effects are superimposed. After combination of these distorted signals at the output of the baseband section of the simulator, these signals modulate the RF carrier that is the test signal for the receiver under test (Figure 1.13).

To make the tests comparable, GSM recommends typical profiles, for example

- rural area (RAx),
- typical urban (TUx), and
- hilly terrain (HTx).

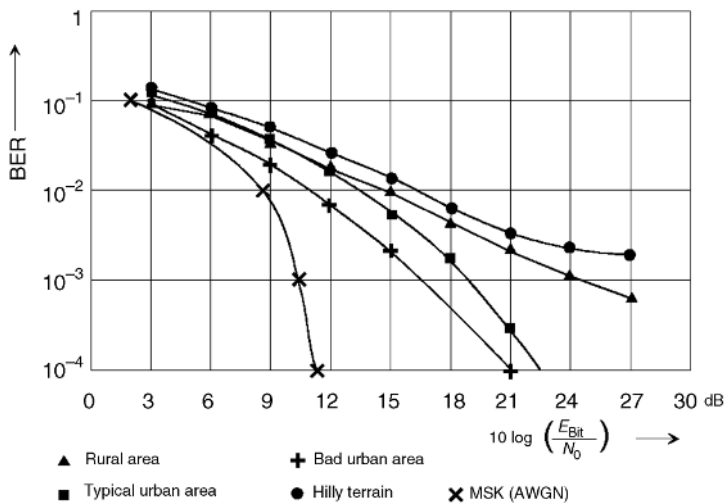


Figure 1.12 BERs after the channel equalizer in different areas.

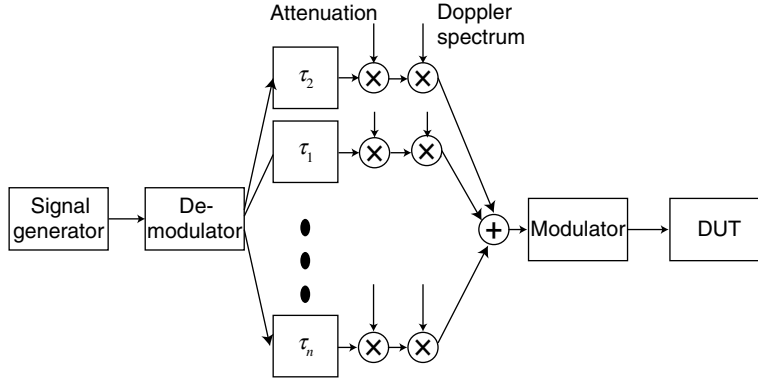


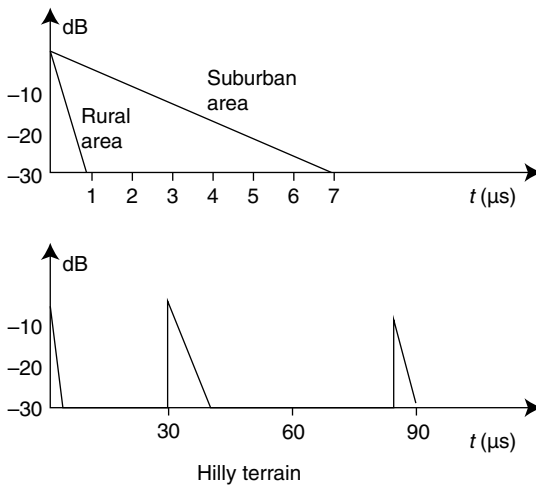
Figure 1.13 Fading simulator.

where number and strengths of the echoes and the Doppler spectra are prescribed (Figure 1.14).

1.3.3 Doppler Effect

Since the mobile receiver and some of the reflecting objects are in motion, the receive frequency is shifted because of the Doppler effect. In the case of single-path reception, this shift is calculated as follows:

$$f_d = \frac{v}{c} f_c \cos \alpha \quad (1.3)$$



$$P_i(t) = \begin{cases} \exp\left(\frac{-t}{0.1}\right) & \text{for } 0 < t < 0.1 \mu\text{s} \\ 0 & \text{elsewhere} \end{cases}$$

Rural
area

$$P_i(t) = \begin{cases} \exp\left(\frac{-t}{1}\right) & \text{for } 0 < t < 7 \mu\text{s} \\ 0 & \text{elsewhere} \end{cases}$$

Suburban
area

$$P_i(t) = \begin{cases} 0.33 \cdot \exp\left(\frac{-t}{0.3}\right) & \text{for } 0 < t < 2 \mu\text{s} \\ 0.5 \cdot \exp\left(\frac{-t}{0.0}\right) & \text{for } 30 < t < 42 \mu\text{s} \\ 0.17 \cdot \exp\left(\frac{-t}{0.0}\right) & \text{for } 80 < t < 85 \mu\text{s} \\ 0 & \text{elsewhere} \end{cases}$$

Hilly terrain

Figure 1.14 Typical landscape profiles.

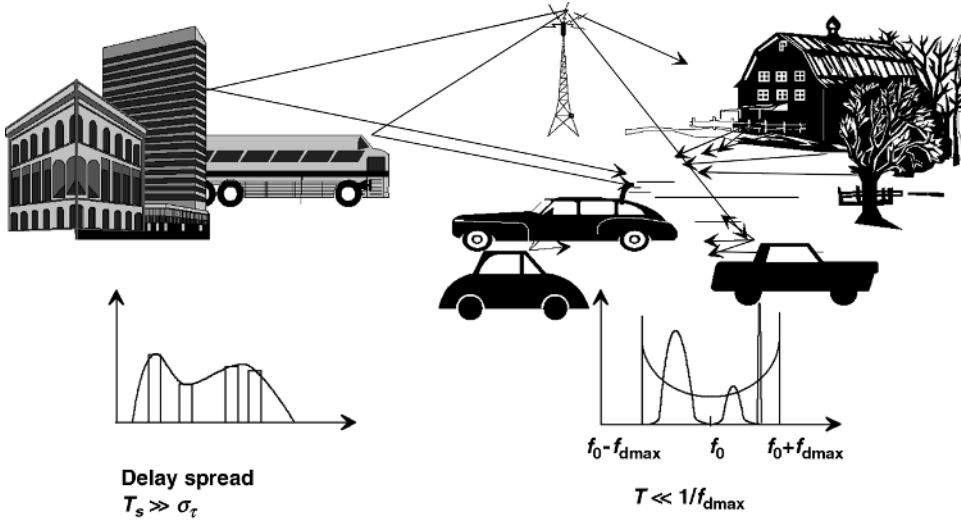


Figure 1.15 Doppler spread.

where v = speed of vehicle, c = speed of light, f = carrier frequency, and α = angle between v and the line connecting transmitter and receiver.

In the case of multipath reception, the signals on the individual paths arrive at the receiving antenna with different Doppler shifts because of the different angles α_i , and the receive spectrum is spread. Assuming an equal distribution of the angles of incidence, the power density spectrum can be calculated as follows:

$$P(f) = \frac{1}{\pi} \frac{1}{\sqrt{f_d^2 - f^2}} \text{ for } |f| < |f_d| \quad (1.4)$$

where f_d = maximum Doppler frequency.

Of course, other Doppler spectra are possible in addition to the pure Doppler shift described above, for example, spectra with a Gaussian distribution using one or several maxima. A Doppler spread can be calculated from the Doppler spectrum analogously to the delay spread (Figure 1.15).

1.3.4 Transfer Function

The FFT value of the channel impulse response is the transfer function $H(f, t)$ of the radio channel, which is also time dependent. The transfer function describes the attenuation of frequencies in the transmission channel. When examining the frequency dependence, it will be evident that the influence of the transmission channel on two sine-wave signals of different frequency becomes greater with increasing frequency difference. This behavior can be adequately described by the coherence bandwidth that is approximately equal to the reciprocal delay spread, that is

$$(\Delta f)_c = \frac{1}{T_d} \quad (1.5)$$

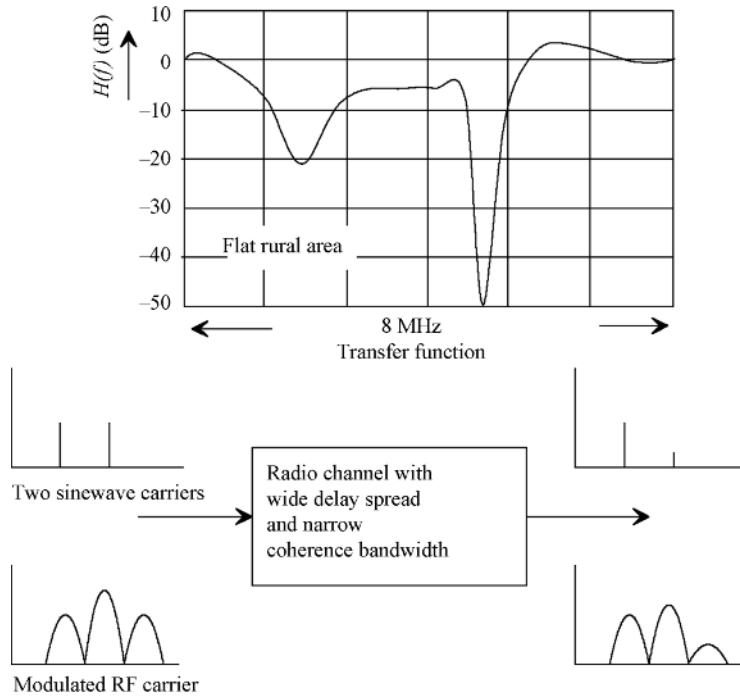


Figure 1.16 Effect of transfer function on modulated RF signals.

If the coherence bandwidth is sufficiently wide and, consequently, the associated delay spread is small, the channel is not frequency selective. This means that all frequencies are subject to the same fading. If the coherence bandwidth is narrow and the associated delay spread wide, even very close adjacent frequencies are attenuated differently by the channel. The effect on a broadband-modulated carrier with respect to the coherence bandwidth is obvious. The sidebands important for the transmitted information are attenuated to a different degree. The result is a considerable distortion of the receive signal combined with a high bit error rate even if the received field strength is high. This characteristic of the radio channel again speaks for the use of narrowband modulation methods (Figure 1.16).

1.3.5 Time Response of Channel Impulse Response and Transfer Function

The time response of the radio channel can be derived from the Doppler spread. It is assumed that the channel rapidly varies at high vehicle speeds. The time variation of the radio channel can be described by a figure, the coherence time, which is analogous to the coherence bandwidth. This calculated value is the reciprocal bandwidth of the Doppler spectrum. A wide Doppler spectrum therefore indicates that the channel impulse response and the transfer function vary rapidly with time (Figure 1.17). If the Doppler spread is reduced to a single line, the channel is time invariant. In other words, if the vehicle has stopped or moves at a constant speed in a terrain without reflecting objects, the channel impulse response and the transfer function measured at different times are the same.

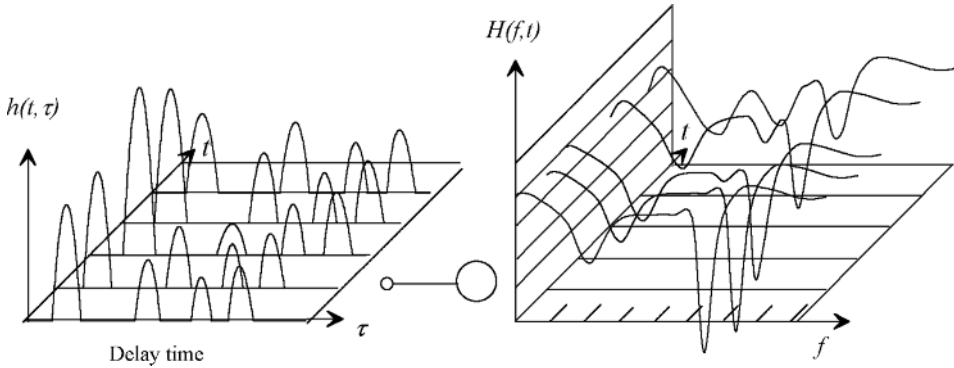


Figure 1.17 Channel impulse response and transfer function as a function of time.

The effect on information transmission will be illustrated in an example. In the case of MPSK modulation using hard keying, the transmitter holds the carrier phase for a certain period of time, that is, for the symbol period T . In the case of soft keying with low-pass-filtered baseband signals for limiting the modulated RF carrier, the nominal phase is reached at a specific time, the sampling time. In both cases, the phase error $\phi_f = f_d T_s$ is superimposed onto the nominal phase angle, which yields a phase uncertainty of $\Delta\phi = 2\phi_f$ at the receiver. The longer the symbol period the greater the angle deviation (Figure 1.18). Considering this characteristic of the transmission channel, a short symbol period of $T_s \ll (\Delta t)c$ should be used. However, this requires broadband modulation methods.

Figure 1.19 shows the field strength or power arriving at the mobile receiver if the vehicle moves in a Rayleigh distribution channel. Since the phase depends on the vehicle position, the receiver moves through positions of considerably differing field strength at different times (time dependence of radio channel). In the case of frequency-selective channels, this applies to one frequency only, that is, to a receiver using a narrowband IF filter for narrowband emissions. As Figure 1.19 shows, this effect can be reduced by increasing the bandwidth of the emitted signal and consequently the receiver bandwidth.

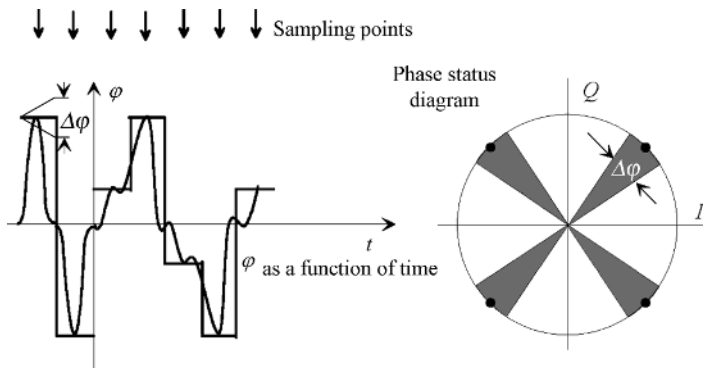


Figure 1.18 Phase uncertainty caused by Doppler effect.

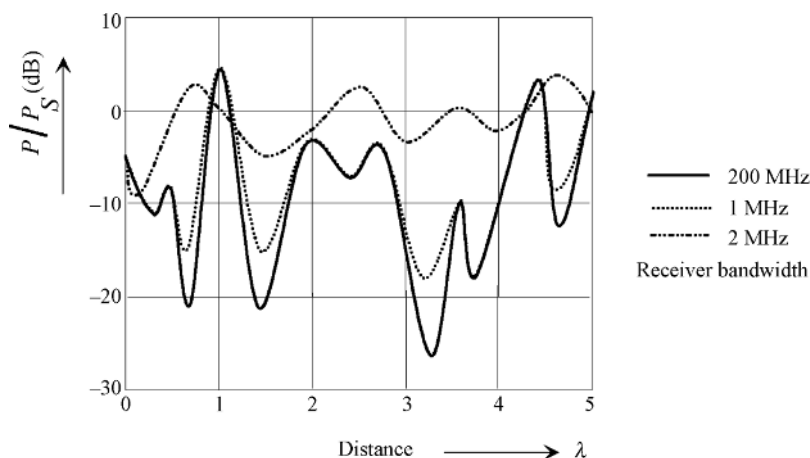


Figure 1.19 Effect of bandwidth on fading.

1.3.6 Lessons Learned

The strongly frequency-selective channel causing inadmissible distortion of the broadband-modulated carrier and channel impulse responses like those expected in a hilly terrain speak in favor of narrowband modulation methods with long symbol periods. In hilly terrain, extensively delayed echoes cause intersymbol interference when broadband modulation with short symbol periods is used. On the other hand, narrowband modulation has the disadvantage that the signals arrive at the receiver considerably attenuated and reception may be interrupted for an indefinite period of time. The burst errors occurring in digital information transmission cannot be corrected even with the most elaborate error protection methods.

These transmission interruptions can be avoided by using broadband modulation methods, but these are sensitive to strongly frequency-selective channels. Since broadband modulation is obtained through the use of short symbol periods, broadband modulation is unsuitable when greatly delayed echoes are expected.

It remains to be defined when a signal is considered a narrowband and when a broadband signal. This question shall be answered with the aid of an example. Apart from extremely narrowband analog modulation methods as are used for sound broadcasting in the longwave, mediumwave, and shortwave bands, FM sound broadcasting transmissions in the VHF bands are narrowband. In the case of digital modulation, this means that transmissions with a rate of 400 kbit/s modulated onto a carrier with a bandwidth efficiency of 1.5 (bit/s)/Hz over a bandwidth of approximately 300 kHz can be regarded as narrowband transmissions according to the definition above. Consequently, a DAB signal with this gross bit rate would be a narrowband signal and suitable for transmission on the radio channel with restrictions only. Based on the experience with conventional FM broadcasting systems in large cities and hilly terrain, this was obvious from the very beginning.

Consequently, it is necessary to find ways for spreading the band artificially without reducing the bandwidth efficiency. This means that a large band must be available for the transmission of several programs, the full bandwidth being used by all the programs without mutual interference.

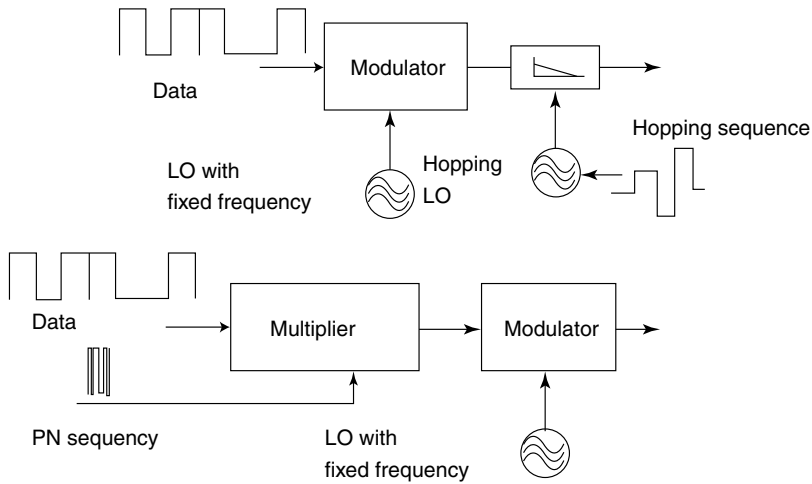


Figure 1.20 Band spreading by frequency hopping and CDMA.

Several approaches can be adopted to tackle the problem (Figure 1.20). One way would be a continuous change of the transmit and receive frequency according to a defined pattern (frequency hopping). This method is used in mobile radio, for instance, but only marginal investigations have been made in this respect for DAB.

Another possibility is to multiply the symbols of the individual programs with digital signals (pseudonoise function) using a much higher bit rate so that a higher symbol rate is obtained. In this case, the different programs are assigned different functions that must be orthogonal to each other (code division multiple access, CDMA). The “chopped” bit streams of the individual programs are modulated onto carriers of identical frequency and the modulated carriers are added. A correlation receiver knowing the pseudonoise function divides the incoming CDMA signal into the individual programs. The disadvantage is obvious. Symbol periods are very short and elaborate means will be required for compensating the intersymbol interference.

A different approach has been chosen for DAB, which does not involve continuous and elaborate channel measurements, so that it would be possible to use favorably priced receivers as are demanded in the field of consumer electronics. The method used for DAB is a multicarrier method where the information to be transmitted is spread onto many carriers using time and frequency interleaving. The terms time and frequency interleaving will be explained in the course of this discussion. The result is a broadband transmission method with long symbol periods. However, certain limitations caused by the Doppler effect will have to be accepted particularly at high carrier frequencies.

1.3.7 Wireless Signal Example: The TDMA System in GSM

1.3.7.1 Frequency Division Multiple Access (FDMA)

In analog radio systems, the trend has always been toward a more efficient utilization of the available frequency spectrum by reducing the channel spacing. The number of radio channels obtained at a channel spacing of 12.5 kHz is of course twice that obtained at 25 kHz. However, any improvement brings about its disadvantage: the narrower the channel spacing,

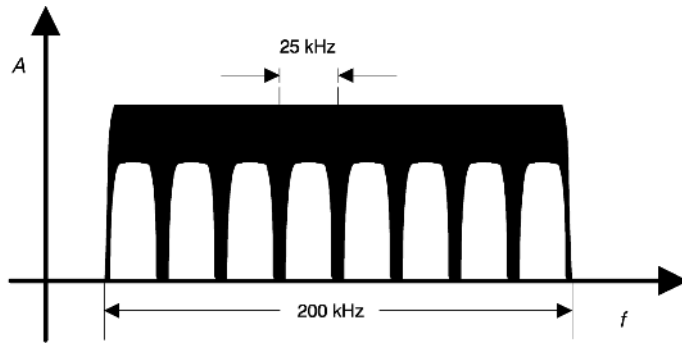


Figure 1.21 Channel spacing in broadband/narrowband systems.

the higher the required frequency accuracy and the lower the possible maximum deviation of the frequency modulation. The latter leads to a poorer transmission quality due to the lower S/N ratio. Furthermore, the gaps between the channels, which must be a number of kilohertz wide for safety reasons, also reduce the available system bandwidth (see Figures 1.21 and 1.22).

The use of an available system spectrum divided into individual frequency channels enables the user to simultaneously access a multitude of different frequencies. This multiple access is called frequency-division multiple access (FDMA). Consequently, all radio systems with a spectrum divided into channels are FDMA systems. At present, the technically useful limit is reached with a channel spacing of 10–12.5 kHz.

Advantages of FDMA.

- Simultaneous access to a given bandwidth by many subscribers.
- Increase in the number of channels through reduction of channel spacing.

Disadvantages of FDMA.

- Higher frequency accuracy required.
- Transmission quality decreasing with reduction of channel bandwidth.
- Better rejection filters required.
- One transmitter/receiver required per channel.

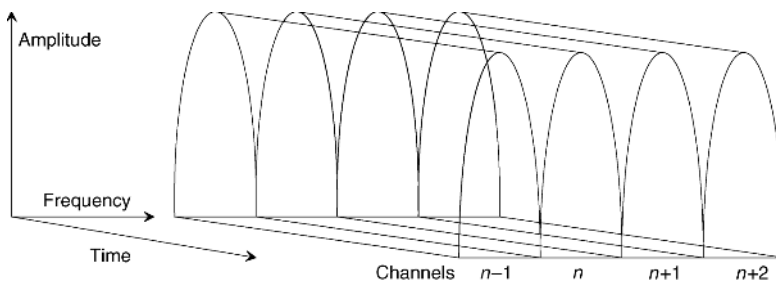


Figure 1.22 Frequency-division multiple access (FDMA).

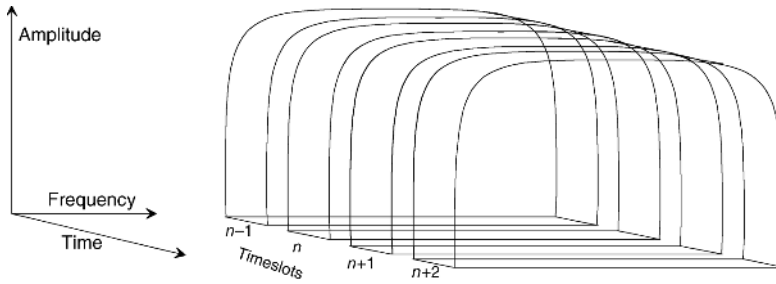


Figure 1.23 Time-division multiple access (TDMA).

1.3.7.2 Time-Division Multiple Access (TDMA)

With TDMA systems, the available bandwidth is divided into considerably fewer, and therefore wider, channels than in FDMA systems. Each of these channels is available to several subscribers quasi-simultaneously (see Figure 1.23). However, a given subscriber can use the whole channel for a very short period (timeslot) only, for the rest of the time, they have no access. This serial access of several users is repeated within a fixed time frame.

Advantages of TDMA.

- Simultaneous use of a specific bandwidth by a great number of subscribers.
- Depending on the number of available timeslots, several subscribers can be served by one transmitter/receiver.
- Transmitter and receiver are not permanently switched on (saves battery power).
- The RF section may carry out other tasks in the intervals between transmission and reception.
- Reduced susceptibility to frequency-selective fading in the case of larger channel bandwidths.

Disadvantages of TDMA.

- Accurate time synchronization of subscribers is required.
- Higher processor capacity is required.
- Broadband modulators are required.

1.3.7.3 Code-Division Multiple Access (CDMA)

The increasing use of low-priced and powerful signal processors allows a less common technique of multiple access to be employed in mass communication systems. In the case of code-division multiple access (CDMA), the whole system bandwidth is available to all subscribers at any time; that is, all send and receive simultaneously, with each using a specific code (Figure 1.24).

Logic “1” represents a certain bit sequence, logic “0” is the inversion of this sequence. The different signals are distinguished in the receiver by means of a cross-correlation of the received signal, which comprises a great number of codes, with the bit sequence expected so that the desired transmission signal can be detected.

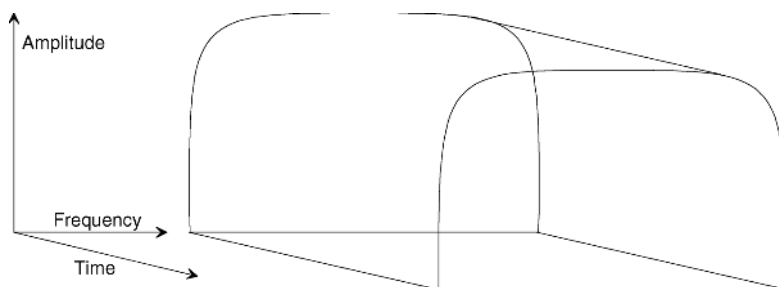


Figure 1.24 Code-division multiple access (CDMA).

Advantages of CDMA.

- Simultaneous use of a specific channel or subband many subscribers.
- Several signals can be received simultaneously by one receiver.
- Reduced susceptibility to frequency-selective fading in the case of large channel bandwidths.
- More subscribers can be served.
- Reduced costs for radio network planning.

Disadvantages of CDMA.

- Accurate time synchronization of subscribers required.
- Fast transmitter power control over a wide dynamic range.

1.3.7.4 TDMA in GSM

RF Data In spite of the competition with other mobile radio systems, a common frequency band could be defined for GSM worldwide. All operators who signed the GSM memorandum of understanding committed themselves to install their GSM systems within the standardized frequency range. The competition for frequencies mainly affects countries using NMT900, the frequency range of which corresponds to the GSM P band. TACS also partly overlaps the GSM P band; the G1 band is completely within the TACS range. Cordless telephones operating in accordance with the CT1 standard also use the upper end of the GSM P band. CT1+ telephones, which had been assigned a frequency range below the P band years ago to protect them against GSM, have now been ousted by the G1 band. See Table 1.1.

Since each frequency channel is divided into eight timeslots, transmitter and receiver operate in an intermittent mode, with the receive and transmit time in the upper and lower channels shifted by three timeslots (Figure 1.25). Although this alternative sending and receiving scheme operates in half-duplex rather than full-duplex operation, the received signal sounds continuous to the user.

1.3.7.5 TDMA Structure

Frame and Multiframe All GSM radio channels are organized in frames of approximately 4.62 ms duration. The frames are continuously repeated. Each frame is divided into

Table 1.1 RF Data for GSM900 and GSM1800

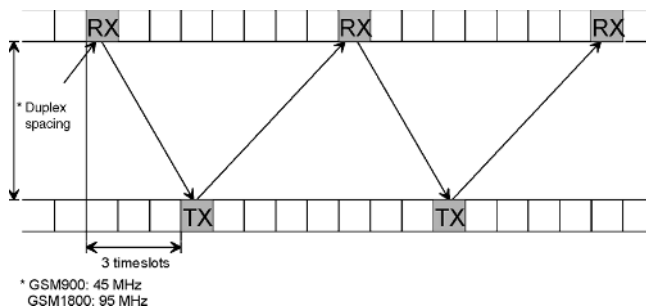
	GSM900		GSM1800
Frequency range	P band	G1 band	
Uplink (MHz)	890–915	880–890	1710–1785
(MS transmitting)			
Downlink (MHz)	935–960	925–935	1805–1880
(BTS transmitting)			
Duplex spacing (MHz)	45	45	95
Spectrum (MHz)	2×25	2×10	2×75
Frequency channels	124	49	374
Channel numbers	1–124	975–1023	512–885
(ARFCN)			
Channel spacing		200 kHz	
Modulation		GMSK with $B \times T = 0.3$	
Data transmission rate		270.833 kbit/s	
Bit duration		$3.69 \mu\text{s}$	

eight timeslots of approximately $577 \mu\text{s}$ each. A timeslot contains an information packet, the burst. Twenty-six-type multiframes are used on all timeslots containing a traffic channel (voice and/or data), 51-type multiframes on all timeslots reserved for control channels. See Figure 1.26.

The TDMA structure uses other frame types above the multiframe level, as shown in Table 1.2.

Table 1.2 Superframes and Hyperframes

Superframe	$= 51 \times 26$ frames $= 1326$ frames $= 6.12$ s
Hyperframe	$= 2048$ superframes $= 2.715.648$ frames $= 3$ h, 29 min, 3.5 s

**Figure 1.25** Duplex spacing of transmission and reception.

Multiframes: 51/26 multiframe

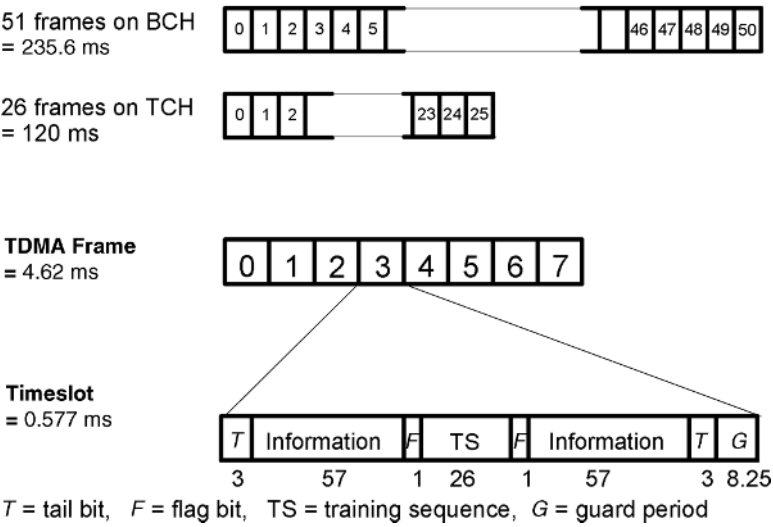


Figure 1.26 Timeslot, frame, and multiframe.

TDMA Timers The frame number within the hyperframe is counted continually so that counting of the TDMA clock restarts after approximately 3.5 h. The frame number therefore represents a time unit in the GSM system. Similar to time counting, where the seconds are combined into minutes, hours, and days, GSM does not count the absolute frame numbers but uses timers instead. These timers are structured as follows as shown in Table 1.3.

The absolute frame number is obtained by a multiplication of the three timers. However, on certain occasions, a short version of the timers is used.

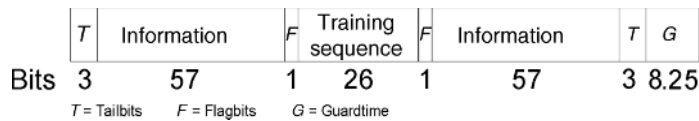
Burst Structures Information between base station and mobile is sent in the timeslots. In each slot, a certain amount of information—that is, a burst—can be transmitted. Normally, the timeslot is occupied by the normal burst (Figure 1.27), which is used for signaling as well as for voice and data transmission.

Each part of the burst serves a specific purpose as described below.

Information Bits The normal burst is able to transmit 2×57 information bits. Since this information can be replaced every 4.62 ms during an ongoing call, the average theoretical

Table 1.3 TDMA Timers

$T1 = FN \text{ div. } (26 \times 51)$	Range	0–2047
$T2 = FN \text{ mod } 26$	Range	0–25
$T3 = FN \text{ mod } 51$	Range	0–50
FN (Frame number)	Range	0....2715647
$FN_{\text{max}} = 51 \times 26 \times 2048 - 1$		

**Figure 1.27** Normal burst.

transmission rate is

$$114 \text{ bits} \div 4.62 \text{ ms} \approx 24.7 \text{ kbit/s} \quad (1.6)$$

At the same time, this rate also represents the maximum transmission rate that can be obtained in the GSM system using one timeslot per transmission per time frame. Consequently, the transmission rate could be increased only if more than one time slot is used for the transmission.

The bit rate is much lower in the control channels; that is, the above transmission rate is only attained by the mobile station if a traffic channel has been set up. In this case, the base station and the mobile station use a signaling channel, which also uses up capacity, in addition to the voice or data channel. Table 1.4 shows the assignment of the theoretically available capacity with a traffic channel set up.

Training Sequence In the middle of the normal burst, a 26-bit training sequence, the bit sequence of which is known to the receiver, is sent. The training-sequence code (TSC) can be one of eight different sequences. These sequences are stored in all receivers, and at the beginning of a transmission, the base transceiver station (BTS) decides on the TSC to be used. The training sequence serves two main purposes: bit synchronization and estimation of channel impulse response.

1.3.7.6 Bit Synchronization

Data transmitted via the air interface are in the asynchronous mode; that is, the receiver has to regenerate the bit clock from the data stream. To enable synchronization in the receiver, the transmitters adds synchronization bits to the information stream. Therefore, in normal data transmission, data telegrams start with a "...10101010..." sequence so that the receiver can regenerate the bit clock. A predefined bit word informs the receiver when the actual information (block synchronization) starts. A receiver synchronized in this way is able

Table 1.4 Transmission Bit Rates

	Information	Error Protection	Total
Traffic channel			22.8 kbit/s
Voice (full-rate)	13.0 kbit/s	9.8 kbit/s	
Data	2.4 kbit/s	20.4 kbit/s	
	4.8 kbit/s	18.0 kbit/s	
	9.6 kbit/s	13.2 kbit/s	
Control channels			0.95 kbit/s
Idle frame			0.95 kbit/s
Total			24.7 kbit/s

to decode the data stream online. The training sequence of the burst has to assume both synchronization tasks. Since it is in the middle of the burst, direct decoding is not possible.

Each burst must first be stored in the receiver and then decoded by postprocessing. The reason for using this method is the second task of the training sequence. Synchronization itself is carried out by means of cross-correlation; that is, the expected training sequence is compared (correlated) to the center of the received burst and to the beginning and end of the training sequence so that the bit clock is also known. A burst containing other than the expected training sequence cannot be synchronized and decoded.

1.3.7.7 Compensation of Multipath Reception

The signal from the transmitter (in Figure 1.28, BTS $\rightarrow$ MS; the same applies also in the opposite direction) arrives at the receiver not only along the direct path but also via various other paths as a result of reflection and diffraction caused by obstacles in the signal path.

Propagation conditions on these additional paths differ from those on the direct path. For instance, we can expect signals traveling via additional paths to exhibit

- longer travel times because of increased path length,
- various strengths, and
- different Doppler shifts.

Because of the different travel times, the signals arrive with a different phase at the receiving antenna. Depending on this phase, components may be canceled—that is, they may totally disappear—or added so that a high-quality signal is received for only a short period of time. RF-level variations are statistically distributed; level shifts due to fading may be as great as 40 dB.

In addition to RF-level fading, another annoying effect is encountered that, uncompensated, would make correct signal decoding rather difficult. Because of the additional distance, the signal travels via the indirect path, the signal arriving at the receiving antenna exhibits time delay of its modulation in addition to variable phase shifts. The total of all channel responses to a single transmitted pulse is called channel impulse response (CIR,

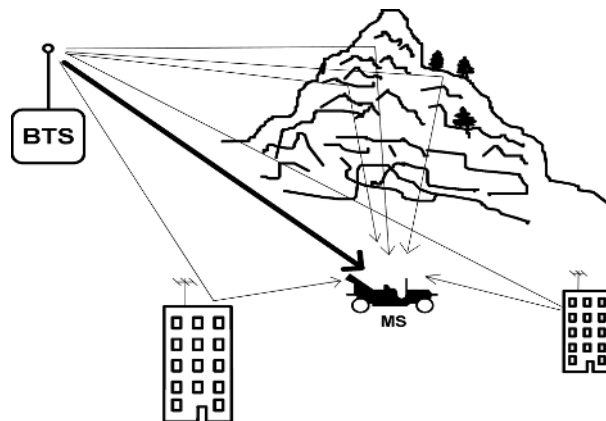


Figure 1.28 Multipath reception due to reflection and diffraction. The base transceiver station (BTS) is transmitting to the mobile station (MS).

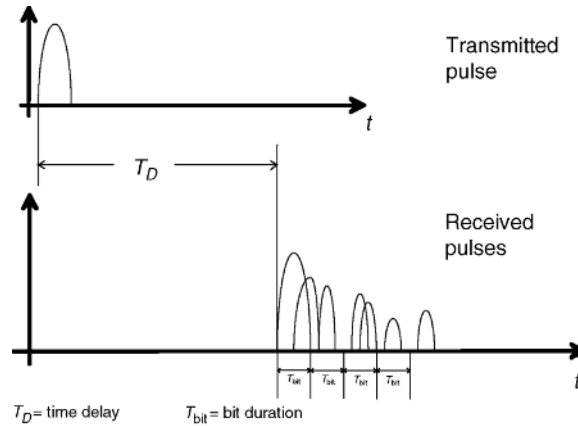


Figure 1.29 Channel impulse response.

Figure 1.29). If the indirect path is only 1 km longer, the GSM echo bit reaches the receiver later than the directly received bit and thus interferes with the next bit received. This intersymbol interference may occur over several bits in succession. With delays of up to 15 μs , differentiating the desired-signal components from echoes becomes more and more difficult. This problem can also be solved with the aid of the training sequence. The echoes on the delayed paths also contain the training sequence. The correlation used for detecting the original training sequence may also be used for detecting the training-sequence echoes as well as their delay and loss. With the aid of this information, the received signal can be corrected by a channel equalizer.

Guard Period Transmission in each timeslot is terminated with a guard period (Figure 1.30) of 8.25 bit periods ($\approx 30 \mu\text{s}$). During this time, the level of the burst must be reduced from nominal to a minimum value (by up to 70 dB) and the burst is modulated with so-called dummy bits (logic 1), that is, no information is transmitted. The user of the next

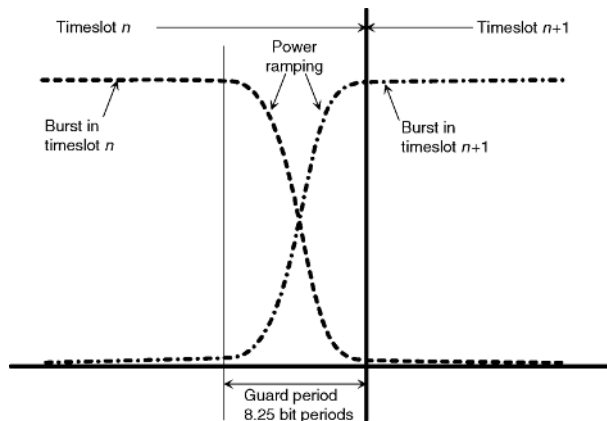


Figure 1.30 Guard period at the end of each timeslot.

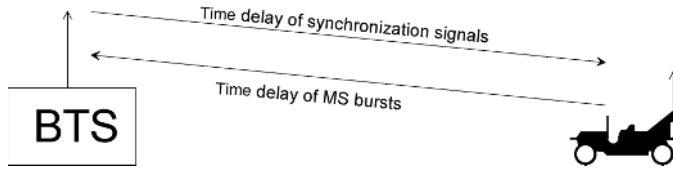


Figure 1.31 Signal delay and its effect.

timeslot should start sending during this guard period so that his burst has reached nominal power when the actual transmission in the timeslot starts. This means that the switching time that cannot be used for transmitting information is used twice.

Delay Correction The integrity of a timeslot depends on whether the subscribers send only during the period assigned to them and otherwise keep quiet. This is only possible when all subscribers are accurately synchronized. For practical reasons, the clock signal is generated by the BTS and the mobile stations synchronize to it. Conflicts with adjacent timeslots may occur in the uplink where several subscribers have to share the same channel. Although several subscribers are addressed in the downlink, signals can only be transmitted by the BTS.

We will now examine the effect of a distance of 10 km between an MS and a BTS (Figure 1.31). Synchronization of the MS is as follows.

- Delay over 10 km = $33.3 \mu\text{s}$ (distance $\div$ velocity of light).
- The sync signals from the BTS requires this time for transmission.
- MS is synchronized by $33.3 \mu\text{s}$ too late.
- MS sends a burst at the correct time from its point of view.
- The burst is sent $33.3 \mu\text{s}$ too late.

Signal delay over a distance of 10 km → the burst requires another $33.3 \mu\text{s}$. From the point of view of the BTS, the burst arrives with a delay of twice the delay time. Transmission cannot be made in the assigned timeslot and interferes with the next one.

The guard period at the end of each burst is only approximately $30 \mu\text{s}$ long and fully used in the example above. The greater the distance between MS and BTS, the greater the effect of the signal delay. The only way to solve this problem is to make the MS send the burst at an earlier time. To do so, the distance between MS and BTS must be known. The BTS determines the distance by means of a delay measurement and informs the MS of the actual delay. As a result, the MS corrects its transmission time so that the signals again arrive time synchronized with the other mobiles at the BTS antenna.

In GSM, this procedure is called timing advance (TA) and is carried out continuously for all active mobile stations. Every 480 ms, a new TA value is sent to all active mobiles. A few limiting values of the GSM system can be deduced from the TA.

The TA is transmitted as a 6-bit word. Numerals from 0 to 63 can be represented with 6 bits. For instance, a TA of 10 means that for an MS from which bursts arrive with a delay of 10 bit periods, transmission must be advanced by this amount. Since the TA is transmitted as an absolute value, a maximum delay of $232.5 \mu\text{s}$ ($63 \times 3.69 \mu\text{s}$) can be signaled and corrected. This maximum double delay corresponds to a distance of approximately 34.9 km.

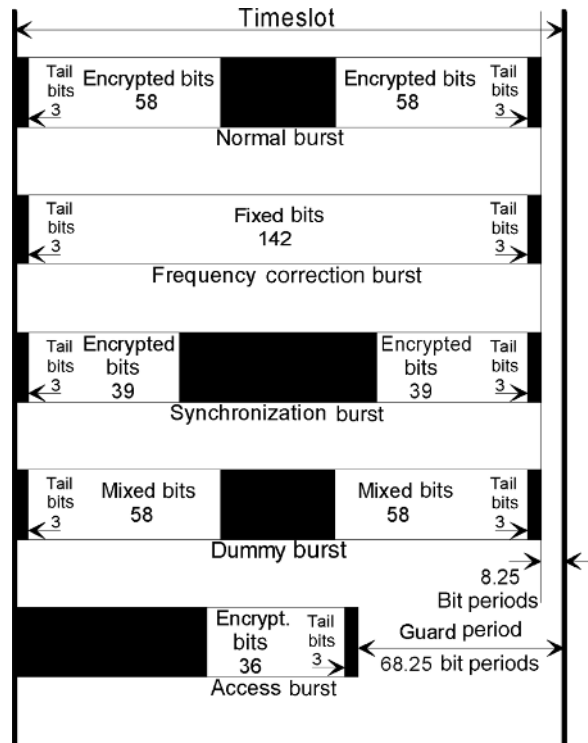


Figure 1.32 Burst types.

With the aid of the TA, the distance between the MS and the BTS can be determined. Since the smallest delay increment is a bit period, the resolution for a distance of one half of the delay corresponds to one bit period and therefore to approximately 550 m.

This synchronization scheme cannot solve an inherent problem. The delay of the very first burst sent by an MS cannot be corrected because, with contact between the MS and BTS yet to be established, the MS has not yet received an appropriate TA value from the BTS. Yet, if the mobile sends a normally timed burst, it may spill over into the next time slot and interfere with transmission from another MS. To avoid causing interference in such cases, the mobile uses a special burst at the beginning of a transmission or when no valid TA has been received. Called the access burst, it is considerably shorter than a normal burst. The next section describes the access burst and other additional burst types in greater detail.

Burst Types In addition to the normal burst described before, other bursts, including the access burst, are available for special purposes (Figure 1.32).

Frequency Correction Burst. The 142 “fixed bits” of the frequency correction burst (FCB) are all set to logic 0. With Gaussian minimum shift keying (GMSK), the type of modulation used in GSM, a stationary carrier frequency deviation, in this case approximately 67.7 kHz, is generated with this burst. The FCB is sent by the BTS only and used by the mobile for synchronization to the carrier frequency and for compensating a possible Doppler shift. It is sent by the BTS every 10

frames (approximately every 46 ms), but only in the timeslot 0 and only on a single (the C0) carrier.

Synchronization Burst. One frame after the frequency correction burst, but also in timeslot 0, the synchronization burst (SB) is transmitted. This burst is sent only by the BTS, and only on the C0 carrier. A particular difference between it and the normal burst is its considerably longer training sequence. Like the 26-bit training sequence of a normal burst, the SB's training sequence is also used for bit synchronization. Due to the great length of the sequence, synchronization can be more exact.

The twice 39 encrypted bits comprise timers T1, T2, and T3 in a coded form, and also the base-station identification code (BSIC). When this "GSM time" is received, the MS is synchronized to the BTS.

Dummy Burst. A BTS must send continuously—that is, in all timeslots—on its C0 carrier because this carrier is used by the MS to find the nearest BTS and for evaluating reception quality. If no normal burst is available for transmission in a timeslot, the BTS sends dummy bursts instead, as the carrier cannot be transmitted without a modulation signal. Only the BTS sends dummy bursts, and only on the C0 carrier.

Access Burst. As already pointed out, the access burst is sent when the MS first calls the BTS as a means of minimizing interference to other MTs while initiating a delay measurement and the determination of the TA. In most cases, the access burst is also sent on the C0 carrier in the uplink direction, but in the case of a handover it may be transmitted on any carrier.

1.3.8 From GSM to UMTS to LTE

With GSM, as with the other second generation (2G) systems, mobile communications became digital. This system was of a completely new quality as compared to its analog predecessors. As an early system, this standard is less complex than newer systems, and easier to comprehend. It still is a good introductory example in a book like this, since it is the basis from which modern standards were developed.

When GSM was developed, computing power was limited, especially regarding mobile devices. Designing the required RFICs, too, was all but an easy task. The standard therefore aimed at taking full advantage of the new possibilities that were offered by the digital world, like privacy protection through encryption, secure billing, and squeezing as many connections into a physical radio channel as possible. But it still is mainly a telephone system, as it offers one dedicated line for each call. Packet-switched data connections were not in the focus—in absence of reasonable mobile computer power and display technology, and considering the low data rates and high costs: a reasonable approach. However, short-message texting (SMS) was build into the system from the start. A low data-rate package-switched signaling channel was used to transfer these few bits.

On the positive side, in the view of an RF designer, GSM is a nice standard. Relying on the GSMK modulation scheme (see Section 1.4.2), the transmitted RF signal is always of constant amplitude. The information is encoded in the phase shift from symbol to symbol. Due to the narrowband phase modulation, the signal did not consume more bandwidth than in AM. These signals were sent out in bursts, as discussed in the previous section.

The advancement of the semiconductor technology enables much more complex systems nowadays. The driving force today is mobile data transmission, mainly access to the internet, e-mail, and social media, but also to entertainment such as stream video. Considering data throughput, voice is becoming negligible.

The third generation (3G) successor of GSM is UMTS. With this system, any connection between the mobile device and the base station is treated as a data session. Instead of defining a fixed number of possible channels, defined in bandwidth and time slots, it uses a more flexible scheme, based on wideband code-division multiple access, which will be defined in a subsequent section.

In future mobile systems, like the long-term evolution (3GPP-LTE), a standard that is currently being deployed, adds numerous new features on the system level. For example, variable bandwidth, enhanced localization, spatial multiplexing, and multiple-input-multiple-output (MIMO) schemes that use multiple receive and transmit antennas in order to enhance data throughput, or maximum ratio combining schemes that reduce bit error rate relying on multiple transmit antennas and a single receive antenna.

An outline addressing most of the details of current and emerging standards beyond GSM are way beyond the scope of this book, and the reader is referred to the literature. Fortunately, a number of good introductory application notes are available, for example, [6–8, 10, 16–18].

1.4 ABOUT BITS, SYMBOLS, AND WAVEFORMS

1.4.1 Introduction

Digital modulation of an RF carrier is the allocation of physically existing RF waveforms to the single elements of an alphabet of logical symbols where the number of allowed waveforms is equal to the number of logical elements of the alphabet (Figure 1.33). The most common alphabet is the binary one with the two logical symbols “0” and “1”, but we will also deal with quaternary, octernary, and hexadecimal alphabets or more generally with M -ary alphabets comprising many more elements when discussing the signal generation with signal generators and dedicated software packages. The waveforms representing these symbols differ from each other by their parameters amplitude $a(t)$, their frequency $f(t)$, and their phase $\phi(t)$.

A modulator therefore is nothing more than a device by which this allocation is performed (Figure 1.34). From a coder it receives the logical symbols and emits at its output the corresponding waveforms $s_i(t)$. The waveform generation may be done by using a set of distinct generators (e.g., two oscillators to generate two signals with different frequencies in the case of binary frequency shift keying), by classical amplitude or frequency modulators or by more sophisticated equipment such as I/Q -modulators for M -ary modulations.

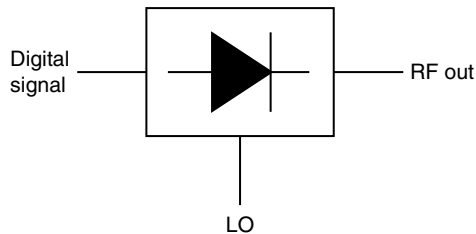


Figure 1.33 At base, digital modulation involves frequency-shifting a baseband digital signal to RF. In practice, the process is more complicated than this because of bandwidth constraints on the resulting RF signal.

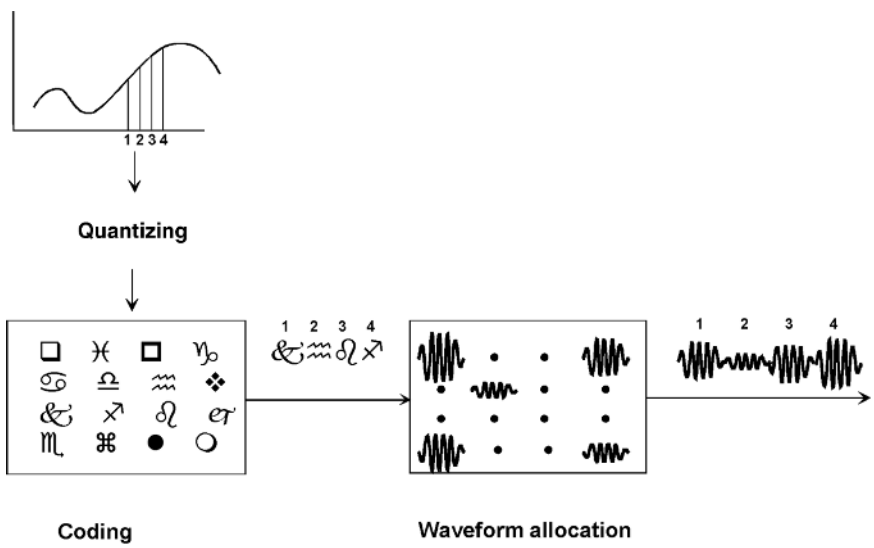


Figure 1.34 Digital modulator.

On their way across the RF channel from the transmitter to the receiver, these waveforms are distorted by noise and other disturbing properties of the RF channel.

The task of the receiver is to interpret the received waveforms $r_i(t)$ and to reallocate the proper logical symbols to them. For this purpose, it is not necessary to reconstruct the original waveforms from the distorted ones (Figure 1.35). The important thing is to find out

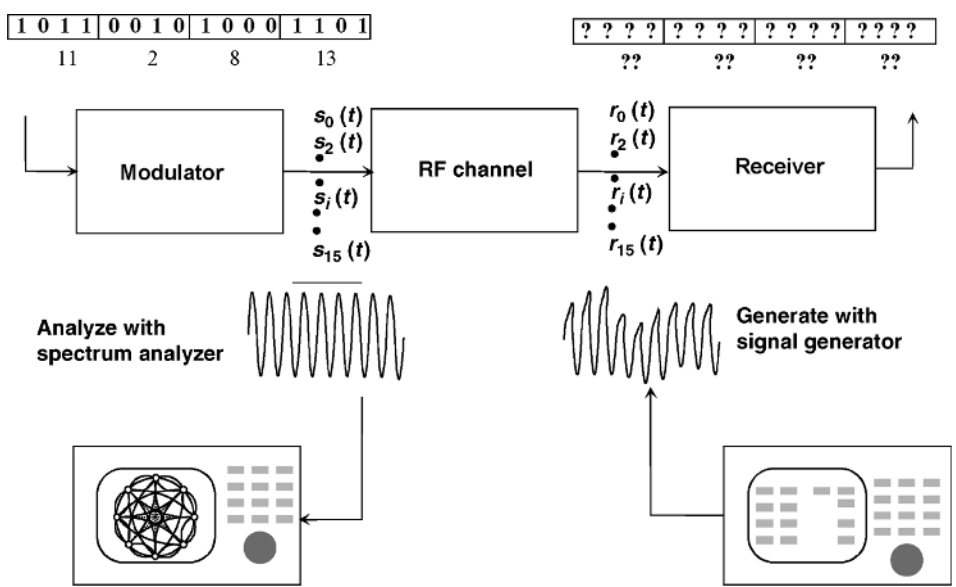


Figure 1.35 The information channel.

which symbol has most probably been sent when a certain signal $r_i(t)$ has been received, a process that is known as maximum likelihood estimation.

For meaningful receiver tests therefore waveforms have to be generated that mimic real, distorted signals to prove the ability of a receiver to tolerate waveform distortions to a certain extent.

1.4.1.1 Representation of a Modulated RF Carrier

The waveform of a modulated RF carrier can be expressed as

$$s(t) = a(t) \cos[2\pi f_c(t)t + \phi(t)] \quad (1.7)$$

and is defined by its amplitude $a(t)$, its carrier frequency $f_c(t)$, and its phase $\phi(t)$. All the three parameters are time variant and may be altered to generate different waveforms to represent logical symbols. If the occupied bandwidth of this modulated carrier is narrow compared to the carrier frequency f_c , we call this signal the RF-bandpass signal (Figure 1.36).

As any frequency variation causes a phase variation and vice versa a phase variation always causes a frequency variation, we can replace any frequency modulation by a corresponding phase modulation. Therefore, we simplify the above equation to

$$s(t) = a(t) \cos[2\pi f_c t + \phi(t)] \quad (1.8)$$

that is, we consider the carrier frequency as a constant and concentrate all frequency and phase variations into the parameter $\phi(t)$.

For our purposes, another representation is more suitable, we will have to look up some trigonometric identities and our formula processor finds that

$$\begin{aligned} a(t) \cos[2\pi f_c t + \phi(t)] &= \cos[\phi(t)]a(t) \cos(2\pi f_c t) \\ &\quad - \sin[\phi(t)]a(t) \sin(2\pi f_c t) \end{aligned} \quad (1.9)$$

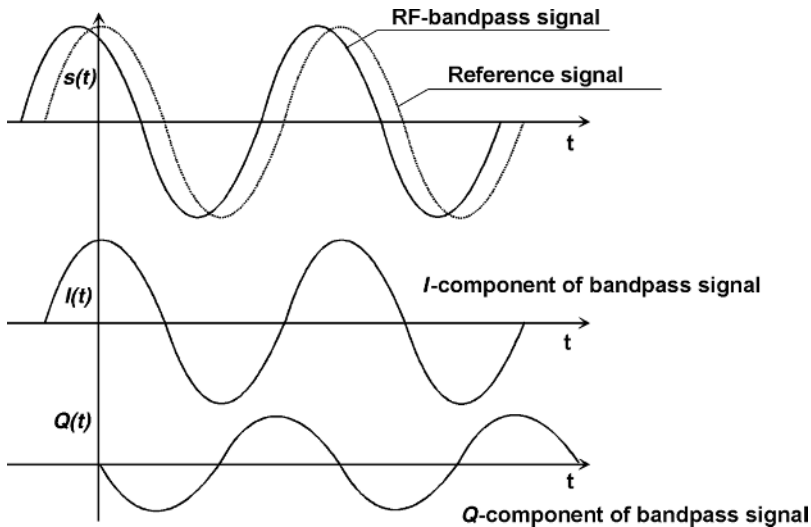


Figure 1.36 The bandpass signal and the I/Q representation of a carrier.

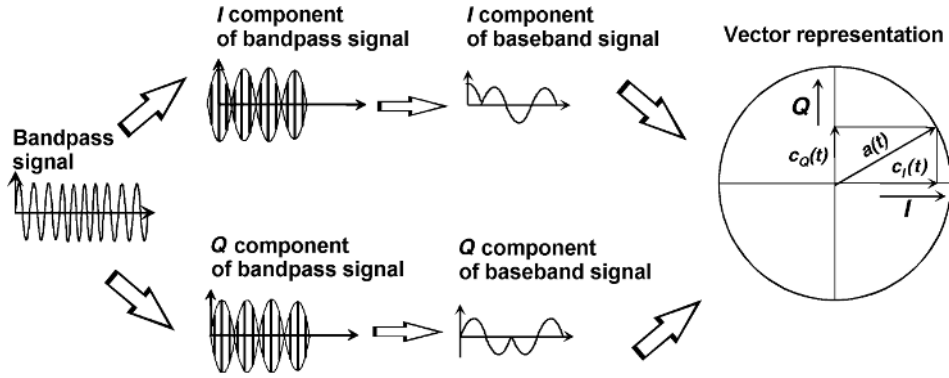


Figure 1.37 Different forms of signal representation.

which we call the I/Q representation of the RF-signal. I/Q means that we have an I (in phase) signal, namely, $\cos[\phi(t)]a(t)\cos(2\pi f_c t)$, and a Q (quadrature) signal, namely, $-\sin[\phi(t)]a(t)\sin(2\pi f_c t)$. These equations help us a lot in understanding an I/Q modulator. Because of the phase difference of 90° between the two carrier components, these are said to be orthogonal to each other.

All the information about the (modulated) carrier with the carrier frequency f_c is contained in the terms

$$c_I(t) = a(t)\cos[\phi(t)] \quad (1.10)$$

$$c_Q(t) = a(t)\sin[\phi(t)] \quad (1.11)$$

and, lazy as we are, we therefore disregard the terms $\cos(2\pi f_c t)$ and $-\sin(2\pi f_c t)$ for further considerations and denote the above signals $c_I(t)$ and $c_Q(t)$ as the components of the complex baseband waveform or baseband signal.

This leads us immediately to the vector representation of the signal, where we consider the two components $c_I(t)$ and $c_Q(t)$ of the complex baseband signal as the time-variant components of a time variant vector with the vector length $a(t)$ and the angle to the I -axis $\phi(t)$. We also get

$$a(t) = \sqrt{c_I^2(t) + c_Q^2(t)} \quad (1.12)$$

$$\phi(t) = \arctan\left(\frac{c_Q}{c_I}\right) \quad (1.13)$$

The vector can be depicted in the I/Q area (Figure 1.37).

Generation of the Modulated Carrier Once we have realized that the modulated carrier can be represented as the sum of its I and Q components, which are the product of the two baseband components with two orthogonal RF carriers of the same frequency, it is easy to understand the hardware of the modulator (Figure 1.38). An unmodulated RF carrier is split up into two equal oscillations $\cos(2\pi f_i)$, one of the two is then shifted by 0.5π , and therefore is described by $-\sin(2\pi f_i)$. The component $\cos(2\pi f_i)$ is multiplied with the I component of the baseband signal $c_I(t)$, the other one, $-\sin(2\pi f_i)$, is multiplied with the

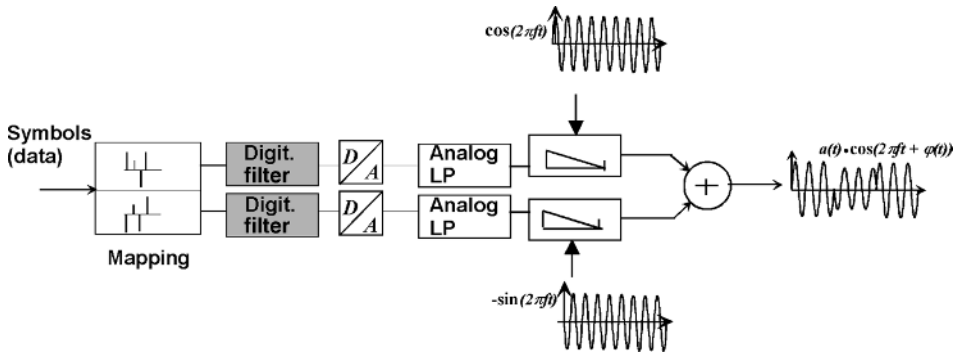


Figure 1.38 Principle of a digital I/Q modulator.

Q component $c_Q(t)$ of the baseband signal. Each multiplication may be performed using a double-balanced mixer. Afterward, the two RF-components are added in a simple power combiner. As it is difficult to shift the carrier by 90° over a broad frequency range, the modulated carrier is generated at an intermediate frequency and then upconverted to the wanted output frequency in a second mixer stage.

The baseband signals are generated by mapping every digital symbol into a pair of digital pulses that are fed to digital baseband filters. The output signal of these filters is D/A converted and smoothed by analog low-pass filters.

Figure 1.39 shows another example where, for a given modulation (MSK or GMSK), the instantaneous phase and then the corresponding cosinusoid and sinusoid, which modulate the two carrier components, are calculated from the data signal.

Digital designs of the modulator also exist, in which the IF-carrier generation, the time-variant phase shift, the multiplication with the baseband signals, and the sum of the components are calculated in a digital signal processor, the output of which is D/A -converted and upconverted to the output frequency in the classical way. A further possibility is the generation of the modulated carrier with direct digital synthesis (DDS), as it is used in the Rohde & Schwarz SME signal generator.

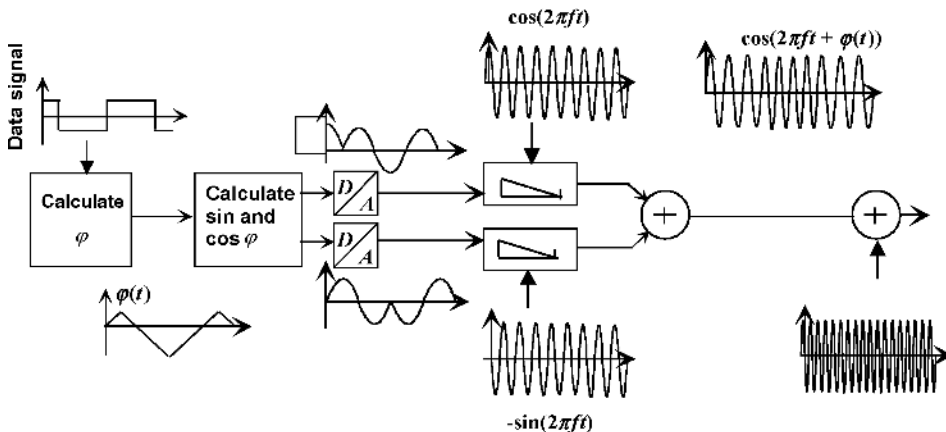


Figure 1.39 I/Q modulation (MSK and GMSK).

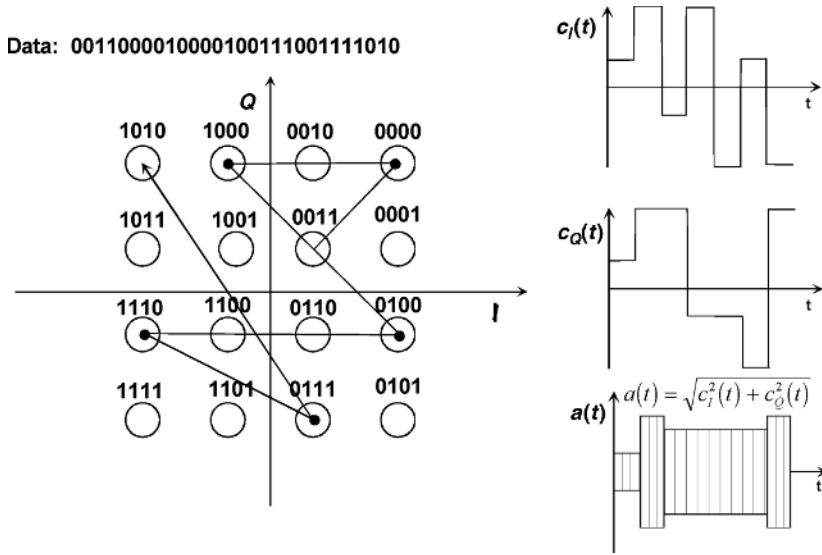


Figure 1.40 Constellation diagram, and baseband and RF signals of 16 QAM.

Mapping the Data into the Baseband Waveforms The next question is, “How do we generate the baseband waveforms $c_I(t)$ and $c_Q(t)$?” There is no general answer to this question, as the generation of the baseband waveforms depends on the type of modulation. The following short descriptions will suffice for the moment.

Linear modulations (all kinds of amplitude and phase-shift keying, and M -ary QAM).

- For binary amplitude and phase shift keying (ASK and BPSK), the data signal itself represented as a unipolar (ASK) or bipolar (BPSK) nonreturn-to-zero (NRZ) signal is the baseband waveform $c_I(t)$; the component $c_Q(t)$ does not exist.
- For M -ary phase shift keying and M -ary quadrature amplitude modulation, N bits are combined to form new symbols that are elements of an alphabet with $M = 2^N$ elements. In the simplest case, every symbol is allocated an I and a Q amplitude during the symbol duration, which is N times the bit duration. The modulating signals $c_I(t)$ and $c_Q(t)$ are then staircase functions, and the modulated carrier has a time-varying envelope with the instantaneous amplitude $a(t)$ (Figure 1.40). Because the steps of the envelope cause unwanted side lobes of the RF spectrum, the baseband signals are filtered to smooth the shape of the RF envelope and reduce the occupied bandwidth of the modulated RF signal.

Nonlinear modulations (frequency-shift keying, minimum shift keying and Gaussian minimum shift keying).

- Despite M -ary frequency shift keying (FSK) could be performed using an I/Q modulator, for this type of modulation much simpler equipment such as a voltage-controlled oscillator is used as a frequency modulator. Figure 1.41 shows an example of quaternary frequency shift keying (4FSK), which also is known as 4 PAM/FM. This term indicates that every two bits are combined to a dibit that is mapped into a baseband

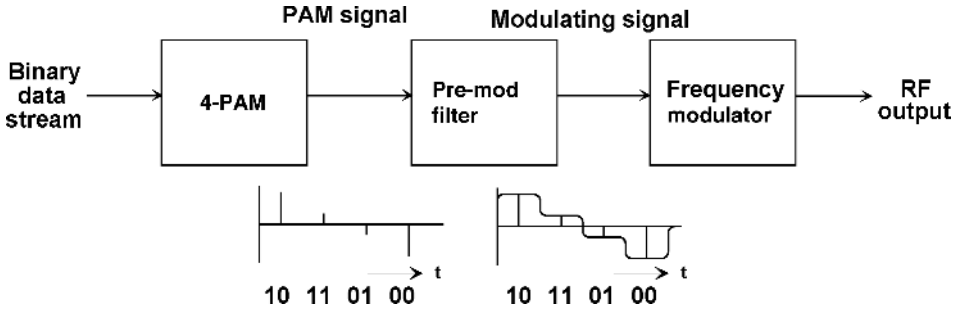


Figure 1.41 4PAM/FSK.

pulse with an amplitude taking on one of four possible levels. The pulse then is shaped by a baseband filter before being fed to the frequency modulator.

- If more precise modulations are required (e.g., MSK and GMSK, which also turn out to be frequency modulations), first the instantaneous phase of the modulated RF carrier is calculated from the data. The corresponding sine and cosine values that form the modulating baseband signals $c_I(t)$ and $c_Q(t)$ are determined from a look-up table. This operation is the reason for the fact that frequency modulation is called a nonlinear modulation.

1.4.1.2 The Spectrum of a Digitally Modulated Carrier

It is the task of any transmission process to occupy as little bandwidth as possible. The absolute lower limit in the baseband is half the symbol rate of the baseband signal, where for M -ary modulation the symbol rate r_{Symbol} is equal to the bit rate divided by $\log_2(M)$. This lower limit is only theoretical as ideal rectangular filters that cannot be realized were necessary. Therefore, in practice, a minimum baseband bandwidth of about $0.75 r_{\text{Symbol}}$ has to be taken into account.

With linear modulation, the occupied bandwidth in the RF range is twice the occupied baseband bandwidth. This follows from the lag theorem, according to which the double-sided spectrum of a time function is shifted from $f = 0$ to the frequency $f = f_c$ when the time function is multiplied with $\cos(2\pi f_c t)$ (Figure 1.42).

Expressing this with formulas, we find

$$c(t) \quad \text{---} \quad C(f) \quad (1.14)$$

$$e^{j2\pi f_o t} c(t) \quad \text{---} \quad C(f - f_o) \quad (1.15)$$

Therefore, if the baseband spectrum is limited by a low-pass filter, the RF spectrum is limited as if it was filtered by an RF bandpass filter with twice the bandwidth of the baseband filter.

Demodulation of Digitally Modulated Carriers The demodulation process is an estimation process. The receiver compares the received waveform with all the possible waveforms

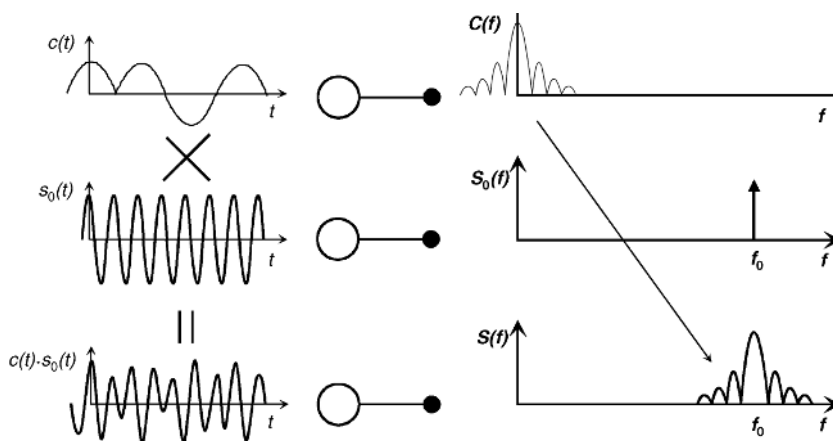


Figure 1.42 Occupied bandwidth in the baseband and the RF range.

and decides which symbol has been received. The waveform allocated is the one that is most similar to the received signal. This process is also called maximum likelihood estimation.

Figure 1.43 shows a possible demodulator. The received signals are cross-correlated with the stored ones in M correlators. High output from a given correlator indicates a match between its stored symbol and a symbol in the incoming signal.

As can easily be seen, such a receiver is quite complex. Therefore, other principles of receivers have been studied. The most often used receiver is based on an I/Q demodulator, the principle of which is shown in Figure 1.44.

The received signal is split into two equal components. One component is multiplied with the reconstructed carrier signal, the other with its orthogonal counterpart, that is, the $\pi/2$ -shifted reconstructed carrier. After low-pass filtering, this multiplication delivers the

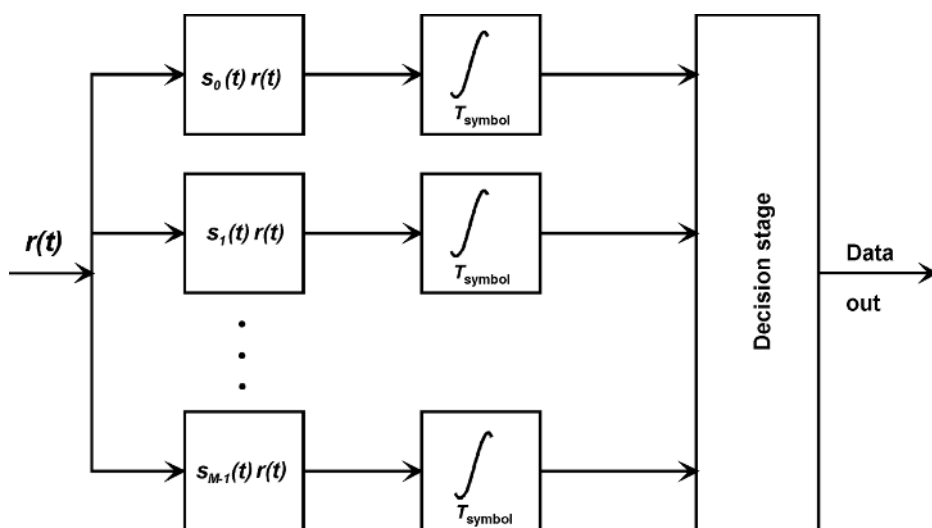


Figure 1.43 Correlation receiver.

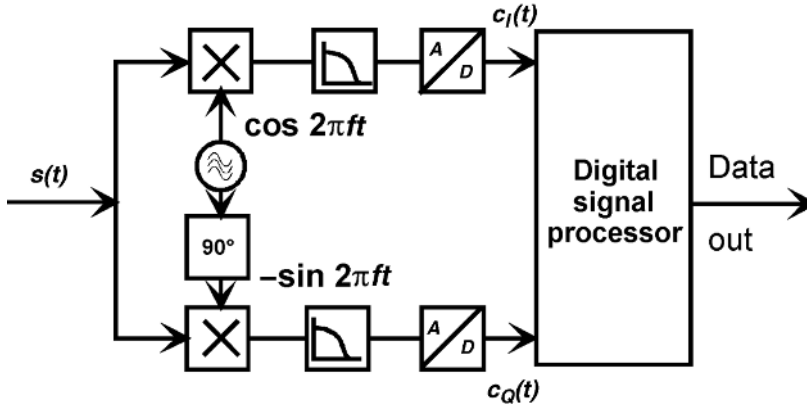


Figure 1.44 Coherent demodulation.

I and the Q components of the baseband signal, which is more or less distorted by interference, multipath, and noise in the radio channel. These demodulated baseband signals then can be A/D converted and treated using simple or sophisticated algorithms to decide which symbols were originally sent. This type of demodulation requires an exact reconstruction of the unmodulated carrier, not only with respect to frequency but also to phase, and is called coherent demodulation. While the frequency recovery is in principle quite simple, the acquisition and tracking of the signal phase requires complex signal processing of the baseband signal.

Coherent demodulation and its accompanying costly signal acquisition and tracking is necessary for all types of phase and quadrature amplitude modulation in which any part of the transmitted information is coded in the absolute phase of the carrier. It only can be avoided when the information is coded in the phase difference of two sequential waveforms. Such modulation schemes are called differential phase modulation. Examples of such modulation include differential binary phase-shift keying (DBPSK) and differential quadriphase-shift keying (DQPSK) and their derivatives.

The following list shows common digital modulation types and the telecom systems that use them.

- BPSK—Satellite radio links, GPS, Inmarsat
- QPSK—Satellite radio links, GPS, IS-95
- OQPSK—IS-95 (reverse link)
- $\pi/4$ -QPSK—NADC, PHS, PDC, TETRA
- GMSK—GSM, DCS 1800
- GFSK (Gaussian FSK)—DECT
- COFDM (coded orthogonal frequency-division multiplexing)—DAB, DVB.

1.4.2 Some Fundamentals of Digital Modulation Techniques

The first waveforms that were used for “digital modulation” were amplitude-shift keying (ASK), similar to on-off-keyed Morse code, and phase-shift keying (similar to FM).

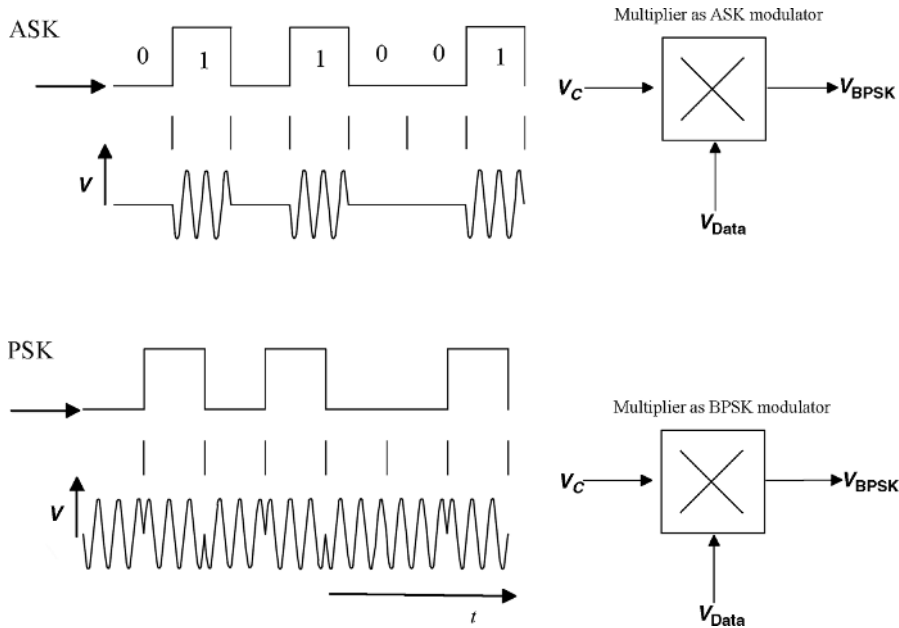


Figure 1.45 Amplitude-shift keying (ASK) and phase-shift keying (PSK) in the time domain.

Figures 1.45 shows the resulting waveforms in the time domain. It is also useful to look at them in the phase domain—the in-phase (I) and quadrature (Q) planes. From Figure 1.46, we see that ASK has only two states: no signal and signal at I (in-phase), while phase-shift keying can have two states: $+I$ and $-I$. In the frequency domain, ASK and BPSK have the output spectra shown in Figure 1.47. These output spectra, with their $(\sin x)/x$ appearance, depend on the rise and decay times and duty cycle of the modulating signal.

BPSK sends one data bit per signal state. Another way to put this is that each of BPSK's two signal states is a symbol that stands for just one bit—0 or 1. Transmitting more than one bit per symbol would allow us to improve our data throughput per unit of time, and this is exactly what is done in quadrature PSK (QPSK), which uses four possible signal states as symbols for two-bit sequences called dibits (Figure 1.48). Figure 1.49 shows a QPSK modulator and the QPSK constellation diagram. Figure 1.50 shows the result of QPSK in the time and frequency domains. Figure 1.51 shows a spectrogram of an actual QPSK emission.

Moving from ASK to BPSK or QPSK results in increased resistance to noise. Figure 1.52 illustrates this by graphing bit error rate versus signal-to-noise ratio (here expressed as $10 \log [E_{\text{bit}} \div N_0]$, where E_{bit} is energy per bit and N_0 is the noise [42]). Figure 1.53 compares the maximum interference voltages for BPSK and QPSK.

Figure 1.54 shows bandwidth requirements and constellation diagrams for BPSK and QPSK. Unfiltered BPSK and QPSK are purely angle-modulated emissions; for each, all symbols are transmitted at the same amplitude. As we will see, however, bandwidth limiting a BPSK or QPSK signal to minimize adjacent-channel interference results in envelope variations that must be preserved through careful circuit design if bit errors are to be minimized.

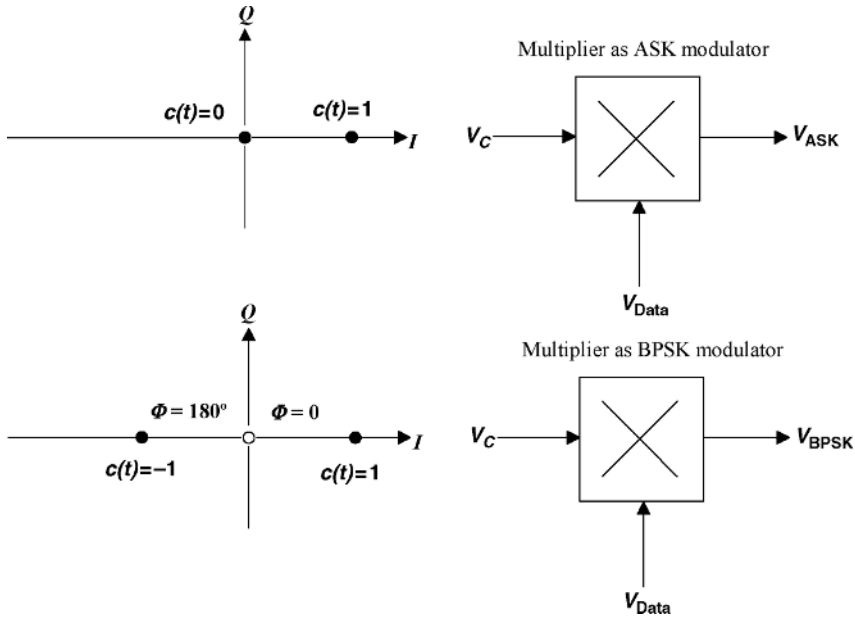


Figure 1.46 ASK and PSK in the I/Q plane.

Further increases in bit rate per symbol are possible with phase-modulation schemes that divide the 360° of each carrier cycle into increasingly smaller segments. Transmitting three bits per symbol, with each symbol spaced from its neighbors by 45° , gives 8-PSK; transmitting four bits per symbol, with each symbol spaced from its neighbors by 22.5° ,

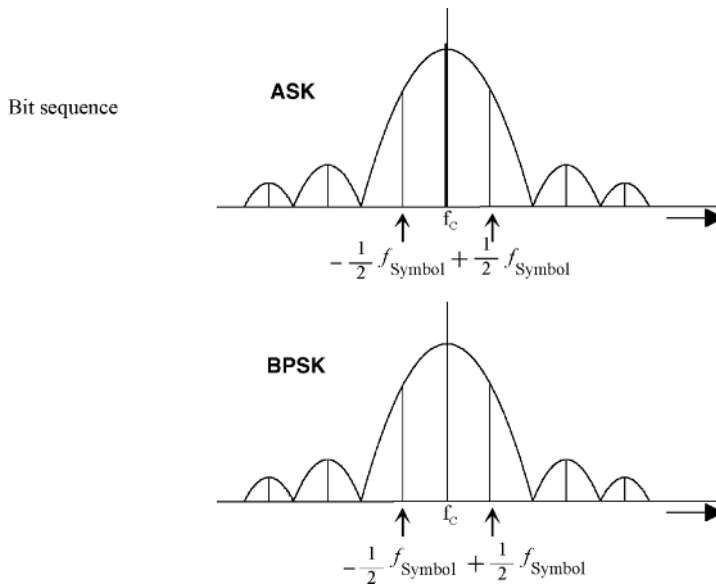


Figure 1.47 ASK and BPSK in the frequency domain.

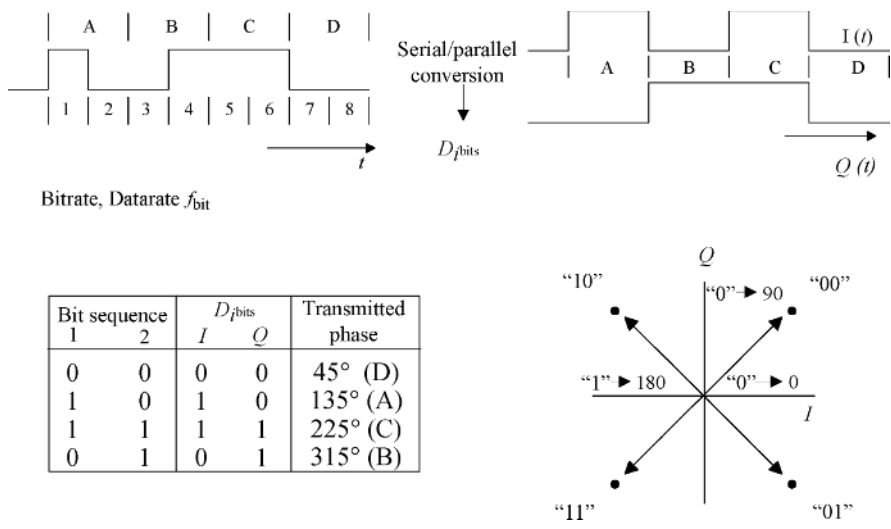


Figure 1.48 Quadrature PSK (QPSK) modulator (I/Q modulator).

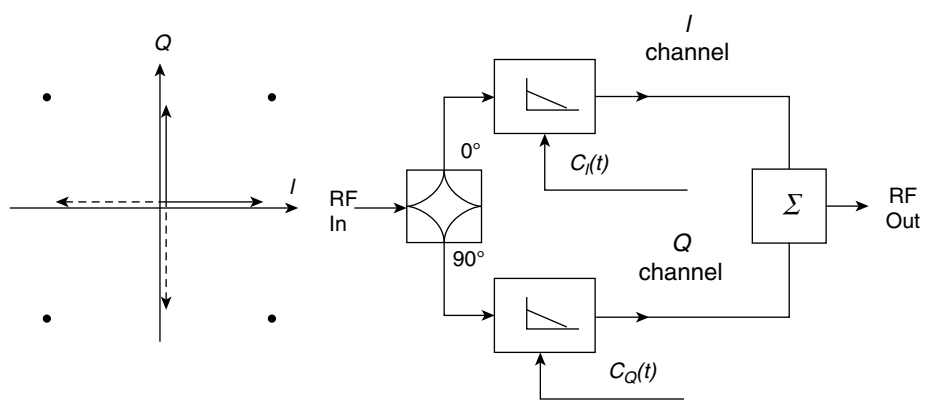


Figure 1.49 QPSK constellation diagram (left) and modulator (right).

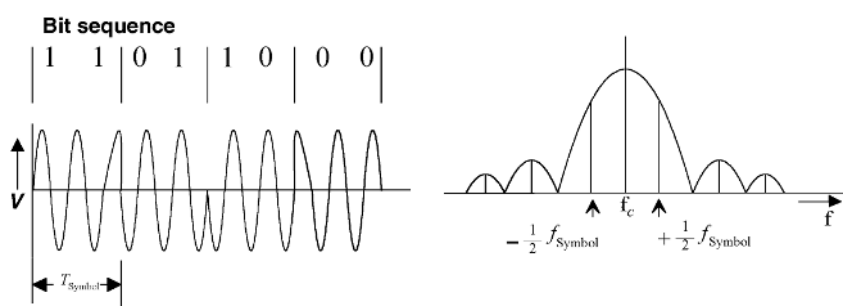


Figure 1.50 Result of QPSK modulation in the time and frequency domains.

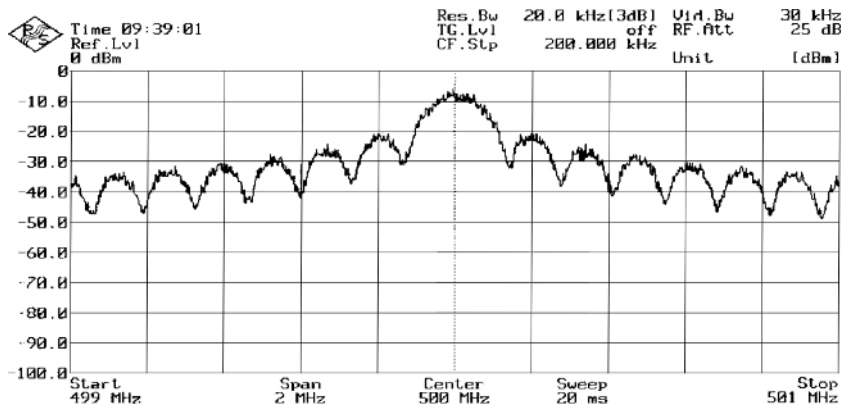


Figure 1.51 QPSK spectrum.

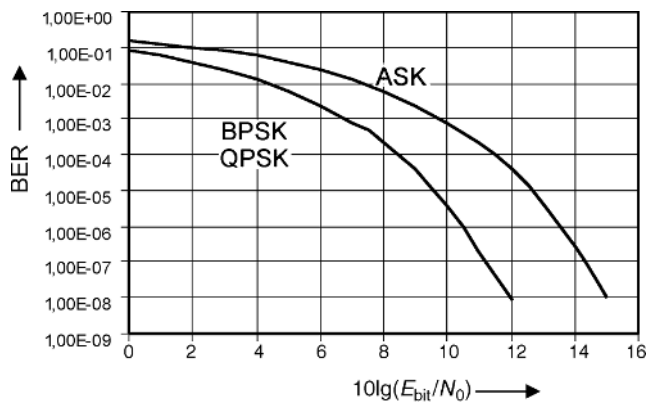


Figure 1.52 Bit error rate (BER) in terms of E_{bit}/N_0 for BPSK, QPSK, and ASK.

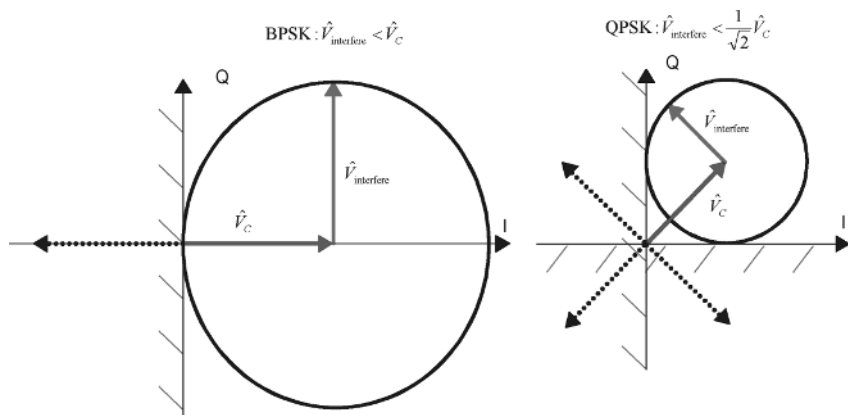


Figure 1.53 Maximum interference voltages for BPSK and QPSK, where $\hat{V}_C$ is carrier voltage of the desired signal and $\hat{V}_{\text{interfere}}$ is the carrier voltage of the interfering signal.

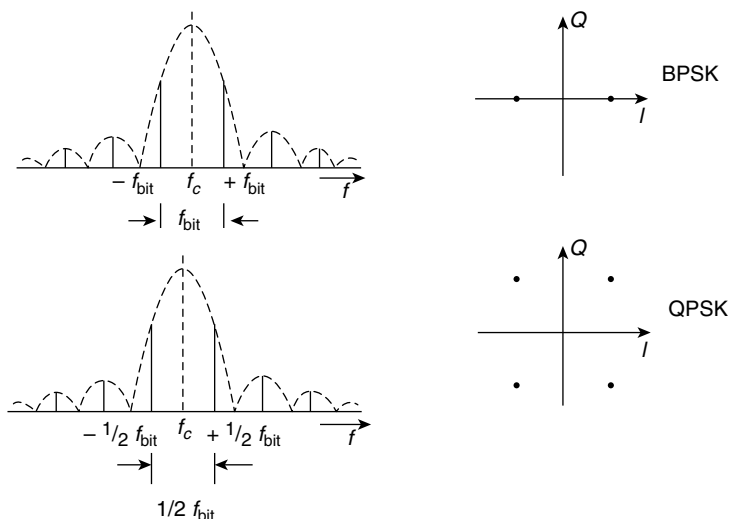


Figure 1.54 Bandwidth requirements (left) and constellation diagrams (right) for BPSK and QPSK. For each emission, all symbols are transmitted at the same amplitude, as is indicated by their equidistance from the constellation origin.

results in 16-PSK. Increasing the data rate by increasing the number of bits per symbol does not give us something for nothing, however; the smaller the phase difference between adjacent symbols, the likelihood that phase shifts resulting from phase noise, multipath reception, and other sources of phase perturbation will cause demodulation errors.

Modulation schemes that differentiate their data symbols through a combination of phase and amplitude shifts are susceptible to disturbances in amplitude in addition to phase. Figure 1.55 shows the modulator and constellation diagram for 16-state quadrature

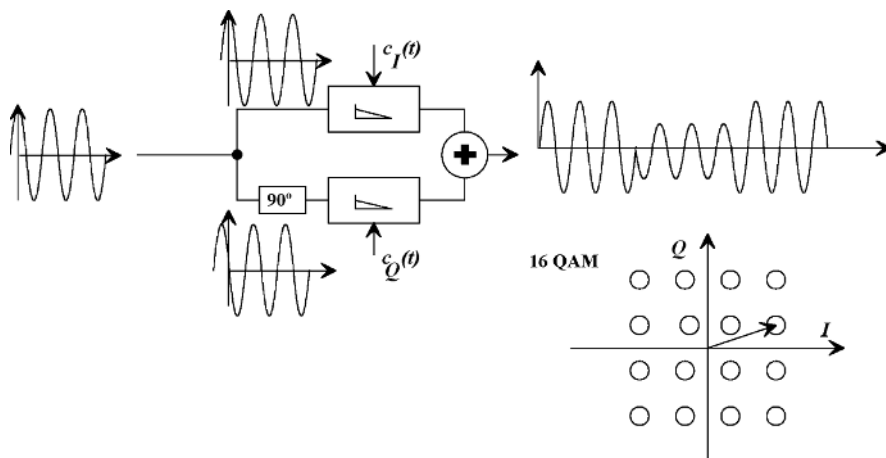


Figure 1.55 Quadrature amplitude modulation (QAM), with constellation diagram for 16-state QAM (16 QAM). Because transitions between QAM symbols involve shifts in phase and amplitude; even unfiltered QAM is a combination of amplitude and angle modulation.

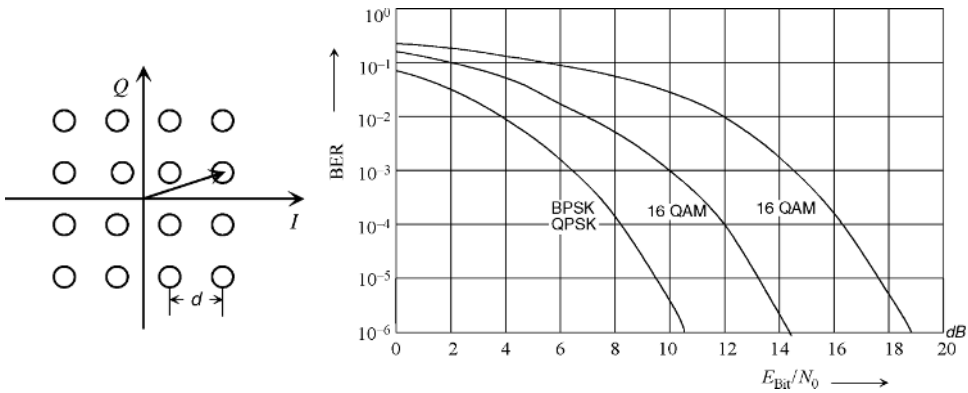


Figure 1.56 Bit error rate versus E_{bit}/N_0 for BPSK/QPSK, 16-QAM and 64-QAM, showing the significantly greater SNR necessary for a given BER as the number of signal states is increased. In 64-QAM, each symbol represents six bits.

amplitude modulation (16-QAM), in which each symbol represents four bits. Figure 1.56 shows bit error rate versus SNR for BPSK/QPSK, and 16-QAM and 64-QAM. QAM systems can support very high bit rates—if sufficient SNR can be maintained and disturbances in signal amplitude and phase can be kept under control [5].

The infinite number of sidebands produced by angle modulation must be reduced by filtering to minimize the spectrum occupied by an emission to just that necessary for communication (Figure 1.57). Considerations of filter realizability and transmitter frequency agility dictate that this filtering must be done at baseband (Figure 1.58). The filter characteristics are optimized in accordance with the modulation scheme used. The filtering is done in using DSP, minimizing component count, and resulting in characteristics that are essentially temperature independent and identical from unit to unit.

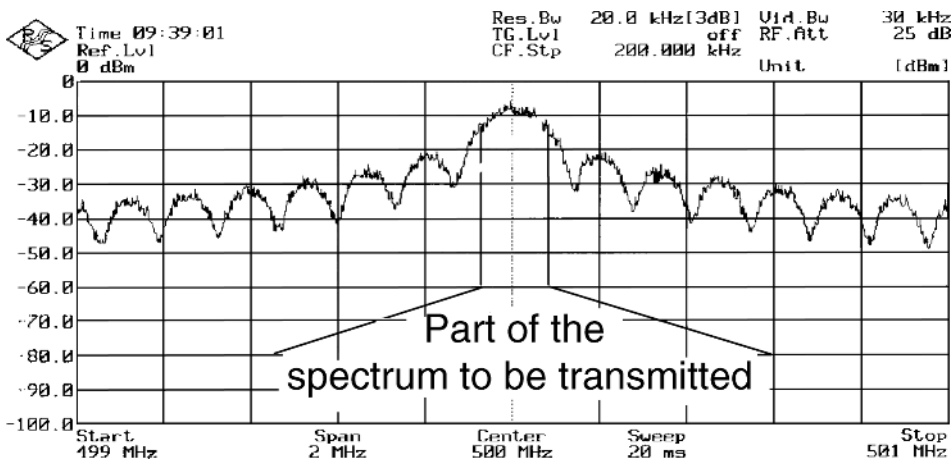


Figure 1.57 QPSK spectrum resulting from pseudorandom binary sequence (PRBS) data. Most of an angle-modulated emission's infinite sidebands are unnecessary for communication and can be removed by filtering—at the cost of introducing amplitude variations that must be sufficiently preserved to keep the bit error rate low.

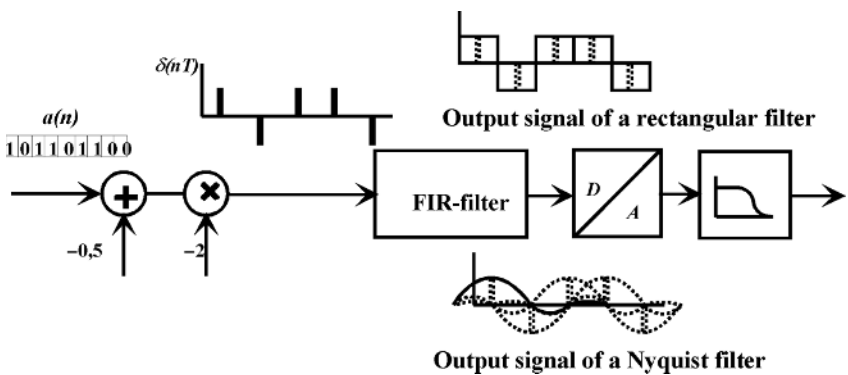


Figure 1.58 Baseband filtering. Figure 1.59 shows the result for QPSK.

Fundamental angle-modulation theory tells us that the envelope of an angle-modulated signal does not vary in amplitude. Such an emission could be amplified in highly nonlinear stages without distortion. But because angle-modulated signals must be band-limited for practical use, and because band-limiting an angle-modulated signal strips away side-band energy that would contribute to its envelope constancy, the emissions used in modern wireless systems, even those that do not involve intentional amplitude modulation, vary in amplitude with modulation. How much an wireless emission's amplitude varies depends on the degree of filtering and the particular modulation scheme used. Various schemes, many of which rely on particular synergies of coding, modulation, and filtering, have been devised for minimizing amplitude variation in digitally modulated signals. Figure 1.60 shows one (offset DQPSK [OQPSK]); Figure 1.61 compares the spectra of three (QPSK, minimum shift keying [MSK], and Gaussian MSK [GMSK]). Detailed discussion of such techniques is more concerned with coding, logic circuitry, and software than with radio hardware, and is therefore beyond the scope of this book. What is important to the circuit designer is that stages handling the emission(s) on which a system depends must be sufficiently amplitude

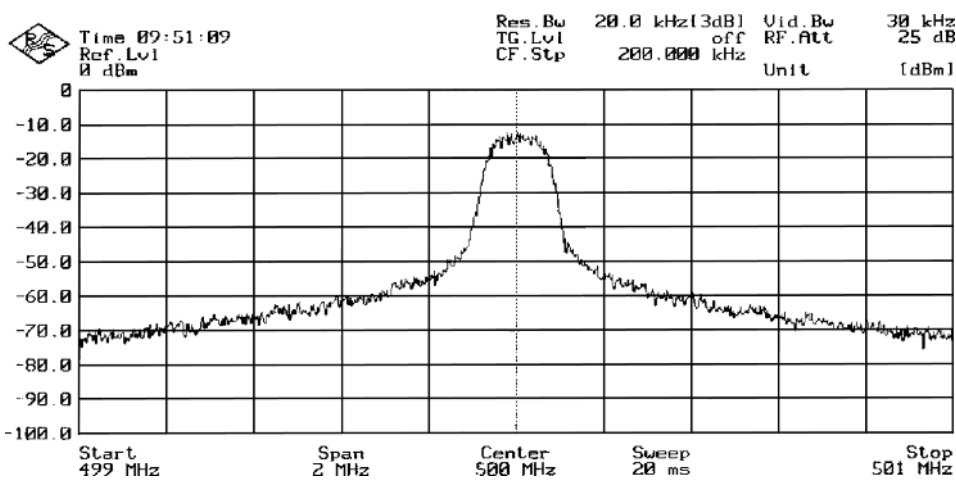


Figure 1.59 Spectrum of a band-limited QPSK signal.

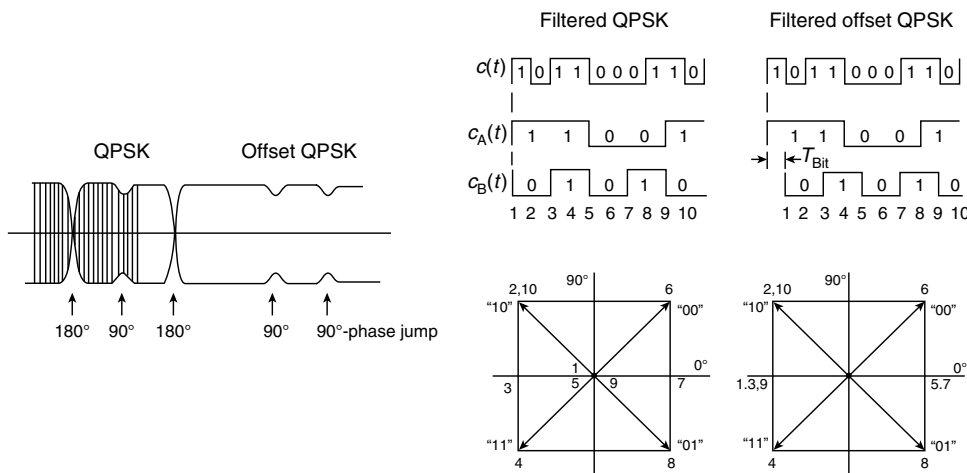


Figure 1.60 Offset QPSK (OQPSK) exhibits reduced envelope variations compared to standard QPSK by time-offsetting the *I* and *Q* channels.

linear to maintain the modulation integrity of the signal and, in the case of transmitters, to keep adjacent-channel interference and other spurious emissions within acceptable levels.

1.4.2.1 Spread-Spectrum and CDMA Modulation Techniques

A spread-spectrum techniques is, by definition, a modulation technique that spreads a base-band signal over a broader bandwidth for transmission—without improving the signal-to-noise ratio (SNR) compared to AM. Common FM or PM therefore are not considered spread spectrum, since increasing the bandwidth yields a reduction in SNR. The two techniques most commonly used for spread spectrum in commercial standards are frequency hopping, and direct-sequence spread spectrum. Let us have a look into these schemes in

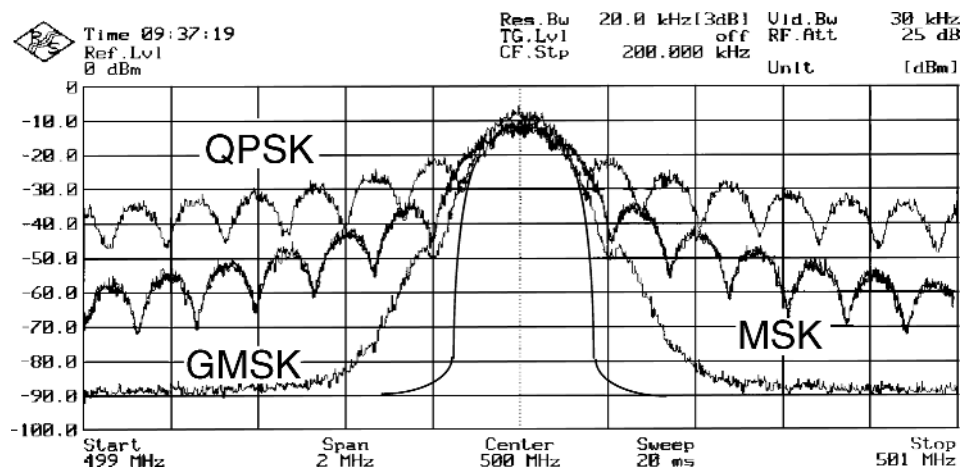


Figure 1.61 Comparison of QPSK, MSK, and GMSK.

order to motivate why it can be advantageous to spread the signal instead of keeping it as narrowband as possible [19, 20].

Frequency Hopping The name already implies how it works: the carrier frequency is not fixed. During transmission, the signal hops between different frequency channels in a pseudorandom sequence that is known only to transmitter and receiver. This approach is advantageous in a hostile environment. The technique is dating back to the 1920s, and was first used as a measure to establish secure military radio links. Without knowledge of the hopping sequence, it is next to impossible in an analog world to track the transmitted signal. The transmission might even be unnoticed by the enemy since the average power for the full band is low and might even be below the total noise power. For one narrow band, on the other hand, the signal only appears for a very short time until it hops away. Another advantage is that it is not easy to disrupt the transmission by jamming. A jammer could practically just block one narrowband channel, while the transmit signal hops around it without experiencing any problem.

Today, frequency hopping is commonly used in standards like Bluetooth that use an unlicensed ISM band. An unlicensed band can be regarded as a quite hostile environment, since many other devices might be active, for example, microwave ovens. Bluetooth uses an adaptive frequency-hopping spread-spectrum technique to ensure good transmission properties. In fact, the hopping sequence is adaptively updated, mainly by removal of channels that provide low-quality transmission, for example, due to fading or due to another device that is sending there.

Direct Sequence Spread-Spectrum DSSS is used in code-division multiple access scenarios (CDMA). In DSSS, the digital baseband signal is multiplied with a fast pseudorandom bit sequence. One bit of information is now expressed by a long series of so-called chips. Since the chip length is a small fraction of the bit length, the respective spectrum is spread accordingly. On the receiver side, the original signal can only be reconstructed, if the received signal is multiplied with the original pseudorandom bit sequence.

This signal is also hard to detect. Without knowledge of the spreading code, it looks like broadband noise. The power level might only be slightly higher or even below the natural noise level. This technique therefore also has its roots in the military sector, and the spreading code was also used to encrypt the transmission.

An important feature of the DSSS is that the received signal is again multiplied with the pseudorandom bit sequence. Figure 1.62 shows a sketch of a spread-spectrum transmission. The original signal is multiplied with the spreading sequence, which can be done through an XOR operation. The figure shows the respective signals in the baseband. Spreading results in a sequence of shorter symbols, thereby the original spectrum is spread. The channel will add white noise, and possibly a narrowband interfering signal.

The received signal is the superposition of these signals. Despreading is performed by applying an XOR operation of the received signal with the original spreading code. The transmitted signal, therefore gets reconstructed. In the time domain, the double XOR operation cancels out, and in the frequency domain, it is despread to its original bandwidth. Any other received signal, on the contrary, is spread, and its power is distributed over a wide band. However, spreading will not affect white noise, which is still white at the same power level after spreading. Therefore, the noise power that defines the SNR of the DSSS transmission equals the noise power within the original bandwidth, no change compared to AM transmission.

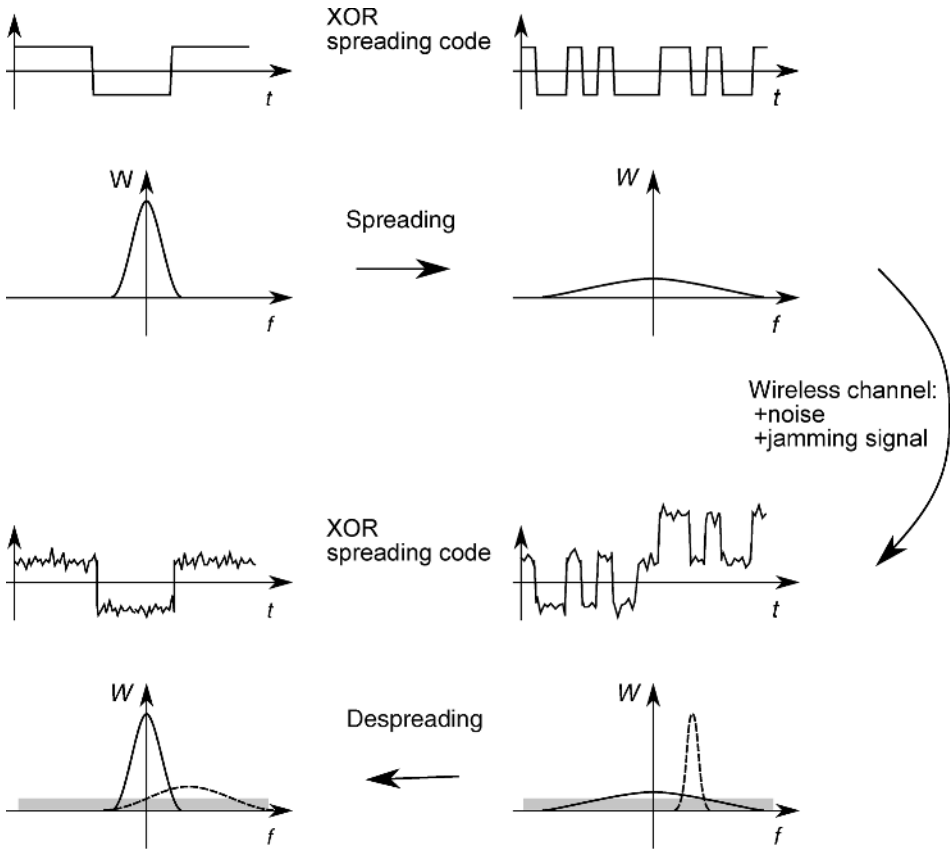


Figure 1.62 Direct-sequence spread spectrum.

Regarding a narrowband jamming signal, a clear advantage over AM is seen. The despread procedure is identical to the spreading procedure for any signal except the desired transmitted signal. A jammer is therefore spread, and its power is distributed over a wide bandwidth. Most of the power is then filtered.

In CDMA, or wideband-CDMA (W-CDMA) systems like UMTS, the different users share the same bandwidth by using different orthogonal spreading codes. Ideally, the cross-correlation between the different bit sequences would be zero, which means that the average of one bit sequence multiplied with another one is equal to zero. In UMTS, these bit sequences are realized through feedback shift registers providing so-called Gold Codes. The codes of this family, however, are not fully orthogonal. The nonzero cross-correlation results in a higher noise floor, depending on the number of parallel users.

W-CDMA can handle multipath propagation that causes reception of delayed echos of the signal. If the spreading code is orthogonal to a delayed copy of itself, all echos will just be discarded when despreading the main signal coherently. However, a rake receiver that basically despreads each of the echos individually, can even take advantage of the multipath propagation. It is, however, required to know in advance when the echos will arrive, which is accomplished through channel estimation techniques.

UMTS mobiles operate in full duplex, they transmit their broadband signal at a quite low-power spectral density. The W-CDMA scheme enables flexible usage of the available bandwidth; more users at a time increase the noise floor for all, which in turn reduces data throughput. But the number of users is not a priori fixed as in pure TDMA or DFMA schemes.

1.4.2.2 Orthogonal Frequency Division Modulation (OFDM) and Single-Carrier Frequency-Division Multiple Access (SC-FDMA)

Orthogonal frequency division modulation (OFDM) starts from a quite simple thought: if very long symbols are transmitted, multipath transmission and echos are not an issue, as long as the difference in delay for the different paths is short compared to the duration of the symbols. This requires a very long duration for a symbol that requires only narrow bandwidth. Thus, for a high-bitrate data transmission, a high number of subchannels is defined, and a high number of bits is transmitted in parallel.

Regarding the different subchannels, it is required that they are independent or orthogonal. OFDM assumes at first a rectangular shape of the bits to be transmitted, which results in a $\sin(x)/x$ -shaped spectrum $S_v(\omega)$ that has periodic zero crossings as follows:

$$S_v(\omega) = \left| \frac{\sin[(\omega - \omega_v)T_S/2]}{(\omega - \omega_v)T_S/2} \right|^2 \quad (1.16)$$

with the bitlength T_S and the subcarrier frequency ω_v . The distance of the zeros in the spectrum defines the channel spacing f_s as follows:

$$f_s = \frac{1}{T_S} \quad (1.17)$$

The subcarrier frequencies therefore are given as $\omega_v = v \cdot 2 \cdot \pi f_s$. Figure 1.63 shows the channel definition for five channels. It is obvious that the channels are only independent at the discrete frequencies f_s ; for all other frequencies, it is not possible to distinguish between the different channels. This channel definition is, however, only processed in the digital domain. Prior to transmission, the signal is inverse Fourier transformed into time domain, bandpass filtered, and converted into analog. The underlying principle is basically the same as the well-known Nyquist criterion for digital data transmission in the time domain. This criterion defines that the bits in a bitstream need to be independent from each other at the sampling time. And assuming an ideal low-pass channel, the bit symbols become $\sin(x)/x$ -shaped, and the time difference between the zeros of one bit signal defines the minimum data rate. OFDM, as stated before, uses basically the same principle, but with time and frequency domain interchanged. OFDM, therefore, requires perfect frequency

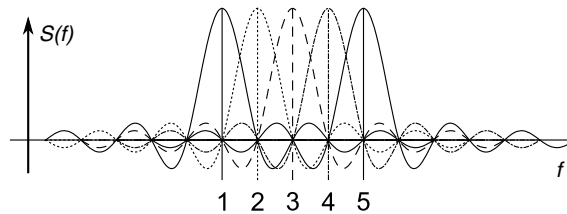


Figure 1.63 OFDM spectrum for five channels.

synchronization between transmitter and receiver, since otherwise the subchannels are no longer orthogonal.

OFDM is inherently robust against multipath transmission due to the long duration of the single bits. In order to completely cancel the impact of multipath effects, a guard period is inserted before each of the symbols, the so-called cyclic prefix. The end of the symbol is copied into this guard period, and all the echo signals are expected to arrive within this time. Before reconstructing the bitstream through FFT, this guard period is discarded. Therefore, in OFDM the signals arriving through the different paths superimpose for each symbol. If they add, it is even able to take advantage of the multipath propagation. And narrowband fading only affects a few subbands, therefore only a few bits are lost without any impact on the other bits transmitted in parallel.

From the point of view of a circuit designer, however, OFDM has a major drawback, that lies in the high number of independent subchannels. The resulting time-domain waveform is almost chaotic and has a high variation in amplitude—the minimum in case that all subchannels transmit the symbol of lowest amplitude, and the maximum being the case, where all channels transmit at maximum amplitude. Even if a clever symbol mapping algorithm is able to reduce the peak-to-average ratio, the time-domain signal's amplitude variation is still the peak-to-average ratio of the single-carrier signal multiplied by the number of subchannels.

Therefore, in order to simplify circuit design, we would prefer a single-carrier modulation due to its lower peak-to-average power. But regarding robustness against multipath propagation, OFDM would be the preferred. Fortunately, it is to a certain extent possible to unite the two approaches. Regarding OFDM, the system looks like this: original signal → IFFT → add cyclic prefix → radio channel → discard cyclic prefix → FFT → equalize → received signal.

However, in a linear system, it is possible to exchange the building blocks without changing the final result. A single-carrier scenario could look like this: original signal → add cyclic prefix → radio channel → discard cyclic prefix → FFT → equalize → IFFT → received signal. The only change is that the IFFT block was moved to after the equalizer. This scheme is called single-carrier frequency-domain equalization (SC-FDE). On the transmitter side, the single-carrier signal is only changed by adding a cyclic prefix to a certain block of data, but no mapping to subcarriers is done. On the receive side, the guard interval is again discarded, which removes the impact of echos on the respective block of data, as in OFDM. Remains frequency-selective fading, which is addressed by equalization in frequency domain. SC-FDE also reduces the requirements regarding frequency synchronization between transmitter and receiver compared to OFDM.

The SC-FDE scheme therefore reduces the requirements on the side of the transmitter power amplifier, since the peak-to-average ratio is the same as for the single-carrier modulation. On the other hand, it requires increased processing power on the receiver side, since an adaptive frequency-domain equalization is required. This scheme is therefore very attractive for the upstream in mobile communication systems, where the mobile unit is transmitting and the base station receives.

In case of LTE, however, the SC-FDE is implemented in a way that additionally allows for multiple access by frequency-domain multiplexing. While in SC-FDE, the full channel bandwidth is dedicated to one transmission, in single-carrier frequency-domain multiple access (SC-FDMA), the different channels defined for equalization are dedicated to different subscribers. It can either be a contiguous block of subchannels or the channels that are interleaved. On the mobile unit's side, it is now required that the single-carrier signal is Fourier

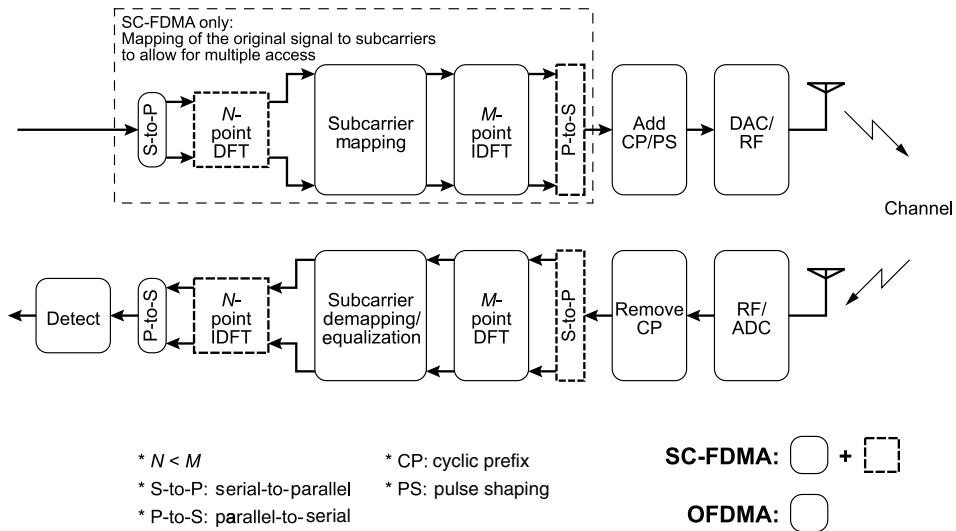


Figure 1.64 SC-FDMA signal flow. Neglecting the dashed box yields a SC-FDE system that transmits a single-carrier signal enhanced by the cyclic prefix and performs the equalization on the receive side in the frequency domain. SC-FDMA in addition first converts the single-carrier signal to frequency domain and maps the discrete spectrum to the available subcarriers (adapted from the public-domain graphic published in the SC-FDMA Wikipedia article).

transformed and shifted to the allowed subchannels. Figure 1.64 shows a block diagram. An N -point FFT is performed first, and the N subchannels are then mapped to a certain subset of M subchannels, unused subchannels are filled with zeros. The IDFT transforms the whole signal back into the time domain for transmission, the guard interval/cyclic prefix is inserted, and the signal finally is converted to analog and transmitted. The receive path is almost unchanged, except that the equalizer needs to be aware of the channel mapping in order to reconstruct the signal.

Further details on the SC-FDMA coding scheme and how it is implemented in LTE can be found in the literature [21, 22].

1.5 ANALYSIS OF WIRELESS SYSTEMS

1.5.1 Analog and Digital Receiver Designs

For the purpose of showing the capability of modern CAD, we now show first an analog receiver and following this an analog/digital receiver with its functionality and performance.

1.5.1.1 Receiver Design Examples

Analog Receiver Design A transmitter modulates information signals onto an RF carrier for the purpose of efficient transmission over a noise filled air channel. The RF receiver's job is to demodulate that information signal while maintaining a sufficient signal-to-noise ratio (SNR). This must be done for widely varying input RF power level and with the presence of noise and interferers.

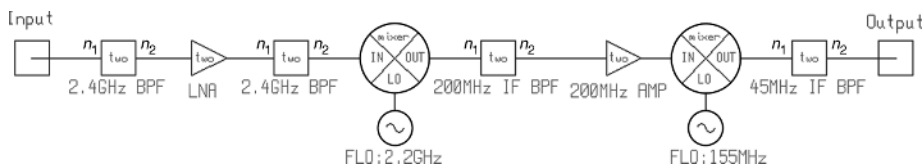


Figure 1.65 Dual-downconversion receiver schematic.

Modern communication standards place requirements on key system specifications such as RF sensitivity and spurious response rejection. These system specifications must then be separated into individual circuit specifications via an accurate overall system model. Symphony can play a major role in modeling systems and in determining the individual component requirements. A 2.4-GHz dual down-conversion system (see Figure 1.65) will be used as an example. The first IF is 200 MHz and the second IF is 45 MHz.

As a signal propagates from the transmitter to the receiver, it is subject to path loss and multipath resulting in extremely low signal levels at the receive antenna. RF sensitivity is a measure of how well a receiver can respond to these weak signals. It is specified differently for analog and digital receivers. For analog receivers, there are several sensitivity measures including minimum discernible signal (MDS), signal-to-noise plus distortion ratio (SINAD), and noise figure. For digital receivers, the typical sensitivity measure is maximum bit error rate (BER) at a given RF level. Typically, a required SINAD at the baseband demodulator output is specified over a given RF input power range. For example, audio measurements may require 12 dB SINAD at the audio output over RF input powers ranging from -110 dBm to -35 dBm. This can then be translated to a minimum carrier-to-noise (C/N) ratio at the demodulator input to achieve a 12 dB SINAD at the demodulator output (see Figure 1.66).

Determining receiver sensitivity requires accurate determination of each component's noise contribution. Modern CAD software, such as Symphony by Ansoft, can model noise from each stage in the system including oscillator phase noise. The C/N ratio can then be plotted as a function of input power (see Figure 1.67).

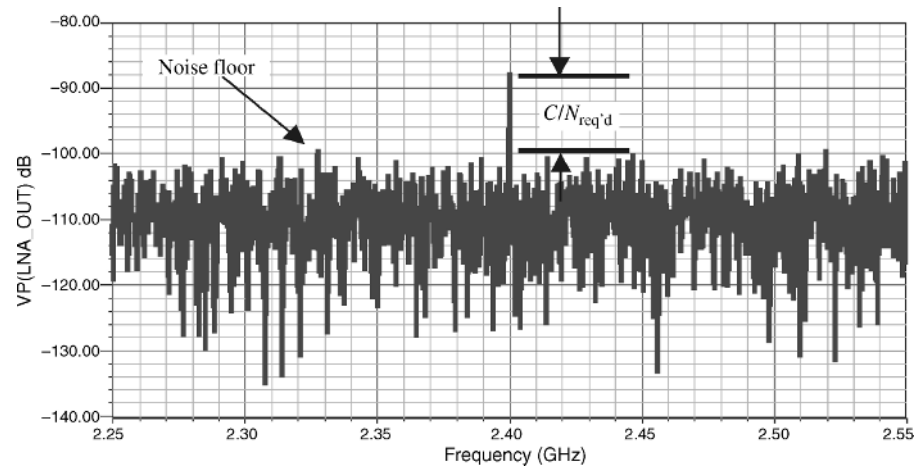


Figure 1.66 Receiver sensitivity measure showing required C/N .

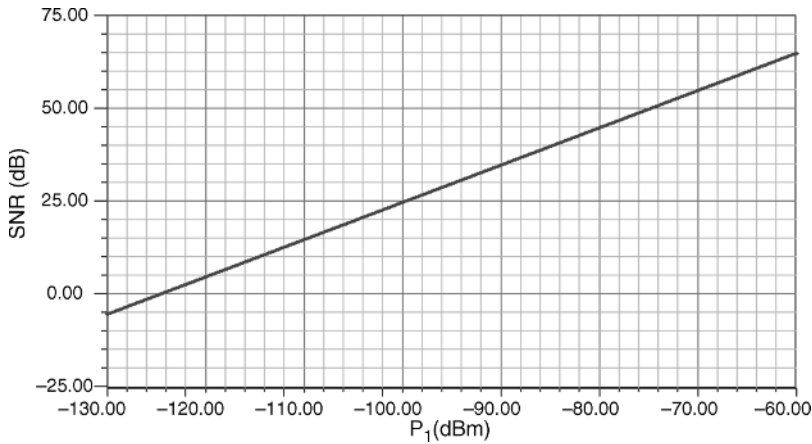


Figure 1.67 C/N ratio versus input RF power level.

This plot enables straightforward determination of the minimum input power level to achieve a certain C/N ratio. Once the sensitivity has been specified, the necessary gain or loss of each component can be determined. A budget calculation (see Table 1.5) can be used to examine the effects of each component on a particular system response.

Another key system parameter is receiver spurious response. The receiver's mixers typically cause the spurious responses. The RF and LO harmonics mix and create spurious responses at the desired IF frequency. These spurious responses can be characterized by the equation [23]

$$\pm m f_{\text{RF}} \pm n f_{\text{LO}} = \pm f_{\text{IF}} \quad (1.18)$$

Some spurious responses, or spurs, can be especially problematic because they may be too close to the intended IF to filter thus masking the actual information bearing signal. These spurious responses and their prediction can be especially troublesome if the receiver is operated near saturation. The analysis can then be refined by including a spur table that predicts the spur level relative to the IF signal. The spur's power at the output of a mixer is calculated using the spur table provided for this mixer. Once generated, each spur is carried

Table 1.5 Budget Calculation for the ΔS_{21} Across Each System Component Accounting for Impedance Mismatches

Names	ΔS_{21} (dB)
2.4 GHz BPF	-0.82
LNA	22.04
2.4 GHz BPF	-0.93
MIXER_1	-5.34
200 MHz IF BPF	-3.85
200 MHz AMP	22.21
MIXER_2	-8.43
45 MHz IF BPF	-0.42

Table 1.6 Spur Output Powers and the Corresponding RF and LO Harmonic Indices

Frequency (MHz)	P_{out} (dBm)
45.00	-26
20.00	-106
65.00	-86
90.00	-76
110.00	-66
135.00	-96
155.00	-44
175.00	-96
180.00	-106
200.00	-56
220.00	-96
245.00	-76
245.00	-46
265.00	-76

through the remainder of the system with all mismatches accounted for. The output power level of each spur is shown in Table 1.6. Another useful output is the RF and LO indices that indicate the origin of each spur.

In addition to sensitivity and spurious response calculations, an analog system can be analyzed in a CAD tool such as Symphony for gain, output power, noise figure, third-order intercept point, IMD (due to multiple carriers), system budget, and dynamic range.

Mixed-Mode MFSK Communication System Next, a mixed-mode, digital and RF, communication system will be described and simulated. In this example, digital symbols will be used to modulate an 8 GHz carrier. The system uses multiple frequency shift keying (MFSK) bandpass modulation with a data rate of 40 Mbps. Convolutional coding is employed as a means of forward error correction. The system includes digital signal processing sections as well as RF sections and channel modeling. Several critical system parameters will be examined including bit error rate (BER). FSK modulation can be described by the equation [24]

$$s_i(t) = \sqrt{\frac{2E}{T}} \cos(\omega_i t + \phi) \quad (1.19)$$

where $i = 1 \dots M$. So the frequency term will have M discrete values with almost instantaneous jumps between each frequency value (see Figure 1.68). These rapid jumps between frequencies in an FSK system lead to increased spectral content.

Figure 1.69 shows a block diagram of the complete system. The system can be split into several major functional subsystems, the baseband modulator, the RF transmitter section, channel model, RF receiver, clock recovery circuitry, and baseband demodulator. Looking specifically at the baseband modulator circuitry (Figure 1.70), a pseudorandom bit source is used and the bit rate is set to 40 MHz.

A convolutional encoder then produces two coded bits per data bit and increases the bit rate to 80 MHz. The purpose of the convolutional encoder is to add redundancy to improve the received BER. A binary-to-M-ary encoder then assigns one symbol to every two bits

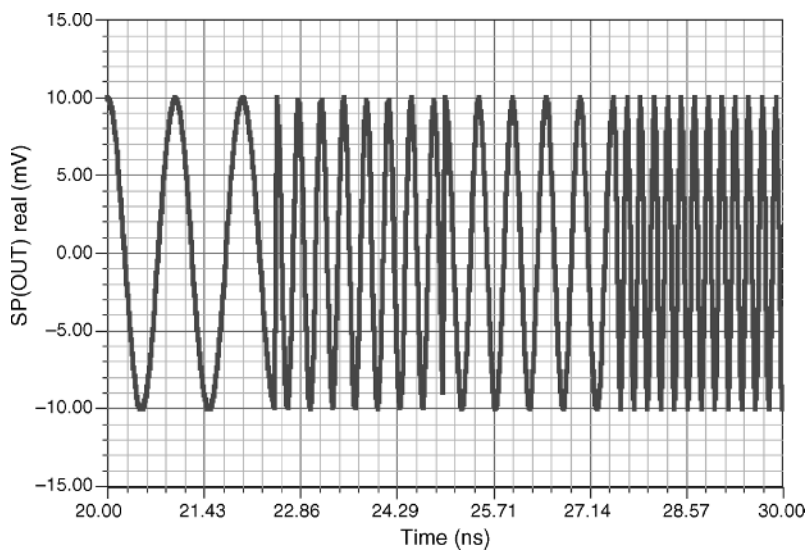


Figure 1.68 4FSK signal generated in Symphony.

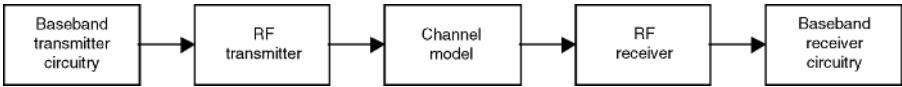


Figure 1.69 Mixed RF/DSP MFSK communication system simulated in Symphony.

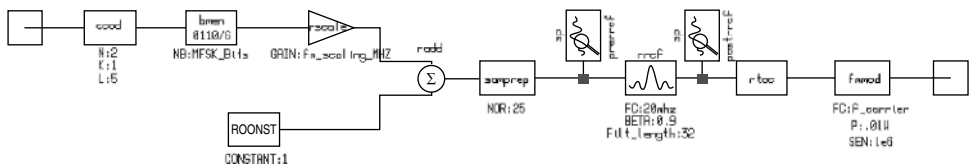


Figure 1.70 MFSK baseband circuitry including frequency modulator.

creating the four levels for the 4FSK effectively halving the bit rate down to 40 MHz again. The signal is then scaled and upsampled. To decrease the bandwidth, a root-raised cosine filter is used to shape the pulses. The filtered signal then serves as the input to the frequency modulator. The RF section (see Figure 1.71) includes the transmitter that modulates the baseband signal onto an 8-GHz carrier.

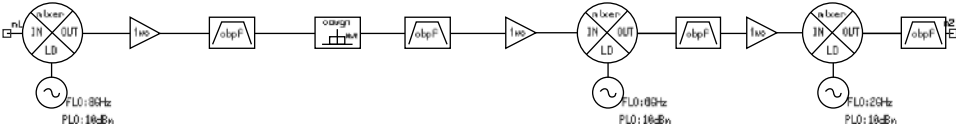


Figure 1.71 RF section including Gaussian noise channel model.

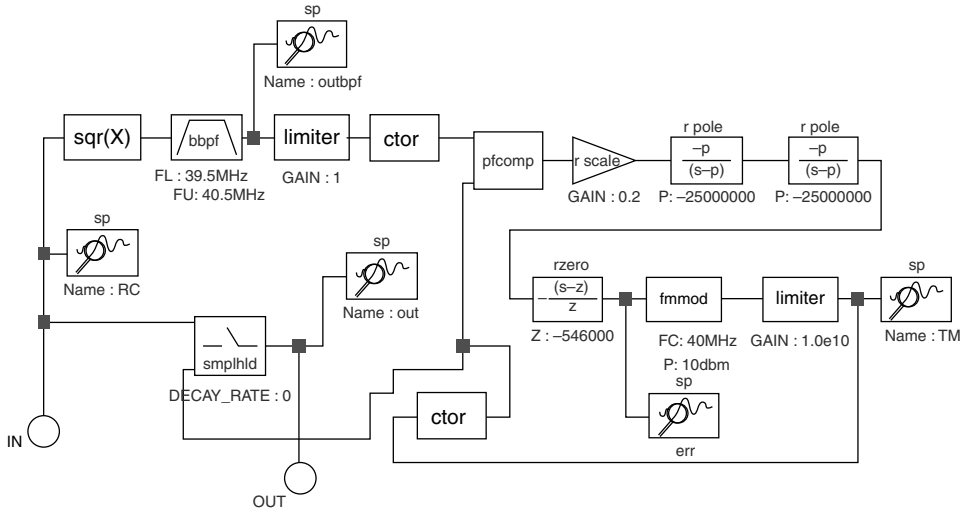


Figure 1.72 Clock-recovery circuitry.

That signal is amplified and then filtered to remove any harmonics. The signal is then passed through an additive white Gaussian noise model. The received signal is filtered, amplified, and downconverted twice to baseband. After carrier demodulation, the signal is then sent through the clock recovery circuitry. Clock recovery is employed in order to ensure that sampling of the received signal is executed at the correct instances. This recovered timing information is then used as a clock signal for sample and hold circuitry. Clock recovery in this system was achieved using a PLL configuration. The schematic for the clock-recovery circuit is shown in Figure 1.72.

At the heart of the clock recovery circuit is the phase comparator element and the frequency modulator. The inputs to the phase comparator consist of a sample of the received data signal and the output of the feedback path that contains the frequency modulator. The frequency modulator reacts to phase differences in its own carrier and the received data signal. The output of the frequency modulator is fed back into the phase comparator whose output is dependent on the phase differential at its inputs. The phase of the frequency modulator continually reacts to the output of the phase comparator and eventually lock is achieved. Once the timing of the received signal is locked onto, the clock that feeds the sample and hold will be properly aligned and correct signal sampling will be assured.

Figure 1.73 shows the received signal (filtered and unfiltered) as well as the recovered clock and the final data. Equalizer circuitry is then used to compensate for channel effects that degrade the transmitted signal. The equalizer acts to undo or adapt the receiver to the effects of the channel. The equalizer employed in this MFSK system is the recursive least square equalizer. The equalizer consists of a filter of N taps that undergoes an optimization in order to compensate for the channel effects. The equalizer depends on a known training sequence in order to adapt itself to the channel. The equalizer model updates the filter coefficients based on the input signal and the error signal (i.e., the difference between the output of the equalizer and the actual desired output). The update (optimization) is based on the recursive least square algorithm. Several equalizers are available in Symphony, including complex least mean square equalizer, complex recursive least square equalizer, least mean square

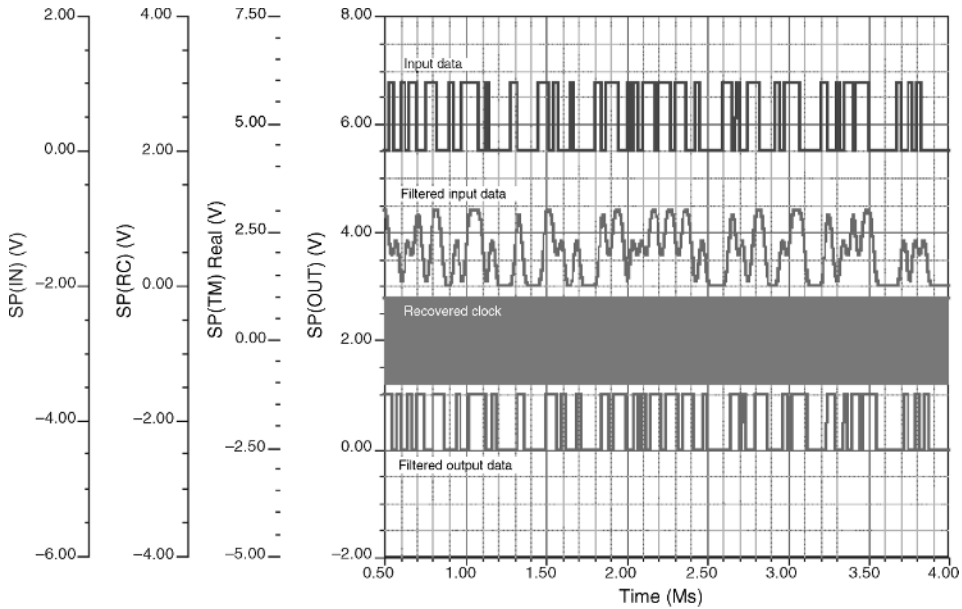


Figure 1.73 Input data, filtered input data, recovered clock, and final output data for the MFSK communication system.

equalizer, recursive least square equalizer, and the Viterbi equalizer. After equalization, the BER of the system is analyzed versus SNR (see Figure 1.74).

1.5.1.2 PLL CAD Simulation

From the clock recovery circuit, it is logical to ask the question about the various frequency sources and their performance. Now to answer this question, we are resorting to Symphony 8.0, where the system simulator allows us to evaluate phase-locked loops. Figure 1.75 shows

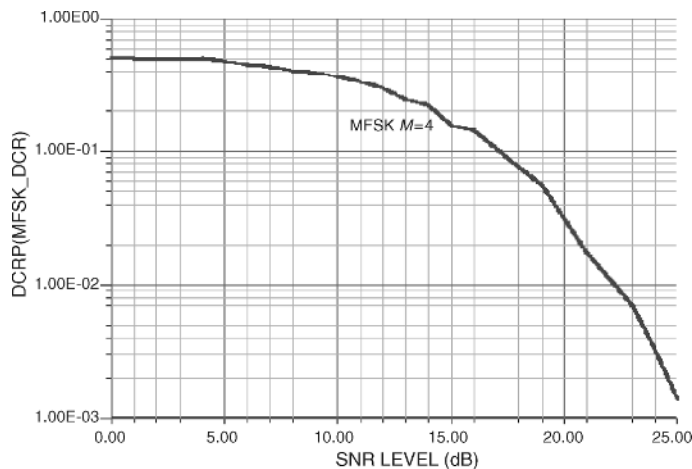


Figure 1.74 BER versus SNR for the MFSK system.

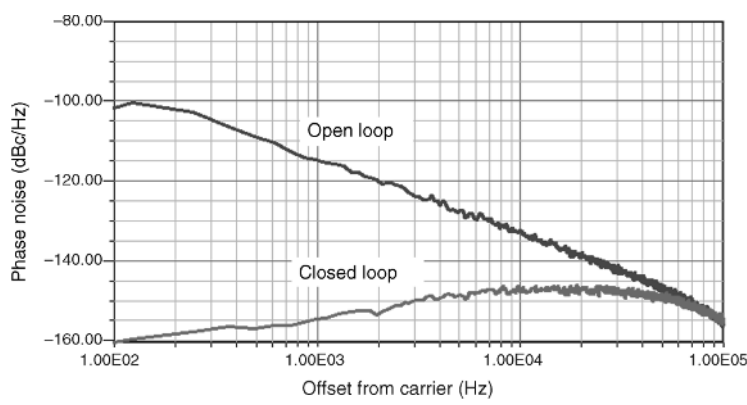


Figure 1.76 Open- and closed-loop phase noise for a CAD-based test phase-locked loop.

Table 1.7 Cellular and Cordless Standards

Cellular Telephone Networks	
Analog	Digital
AMPS Advanced Mobile Phone System	IS-54 North American Digital Cellular
TACS Total Access Communication System	IS-95 North American Digital Cellular
NMT Nordic Mobile Telephone	GSM Global System for Mobile Communications
	PDC Personal Digital Cellular
Cordless Telephone Networks	
Analog	Digital/PCN
CTO Cordless Telecom 0	CT2/CT2+ Cordless Telecom 2
JCT Japanese Cordless Telecom	DECT Digital European Cordless Telecom
CT1/CT1+ Cordless Telecom 1	PHS Personal Handy Phone System
	DCS-1800

*Introduction to the SA900*¹

The SA900 (Figure 1.79) is a truly universal in-phase and quadrature (*I/Q*) radio transmitter that can perform many types of analog and digital modulation including AM, FM, SSB, QAM, BPSK, QPSK, FSK, and so on. It is a highly integrated system that saves space

¹Based on portions of the Philips Semiconductors/Signetics RF Communications Products Application Note AN1892, "SA900 I/Q transmit modulator for 1 GHz applications," August 20, 1997. Used with permission.

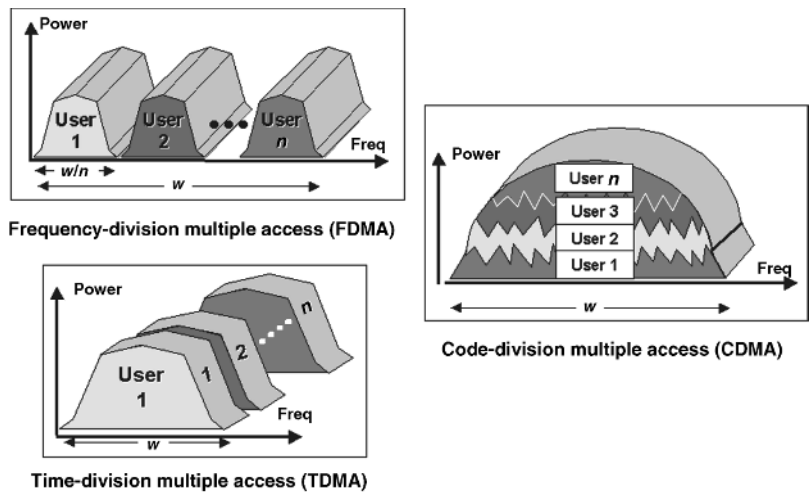


Figure 1.77 Multiplexing schemes.

and cost for the manufacturers producing cellular and wireless products. The device allows baseband signals to directly modulate the I/Q carriers, which are generated by internal phase shift network, in the 1-GHz range, and to maintain good linearity required for linear modulation scheme (e.g., $\pi/4$ -DQPSK). It contains an on-chip frequency divider, phase detector, and VCO, which can be built into a PLL frequency synthesizer to create a transmit offset frequency. Its unique internal design allows frequency conversion without having an external image rejection filter for eliminating the sum term after mixing. The SA900 meets

Table 1.8 Parameters of Cellular and Cordless Systems

Characteristics	Cellular Telephones			Cordless Telephones		
	IS-54	IS-95	GSM	CT2	DCS1800	DECT
Band (downlink)	869–894	869–894	935–960	864–868	1805–1880	1880–1900
Band (uplink)	824–849	824–849	890–915	864–868	1710–1785	1880–1900
Bandwidth	50 MHz	50 MHz	50 MHz	2 MHz	150 MHz	20 MHz
Channelization	TDMA/FDMA	CDMA/FDMA	TDMA/FDMA	FDMA	TDMA/FDM	TDMA
Channel spacing	30 KHz	1250 kHz	200 KHz	100 kHz	200 kHz	1728 kHz
Channels/carrier	3	55–62?	8	1	16	12
Number of channels	832	20	124	40	750	10
	(3 users/ch.)	(798 users/ch.)	(8 users/ch.)		(16 users/ch.)	(12 users/ch.)
Duplex method	FDD	FDD	FDD	TDD	FDD	TDD
Channel bit rate	48.6 kbps	1.2288 Mb/s	271 kbps	72 kbps	271 kbps	1152 kbps
Speech codec	VSELP	CELP	RPE-LTP	ADPCM	RPE-LTP	ADPCM
Bit rate (voice)	8 kbps	1.2–9.6 kbps	13 kbps	32 kbps	13 kbps	32 kbps
Modulation	$\pi/4$ DQPSK	BPSK/OQPSK	GMSK	FSK	GMSK	GMSK
Mobile peak power	0.6–3 W	0.2–2 W	2–20 W	10 mW	0.25–2 W	250 mW
Mobile average power	0.6–3 W	0.2–2 W	0.25–2.5 W	5 mW	0.03–0.25 W	10 mW
Cell radius	30 miles	30 miles	1–5 miles	?	?	40–140 m
Cluster size (min)	7	1	3	N/A	3	N/A

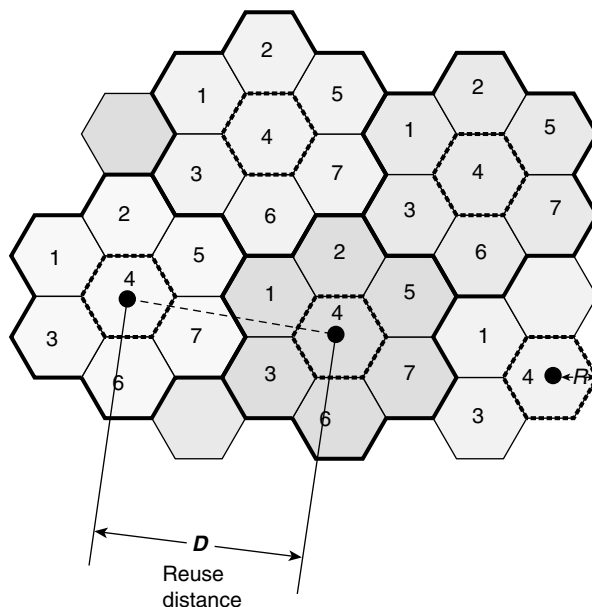


Figure 1.78 Cellular structure. $W = N \times B_C$, where W is the full bandwidth and B_C is the bandwidth per cell. The number of channels/cell = B_C/B_U , where B_U is bandwidth per user (FDMA) or N users (TDMA). $D = R\sqrt{3N}$.

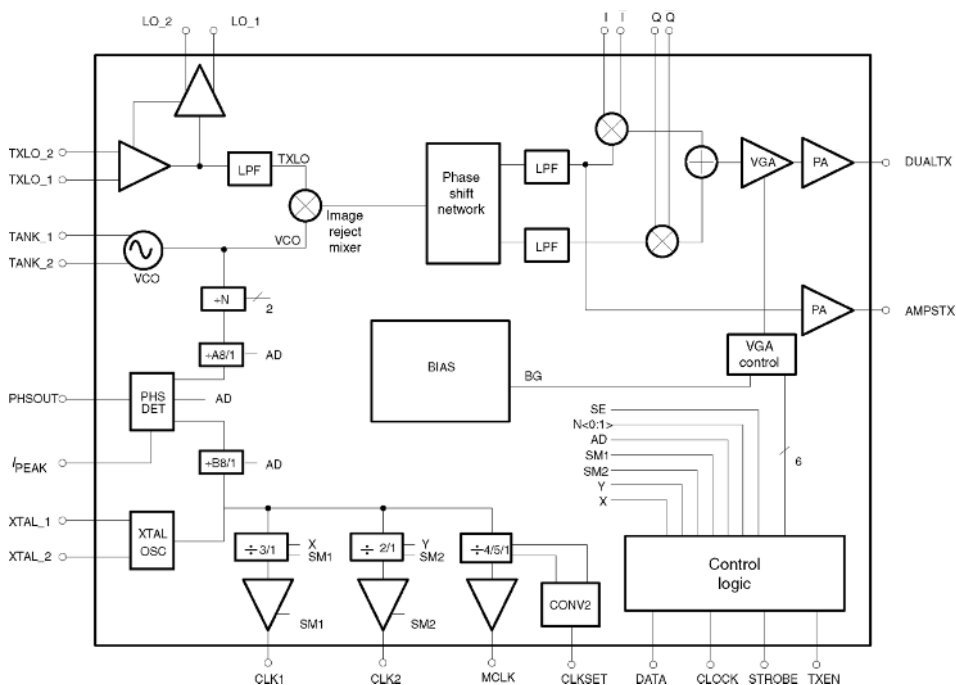


Figure 1.79 SA900 transmit modulator.

the specifications required by the IS-54, the industry standard for North America Digital Cellular (NADC) system. This application note reviews the basic concept of I/Q modulation and discusses the key points when designing the SA900 for an RF transmitter.

I/Q Modulation Any bandpass RF signals can be represented in polar form by

$$s(t) = A(t) \cos[\omega_c t + \phi(t)] \quad (1.20)$$

where $A(t)$ is the signal envelope and $\phi(t)$ is the phase. By using the trigonometric identities, we can represent (1.20) in rectangular form by

$$\begin{aligned} s(t) &= I(t) \cos[\omega_c t] - Q(t) \sin[\omega_c t], \\ I(t) &= A(t) \cos[\phi(t)] \\ Q(t) &= A(t) \sin[\phi(t)] \end{aligned} \quad (1.21)$$

Since the baseband signals $I(t)$ and $Q(t)$ modulate two exactly 90° out-of-phase carriers $\cos(\omega_c t)$ and $-\sin(\omega_c t)$, respectively, we call the system implementing (1.21) an in-phase and quadrature (I/Q) modulator. Figure 1.80 shows the mathematics and hardware implementation of an I/Q modulator.

The local oscillator, usually a VCO within a PLL, generates the carrier and is split into two equal signals. One goes directly into a double-balanced mixer to form the I -channel and the other one goes into the other mixer via a 90° phase shifter (realized by passive elements) to provide the Q -channel. The baseband signals $I(t)$ and $Q(t)$, either analog or digital in nature, modulate the carrier to produce the I and Q components, which are finally combined to form the desired RF transmitting signal. Since any RF signal can be represented in the I/Q form, any modulation scheme can be implemented by an I/Q modulator.

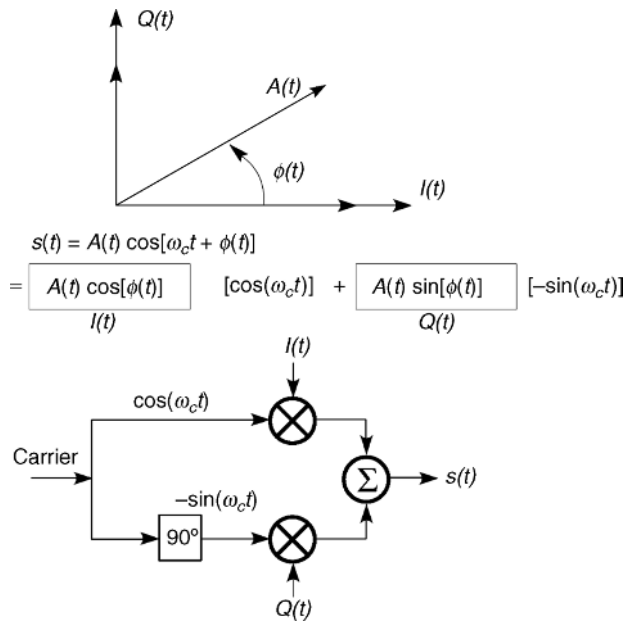


Figure 1.80 Mathematical representation and hardware implementation of I/Q modulator.

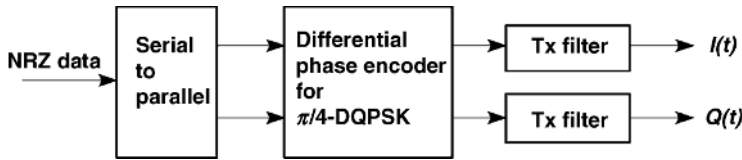


Figure 1.81 QPSK and $\pi/4$ -DQPSK baseband generator.

1.5.2.1 Linear Digital Modulation

Linear digital modulation techniques depend on varying the phase and/or magnitude of an analog carrier according to some digital information: ones and zeros. This digital information can be the output of an analog-to-digital converter (e.g., voice codec) or it can be digital data in some standard formats (e.g., ASCII). The most popular digital signaling format is non-return-to-zero (NRZ), where 1s and 0s are converted into signal with amplitude of 1 and -1 , respectively, in a symbol duration. Since NRZ signal has infinite bandwidth, transmit filters have to be used to limit the spectral spreading. To ensure each NRZ symbol does not smear into its neighbors due to low-pass filtering and channel distortion causing intersymbol interference (ISI), the frequency response of the low-pass filter has to satisfy Nyquist criteria. One example of this type of filter is the linear phase square-root-raised cosine filter. Together with the same type of filter for receive low-pass filtering, the signal is guaranteed ISI-free in a Gaussian environment. One straightforward technique of transmitting these band-limited signals through communication channels would be applying it directly to the mixer of the I channel to generate the RF signal. This is known as binary phase shift keying (BPSK), where the phase of the carrier is shifted 180° to transmit a data change from 0 to 1 or 1 to 0.

Quadrature or quaternary phase shift keying (QPSK) is a much more common type of modulation scheme used in mobile and satellite communications. It has four possible states (90° apart) and each of them represents two bits of data. Figure 1.81 shows the baseband generator for QPSK (without the differential phase encoder). NRZ data bits go through the serial-to-parallel converter (see Figure 1.82) and are mapped in accordance to some rules to generate I and Q values. The generic rule will be the values of I and Q components are 1 and 1 for the data bits “11” (45°) and -1 and -1 for the data bits “00” (-135°). These discrete signals have to be band limited by Nyquist low-pass filters to be ISI free.

A more sophisticated way of mapping results in $\pi/4$ -DQPSK (D for differential encoding), which is chosen for North America Digital Cellular (IS-54), Personal Digital Cellular (PDC) in Japan, and Personal Handy Phone System (PHS) in Japan. In this scheme, consecutive pairs of bits are encoded into one of the four possible phases: $\pi/4$ for “11,” $3\pi/4$ for “01,” $-3\pi/4$ for “00,” and $-\pi/4$ for “10.” However, unlike the previous case that “11” is always $\pi/4$ and “00” is always $-3\pi/4$, the encoded phases are the degrees that the carrier has to shift at each sampling instances. Thus, the information is contained in the phase difference (differential) instead of absolute phase for $\pi/4$ -DQPSK.

A better way to tell the difference between QPSK and $\pi/4$ -DQPSK is by looking at the signal constellation diagram, shown in Figure 1.83, which displays the possible values of I and Q vectors and change of states. Constellation diagram is also known as phase diagram because it shows the phase of the carrier at the sampling point. Note that the phases of QPSK are assigned for every two bits of data; therefore, it can transmit twice as much information as BPSK in a given bandwidth, that is, more bandwidth efficient. 8-PSK is another type of

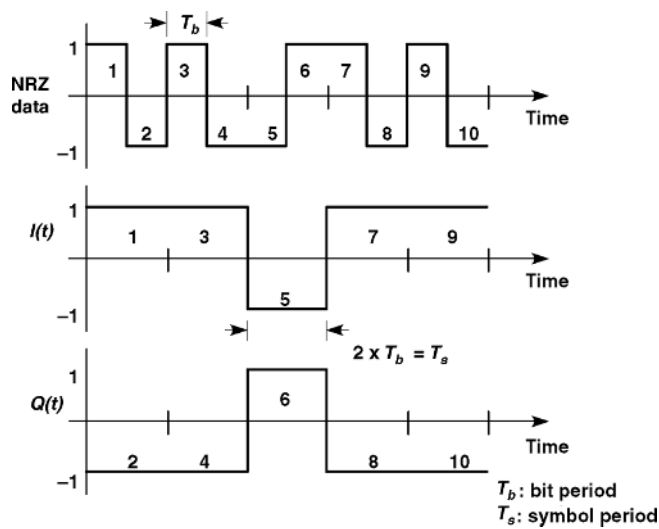


Figure 1.82 Serial-to-parallel conversion.

modulation used for high-efficiency requirements. It maps three bits into eight phases, 45° apart, in the constellation. More spectrally efficient modulation can be created by mapping more bits into one phase at each sampling point. However, as you put more dots in the signal constellation, the signal susceptibility to noise is lower because the decision distance is shorter (dots are closer). Then, it requires higher carrier-to-noise (C/N) ratio to maintain the same bit error rate (BER).

One common misconception is that since $\pi/4$ -DQPSK has eight states in the constellation, it is just another type of 8-PSK. Note that at every sampling instant, the carrier of $\pi/4$ -DQPSK is only allowed to switch to one of the four possible states (see Figure 1.83). So, we still have two data bits that get encoded into four phases. Thus, it has the same spectral efficiency as QPSK for the same carrier power. The reason for using this modulation scheme is twofold. First, the envelope fluctuation, which causes spectral spreading due to nonlinearity of transmitter and amplifier, is reduced because the maximum phase

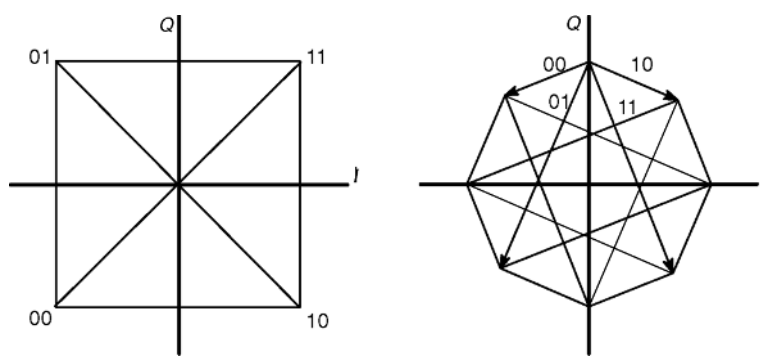


Figure 1.83 Signal constellation of QPSK and $\pi/4$ -DQPSK.

shift is 135° instead of 180° . Second, the signal can be demodulated noncoherently, which simplifies the receiver circuitry by eliminating the need for carrier recovery.

1.5.2.2 Digital and Analog FM

Another family of digital modulation is categorized by frequency change of the carrier instead of phase and/or amplitude change. One of them is frequency shift keying (FSK), where the carrier switches between two frequencies. FSK is also known as digital FM because it can be generated by feeding the NRZ data stream into an analog VCO. FSK appears as a unit circle in the signal constellation because the RF signal envelope is constant and the phase is continuous. Baseband filtering is usually applied for FSK to limit the RF bandwidth of the signal so that more channels can fit into a given frequency band.

One common modulation of this type is known as Gaussian minimum shift keying (GMSK), which is used for GSM and some other wireless applications. GMSK can be generated by following its definition: band-limit the NRZ data stream by a Gaussian low-pass filter, then modulate a VCO with modulation index ($2 \times$ frequency deviation/bit rate) set to 0.5. In other words, the single-sided frequency deviation is one fourth of the bit rate ($\Delta f = R/4$).

Another way of generating GMSK is by I/Q modulator. Referring back to (1.21), any RF signal can be split into I and Q components. Unlike the QPSK mentioned before, baseband $I(t)$ and $Q(t)$ are not discrete points for FM signals; rather, they are continuous functions of time. The way to produce FM is shown in Figure 1.84. We first store all the possible values of $\cos[\phi(t)]$ and $\sin[\phi(t)]$ in a ROM lookup table, which will be addressed by the incoming data to generate the I and Q samples. The output data from the ROM is then applied to D/A converters, after low-pass filtering for signal smoothing, to produce the analog baseband I and Q signals. This method guarantees the modulation index to be exactly 0.5, which is required for coherent detection of GMSK (e.g., the GSM system). The same I/Q principle can also be applied to generating analog FM signals.

1.5.2.3 Single Sideband AM (SSB-AM)

AM signals can be divided into three types: the conventional AM, double sideband suppressed carrier AM (DSB-AM), and single sideband suppressed carrier AM (SSB-AM). The first type is not attractive because for 100% modulation, two thirds of the transmit signal power appears in the carrier, which itself conveys none of the information added by modulation. By using a balanced mixer (e.g., a Gilbert cell), one can generate DSB-AM, where the carrier is totally suppressed and only the upper and lower sidebands are present. However, this is still not the best because the information is transmitted twice, once in each sideband. To further increase the efficiency of transmission, only one sideband is needed to deliver the information. The SSB-AM can be generated by an I/Q modulator with the baseband information feeding the modulator (by quadrature), as shown in Figure 1.85. This modulation technique can greatly reduce the bandwidth of the signal and allows more signals to be transmitted in a given frequency band. This topic is discussed in detail in Ref. [26].

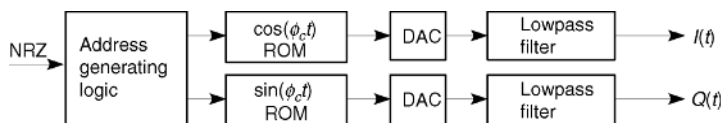


Figure 1.84 Digital FM (e.g., GMSK) baseband I/Q generator.

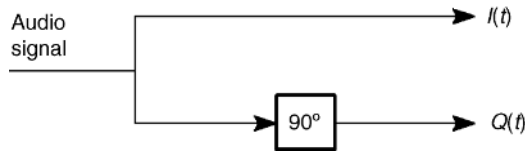


Figure 1.85 Baseband processing for SSB-AM.

System Architecture There are usually two schemes, the dual conversion and direct conversion, used for implementing transmit modulators. Dual conversion is simpler to implement by modulating an oscillator at lower frequency and then upconverting to the carrier frequency. This scheme, however, is more expensive due to the need for additional filtering and more PC board space. By using only one mixer, direct conversion requires fewer components but is harder to implement.

The problems that direct conversion suffers are carrier leakage and modulated signal coupling. Poor RF isolation of the surface mount packages will allow the carrier to be present at the transmitter output thus making it difficult to have -40 dBc carrier suppression. In addition to that, modulated RF signal would couple back to the oscillator (usually a VCO in a PLL synthesizer loop) and cause modulation distortion.

Based on the concept of dual conversion, the SA900 uses an image-reject mixer to eliminate the need for IF filtering and allow monolithic integration. The transmit carrier (LO) is downconverted by the frequency synthesized by the on-chip VCO, which operates from 90 to 140 MHz. This LO is then modulated by the baseband I/Q signals to obtain a complex modulation scheme. The image (sum term) after mixing and LO is sufficiently suppressed by the image rejection mixer. Any residual amounts can be further suppressed by an external duplex filter.

Figures 1.86 and 1.87, respectively, show how the SA900 can be used in frequency division duplex (FDD) and time division duplex (TDD) transceivers. Note that the LO for both systems is running at a frequency that is higher than the transmit frequency, thus minimizing carrier leakage. In the FDD system, only one external VCO is required for generating both transmit and receive LO when using the SA900.

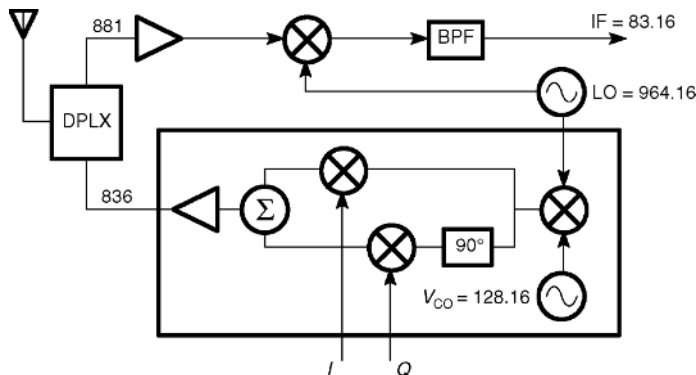


Figure 1.86 FDD system using SA900.

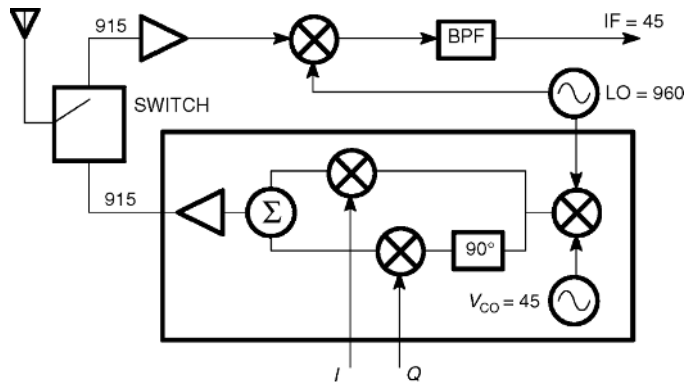


Figure 1.87 TDD system using SA900.

Figure 1.88 shows the IS-54 front-end chip set that consists of the SA601, SA7025, SA900, and SA637. This receiver architecture (SA637) supports a digital magnitude/phase baseband demodulator. An alternate configuration will be using the SA606 FM/IF receiver in conjunction with an external I/Q demodulator IC. Table 1.9 shows the possible configurations for the IS-54 handsets using the SA900 as transmitters.

1.5.2.4 Designing with the SA900

Baseband I/Q Inputs The baseband modulation inputs are designed to be driven differentially for the SA900 to operate at its best. The I and Q inputs should have a dc offset of $V_{CC}/2$, which is externally provided by common DSP chips. If all four inputs are biased from the same source, the device can tolerate ± 0.5 V dc error; however, inaccuracy of dc

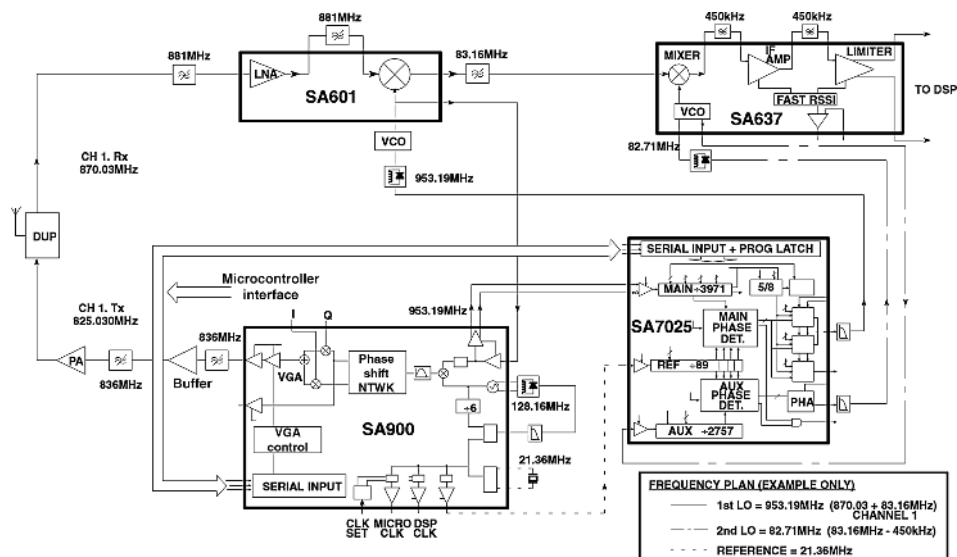


Figure 1.88 IS-54 front-end chipset from Philips.

Table 1.9 Possible Configurations for IS-54 Handsets

Rx 1st IF (MHz)	On-Chip VCO Frequency (MHz)	On-Chip + N Value (MHz)	Crystal Frequency (MHz)
83.16	128.16	6	21.36
71.64	116.64	6	19.44
45	90	6	15
84.6	129.6	9	14.4

bias between I_1/I_2 or Q_1/Q_2 causes reduced suppression of the carrier. Thus, it is important to have a well-regulated dc supply for I and Q signal biasing. The bandwidth of the inputs is much higher than the specified 2 MHz. Approximately 2 dB of power loss will be experienced if the I and Q inputs are 50 MHz.

The SA900 generates a minimum of 0 dBm of power to a 50- Ω load when the amplitude of the I and Q signals are 400 mV P-P. The output power will decrease by 6 dB for every 50% decrease in I/Q amplitude. Single-ended I and Q sources can be used but are not recommended due to the degradation in carrier suppression (more than 10 dB compared to differential). In addition, the entire noise performance of the device will suffer. $V_{CC}/2$ should be applied to I_2 and Q_2 pins if the part is driven single endedly.

Transmit Local Oscillator The transmit local oscillator path consists of a TXLO input buffer, LO output buffer, VCO, image rejection mixer, and phase shift network. Together with a few external components, this section provides the I and Q carrier for modulation.

The TXLO inputs and LO outputs are designed to be used in an external PLL that synthesizes different frequencies for channel selection. The RF signal being generated is fed into TXLO inputs and then comes out of LO outputs to complete the system synthesizer loop. The TXLO inputs are differential in nature and have a VSWR of 2:1 with input impedance of 50 Ω . Single-ended sources can be used by ac grounding the TXLO_2, as done on the demo board. This signal should also be ac coupled into the TXLO_1. The frequency range for these inputs is from 900 to 1040 MHz while the input power should be between -10 and -13 dBm. The output level will be changed significantly if the input level is below -25 dBm.

The output power of the LO-buffered signal changes by about 2 dB when the SA900 is in a different mode of operation. Typical values are -13.5 dBm and -15.5 dBm for DUAL mode and STANDBY mode, respectively.

The 90° phase shift network, realized by RC networks, is capable of operating over a wide frequency range. Even though their frequency characteristics are optimized for cellular band, the part can also be used in other applications in a different band. In such cases, designers have to test the part experimentally to find out the performance, such as sideband suppression, carrier suppression, and image rejection.

Crystal Oscillator The crystal oscillator (XTAL_1 and XTAL_2 pins) is used to provide reference frequency between 10 and 45 MHz for the phase detector and the three on-chip clocks. It can be configured as a crystal oscillator using external crystal and capacitors or it can be driven by an external source. In the latter case, pin XTAL_2 can be left floating. Information regarding crystal oscillator design can be found in Ref. [27].

VCO The VCO, together with the phase detector, the divider, and external low-pass filter, can form a PLL for the transmit offset frequency. The image-rejected mixer down converts

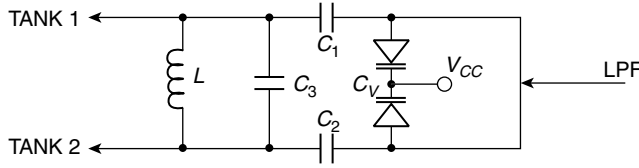


Figure 1.89 VCO tank circuit.

the TXLO signal to the RF carrier by the amount of VCO frequency. Thus, the TXLO frequency should be the desired channel frequency plus the IF offset generated by the VCO. Note that the part will not function if the VCO section is not used.

The VCO is designed for generating IF frequency between 90 and 140 MHz. Together with an external varactor diode and resonator, it can be configured as an oscillator shown in Figure 1.89.

The resonant frequency of such a circuit is

$$f_{\text{VCO}} = \frac{1}{2\pi\sqrt{LC_T}} \quad (1.22)$$

where $C_T = (C_1 || C_2 || C_V) + C_3$. C_V is a varactor diode, which changes capacitance as the voltage across it varies.

Calculation:

$$C_1 = C_2 = 33 \text{ pF}$$

$$C_3 = 5.6 \text{ pF}$$

$$C_V = 33.5 \text{ pF at } 2.5\text{V}$$

$$L = 100 \text{ nH}$$

$$C_T = 5.6 + (1/33 + 1/33 + 1/33.5) - 1 = 16.7 \text{ pF}$$

$$f_{\text{vco}} = \frac{1}{2\pi} (100 \cdot 10^{-9} \cdot 16.7 \cdot 10^{-12})^{0.5} = 123 \text{ MHz}$$

On the demo board, a 1:1 ratio RF transformer is also included to allow single-ended external source driving differential inputs when the VCO is not used.

When designing the VCO, careful PCB layout has to be made. Traces have to be short to avoid the parasitic capacitance and inductance that may cause unwanted oscillation. Referring to (1.22), there is a large combination of L and C_T values that will give the same resonant frequency. If undesired spurs are found in the design due to PCB layout, experimenting with a different set of LC values may sometimes solve the problem.

Output Impedance Matching The equivalent output impedance at the DUALTX pin is approximately equal to 600Ω in parallel with 2 pF at 830 MHz . It has to be matched properly to generate maximum power into a $50\text{-}\Omega$ load (e.g., a SAW filter). Figure 1.90 shows the recommended matching network. The shunt inductor (L_1) is used to provide maximum swing at the output (short at dc) and also provide reactance to make the real impedance $50\text{-}\Omega$ looking into the matching network. The remaining negative reactance is canceled by the series inductor (L_2). The values used on the demo board can be used as

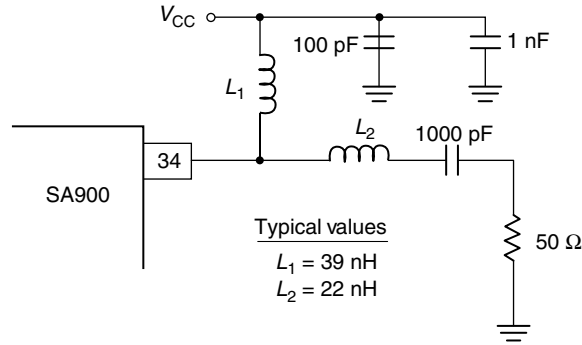
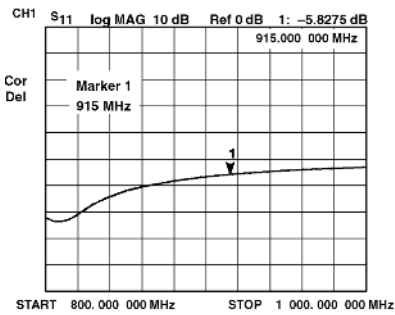


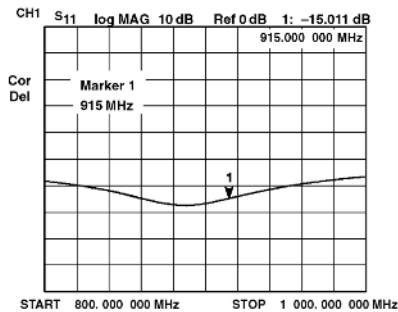
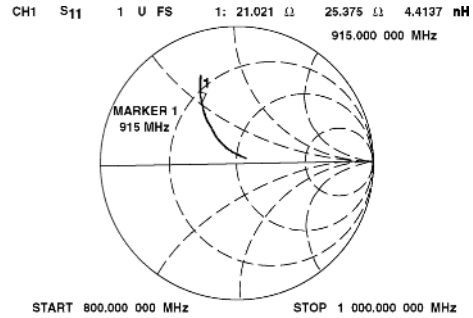
Figure 1.90 DUALTX output matching network.

a reference but may not be suitable if a different layout is implemented. The two shunt capacitors are included to bypass the high-frequency RF signal, avoiding direct coupling into V_{CC} . The series ac coupling capacitor is used to maintain the proper bias for the output stage. Their values are big enough to be left out in impedance matching calculation.

Using a network analyzer to measure the S characteristic is necessary for obtaining optimum matching, which generates maximum output power. Figure 1.91a, b shows how to match the output impedance to a $50\text{-}\Omega$ load at 915 MHz . First, calibrate the network analyzer



(a) S_{11} Response and Smith Chart when $L_2 = 22 \text{ nH}$



(b) S_{11} Response and Smith Chart when $L_2 = 15 \text{ nH}$

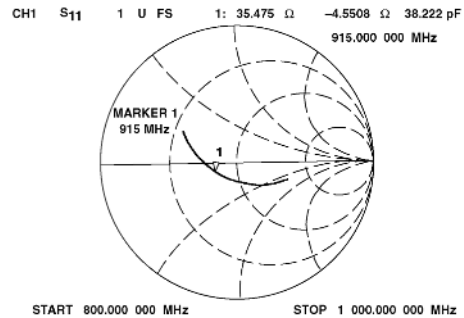


Figure 1.91 SA900 output matching using S parameters.

Table 1.10 Possible SA900 Clock Outputs

CLK1	Divide by 3	$X(\text{bit } 18) = 1$
	Divide by 1	$X(\text{bit } 18) = 0$
CLK2	Divide by 2	$Y(\text{bit } 19) = 1$
	Divide by 1	$Y(\text{bit } 19) = 0$
MCLK	Divide by 4	CLKSET pin = V_{CC}
	Divide by 5	CLKSET pin = $V_{CC}/2$
	Divide by 1	CLKSET pin grounded

to the DUALTX SMA connector on the demo board. Then, short the point where the series inductor is located and use the DELAY feature of the network analyzer to move the point of reference in the Smith Chart to the leftmost point. Now the network analyzer is calibrated to the beginning of the matching network, not just the SMA connector. The frequency response (Figure 1.91a) shows that the “dip” is around 830 MHz, the frequency where the board was originally matched. The Smith Chart shows that it requires less inductance to bring the marker to the center of the chart ($50\ \Omega$). By using a 15-nH series inductor, the “dip” was moved closer to 915 MHz ($-15\ \text{dB}$) and a better matching is achieved (Figure 1.91b).

On-Chip Clocks The crystal oscillator is buffered to provide three external clock signals: CLK1, CLK2, and MCLK. Table 1.10 shows the divide ratio and the controlling mechanism.

CLK1 is usually used for the system synthesizer (e.g., SA7025) reference. Since MCLK is active all the time, it is ideal for providing the master clock for the microcontroller. When the device is in STANDBY mode, CLK1 and MCLK provide the clock signals necessary for receiving RF signals. CLK2 can also be used as a clock for digital signal processing (DSP) chip.

Modes of Operation The SA900 is intended for either AMPS mode (analog cellular) or DUAL mode (digital cellular, IS-54) operation. When the device is running in AMPS mode, the I/Q modulator, variable gain amplifier (VGA), and phase shifter are disabled. The fixed-gain amplifier is powered up during AMPS mode operation. However, since the divide ratio is too low (6, 7, or 8), the comparison frequency of the on-board PLL is too high, making it very difficult for the loop bandwidth to be less than 300 Hz for analog frequency modulation. The device includes two power saving modes of operation that disable partial circuitry to reduce the power consumption of the overall chip. The SLEEP mode disables all the circuitry except the master clock (MCLK pin) of the SA900. The STANDBY mode shuts down everything except the TXLO buffer, MCLK, and CLK1, which allows the system synthesizer (e.g., SA7025) to continue running. These two power saving modes are common to both AMPS and DUAL mode operation. The SA900 draws 60 mA in DUAL mode, reduced to 3 mA and 8 mA in SLEEP and STANDBY modes, respectively.

TXEN pin is for hardware powering down the modulator and synthesizer. The falling edge of the signal disables the modulator and synthesizer while the rising edge enables the modulator. To power down the synthesizer using software, send a data word with SE bit set to “0” (“1” for enable). The synthesizer will be disabled right after the strobe signal is transmitted. Either SE or TXEN going low will turn off the synthesizer. This operation is common to both AMPS and DUAL mode.

Performance of the SA900

Performance Criteria Since the I/Q modulator is a universal transmitter, measuring only the frequency stability and modulation index of a generated FM signal would not be useful for other modulation schemes. Measurement parameters should be general enough so that they can represent the performance of modulators when applying different types of modulation and allow fair comparisons among different I/Q modulators. Based on this idea, two measurement techniques, in-phase modulation and quadrature modulation, are used for evaluating I/Q modulators.

The in-phase modulation relies on injecting two equal frequencies and phase signals at f_{mod} into the I and Q inputs. The result of this modulation is two sidebands appearing at f_{mod} offset from the carrier, with the carrier totally suppressed. This is also known as double-sideband (DSB) conversion. The quadrature modulation requires two equal frequencies (but 90° out-of-phase signals) being injected into the I and Q inputs. The result is a single-sideband suppressed carrier (SSB-SC) signal with either the upper or the lower sideband at f_{mod} carrier offset being suppressed. This is also known as single-sideband (SSB) upconversion. Figure 1.92 summarizes these two tests.

In a practical system, imperfection of an I/Q modulator is directly related to these two measurements. Sideband and carrier suppression from the quadrature modulation test will show the amount of gain imbalance, phase imbalance, and dc offset. On the other hand, intermodulation product suppression from the in-phase modulation test will show the linearity of an I/Q modulator. When making measurements, it is important to have well-balanced I and Q baseband modulating signals for measurement since the signal imperfection will translate into degradation in sideband and carrier suppression.

Performance Graphs In making those measurements for the demo board, the following parameters were used.

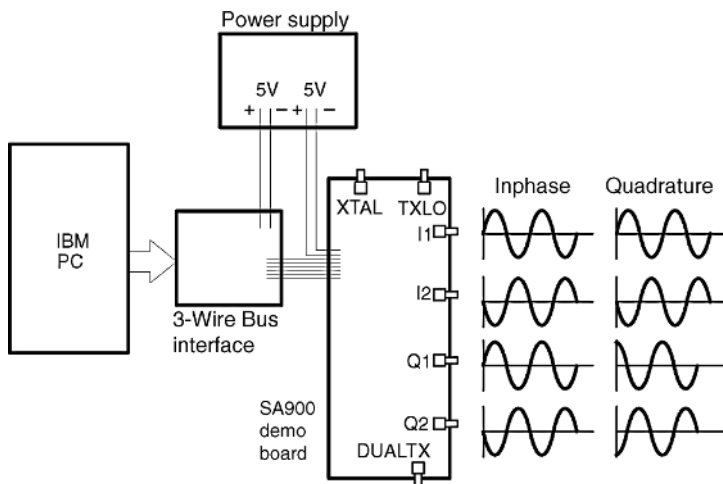


Figure 1.92 In-phase and quadrature modulation test.

In-phase modulation.

PIN 43 $I_1 = 400$ mV P-P, dc = $V_{CC}/2$ at 200 kHz, Phase = 0°

PIN 42 $I_2 = 400$ mV P-P, dc = $V_{CC}/2$ at 200 kHz, Phase = 180°

PIN 41 $Q_1 = 400$ mV P-P, dc = $V_{CC}/2$ at 200 kHz, Phase = 0°

PIN 40 $Q_2 = 400$ mV P-P, dc = $V_{CC}/2$ at 200 kHz, Phase = 180°

Quadrature modulation.

PIN 43 $I_1 = 400$ mV P-P, dc = $V_{CC}/2$ at 200 kHz, Phase = 0°

PIN 42 $I_2 = 400$ mV P-P, dc = $V_{CC}/2$ at 200 kHz, Phase = 180°

PIN 41 $Q_1 = 400$ mV P-P, dc = $V_{CC}/2$ at 200 kHz, Phase = 90°

PIN 40 $Q_2 = 400$ mV P-P, dc = $V_{CC}/2$ at 200 kHz, Phase = 270°

Figure 1.96a and b illustrates what the typical output spectrum would be if in-phase and quadrature modulation were applied to an I/Q modulator. Quadrature modulation will produce lower sideband (LSB) or upper sideband (USB) signal, depending on the phase angle between the I and Q signals. The SA900 was designed to have USB suppressed when the I signal is leading the Q signal. The undesired signals are carrier breakthrough and the harmonic products of the baseband modulating signals sitting at $f_c \pm n \cdot f_{\text{mod}}$, where n is an integer ≥ 2 .

Referring to Figure 1.93a, the output power is 1.3 dBm (cable loss = 0.7 dB) for the LSB while better than -38 dBc of carrier, sideband, and harmonics suppression is measured. The USB better than -26 dBc implies that the residual AM of the transmit signal is better than 5%, a requirement of the IS-54 specification.

In-phase modulation test will generate both LSB and USB. Beside these two tones, the carrier breakthrough and the harmonics, intermodulation (IM) products will all appear at the output. The odd IM products are dominant, and they satisfy the following rules.

Let $f_1 = f_c - f_{\text{mod}}$, $f_2 = f_c + f_{\text{mod}}$.

Third-order IM: $2f_1 - f_2 = f_c - 3f_{\text{mod}}$, $2f_2 - f_1 = f_c + 3f_{\text{mod}}$

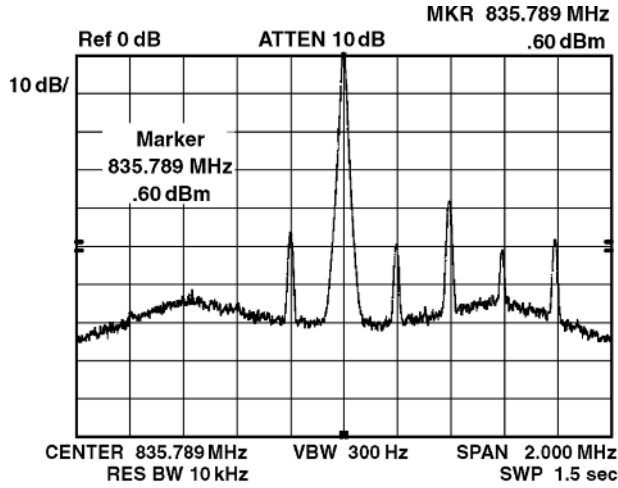
Fifth-order IM: $3f_1 - 2f_2 = f_c - 5f_{\text{mod}}$, $3f_2 - 2f_1 = f_c + 5f_{\text{mod}}$

Seventh order IM: $4f_1 - 3f_2 = f_c - 7f_{\text{mod}}$, $4f_2 - 3f_1 = f_c + 7f_{\text{mod}}$

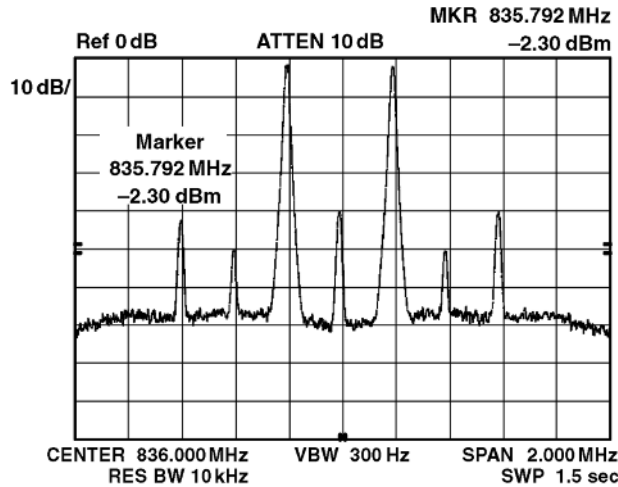
Referring to Figure 1.93b, both LSB and USB are -1.6 dBm (cable loss = 0.7 dB) in power, which is 3 dB less than the measured power for the quadrature modulation test. The IM_3 is better than -35 dBc. IM products of much higher orders are totally suppressed.

Amplitude and Phase Imbalance Both amplitude and phase imbalance (error) of an I/Q modulator can be calculated directly from the SSB performance plots. Assume phase error equals ϕ radian and amplitude error equals K , the sideband suppression, X , in dBc can be expressed as follows (see the I/Q Modulator Equations section for derivation):

$$\text{SSB suppression, } X(\text{dBc}) = 10 \log \left(\frac{K^2 + 2 \times K \times \cos(\phi) + 1}{K^2 - 2 \times K \times \cos(\phi) + 1} \right) \quad (1.23)$$



(a) Quadrature test (SSB upconversion)



(b) In-phase test (DSB upconversion)

Figure 1.93 Typical SA900 output spectra.

Collecting the like terms and expressing ϕ in terms of K and X , it becomes

$$\phi = \cos^{-1} \left(\frac{10^{X/10} \times K^2 + 10^{X/10} - 1 - K^2}{2 \times K + 2 \times K \times 10^{X/10}} \right) \quad (1.24)$$

For a given X , there will be a set of ϕ and K that satisfies (1.24). We can represent this relationship graphically, as shown in Figure 1.94. The contours show the phase and amplitude errors for SSB suppression, X , from -44 to -26 dBc. When X equals -40 dBc, phase error is less than 1.2° with an amplitude error of 0 dB. By the same token, the amplitude error is less than 0.2 dB with a 0° phase error.

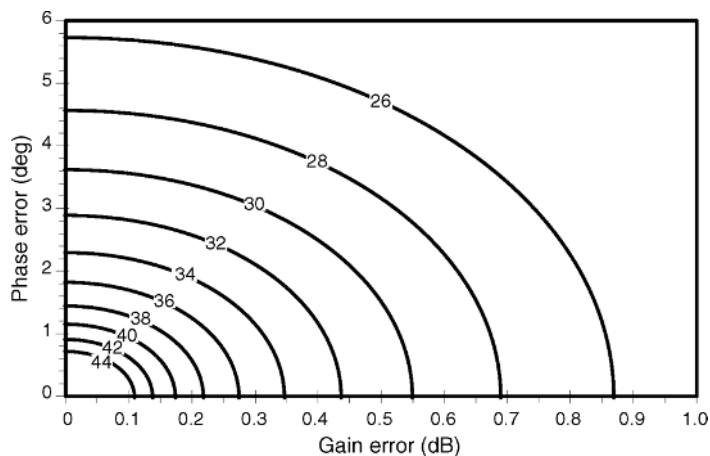


Figure 1.94 SSB suppression contours.

Spectral Mask To fully characterize the performance of an I/Q modulator, measurements of the power spectral density of various digital modulation schemes have to be made. Figure 1.95a and b shows the measured spectral masks of IS-54 and PDC standards, which designate $\pi/4$ -DQPSK as the modulation format.

GMSK is a digital modulation scheme widely used for wireless and mobile communications. Figure 1.95c shows the spectral mask of the modulation format required by GSM, the digital cellular standard in Europe. At 200 kHz and 300 kHz carrier offset, the power of the signal is suppressed by 46 dB and 58 dB, respectively, which is well within the GSM specification.

Power ON Time The power ON time for the SA900 is mainly determined by the loop bandwidth of the on-board PLL frequency synthesizer. It can be measured by using the HP 53310A modulation domain analyzer set to the EXTERNAL TRIGGERED mode. The STROBE signal from three-wire bus is used to trigger the equipment. Figure 1.96 shows that the part can be powered up and locked in about 62 μ s.

1.5.2.5 ISM Band Application

The FCC has assigned three bands for ISM type of application. The one below 1 GHz is from 902 to 928 MHz. This band becomes very attractive because users are allowed, without having a license, to transmit up to 1 W of power when frequency hopping or direct sequence CDMA is used. The wide bandwidth nature of the SA900 fits well into this application. Figures 1.97a and b are the output spectrum of the SA900 showing how well the image reject mixer works. A common IF (45 MHz) was chosen to be the offset frequency, and then injected externally into the VCO pins. The closest images are sitting at 45 MHz apart and are better than -36 dBc.

I/Q Modulator Equations Assume an imperfect I/Q modulator with gain error, K , and phase error, ϕ , modulated by quadrature I/Q signals (SSB upconversion) ω_m . Then the

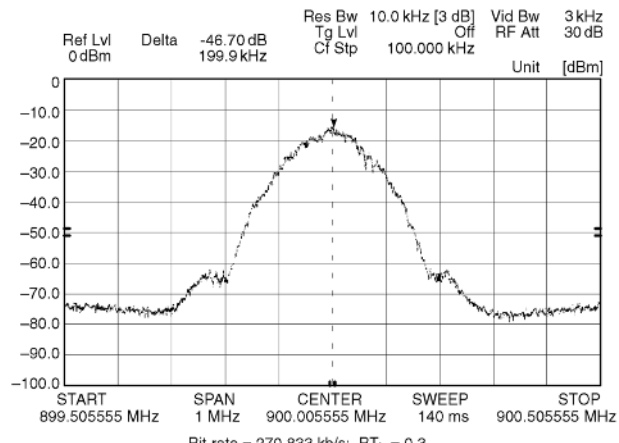
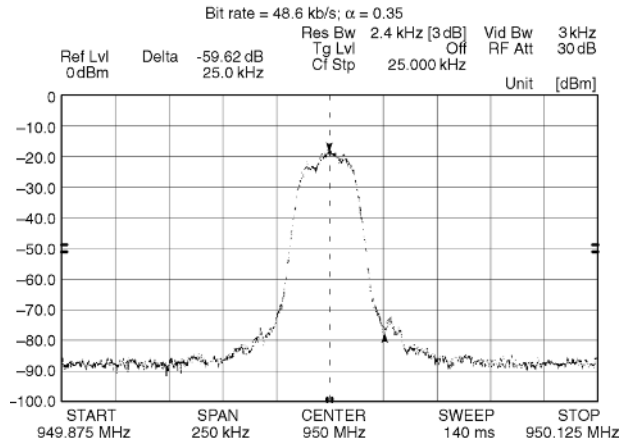
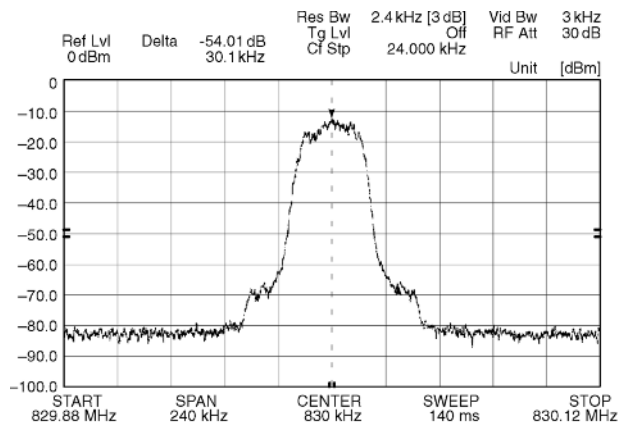


Figure 1.95 SA900 ADC, PDC, and GSM outputs.

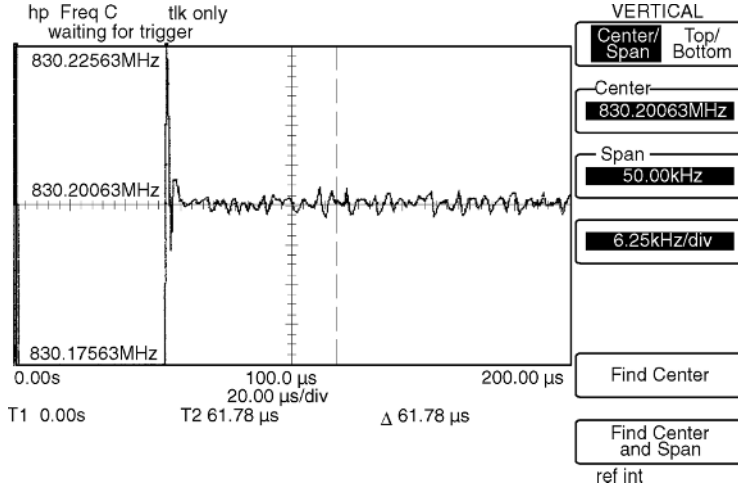


Figure 1.96 Power ON time measurement.

signal, $s(t)$, at the output of the I/Q modulator becomes

$$s(t) = K \cos(\omega_c t + \phi) \cos(\omega_m t) - \sin(\omega_c t) \cos(\omega_m t + 90^\circ) \quad (1.25)$$

Using trigonometric identity and letting $\omega_c - \omega_m = A$ and $\omega_c + \omega_m = B$, we obtain,

$$s(t) = \frac{K}{2} \cos[A t + \phi] + \frac{K}{2} \cos[B t + \phi] + \frac{1}{2} \cos[A t] - \frac{1}{2} \cos[B t] \quad (1.26)$$

Assume that the information is in LSB—that is, A—and the spur is the USB, that is, B, we have

$$\text{Signal} = \frac{K}{2} \cos A \cos \phi + \frac{1}{2} \cos A - \frac{K}{2} \sin A \sin \phi \quad (1.27)$$

$$\text{Noise} = \frac{K}{2} \cos B \cos \phi + \frac{1}{2} \cos B - \frac{K}{2} \sin B \sin \phi \quad (1.28)$$

To find the power, we have to evaluate the envelope (amplitude) of these two signals. Recall that for any given bandpass signal in rectangular form

$$\text{Bandpass signal} = X \cos \omega t - Y \sin \omega t$$

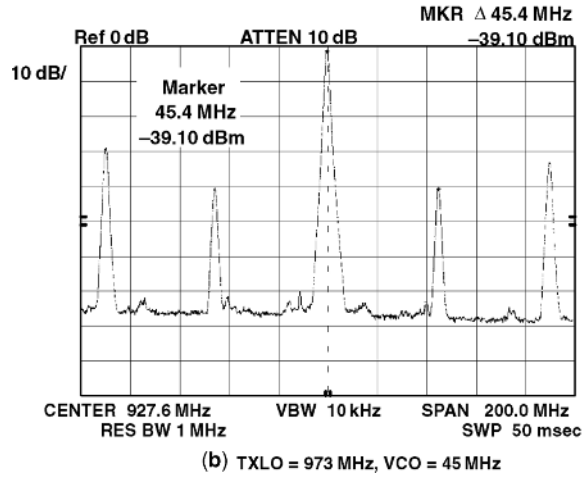
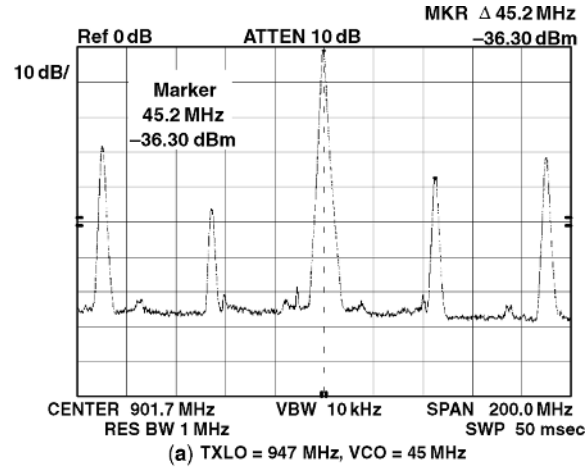


Figure 1.97 Output spectrum of the SA900 in the ISM band.

and the envelope is

$$\text{Envelope} = \sqrt{X^2 + Y^2}$$

Therefore, from (1.27) and (1.28)

$$\text{Signal} = \left[\left(\frac{K}{2} \cos \phi + \frac{1}{2} \right)^2 + \left(\frac{K}{2} \sin \phi \right)^2 \right]^{0.5} \quad (1.29)$$

$$\text{Noise} = \left[\left(\frac{K}{2} \cos \phi - \frac{1}{2} \right)^2 + \left(\frac{K}{2} \sin \phi \right)^2 \right]^{0.5} \quad (1.30)$$

Finally, the S/N ratio can be found by taking $20 \log$ the ratio of (1.29) and (1.30):

$$S/N = 10 \log \left(\frac{K^2 + 2K \cos \phi + 1}{K^2 - 2K \cos \phi + 1} \right) \quad (1.31)$$

1.6 BUILDING BLOCKS

Our earlier block diagram already referred to a variety of integrated circuits that can be used to build a wireless system—in particular, a portable telephone. If we open any of the earlier-mentioned magazines and go through the advertising section, we will find many suppliers of subsystems/components that fit this requirement.

They can be split into various technologies. The following table is a product overview of next-generation advances in wireless technology taken from an ad of Stanford Microdevices, one of the very aggressive component/subsystem companies.

- New SX amplifiers features 2.4-GHz MMICs
- GaAs heterojunction bipolar transistor (HBT) discrete transistors (100 mW to 10 W)
- InP/GaAs (indium phosphate/gallium arsenide) high-linearity gain blocks
- Millimeter-wave product line featuring ICs for LMDS
- Silicon-germanium product line (gain blocks and low-noise amplifiers)
- SXQ power modules for cellular and PCS applications
- Power modules featuring 4- to 25-W power amplifiers for land mobile amplifiers
- 30- to 120-W silicon LDMOS power transistors for PCS, CDMA, W-CDMA
- SAW files featuring integrated transceiver modules
- Electro-optic and datacomm products.

In the beginning of this chapter, we looked at block diagrams and now into building blocks. Most manufacturers will try to move to the highest possible integration while avoiding external components. Figure 1.98, a test circuit for Motorola's MC13109 cordless telephone subsystem IC, shows the level of complexity needed to test such an IC with all its functionality.

Another key issue at those frequencies is to really know a component's frequency dependencies. Figure 1.99 shows a good example. Typical components are capacitors and inductors as reactive elements, and resistors as passive elements. Depending on the type of integration, one may also have to look into the noise performance of integrated resistors. To look into this feature is best discussed with the foundries. Along these lines, the dielectric material on which these surface-mounted components are mounted plays a big role and must be addressed individually.

Besser Associates of Los Altos, California, among other firms, offers a very nice one-day short course titled "Wireless Circuit Components: Measurements, Models and Data Extraction."

Table 1.11 presents a sampling of RFICs for wireless applications. These types of ICs, which were taken from Motorola are made in similar form by many companies. The selection guide is useful in starting off with a design that is less integrated, therefore allowing the designer to have access to portions of the circuit that later "disappear" inside the IC.

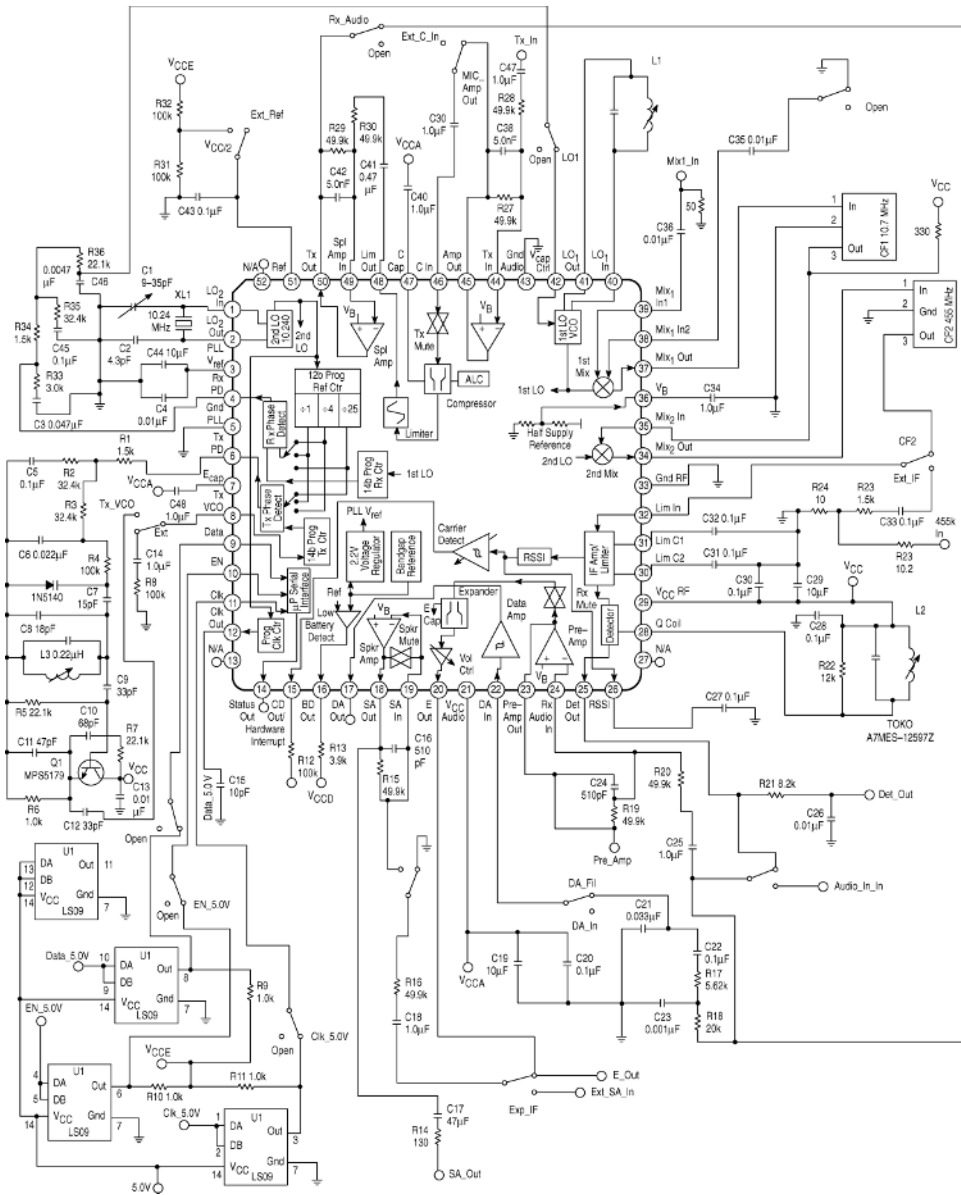


Figure 1.98 MC13109FB universal cordless telephone subsystem IC.

1.7 SYSTEM SPECIFICATIONS AND THEIR RELATIONSHIP TO CIRCUIT DESIGN

Wireless communication involves a large range of signal powers—from levels on the order of 10–18 W (at the receiver input) to 102 W (at a base-station transmitter output). A receiver must be able to demodulate signals that have been attenuated billions of times through propagation; a transmitter must be able to produce a properly modulated signal at a frequency

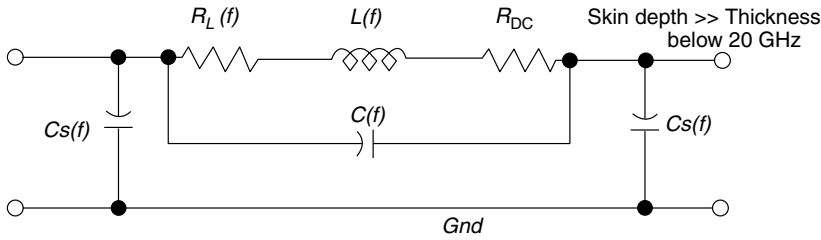


Figure 1.99 Equivalent model of a film resistor above a ground plane.

suitable for propagation, at a level high enough to overcome worst-case propagation losses and provide a useful signal at the receiver. Gain is therefore an essential attribute of wireless systems. Because no single active device can provide all the gain required for transmission or reception, we must distribute the gain among multiple stages, designing each for optimum performance across the power span it bridges.

Two inescapable realities impose limits on the gain and absolute power output, we may achieve with a given circuit. All real electrical and electronic networks generate noise to some degree, and all real electrical and electronic networks distort the signals applied to them to some degree. A signal weaker than a circuit's inherent noise cannot be amplified by that circuit because it remains indistinguishable from the noise. A signal that exceeds the power-handling capability of the circuit to which it is applied may be degraded, even rendered useless, by the resulting distortion. The following sections examine issues particular to system noise and linearity performance.

1.7.1 System Noise and Noise Floor

Assuming that a system's gain is sufficient, the weakest signal that may be processed satisfactorily, a figure of merit referred to as noise floor or (in receivers) minimum detectable signal (MDS), is limited by thermal noise, assumed to be equal to the noise power available from a resistor at 290 K (about 17°C or 62°F), an arbitrary reference value near standard room temperature. The noise power is equal to

$$P_n = kTB \quad (1.32)$$

where P_n is the noise power, k is Boltzmann's constant ($1.38 \cdot 10^{-23}$ W/K), T is the temperature in kelvins, and B is the bandwidth (in Hertz) in which the noise appears. For $T = 290$ K, P_n is, therefore, $4.00 \cdot 10^{-18}$ W, or -174 dBm in a 1-Hz bandwidth. Increasing B to a value suitable for digital communications, such as 160 kHz for a GSM system, admits more noise to the network, raising the minimum noise power against which an incoming signal must compete to -122 dBm.

If the noise figure and bandwidth are known, the system noise floor can be calculated using the equation

$$\text{Noise floor} = -174 \text{ dBm} + NF + 10 \log B \quad (1.33)$$

The trouble with this equation is that the “integrated” bandwidth depends so much on the selectivity shape factor, which is not always known. Figure 1.100 shows the translation of the bandwidth of a single tuned circuit with its Gaussian shape into its rectangular equivalent. The transformation is done by sizing the rectangle such that $A' = A$ and $B' = B$; when this is true, the area of the rectangular equals the area under the curve.

Table 1.11 Sample RF Component/IC Selector Guide

RFICs										
Downconverters										
Device	RF Freq. Range (MHz)	Supply Volt. Range (V/dc)	Supply Current (mA) (Typ)	LNA Gain (dB) (Typ)	LNA NF (dB) (Typ)	Mixer Conv. Gain (dB) (Typ)	Mixer NF(dB) (Typ)	Package	System Applicability	
MC13142	dc to 1800	2.7–6.5	13.5	17	1.8	–3.0	12	SO-8	ISM, Cellular, PCS	
MRFC1502	1575	4.5–5.5	52	20	–	45	9.5	LQFP-48	GPS	
MRFC1814	1800 to 2000	2.7–4.5	10	17	2.5	8.0	10	TSSOP-16	DCS1800, PCS, PHS	
Upconverters/Exciters										
Device	RF Freq. Range (MHz)	Supply Volt. Range (V/dc)	Supply Current (mA) (Typ)	Standby Current (Typ)	Conv. Gain (dB) (mA) (Typ)	Output IP3 (dBm) (Typ)	Package	System Applicability		
MRFC0954	800 to 1000	2.7–5.0	65	5.0	31	28	TSSOP-20EP	CDMA, TDMA, ISM		
MRFC1813	1700 to 2000	2.7–4.5	25	0.1	15	11	TSSOP-16	DCS1800, PCS		
MRFC1854	1700 to 2000	2.7–5.0	70	5.0	31	23	TSSOP-20EP	CDMA, TDMA, PCS		
Power Amplifiers										
Device	Freq. Range (MHz)	Supply Volt. Range (V/dc)	Saturated P _{out} (dBm) (Typ)	PAE% (Typ)	Gain P _{out} /P _{in} (dB) (Typ)	Package	System Applicability			
MRFC0913	800 to 1000	2.7–7.5	35	50	25	PFP-16	GSM			
MRFC0917	800 to 1000	2.7–5.5	34.5	45	22.5	PFP-16	GSM			
MRFC0919	800 to 1000	3.0–5.5	35.3	48	32.3	TSSOP-16EP	GSM			
MRFC1805	1500 to 2200	2.7–5.0	25	28	20	TSSOP-16	PHS, DECT, PCS			

(continued)

Table 1.11 (Continued)*Power Amplifiers*

Device	Freq. Range (MHz)	Supply Volt. Range (V/dc)	Saturated P_{out} dBm (Typ)	PAE% (Typ)	Gain P_{out}/P_{in} dB (Typ)	Package	System Applicability
MRFIC1807	1500 to 2200	3.0–5.0	26.8	35	8.0	SO-16	DECT, PCS
MRFIC1817	1700 to 2000	2.7–5.0	33.5	42	30.5	PFP-16	DCS1800, PCS
MRFIC1818	1700 to 2000	2.7–6.0	34.5	42	31.5	PFP-16	DCS1800, PCS
MRFIC1819	1700 to 2000	3.0–5.0	33	40	27	TSSOP-16EP	DCS1800, PCS
MRFIC1856	800 to 1000	3.0–5.6	32	50	32	TSSOP-20EP	TDMA, CDMA, AMPS
	1700 to 2000		30	35	30		TDMA, CDMA, PCS
MRFIC2006	500 to 1000	1.8–4.0	15.5	25	23	SO-8	Cellular, ISM, CT2

*RF Building Blocks**Amplifiers*

Device	RF Freq. Range (MHz)	Supply Volt. Range V/dc	Supply Current mA (Typ)	Standby Current μ A(Typ)	Small Signal Gain dB (Typ)	Output IP3 dBm (Typ)	NF dB (Typ)	Package	System Applicability
MC13144	100 to 2000	1.8–6.0	8.5	1	17	–5.0	1.4	SO-8	ISM, PCS, Cellular
MRFIC0915	100 to 2500	2.7–5.0	2.0	–	16.2	4.0	1.9	SOT-143	ISM, PCS, Cellular
MRFIC0916	100 to 2500	2.7–5.0	4.7	–	18.5	11	1.9	SOT-143	ISM, PCS, Cellular
MRFIC0930	800 to 1000	2.7–4.5	8.5	20	19	10	1.7	SO-8	GSM, AMPS, ISM
MRFIC1808DM	1700 to 2100	2.7–4.5	5.0	8.0	18	13	1.6	Micro-8	DCS1800, PCS
MRFIC1830	1700 to 2100	2.7–4.5	9.0	20	18.5	8.5	2.1	Micro-8	DCS1800, PCS
MRFIC1501	1000 to 2000	3.0–5.0	5.9	–	18	10	1.1	SO-8	GPS

Mixers

Device	RF Freq. Range (MHz)	Supply Volt. Range (V/dc)	Supply Current mA (Typ)	Standby Current μ A (Typ)	Conv. Gain dB (Typ)	Input IP3 dBm (Typ)	Package	System Applicability
MC13143	DC to 2400	1.8–6.0	4.1	–	–2.6	16	SO-8	ISM, PCS, Cellular

Frequency Synthesis

PLL Synthesizers

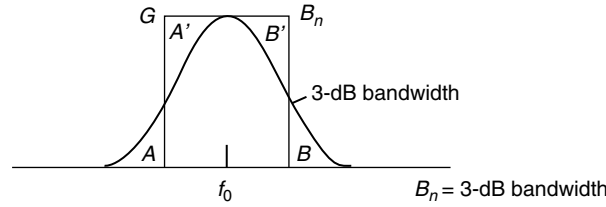
Frequency (MHz)	Supply Voltage (V)	Nominal Supply Current (mA)	Phase Detector	Standby	Interface	Device	Suffix/ Case
20 @ 5.0 V	3.0–9.0	7.5 @ 5 V	Single-ended 3-state, double-ended	No	Parallel	MC145151-2	DW1751F
			Double-ended			MC145152-2	DW/751F
			Single-ended 3-state, double-ended			MC145157-2	DW/751G
60 @ 3.0 V	2.5–5.5	3 @ 3 V	Two single-ended 3-state	Yes	Serial	MC145158-2	DW/751G
						MC145162 ^a	P/648, D/751B
						MC145162-1 ^a	D/751B
100 @ 3.0 V 185 @ 5.0 V	2.7–5.5	2 @ 3 V 6 @ 5 V	Single-ended 3-state, double-ended	No		MC145170-2	P/648, D/751B, DT/948C
						MC145181 ^a (46a)	FTA873C
550,60	1.8–3.6	3	Loop 1 = Current source/sink/float Loop 2 = Three-state	Yes			
1000	2.7–5.5	4.25	Current source/sink/float	No	Parallel	MC12181	D/751B
1100	4.5–5.5	7 @ 5 V	Current source/sink/float, double-ended	Yes	Serial	MC145191	F/751J, DT/948D
1100	2.7–5	6 @ 2.7 V				MC145192	F/751J, DT/948D

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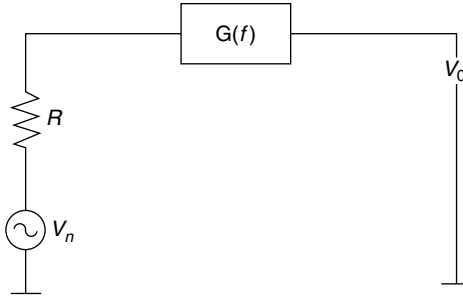
Table 1.11 (Continued)

Frequency (MHz)	Supply Voltage (V)	Nominal Supply Current (mA)	Phase Detector	Standby	Interface	Device	Suffix/ Case
1100	2.7–5.5	12	Two current source/sink/float, double-ended			MC145220 ^a	F/803C, DT/9480
1200, 550	1.8–3.6	4	Loop 1 = current source/sink/float Loop 2 = three-state			MC145225 ^a (46a)	FTA/873C
2000	4.5–5.5	12 @ 5 V	Current source/sink, double-ended			MC145201	F/751J, DT/948D
2000	2.7–5.5	4 @ 3 V				MC145202	F/751J, DT/9480
2200, 550	1.8–3.6	5	Loop 1 = current source/sink/float Loop 2 = Three-state			MC145230 ^a (46a)	FTA/873C
2500	2.7–5.5	9.5	Current source/sink/float with dual outputs	No	None	MC12210	D/751B, DT/948E
2800	4.5–5.5	3.5	Current source/sink/float			MC12179	D/751

^aDual PLL.



$$G(f) = \left| \frac{V_0(f)}{V_n(f)} \right|^2$$



$$\begin{aligned} V_0^2 &= \int_0^\infty 4\kappa T_0 R G(f) dF \\ &= 4\kappa T_0 R \int_0^\infty G(f) dF \\ B_n &= \frac{1}{G} \int_0^\infty G(f) dF \\ B_n &= \text{noise bandwidth} \end{aligned}$$

Figure 1.100 Graphical and mathematical explanation of the noise bandwidth from a comparison of the Gaussian-shaped bandwidth to the rectangular filter response.

Signal-to-Noise Ratio (S/N , SNR) and Sensitivity Successful radiocommunication depends on the achievement of a specified minimum ratio of signal power to noise power, expressed in decibels, at the output of the receiver. The input voltage, expressed in absolute units or decibels relative to a microvolt ($\text{dB}\mu\text{V}$), necessary to achieve a particular signal-to-noise ratio in a particular bandwidth may be specified as a figure of merit called sensitivity. Because techniques used to measure S/N actually measure the ratio of signal-plus-noise to noise, specifications may refer to (or imply) $S + N/N$ or $(S + N)/N$ rather than S/N . The difference between $(S + N)/N$ and S/N becomes negligible at high ratios of signal to noise; even at an SNR of 10 dB—a common value—the difference is only 0.46 dB [28].

Most receivers are designed to operate optimally when connected to an antenna system of a specified impedance (commonly $50\ \Omega$), but relatively few receivers exhibit this design load impedance at their input terminals; that is, they are not designed for a conjugate input match. It is therefore customary to specify sensitivity in terms of “open circuit” voltage—the signal voltage that, with the receiver’s antenna input terminated in its design antenna impedance, results in the desired ratio of signal to noise. If, as is commonly the case, the

input voltage for a given SNR is determined using instrumentation calibrated in terms of closed-circuit voltage—that is, in terms of voltage across a load resistance equal to the instrument's source resistance—the voltage indicated will be 1/2 the open-circuit value for the SNR specified [29]. By convention, the open-circuit measurement condition is indicated by a sensitivity specification in volts of electromotive force (EMF). Specifying sensitivity in terms of available signal power (usually decibels relative to 1 mW or dBm) eliminates this open/closed-circuit confusion.

SINAD Ratio Extending the measurement of signal-plus-noise to noise to include distortion results in a figure of merit called SINAD (signal-plus-noise-and-distortion), commonly applied to FM receivers

$$\text{SINAD} = 10 \log_{10} \frac{S + N + D}{N + D} \quad (1.34)$$

where SINAD is in decibels, S is signal power, N is noise power, and D is distortion power. At a SINAD ratio of 12 dB—a common specification—the noise-and-distortion power is 25% that of the desired signal. As is true of $(S + N)/N$, SINAD closely approximates S/N at high ratios of signal to noise.

Bit Error Rate and Noise For digital systems, signal-to-noise ratio and bit error rate are related. As introduced earlier, depending on the waveform, coding, and filtering, different BERs are related to particular SNRs (Figure 1.101). The SNR also depends on the interference from the synthesizer, as will be seen in the section on adjacent channel power ratio (ACPR). Since a digital transmitter is always involved, in the on-the-air testing, the laboratory measurement of SINAD cannot quite be correlated. At the IF level, we find the handover point between the transceiver's analog front end and its digital portion, and a standard noise-figure test setup, such as the one by Agilent, is still a valuable instrument for evaluating noise figure and signal-to-noise ratio.

Here is an example for a digital measurement. Assume that we use a test generator like the SMIQ to evaluate a front end having an 8.8-dB noise figure. A 384-kbps $\pi/4$ DQPSK digital modulation signal (−5 dBm) is applied to the transceiver front end. Testing at 2.47 GHz,

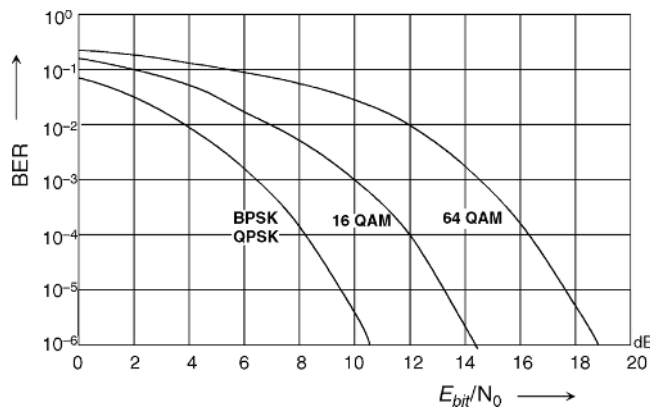


Figure 1.101 Bit error rate versus E_{bit}/N_0 for BPSK/QPSK, 16-QAM and 64-QAM, showing the significantly greater SNR necessary for a given BER as the number of signal states is increased.

we find that the error vector magnitude (EVM; defined later in this section) is about 3.89% in transmit and about 1.37% in receive. To determine the receive sensitivity of the RF front end, the signal from the digitally modulated generator (at -95 dBm) is applied to the RF front end in receive mode. The measured EVM at 2.47 GHz is now 20%. By using the equation

$$S/N = -20 \log(\text{EVM}) \quad (1.35)$$

we see that the ratio that corresponds to an EVM of 20% is 14 dB.

Since

$$S/N = \frac{(E_S R_S)}{N_0 B} \quad (1.36)$$

where E_S is the energy of a symbol, R_S is the symbol rate, N_0 is the noise power density, and B is the bandwidth. Assuming a $\pi/4$ DQPSK signal and ratio of 14 dB, it can be determined that

$$E_S/N_0 = S/N(B/R_S) = 17 \text{ dB} \quad (1.37)$$

Assuming that the BER is 10^{-5} , the required E_S/N_0 is 10 dB. Hence, from the definition of the sensitivity

$$\begin{aligned} \text{Sensitivity} &= -144 + NF + R_S + E_S/N_0 \\ &= -144 + 8.8 + 10 \log(384/2) + 10 \\ &= -102 \text{ dBm} \end{aligned} \quad (1.38)$$

The value -144 is derived from kT_0 by taking the bandwidth (1 kHz) into consideration.

The result obtained at (1.38) is consistent with the obtained E_S/N_0 of 17 dB from the measurement at -95 dBm input power (since $-102 + [17 - 10] = -95$). Hence, it was determined that the receive sensitivity of the RF front end is approximately -102 dBm. We can then determine the dynamic range (DR) from

$$\text{DR} = P_{-1\text{dB}} - \text{Sensitivity} = -21 - 102 = 81 \text{ dB} \quad (1.39)$$

Noise Factor and Noise Figure The noise floor predicted by (1.32) cannot be achieved and maintained in any real network or system of networks because all real networks generate noise. Determining how closely the SNR achieved at a given input level approaches the SNR achievable at that input level in a noiseless network is therefore of high interest to the circuit and system designer. The degree to which a network's noise contribution degrades the noise floor predicted by (1.32) is evaluated by its noise factor (F), which is expressed as the ratio

$$F = \frac{N_{\text{in}} + N_{\text{added}}}{N_{\text{in}}} \quad (1.40)$$

where F is noise factor, N_{in} is the noise power available from the source and N_{added} is the noise power added by the network, with both powers determined in the same bandwidth. Expressing this ratio in decibels ($10 \log_{10} F$) returns noise figure (NF), a bandwidth-independent figure of merit of great value in evaluating the noise performance of networks and communication systems. We can also express NF as the ratio of the network's input

SNR to its output SNR:

$$NF = 10 \log_{10} \left[\frac{(S_{in}/N_{in})}{(S_{out}/N_{out})} \right] \quad (1.41)$$

where NF is noise figure in decibels, S is signal power, and N is noise power, with the input and output values of these quantities signified by the subscripts and all powers determined in the same bandwidth. The noise figure of an ideal noiseless network is 0 dB; for all real, noisy networks, NF is positive. The NF of a lossy passive device is equal to its insertion loss.

Noise figure can also be defined for antennas and antenna systems:

$$NF_{ant} = 10 \log_{10} \frac{N_t + N_{ant}}{N_t} \quad (1.42)$$

where NF_{ant} is the antenna system noise figure in decibels, N_t is the antenna's system's thermal noise power, as defined in (1.32), and N_{ant} is the total noise power picked up by the antenna system. From the lower end of the radio spectrum, and decreasingly up to approximately 400 MHz, noise intercepted by an antenna system from atmospheric, man-made, and galactic sources will dominate NF_{ant} (Figure 1.102), and N_t can be considered as equivalent to the noise power of a resistor at 290 K. Atmospheric noise subsides above 40 MHz; from this region to perhaps 700 MHz, N_t is still largely negligible, with noise from man-made and/or sky sources largely determining an antenna's noise figure. At these frequencies, an antenna's directivity and orientation relative to noise sources play a critical role in determining its noise figure; a strongly directive antenna located in a city suburb, for instance, exhibits a significantly higher noise figure when pointed toward the city center than it exhibits when pointed toward a more sparsely populated area. At frequencies above about 700 MHz, the RF environment is generally quieter, allowing receiver NF , not antenna NF , to more routinely limit a wireless receiver's sensitivity. At these frequencies, the concept

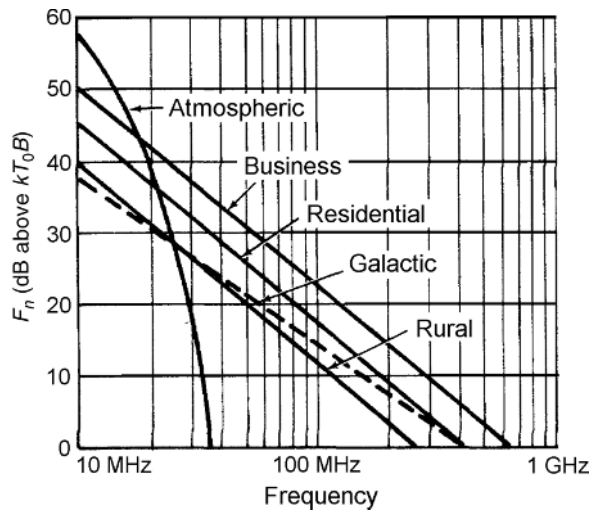


Figure 1.102 The higher the frequency, the more quiet the RF environment becomes, although the noise profile of specific sources may contradict this general rule.

of noise temperature is commonly used to evaluate the quietness of a receiver, its antenna system, and its RF environment. Noise temperature is particularly useful in designing systems for space communication, an antenna for which may be cooled, through radiation of a portion its own noise into the RF-cold sky, to a noise temperature an order of magnitude below 290 K.

Noise Figure of Cascaded Networks The noise figure of two networks in cascade may be determined from

$$\text{NF}_{\text{total}} = 10 \log_{10} \left(F_1 + \frac{F_2 - 1}{G_1} \right) \quad (1.43)$$

where NF is noise figure in dB, F_1 is the noise factor of the first network, F_2 is the noise factor of the second network, and G_1 is the gain (as a numerical ratio, not in dB) [30]. The noise figure of a system with more than two stages can be evaluated through repeated iterations of (1.43). Note that (1.43) assumes two conditions: (1) that F_1 and F_2 are determined in the same bandwidth, and (2) that the networks' input and output terminations are resistive—a condition that is commonly not true of RF amplifiers optimized for lowest noise. Equation (1.43) can be expanded to account for bandwidth, but accounting for complex terminations requires the use of noise correlation matrix techniques as described in Ref. [28].

1.7.2 System Amplitude and Phase Behavior

If we could build electronic systems that were absolutely amplitude- and phase-linear, radiocommunication system design would be greatly simplified. An amplifier designed for a power gain of 10 dB, for example, would merely increase the magnitude of all signals at its input by a factor of 10, regardless of their frequencies, while perfectly maintaining their relative phases. But all real electrical and electronic networks, even those designed (or supposed) to be amplitude- and phase-linear, exhibit amplitude and phase nonlinearity to some degree, just as they all generate noise to some degree.

The effects of amplitude nonlinearity, generically referred to as nonlinear distortion, include the generation, through harmonic distortion and/or intermodulation distortion (IMD), of output signals at frequencies not present at a system's input. Nonlinear distortion also results in gain compression—changes in system gain with changes in input-signal level. By convention, when workers in electronics refer to or consider a network's "linearity," they usually mean its amplitude linearity; likewise, by "distortion" they usually mean nonlinear distortion.

The effects of phase or frequency nonlinearity are generically referred to as linear distortion because they occur independently of signal amplitude and polarity. We often intentionally apply linear distortion through filtering, which modifies the amplitude relationships among existing spectral components of a signal without creating any new frequencies. Another linear-distortion effect, phase or delay distortion, results in the delay of signals of differing frequencies by differing amounts of time. In a system where signal phase conveys information, as is true of most wireless links, phase distortion can seriously degrade communication.

The fact that all real networks are amplitude- and angle-nonlinear to some degree means that all real networks modify the amplitude and angle characteristics of the signals they handle. What is perhaps less obvious is that subjecting a signal to amplitude and angle nonlinearities causes "crosstalk" between its amplitude and angle characteristics.

For example, through a nonlinear distortion effect called AM-to-PM conversion, changes in a signal's amplitude result in changes in its phase. Filtering an angle-modulated signal produces the reverse effect; deprived by the filter's selectivity of spectral components necessary to maintain its envelope constancy, it emerges from the filter as a combination of angle and amplitude modulation.

Spectral Considerations of Analog and Digitally Modulated Signals Many wireless modulation schemes exist and even more are proposed; all use angle (usually phase) modulation or a combination of phase and amplitude modulation using an emission format engineered to achieve multiple goals of information throughput, robustness, spectral and hardware efficiency, and reproducibility. Digital modulation is standard in mainstream wireless applications because it allows increased channel capacity, and immunity to noise and distortion, compared to analog systems. At the circuit-design level, the particulars of whether, or to what degree, a modulation scheme is AM, PM, analog, or digital, matters less than the actual angle and amplitude characteristics of the signal(s) involved, and the tolerances within which these characteristics can be expected and allowed to vary.

A system's amplitude linearity is of major concern because of its relation to energy efficiency, receiver dynamic range, and transmitter spectral purity. Energy efficiency, important because many wireless applications use battery power, generally decreases inversely with amplitude linearity. Yet, sufficient amplitude linearity must be guaranteed in the front-end circuitry of wireless receivers subjected to multiple strong signals, and in all wireless receiver and transmitter stages handling variable-envelope signals.

Some digital modulation schemes, designed to be distortion-tolerant, produce constant-envelope signals; Gaussian minimum shift keying (GMSK) and Feher's quadrature phase-shift keying (FQPSK) are examples of these. Such signals can be processed in highly nonlinear circuitry—for example, an RF power amplifier operated in saturation to maximize efficiency—without degrading their spectral composition. Other digital modulation schemes result in the emission of, ideally, of a single PM sideband, with the carrier and all other sideband components suppressed. Such a signal exhibits, of necessity, phase and amplitude variations; the IS-54 cellular standard's $\pi/4$ differential quadrature phase-shift keying ($\pi/4$ DQPSK) scheme, which results in envelope fluctuations of 3–6 dB, exemplifies this [31]. Intermodulation distortion among such a signal's spectral components can generate products that fall outside the bandwidth painstakingly controlled in the modulation process. This spectral regrowth must be minimized to prevent adjacent-channel interference (Figure 1.103).

For a better understanding, Figure 1.104 shows the various channels and the energy levels associated with them.

Amplitude Linearity Issues and Figures of Merit A network's amplitude nonlinearity can be characterized by the expansion

$$y = k_1 f(x) + k_2 [f(x)]^2 + k_3 [f(x)]^3 + \text{higher-order terms} \quad (1.44)$$

where y represents the output, the coefficients k_n represent complex quantities whose values can be determined by an analysis of the output waveforms, and $f(x)$ represents the input. Even though all practical networks exhibit amplitude nonlinearity, we can (and often do) refer to many networks as “linear.” We say this of networks that are sufficiently amplitude-linear for our purposes—for example, weakly nonlinear networks in which small-signal

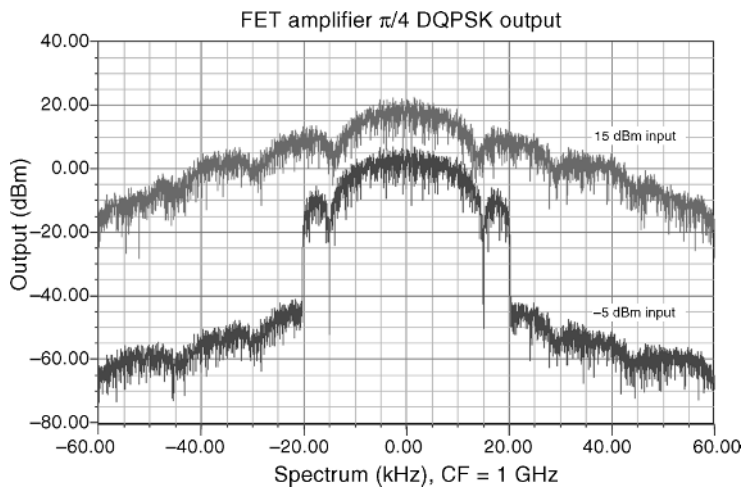


Figure 1.103 Spectral regrowth results when a variable-envelope emission, $\pi/4$ DQPSK in this case, is subjected to significant nonlinear distortion. This graph shows the simulated performance of a MESFET power amplifier operating at 1 GHz; the amplifier is 6 dB into compression when driven at 15 dBm.

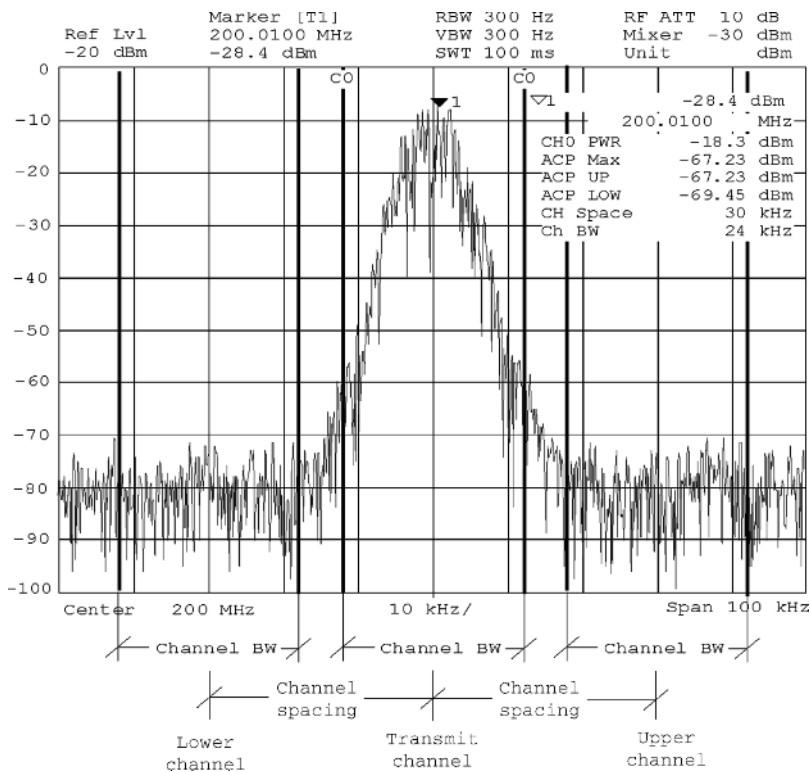


Figure 1.104 NADC signal and parameters, including channel spacing and channel bandwidth.

operation is assumed even though the signal levels involved are sufficient to cause slight distortion. For many practical purposes, the first three terms of (1.44) adequately describe such a network's nonlinearity:

$$y = k_1 f(x) + k_2 [f(x)]^2 + k_3 [f(x)]^3 \quad (1.45)$$

In adopting this simplification, we also assume that the nonlinearity is frequency independent—that is, the network has sufficient bandwidth to allow all the products predicted by (1.45) to appear at its output terminals unperturbed [32].

When multiple signals are present in a network, even weak nonlinearity can result in profound consequences. To illustrate this, we will let $f(x)$ consist of two sinusoidal signals:

$$f(x) = A_1 \cos \omega_1 t + A_2 \cos \omega_2 t \quad (1.46)$$

We will assume that ω_1 and ω_2 are close enough so that the coefficients k_i can be considered equal for both signals. We will also assume for simplicity that all the k_i are real. If (1.45) describes the network's response to an input $f(x)$, the response will be

$$\begin{aligned} y &= k_1(A_1 \cos \omega_1 t + A_2 \cos \omega_2 t) + k_2(A_1 \cos \omega_1 t + A_2 \cos \omega_2 t)^2 \\ &\quad + k_3(A_1 \cos \omega_1 t + A_2 \cos \omega_2 t)^3 \\ &= k_1(A_1 \cos \omega_1 t + A_2 \cos \omega_2 t) \\ &\quad + k_2 \left[A_1^2 \frac{1 + \cos 2\omega_1 t}{2} + A_2^2 \frac{1 + \cos 2\omega_2 t}{2} + A_1 A_2 \frac{\cos(\omega_1 + \omega_2)t + \cos(\omega_1 - \omega_2)t}{2} \right] \\ &\quad + k_3 \left\{ A_1^3 \left(\frac{\cos \omega_1 t}{2} + \frac{\cos \omega_1 t}{4} + \frac{\cos 3\omega_1 t}{4} \right) + A_2^3 \left(\frac{3 \cos \omega_2 t}{4} + \frac{\cos 3\omega_2 t}{4} \right) \right. \\ &\quad + A_1^2 A_2 \left[\frac{3}{2} \cos \omega_2 t + \frac{3}{4} \cos(2\omega_1 - \omega_2)t + \frac{3}{4} \cos(2\omega_1 + \omega_2)t \right] \\ &\quad \left. + A_2^2 A_1 \left[\frac{3}{2} \cos \omega_1 t + \frac{3}{4} \cos(2\omega_2 + \omega_1)t + \frac{3}{4} \cos(2\omega_2 - \omega_1)t \right] \right\} \end{aligned} \quad (1.47)$$

The k_1 term of equation (1.47) represents the results of amplitude-linear behavior. No new frequency components have appeared; the two sine waves have merely been “rescaled” by k_1 .

The second- and third-order terms of (1.47) represent the effects of harmonic distortion and intermodulation distortion. Second-order effects include second-harmonic distortion (the production of new signals at $2\omega_1$ and $2\omega_2$) and IMD (the production of new signals at $\omega_1 + \omega_2$ and $\omega_1 - \omega_2$). Third-order effects include gain compression, third-harmonic distortion (the production of new signals at $3\omega_1$ and $3\omega_2$), and IMD (the production of new signals at $2\omega_1 \pm \omega_2$ and $2\omega_2 \pm \omega_1$).

Gain Compression Gain compression occurs when a network cannot increase its output amplitude in linear proportion to an amplitude increase at its input; gain saturation occurs when a network's output amplitude stops increasing (in practice, it may actually decrease) with increases in input amplitude. We can deduce from (1.47) that the amplitude of the $\cos \omega_1 t$ signal has become

$$A'_1 = k_1 A_1 + k_3 \left(\frac{3}{4} A_1^3 + \frac{3}{2} A_1 A_2^2 \right) \quad (1.48)$$

Because k_3 will normally be negative, a large signal $A_2 \cos \omega_2 t$ can effectively mask a smaller signal $A_1 \cos \omega_1 t$ by reducing the network's gain. This third-order effect, known as blocking or desensitization when it occurs in a receiver, is a special case of gain compression. The presence of additional signals results a greater reduction in gain; the gain reduction for each signal is a function of the relative levels of all signals present. A receiver's blocking behavior may be characterized in terms of the level of off-channel signal necessary to reduce the strength of an in-passband signal by a specified value, typically 1 dB; alternatively, the decibel ratio of the off-channel signal's power to the receiver's noise-floor power may be cited as blocking dynamic range. Desensitization may be also characterized in terms of the off-channel-signal power necessary to degrade a system's SNR by a specified value.

Multiple signals need not be present for gain compression to occur. If only one signal is present, the ratio of gain with distortion to the network's idealized (linear) gain is

$$A'_1 = \frac{k_1 + k_3 \left(\frac{3}{4} A_1^2 \right)}{k_1} \quad (1.49)$$

This is referred to as the single-tone gain-compression factor. Figure 1.105 shows how the k_3 term causes a network's gain to deviate from the ideal. The point at which a network's power gain is down 1 dB from the ideal for a single signal is a figure of merit known as the 1-dB compression point ($P_{-1\text{dB}}$). Many networks (including many receiving and low-level transmitting circuits, such as low-noise amplifiers, mixers, and IF amplifiers) are usually operated under small-signal conditions—at levels sufficiently below $P_{-1\text{dB}}$ to maintain high linearity. As we will see, however, some networks (including power amplifiers for wireless systems) may be operated under large-signal conditions—near or in compression—to achieve optimum efficiency at some specified level of linearity. Figure 1.106 shows what happens when a digital emission that uses amplitude to convey information is subjected to amplitude compression.

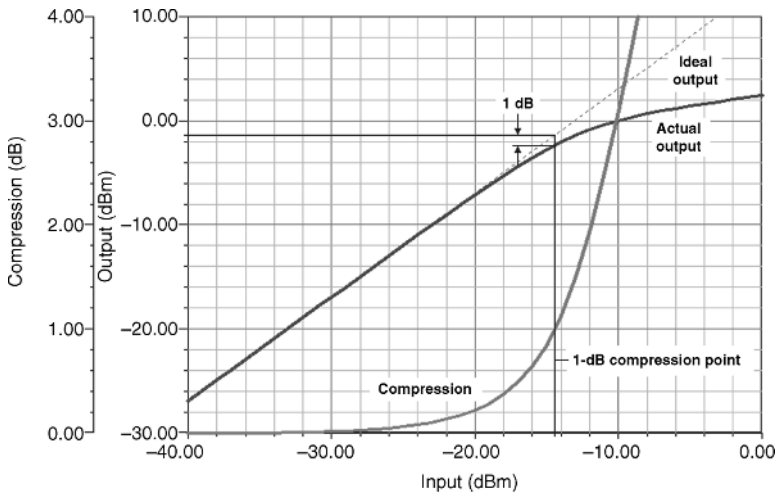


Figure 1.105 The power level at which a network's power output is down 1 dB relative to that of its ideally linear equivalent is a figure of merit known as the 1-dB compression point ($P_{-1\text{dB}}$). The 1-dB compression point can be expressed relative to input power ($P_{-1\text{dB},\text{in}}$) or output power ($P_{-1\text{dB},\text{out}}$). For the amplifier simulated here, $P_{-1\text{dB},\text{in}} \approx -14.5 \text{ dBm}$ and $P_{-1\text{dB},\text{out}} \approx -1.3 \text{ dBm}$.

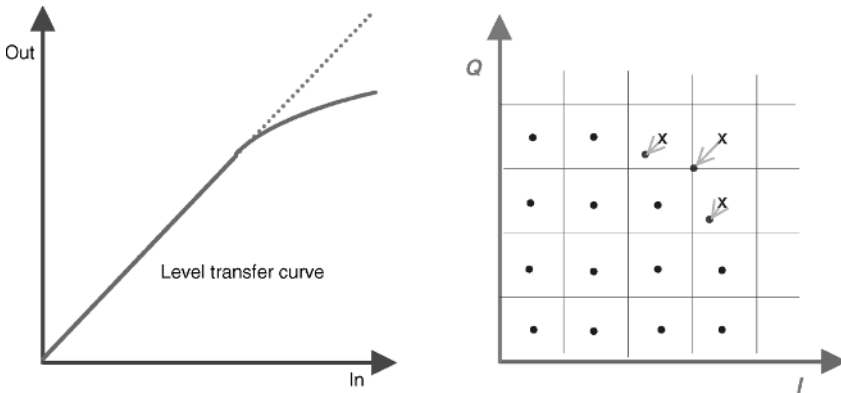


Figure 1.106 Influence of differential amplitude error (compression) on a QAM constellation.

Intermodulation The new signals produced through intermodulation distortion (IMD) can profoundly affect the performance even of systems operated far below gain compression (Figure 1.107). IMD products of significant power can appear at frequencies remote from, in and/or near the system passband, resulting in demodulation errors (in reception) and interference to other communications (in transmission). Where an IMD product appears relative to the passband depends on the passband width and center frequency, the frequencies of the signals present at the system input, and the order of the nonlinearity involved. These factors also determine the strength of an IMD product relative to the desired signal.

Second-order IMD (IM_2) results, for an input consisting of two signals ω_1 and ω_2 , in the production of new signals at $\omega_1 + \omega_2$ and $\omega_1 - \omega_2$; third-order IMD (IM_3) results, for an input consisting of two signals ω_1 and ω_2 , in the production of new signals at $2\omega_1 \pm \omega_2$ and $2\omega_2 \pm \omega_1$.

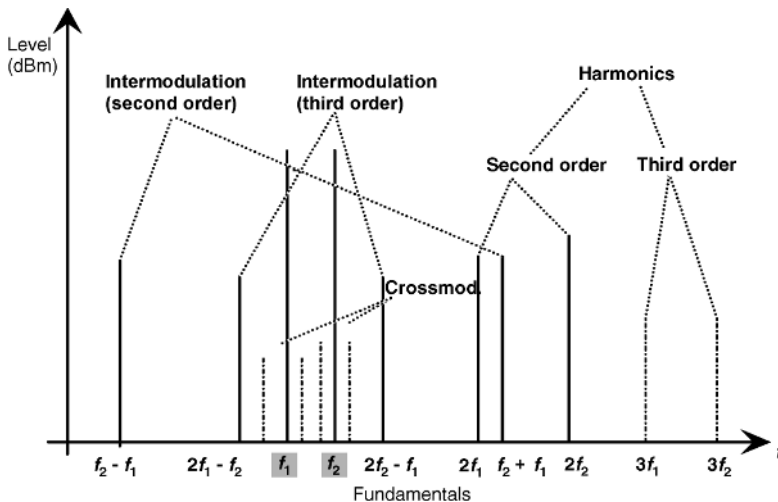


Figure 1.107 Relationships between fundamental and spurious signals, including harmonics and products of intermodulation.

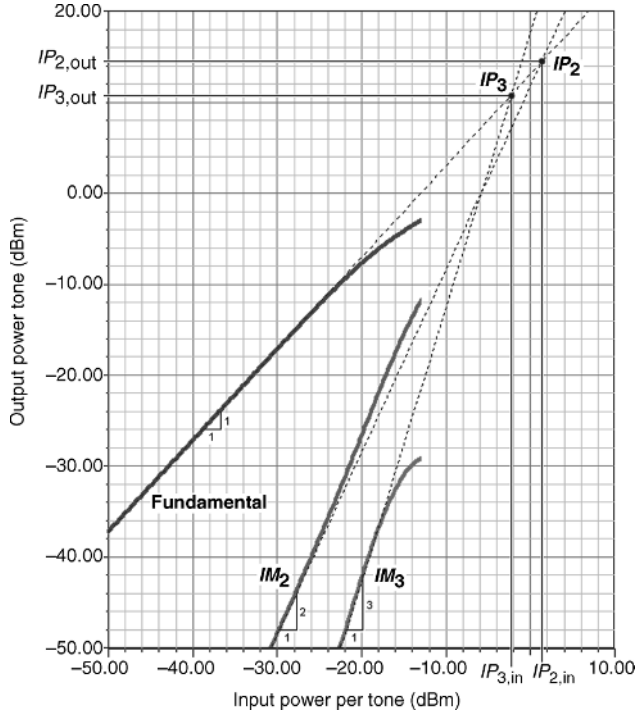


Figure 1.108 The level at which the power of one of a network's IM products equals that of the network's linear output is a figure of merit known as the intermodulation intercept point (IP). The intercept point for a given IM order n can be expressed, and should always be characterized, relative to input power ($IP_{n,in}$) or output power ($IP_{n,out}$); the IP_{in} and IP_{out} values differ by the network's linear gain. For the amplifier simulated here, $IP_{2,in} \approx 1.5$ dBm, $IP_{2,out} \approx 14.5$ dBm, $IP_{3,in} \approx -2.3$ dBm, and $IP_{3,out} \approx 10.7$ dBm. Each curve depicts the power in one tone of the response evaluated.

Under small-signal conditions—that is, at levels well below compression—the power of an IM_2 product varies by 2 dB, and the power of an IM_3 product varies by 3 dB, per decibel change in input power level. This allows us to derive a network figure of merit, the intermodulation intercept point (IP), for a given IM order by extrapolating a network's linear and IM responses to their point of intersection (Figure 1.108)—the point at which their powers would be equal if compression did not occur. Because of the system noise and/or intermodulation distortion products, there is a minimum discernible signal (MDS) that limits the dynamic range at the lower end. Theoretically, Figure 1.108 should show a noise floor or IMD-spur floor for a given input signal that represents a lower limit below which signals cannot be detected. The intercept point for a given IM order n can be expressed, and should always be characterized, relative to input power ($IP_{n,in}$) or output power ($IP_{n,out}$); the IP_{in} and IP_{out} values differ by the network's linear gain. For equal-level test tones, IP_n , in can be determined by

$$IP_{n,in} = \frac{nP_A - P_{IMn}}{n - 1} \quad (1.50)$$

where n is the order, P_A is the input power (of one tone), P_{IMn} is the power of the IM product, and IP is the intercept point. The intercept point for cascaded networks can be

determined from

$$IP_{2,\text{in}} = \frac{1}{\left(\frac{1}{\sqrt{IP_1}} + \frac{G}{\sqrt{IP_2}}\right)^2} \quad (1.51)$$

for IP_2 and from

$$IP_{3,\text{in}} = \frac{1}{\frac{1}{IP_1} + \frac{G}{IP_2}} \quad (1.52)$$

for IP_3 . In both equations, IP_1 is the input intercept of Stage 1 in watts, IP_2 is the input intercept of Stage 2 in watts, and G is the gain of Stage 1 (as a numerical ratio, not in decibels). Both equations assume the worst-case condition, in which the distortion products of both stages add in-phase.

The ratio of the signal power to the IM-product power, the distortion ratio, can be expressed as

$$R_{dn} = (n - 1) [IP_{n(\text{in})} - P_{(\text{in})}] \quad (1.53)$$

where n is the order, R_{dn} is the distortion ratio, $IP_{n(\text{in})}$ is the input intercept point, and $P_{(\text{in})}$ is the input power of one tone.

Discussions of IMD have traditionally downplayed the importance of IM_2 because the incidental distributed filtering contributed by the tuned circuitry once common in radiocommunication systems was usually enough to render out-of-passband IM_2 products caused by in-passband signals, and in-passband IM_2 products caused by out-of-passband signals, vanishingly weak compared to fundamental and IM_3 signals. In broadband systems that operate at bandwidths of an octave or more, however, in-passband signals may produce significantly strong in-passband IM_2 and second-harmonic products. In such applications, balanced circuit structures (such as push-pull amplifiers and balanced mixers) can be used to minimize IM_2 and other even-order nonlinear products.

As with IM_2 , which IM_3 products are important depends on the spacing of the signals involved and the relative width of the system passband. If ω_1 and ω_2 are of approximately the same frequency, the additive products $2\omega_1 + \omega_2$ and $2\omega_2 + \omega_1$ will be outside the passband of a narrowband system. The subtractive products $2\omega_1 - \omega_2$ and $2\omega_2 - \omega_1$, however, will likely appear near or within the system passband. The IM_3 performance of any network subjected to multiple signals is therefore of critical importance, and an array of IM_3 -related, sometimes application-specific, figures of merit has evolved as a result.

Dynamic Range As we have seen, thermal noise sets the lower limit of the power span over which a network can operate. Distortion—that is, degradation by distortion of the signal’s ability to convey information—sets the upper limit of a network’s power span. Because the power level at which distortion becomes intolerable varies with signal type and application, a generic definition has evolved; the upper limit of a network’s power span is the level at which the power of one IM product of a specified order is equal in power to the network’s noise floor. The ratio of the noise-floor power to the upper-limit signal power is referred to as the network’s dynamic range (DR), often more carefully characterized as two-tone IMD dynamic range, which, when evaluated with equal-power test tones, is a figure of merit commonly used to characterize receivers. The MDS relative to the input, as

already defined, is

$$\text{MDS}_{\text{in}} = kTB + 3 \text{ dB} + \text{NF}$$

When $IP(n)$ in and MDS are known, IMD DR can be determined from

$$\text{DR}_n = \frac{(n-1)[IP_{n(\text{in})} - \text{MDS}_{\text{in}}]}{n} \quad (1.54)$$

where DR is the dynamic range in decibels, n is the order, $IP_{(\text{in})}$ is the input intercept power in dBm, and MDS is the minimum detectable signal power in dBm. The so-called spurious-free dynamic range (SFDR or DRSF) is calculated from

$$\text{DR}_{\text{SF}} = \frac{2}{3} (IP_3 - 174 \text{ dBm} + \text{NF} + 3 \text{ dB})$$

This equation allows us to determine how to measure the spurious-free dynamic range. This is done by applying the two-tone signals (in the case of IP_3) and increasing the two signals to the point where the signal-to-noise ratio deteriorates by 3 dB or, if the measurement is done relative to MDS, where the noise floor rises by 3 dB. The factor 2/3 is derived from the fact that the levels of IM_3 outputs increase 3 dB for 1 dB of input increase. This definition of dynamic range now is referenced to a noise figure rather than a minimum level in dBm, and is therefore independent of bandwidth. (By choosing smaller bandwidths (1 kHz instead of 10 kHz), a dynamic range measurement can be made to look better. Biasing the specification on noise figure directly avoids this problem.)

Triple-Beat Distortion and Cross-Modulation $P_{-1\text{dB}}$ is a single-tone figure of merit; blocking, intercept point and dynamic range evaluate two-tone behavior. For networks that must handle AM and composite (AM and angle modulation) signals very linearly, such as television transmitters and cable TV distribution systems, a three-tone figure of merit called triple-beat distortion has gained acceptance. Signals at ω_1 and ω_2 (closely spaced) and ω_3 (positioned far away from ω_1 and ω_2) are applied to the network under test, at levels, frequencies, and spacings that vary with the application. One triple-beat distortion figure of merit is the ratio, expressed in decibels, of the IM product at $\omega_3 + (\omega_2 - \omega_1)$ to one of the network's linear outputs at a specified output level. Alternatively, the triple-beat figure of merit may express the network output level at which a specified triple-beat ratio occurs.

Triple-beat distortion is the mechanism underlying cross-modulation, a form of intermodulation in which one or more AM signals present in a network amplitude-modulate all signals present in the network [33]. Angle-modulation-based wireless systems are largely immune to such effects.

Figures 1.108 and 1.109 graph the results of gain compression, two-tone intermodulation, cross-modulation, and triple-beat testing on a wideband (5–1000 MHz) amplifier.

Noise Power Ratio Triple-beat testing is one way of improving on two-tone testing as a means of evaluating a network's intermodulation behavior in the presence of multiple signals. Another figure of merit, noise power ratio (NPR), uses thermal noise as a test signal. The test measures the introduction, by IM, of noise into a quiet slot created by the insertion of a band-stop filter, equal in stopband width to the width of the measurement channel, between the noise generator and the network under test (Figure 1.110).

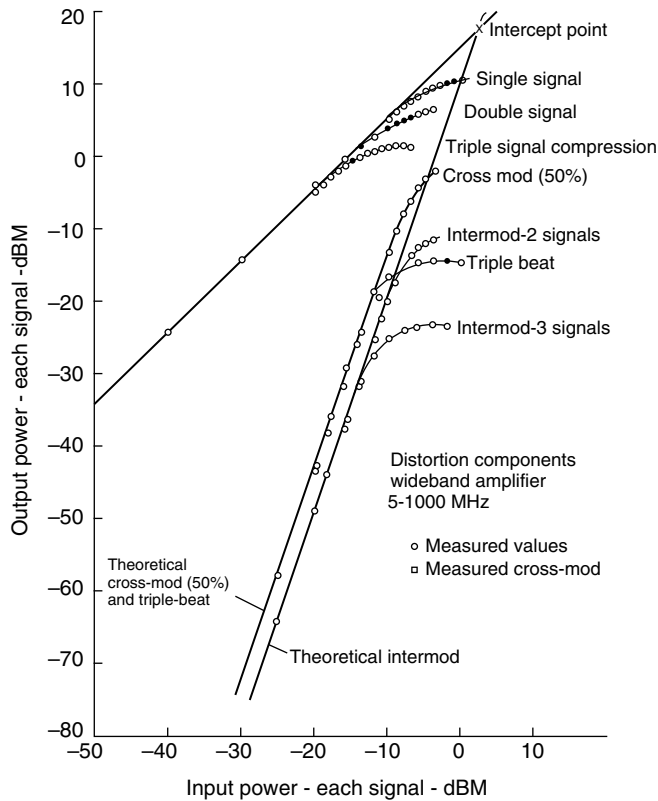


Figure 1.109 Measured distortion components in a wideband (5–1000 MHz) amplifier. Figure 1.108 shows a magnified view of the gain-compression region [34].

Large-Signal Effects Except for $P_{-1\text{dB}}$, the figures of merit discussed so far evaluate amplitude nonlinearity under small-signal conditions. At input powers that drive a network into gain compression and saturation, IM products of odd orders higher than 3 become significant, and curves, dips, and nulls appear in their characteristics (Figure 1.111). Phase shifts related to nonlinear (primarily voltage-dependent) capacitances in solid-state devices are one cause of these effects. Under such conditions, a network may exhibit hysteresis, with its behavior at any given instant depending not only on the voltage or current applied to it but also on its recent history [35].

AM-to-PM Conversion The nonlinear distortion effects we have discussed so far can be termed AM-to-AM distortion—distortion that, to a degree that depends on the amplitude of the signal(s) applied to the network, results in changes in the network's gain, and/or production of signals at new frequencies. AM-to-PM distortion can also occur. As a network nears saturation, part of its driving signal goes into shifting the bias point(s) of its active device(s), changing their drive-dependent reactances and shifting the phase of the output signal relative to its value at input levels below compression (Figures 1.112 and 1.113). This effect, AM-to-PM conversion, can cause incidental phase modulation that degrades the performance of digital communication systems.

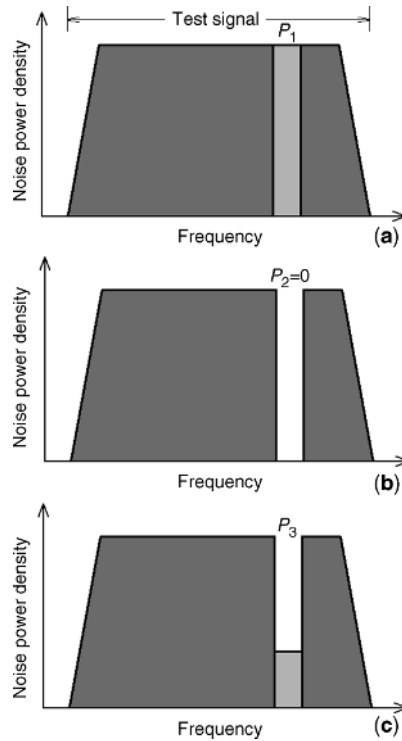


Figure 1.110 Determining a network's noise power ratio (NPR) involves the application of a test signal consisting of thermal noise [34]. The reference measurement-channel noise power, P_1 , is then measured (a). Next, a stop-band filter is placed between the noise generator and network under test to keep the test signal out of the measurement channel (b). Assuming sufficient filter attenuation, if the network were absolutely noiseless and linear, the ideal noise power in the measurement channel, P_2 , would then be zero. In practice, the network's own thermal noise and intermodulation between noise components outside the measurement channel result in an actual measurement-channel noise power (P_3) greater than zero. The noise power ratio equals P_1/P_3 .

Spectral Regrowth and Adjacent-Channel Power Ratio Spectral regrowth occurs largely as a result of third, fifth, and seventh-order IMD in power amplifiers operated near or in compression—at power levels where hysteretic IM effects result in poor agreement between measured behavior and predictions based on small-signal IM figures of merit [36]. We therefore evaluate the impact of spectral regrowth more directly, using a figure of merit called adjacent channel power ratio (ACPR). ACPR measurement techniques that incorporate memory can be used to increase ACPR predictions for networks with that exhibit saturation hysteresis [37]. Figure 1.114 shows the critical relationship between compression, power-added efficiency, and ACPR in a MESFET power amplifier.

Phase Response Issues and Figures of Merit We have already seen how large-signal nonlinear distortion can result in amplitude-dependent phase shifts through AM-to-PM conversion. Because phase linearity is critical at all signal levels in PM systems, especially those using digital modulation, we must also consider linear distortion in evaluating networks used in wireless systems.

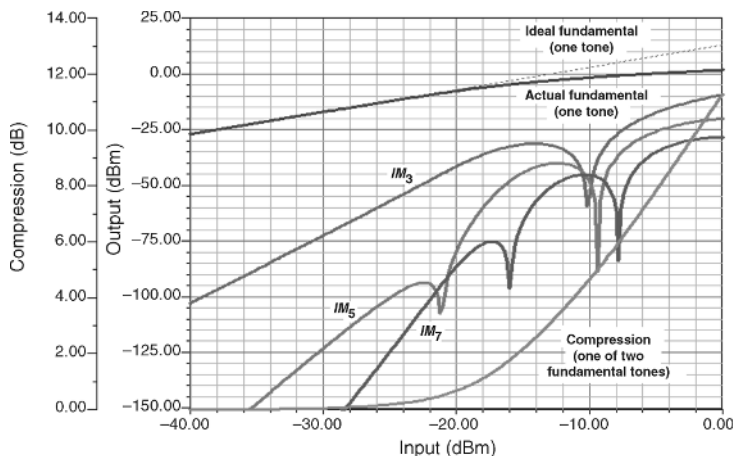


Figure 1.111 As a network is driven into compression, IM products at odd orders higher than 3 become significant, and phase shifts in power-dependent device capacitances cause curves and dips in the IM characteristics. The onset of these departures from IM-response linearity occurs at generally lower input power levels for higher IM orders; their severity, and their position on the IM curves, differs among the various products of a given order and varies with network topology and tone spacing. Figures of merit based on straight-line IM responses fail to usefully predict nonlinear network behavior under these conditions. This graph shows the simulated performance of a single-BJT broadband amplifier driven by two equal-amplitude tones at 10 and 11 MHz.

Differential Group Delay Very frequency-selective network subjects signals passing through it to some degree of time delay. Ideally, this delay, also known as group or envelope delay, does not vary with frequency; that is, the network's phase-shift versus frequency response is monotonic and linear. In practice, a network's time delay varies across its

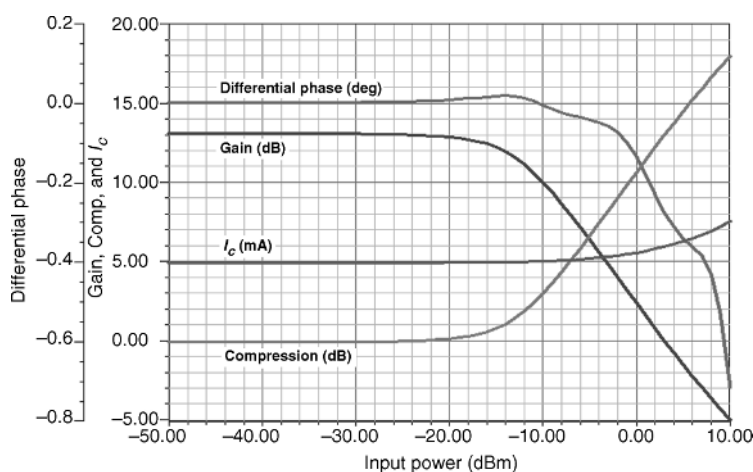


Figure 1.112 Driving a network into compression and saturation shifts the bias point(s) of its active device(s), changing their drive-dependent reactances and shifting the phase of the output signal relative to its value at input levels below compression. This graph shows the simulated performance of a single-BJT broadband amplifier driven by a single tone at 10 MHz.

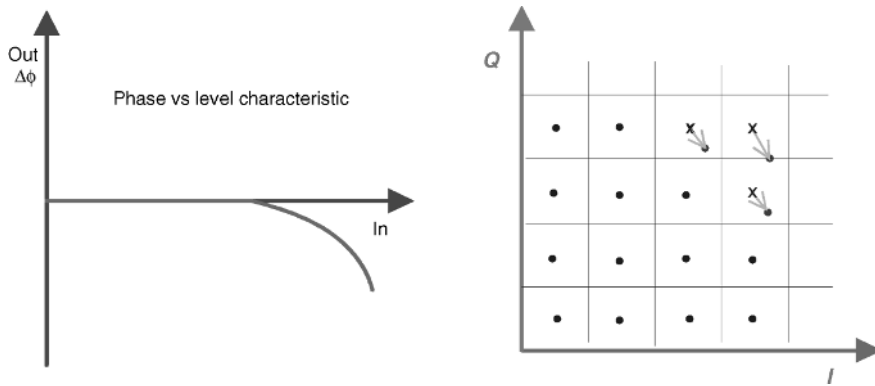


Figure 1.113 Influence of differential phase error (AM-to-PM conversion) on a QAM constellation.

passband, transition bands, and stopbands, exhibiting curvature, ripple, and transition-band peaks (Figure 1.115). The network's differential group delay—its group-delay spread—is therefore of considerable importance. This is especially so in digitally modulated systems, where the resulting phase distortion can cause errors in modulation and demodulation.

Effects of Phase Noise As Chapter 5 will discuss in detail, the phase of an oscillator's output signal is subject to random phase variations (Figures 1.116 and 1.117). Called phase noise, this effect is often quantified as the decibel ratio of the phase noise power in a single (the upper or lower) phase-noise sideband, in a 1-Hz bandwidth centered at a specified frequency offset from the oscillator carrier, to the carrier power (Figure 1.118); alternatively, it may be specified in degrees rms. A microwave voltage-controlled oscilla-

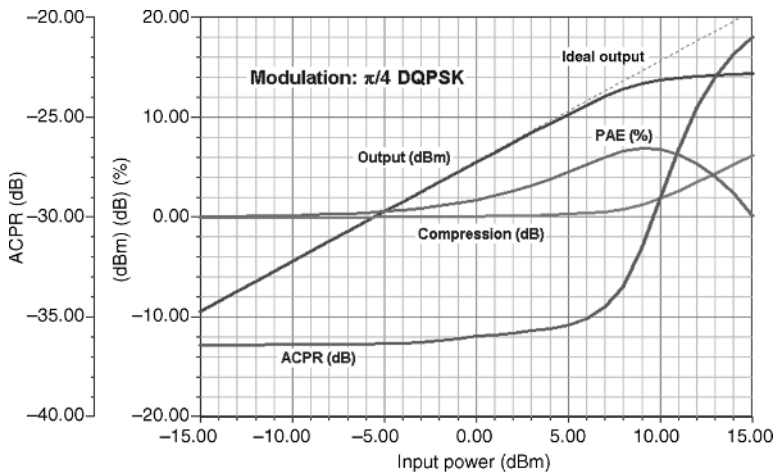


Figure 1.114 Keeping adjacent-channel interference under control can involve a critical trade-off between a wireless transmitter's power-added efficiency (PAE) and ACPR, as shown in this graph of the simulated behavior of the 1-GHz MESFET power amplifier introduced in Figure 1.103. As the amplifier is driven into compression, a peak occurs in the PAE response—in this case, at $P_{-1\text{dB}}$ —and ACPR rises sharply.

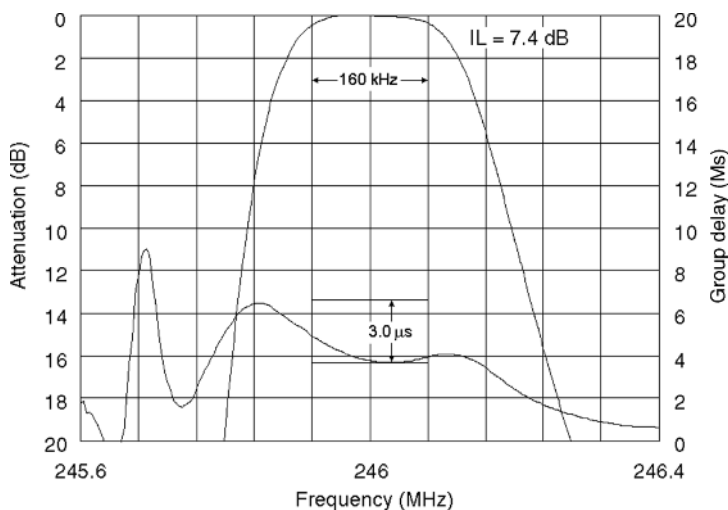


Figure 1.115 Close-in amplitude and group delay responses for a 246-MHz SAW filter designed for GSM applications [38]. This filter is well within its $3.0\ \mu\text{s}$ differential group delay specification across its passband (160 kHz at $-3\ \text{dB}$); the peaks just outside the passband limits are characteristic of a network's transition-band phase response.

tor, for instance, might exhibit an SSB phase noise of $-95\ \text{dBc/Hz}$ at 10 kHz. Oscillator phase noise may manifest itself, through a mechanism known as reciprocal mixing, as the emission of unacceptably strong noise outside a transmitter's occupied bandwidth or as an increase in receiver noise floor. Phase noise may also directly introduce phase errors that result in modulation and demodulation errors.

Because the oscillators used for frequency translation in wireless systems are usually embedded in phase-locked loops, their phase-noise characteristics differ from those of “bare” oscillators as shown in Figures 1.119 and 1.120. Figures 1.121 and 1.122 show the measured phase noise of the Rohde & Schwarz SMY and SMIQ signal generators.

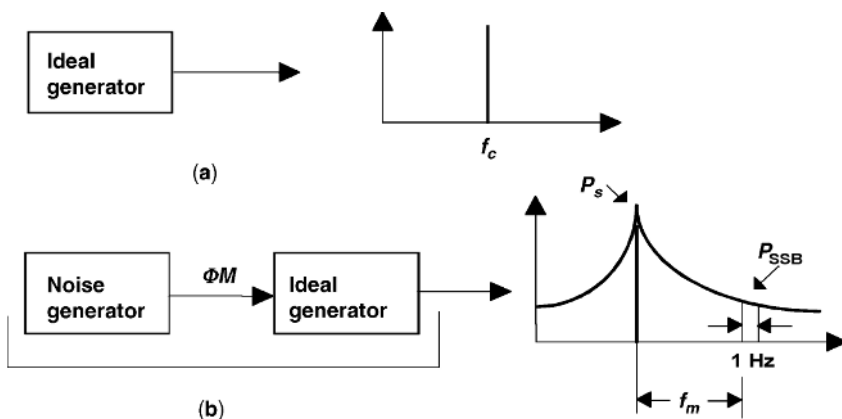


Figure 1.116 SSB phase noise. An ideal signal generator (a) would produce an absolutely pure carrier. A real signal generator (b) acts like an ideal generator driven by a noise generator, producing a noise-modulated carrier.

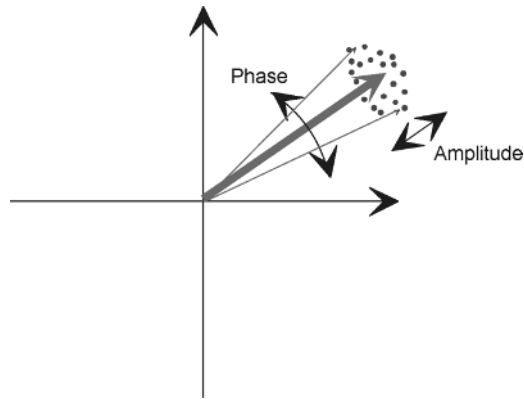
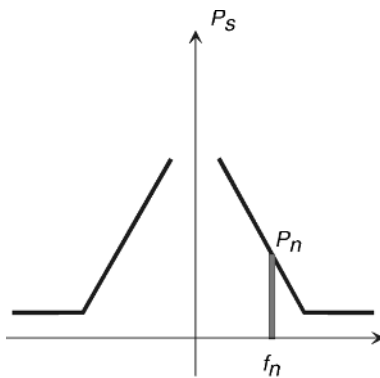


Figure 1.117 Oscillator noise can be split into amplitude and phase components.



$$L(f) = 10 * \log \frac{P_n}{P_s}$$

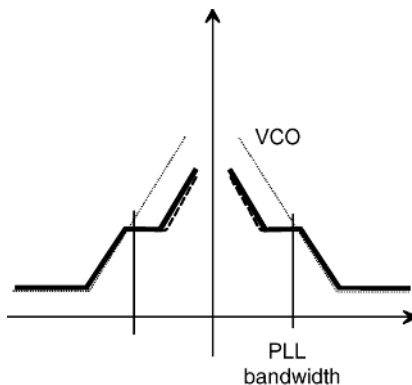
P_n = Sideband noise in 1-Hz bandwidth
at offset frequency f_n

P_s = Total signal power

Single-sideband phase noise is
expressed in dBc measured in
a bandwidth of 1 Hz [dBc(Hz)]
at offset frequency f_n .

Figure 1.118 Phase-noise calculation.

The SMY is a low-cost signal source, while the SMIQ is a very high-performance signal generator capable of being programmed for all digital modulations; therefore, their PLL systems exhibit different phase noise versus frequency responses as the measured results show.



Phase noise is controlled by
reference phase noise within
PLL bandwidth and follows
VCO noise outside PLL bandwidth

Figure 1.119 Phase noise of an oscillator controlled by a phase-locked loop.

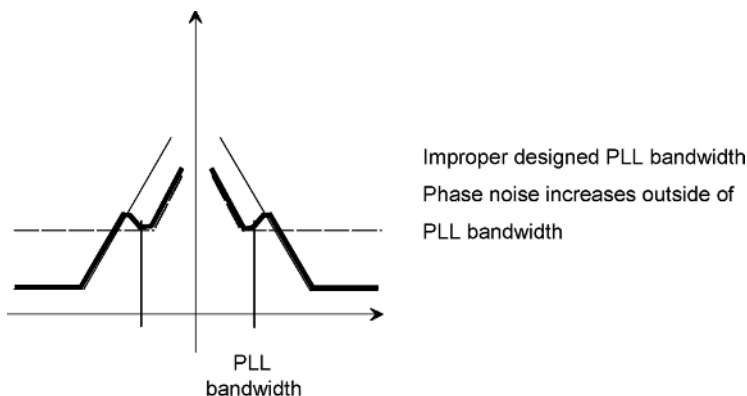


Figure 1.120 Effect of improper loop-filter design.

Reciprocal Mixing In reciprocal mixing, incoming signals mix with LO-sideband energy to produce IF output (Figure 1.123). Because one of the two signals is usually noise, the resulting IF output is usually noise. (Reciprocal mixing effects are not limited to noise; discrete-frequency oscillator sideband components, such as those resulting from crosstalk to or reference energy on a VCO's control line, or the discrete-frequency spurious signals endemic to direct digital synthesis, can also mix incoming signals to IF.) In practice, the resulting noise-floor increase can compromise the receiver's ability to detect weak signals and achieve a high IMD dynamic range; on the test bench, noise from reciprocal mixing may invalidate desensitization, cross-modulation, and IM testing by obscuring the weak signals that must be measured in making these tests.

Figure 1.124 shows a typical arrangement of a dual-conversion receiver with local oscillators. The signal coming from the antenna is filtered by an arrangement of tuned circuits referred to providing as input selectivity. For a minimum attenuation in the passband, an operating bandwidth of

$$B = \frac{f}{\sqrt{2} \cdot Q_L}$$

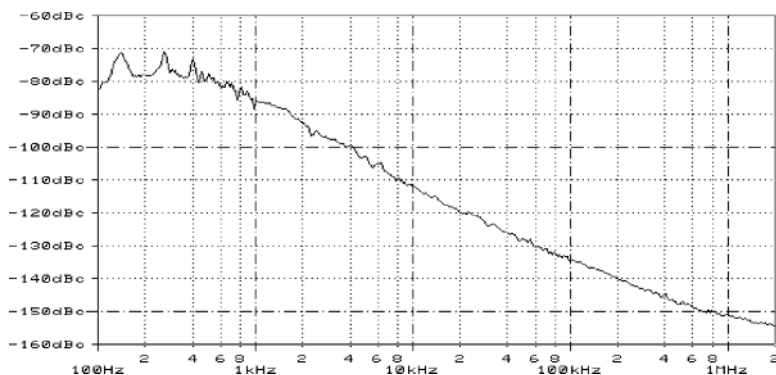


Figure 1.121 Measured phase noise of the Rohde & Schwarz SMY signal generator at 1 GHz. This signal generator has no provision for digital modulation, and therefore shows the best possible phase noise in its class.

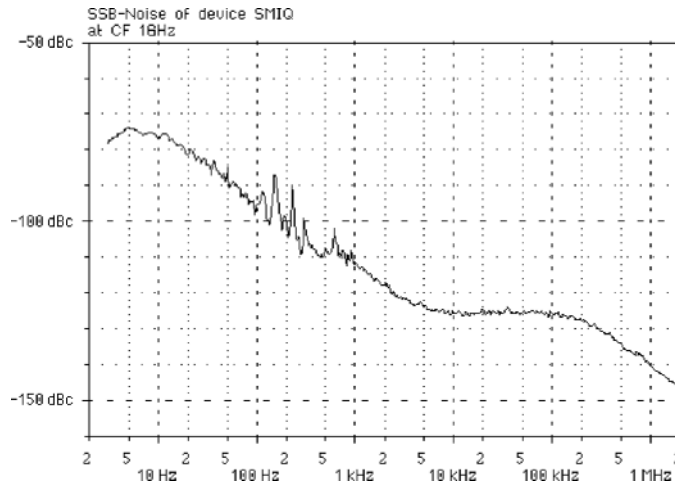


Figure 1.122 Measured phase noise of the Rohde & Schwarz SMIQ signal generator at 1 GHz. This signal generator is optimized for all digital modulation capabilities and can be configured via appropriate programming. Above 10 kHz, the influence of the wideband loop becomes noticeable; above 200 kHz, the resonator Q takes over.

This approximation formula is valid for the insertion loss of about 1 dB due to loaded Q .

The filter in the first IF is typically either a SAW filter (in the frequency range from 500 MHz to 1 GHz) or a crystal filter (45–120 MHz). Typical insertion loss is 6 dB. Since these resonators have significantly higher Q than LC circuits, the bandwidth for the first

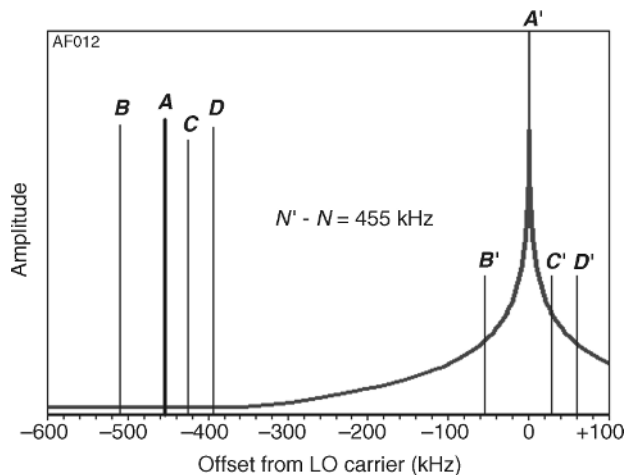


Figure 1.123 Reciprocal mixing occurs when incoming signals mix energy from an oscillator's sidebands to the IF. In this example, the oscillator is tuned so that its carrier, at A' , heterodynes the desired signal, A , to the 455 kHz as intended; at the same time, the undesired signals B , C , and D mix the oscillator noise-sideband energy at B' , C' , and D' , respectively, to the IF. Depending on the levels of the interfering signals and the noise-sideband energy, the result may be a significant rise in the receiver noise floor.

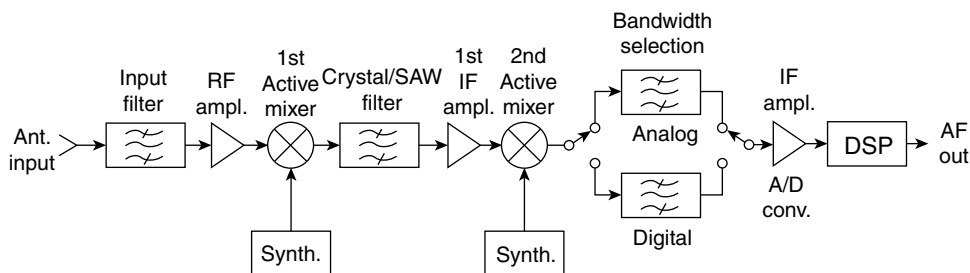


Figure 1.124 Block diagram of an analog/digital receiver showing the signal path from antenna to audio output. No AGC or other auxiliary circuits are shown. This receiver principle can be used for all types of modulation, since the demodulation is done in the DSP block.

IF will vary from ± 5 kHz to ± 500 kHz. It is now obvious that the first RF filter does not protect the first IF because of its wider bandwidth. For typical communication receivers, IF bandwidths from 150 Hz to 1 MHz are found; for digital modulation, the bandwidth varies roughly from 30 kHz to 1 MHz. Therefore, the IF filter in the second IF has to accommodate these bandwidths, otherwise, the second mixer easily gets into trouble from overloading. This also means that both synthesizer paths must be of low-noise and low-spurious design. The second IF of this arrangement (Figure 1.124) can be either analog or digital, or even zero-IF. In practical terms, there are good reasons for using IFs like 50/3 kHz, as in hi-fi audio equipment, with DSP processing at this frequency (50/3 kHz) using the low-cost modules found in mass-market consumer products.

The following two pictures (Figures 1.125 and 1.126) show the principle of selectivity measurement both for analog and digital signals. The main difference is that the occupied bandwidth for the digital system can be significantly wider, and yet both signals can be interfered with by either a noise synthesizer/first LO or a synthesizer that has unwanted spurious frequencies. Such a spurious signal is shown in Figure 1.125. In the case of Figure 1.125, the

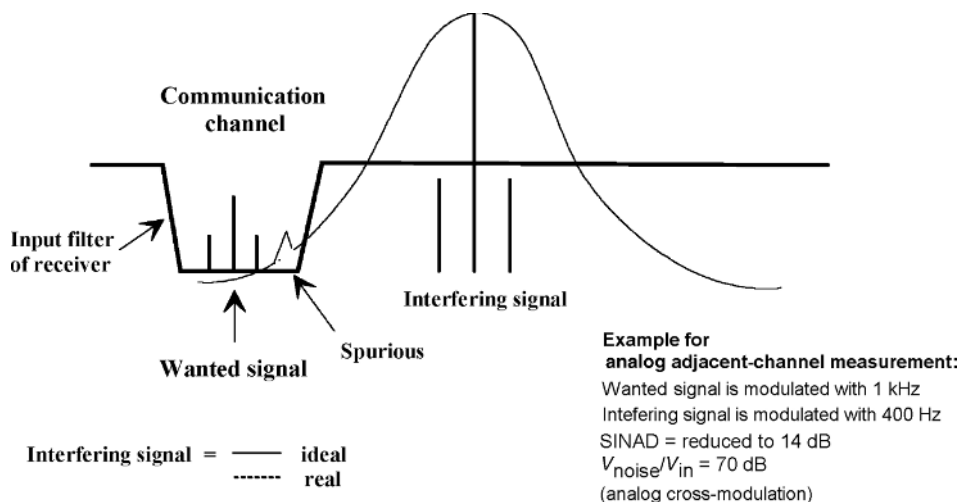


Figure 1.125 Principle of selectivity measurement for analog receivers.

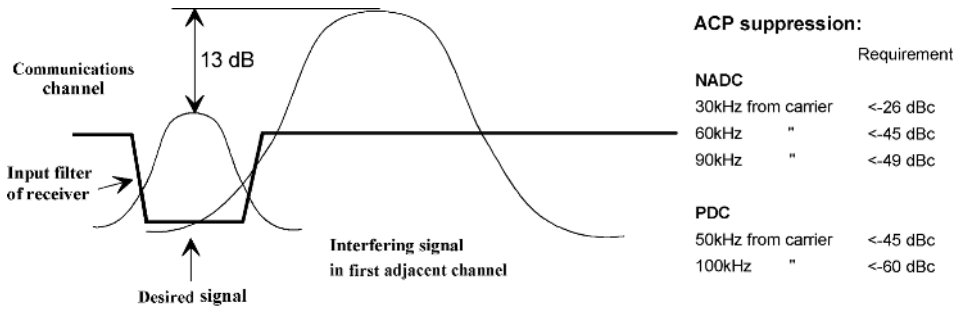


Figure 1.126 Principle of selectivity measurement for digital receivers.

analog adjacent-channel measurement has some of the characteristics of cross-modulation and intermodulation, while in the digital system, the problem with the adjacent-channel power suppression in modern terms is more obvious. Rarely has the concept of adjacent-channel power (ACP) been used with analog systems. Also, to meet the standards, signal generators have to be used that are better, with some headroom, than the required dynamic measurement. We have therefore included in Figure 1.126 the achievable performance for a practical signal generator—in this case, the Rohde & Schwarz SMHU58.

Because reciprocal mixing produces the effect of noise leakage around IF filtering, it plays a role in determining a receiver's dynamic selectivity (Figure 1.127). There is little value in using IF filtering that exhibits more stopband rejection than a value 3–10 dB greater than that results in an acceptable reciprocal mixing level.

Although additional RF selectivity can reduce the number of signals that contribute to the noise, improving the LO's spectral purity is the only effective way to reduce reciprocal mixing noise from all signals present at a mixer's RF port.

Factoring in the effect of discrete spurious signals with that of oscillator phase noise can give us the useful dynamic range of which an instrument or receiver is capable (Figure 1.128).

In evaluating the performance of digital wireless systems, we are interested in determining a receiver's resistance to adjacent-channel signals.

Phase Errors In PM systems, especially those employing digital modulation, oscillator phase noise contributes directly to modulation and demodulation errors by introducing random phase variations that cannot be corrected by phase-locking techniques (Figure 1.129). The greater the number of phase states a modulation scheme entails, the greater its sensitivity to phase noise.

When an output signal is produced by mixing two signals, the resulting phase noise depends on whether the input signals are correlated—all referred to the same system clock—or uncorrelated, as shown in Figure 1.130.

Error Vector Magnitude Several earlier figures (Figures 1.106, 1.113, and 1.129) have shown how particular sources of amplitude and/or phase error can shift the values of a digital emission's data states toward decision boundaries, resulting in increased BER due to intersymbol interference. Figures 1.131–1.133 show three additional sources of such errors.

A figure of merit known as error vector magnitude (EVM) has been developed as sensitive indicator of the presence and severity of such errors. An emission's error vector magnitude

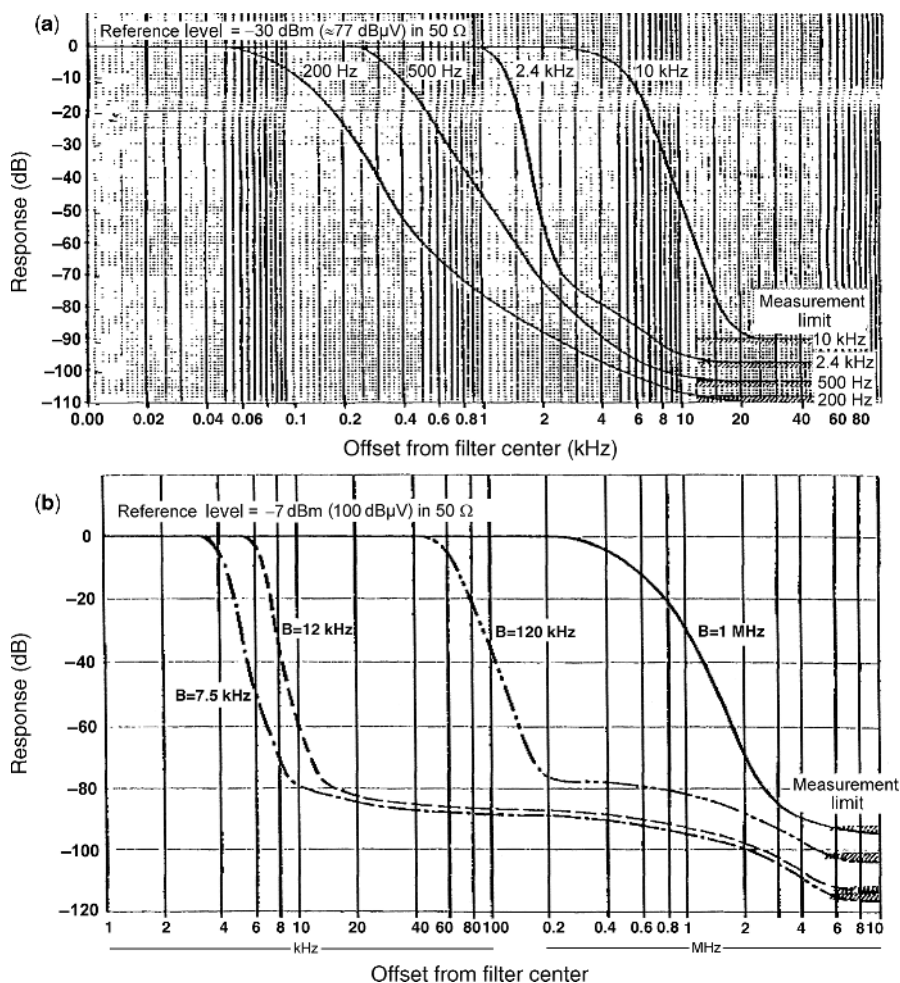


Figure 1.127 Dynamic selectivity versus IF bandwidth for (a) the Rohde & Schwarz ESH-2 test receiver (9 kHz to 30 MHz) and (b) the Rohde and Schwarz ESV test receiver (10 MHz to 1 GHz). Reciprocal mixing widens the ESH-2's 2.4 kHz response below -70 dB (-100 dBm) at (a) and the ESV's 7.5, 12, and 120 kHz responses below approximately -80 dB (-87 dBm) at (b).

is the magnitude of the phasor difference as a function of time between an ideal reference signal and the measured transmitted signal after its timing, amplitude, frequency, phase, and dc offset have been modified by circuitry and/or propagation. Figure 1.134 illustrates the EVM concept.

1.8 TESTING

1.8.1 Introduction

Testing of digital circuits deviates from the typical analog measurements, and yet the analog measurements are still necessary and related. As an example, the second-generation cellular

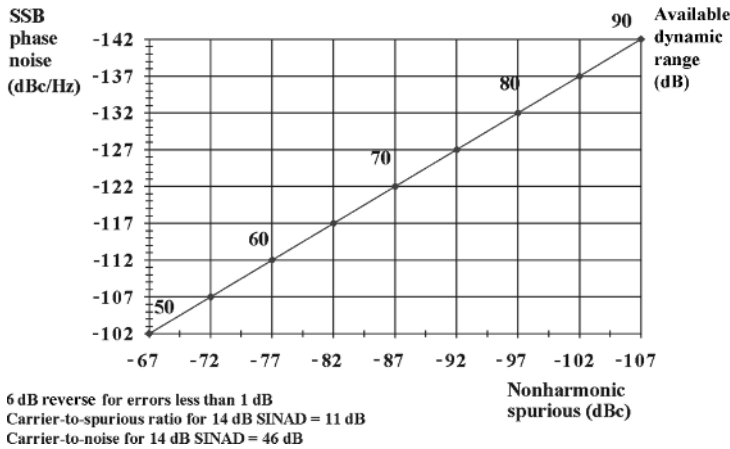


Figure 1.128 This graph shows the available dynamic range, which is determined either by the masking of the unwanted signal by phase noise or by discrete spurs. As far as the culprit synthesizer is concerned, it can be either the local oscillator or the effect of a strong adjacent-channel signal that takes over the function of the local oscillator.

telephones standards really have addressed only the transfer of voice and were just beginning to look into the transfer of data, mostly in the form of SMS, text messages limited to 140 characters. Current 3G standards with adaptive bandwidth do address high-speed data and video. Since the major use of 2G phones is voice, information such as signal-to-noise ratio as a function of many things is important. In particular, because of the Doppler effect and the use of digital rather than analog signals, where the phase information is significant, the designer ends up using coding schemes for error-correction—specifically, forward error correction (FEC). The signal-to-noise ratio, as we know it from analog circuits, now determines the bit

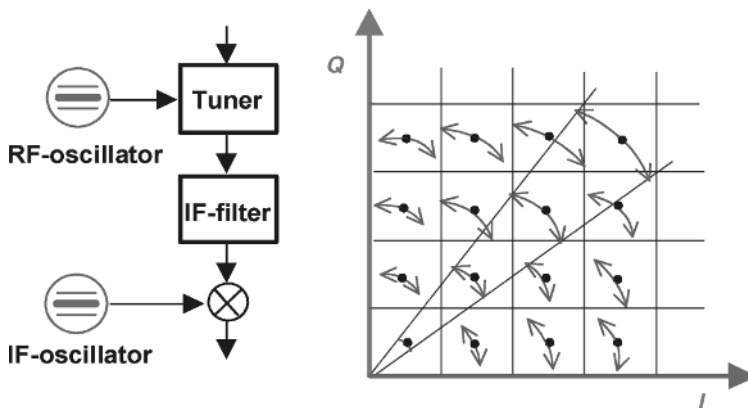
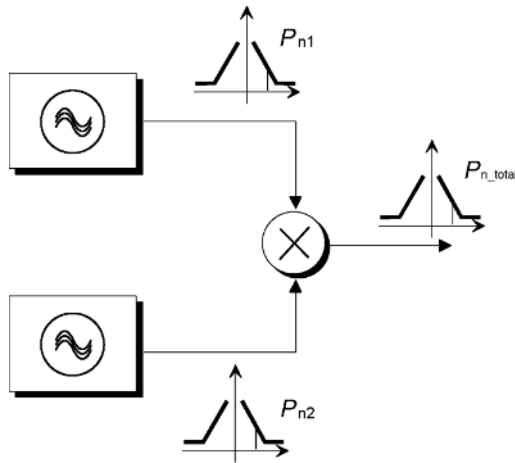


Figure 1.129 Phase noise is critical to digitally modulated communication systems because of the modulation errors it can introduce. Intersymbol interference (ISI), accompanied by a rise in BER, results when state values become so badly error blurred that they fall into the regions of adjacent states. This drawing depicts only the results of phase errors introduced by phase noise; in actual systems, thermal noise, AM-to-PM conversion, differential group delay, propagation, and other factors may also contribute to the spreading of state amplitude and phase values.



Uncorrelated signals:
Noise sideband power is added

$$P_{n_total} = P_{n1} + P_{n2}$$

$$\Delta\mathcal{L}(f) = 10 * \log \frac{P_{n1} + P_{n2}}{P_s}$$

Correlated signals:
Noise sideband voltage is added
or subtracted

$$V_{n_total} = V_{n1} \pm V_{n2}$$

$$\Delta\mathcal{L}(f) = 20 * \log \frac{V_{n1} \pm V_{n2}}{V_s}$$

Figure 1.130 The noise sideband power of a signal that results from mixing two signals depends on whether the signals are correlated—referenced to the same system clock—or uncorrelated.

error rate (BER), and its tolerable values depend on the type of modulation used. Because of correlation, it is possible to “rescue” a voice with a bit error rate of 10^{-4} , but for higher data rates with little, if any, correlation, we are looking for BERs down to 10^{-9} . The actual bit error rate depends on the type of filtering, coding, modulation, and demodulation. In several earlier figures (Figures 1.52 and 1.56), we related BER to signal-to-noise ratio; this is a key issue in receiver design.

Another problem in receivers that has to do with the transmitter of a second station is adjacent-channel power ratio (ACPR). Given the fact that a transmitter handling digital modulation delivers its power in pulses, its transmission may affect adjacent channels by producing transient spurious signals similar to what we call splatter in analog SSB systems. This is a function of the linearity of the transmitter system all the way out to the antenna, and forces most designers to resort to less-efficient Class A operation. As possible alternatives,

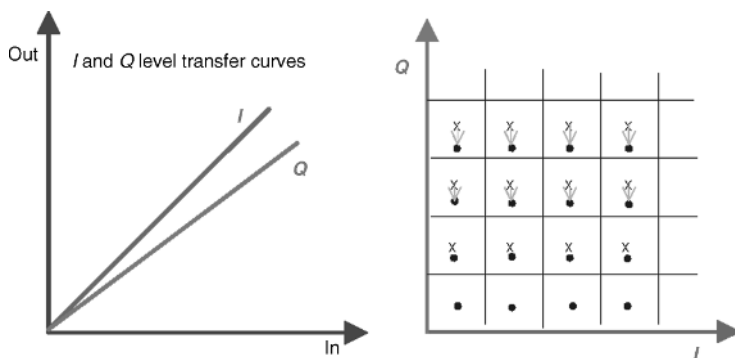


Figure 1.131 Effect of gain imbalance between *I* and *Q* channels on a data signal's phase constellation.

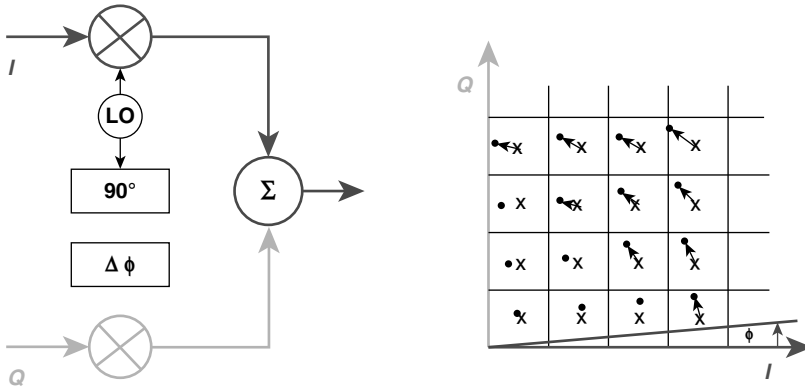


Figure 1.132 Effect of quadrature offset on a data signal's phase constellation.

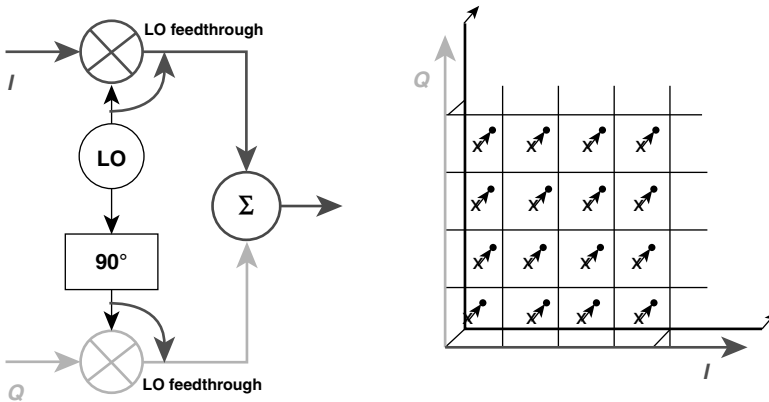


Figure 1.133 Effect of LO-feedthrough-based I/Q offset on a data signal's phase constellation.

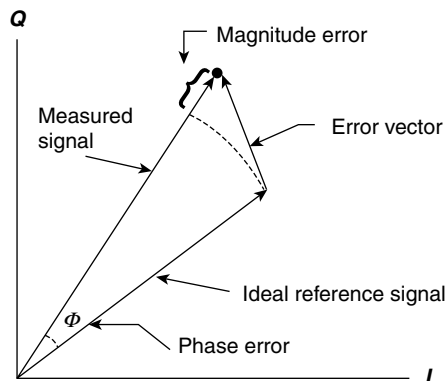


Figure 1.134 Representing errors in a digitally modulated signal's in-phase (I) and quadrature (Q) values relative to the ideal as a single error vector allows us to derive the resulting error vector magnitude (EVM), a sensitive indicator of transmission quality [39].

some researchers are looking into Classes D and E modulation, more about which is found in the power amplifier section of Chapter 3.

It is actually not uncommon to do many linear measurements, and then by using correlation equations translate these measured results into their digital equivalents.

Therefore, the robustness of the signal as a function of antenna signal at the receiver site, constant or known phase relationships, and high adjacent power ratios will provide good transfer characteristics.

1.8.2 Transmission and Reception Quality

The acoustic transmission and reproduction quality of a mobile phone is a highly important characteristic in everyday use. Accurate and reproducible measurements of a cell phone's frequency response cannot be achieved with static sinewave tones (SINAD measurements) because of coder and decoder algorithms. In this case, test signals simulating the characteristic of the human voice—that is, tones that are harmonic multiples of the fundamental, are required. A so-called vocoder is used to produce the lowest possible data rate. Instead of the actual voice, only the filter and fundamental parameters required for signal reconstruction are transmitted. Particularly, in the medium and higher audio frequency ranges, the static sinusoidal tones become a more or less stochastic output signal. For example, if a tone of approximately 2.5 kHz is applied to the telephone at a constant sound pressure, the amplitude of the signal obtained at the vocoder output varies by approximately 20 dB. In type-approval tests, where highly accurate measurements are required, the coder and decoder are excluded from the measurement.

Whether the results obtained for the fundamental are favorable depends on how far the values coincide with the clock of the coding algorithm. Through a skillful choice of fundamental frequencies, test signals with overlapping spectral distribution can be generated, giving a sufficient number of test points in subsequent measurements at different fundamental frequencies so that a practically continuous frequency response curve is obtained. Evaluation can be done by means of FFT analysis with a special window function and selection of result bins. The results are sorted and smoothed by the software and display in the form of a frequency response curve. Depending on the measurement, a program calculates the sending or receiving loudness rating in line with CCITT P.79 and shows the result.

Acoustic measurements relevant to GSM 11.10 include the following.

- Sending frequency response
- Sending loudness rating
- Receiving frequency response
- Receiving loudness rating
- Sidetone masking rating
- Listener sidetone rating
- Echo loss
- Stability margin
- Distortion sending
- Distortion receiving
- Idle channel noise receiving
- Idle channel noise sending.



Figure 1.135 The Rohde & Schwarz CMW 500 Wideband Radio Communication Tester can be used to measure all the relevant specifications of cellular telephones.

There are two categories of testing. One is a full-compliance test for all channels and all combinations of capabilities, and the other is a production tester with evaluation of the typical characteristic data. Both Agilent and Rohde & Schwarz are the leading companies in this area, and as the technology develops further, different equipment will be made available. Figure 1.134 shows a Rohde & Schwarz radiocommunication tester capable of evaluating cellular systems. Table 1.12 shows the typical specifications required to make the appropriate cellular telephone measurements.

The test setup must be capable of measuring key IS-95 parameters, using the following.

- Frame errors
- Waveform quality
- Error vector magnitude
- Phase error
- Magnitude error
- Carrier feedthrough
- Frequency accuracy
- Power measurements
- Base station signaling for mobile testing.

1.8.3 Base Station Simulation

Simulating a CDMA (IS-95) base station involves the following.

- Synchronization of mobile
- Registration of mobile
- Incoming/outgoing call origination
- Handoff.

Table 1.12 Technical Specification of the Rohde & Schwarz CMW 500

General Technical Specifications		
<i>RF Generator</i>		
Frequency range		70 MHz to 3300 MHz Up to 6000 MHz with the R&S [®] CMW-KB036 option
Frequency resolution		0.1 Hz
Frequency uncertainty		Sema as timebase + frequency resolution
Output level range		
RF1 COM, RF2 COM	70 MHz to 100 MHz	
	Continuous wave (CW)	–130 dBm to –15 dBm
	Peak envelope power (PEP)	Up to –15 dBm
	Overranging (PEP)	Up to –10 dBm
	100 MHz to 3300 MHz	
	Continuous wave (CW)	–130 dBm to –5 dBm
	Peak envelope power (PEP)	Up to –5 dBm
	Overranging (PEP)	Up to 0 dBm
	3300 MHz to 6000 MHz	
	Continuous wave (CW)	–120 dBm to –15 dBm
	Peak envelope power (PEP)	Up to –15 dBm
	Overranging (PEP)	Up to –10 dBm
	Maximum input DC level	0 V DC
RF1 OUT	70 MHz to 100 MHz	
	Continuous wave (CW)	–120 dBm to –2 dBm
	Peak envelope power (PEP)	Up to –2 dBm
	Overranging (PEP)	Up to +3 dBm
	100 MHz to 3300 MHz	
	Continuous wave (CW)	–120 dBm to +8 dBm
	Peak envelope power (PEP)	Up to +8 dBm
	Overranging (PEP)	Up to +13 dBm
	3300 MHz to 6000 MHz	
	Continuous wave (CW)	–110 dBm to –2 dBm
	Peak envelope power (PEP)	Up to –2 dBm
	Overranging (PEP)	Up to +3 dBm
	Maximum input DC level	0 V DC
Output level uncertainty	In temperature range +20 °C to +35 °C, no overranging	
RF1 COM, RF2 COM	Output level > –120 dBm	
	70 MHz to 100 MHz	< 1.2 dB ^a
	100 MHz to 3300 MHz	< 0.6 dB ^a
	3300 MHz to 6000 MHz	< 1.2 dB ^a
RF1 OUT	Output level > –110 dBm	
	70 MHz to 100 MHz	< 1.6 dB ^a
	100 MHz to 3300 MHz	< 0.8 dB ^a
	3300 MHz to 6000 MHz	< 1.6 dB ^a
Output level uncertainty	In temperature range +5 °C to +45 °C, no overranging	
RF1 COM, RF2 COM	Output level > –120 dBm	
	70 MHz to 100 MHz	< 2.0 dB ^a
	100 MHz to 3300 MHz	< 1.0 dB ^a
	3300 MHz to 6000 MHz	< 2.0 dB ^a
RF1 OUT	Output level > –110 dBm	
	70 MHz to 100 MHz	< 2.0 dB ^a
	100 MHz to 3300 MHz	< 1.0 dB ^a
	3300 MHz to 6000 MHz	< 2.0 dB ^a

Table 1.12 (Continued)

Output level linearity with fixed RF output attenuator setting	In temperature range +20 °C to +35 °C, GPRF generator list mode, Level range 0 dB to –30 dB	
RF1 COM, RF2 COM	No overranging	< 0.2 dB, typ. < 0.1 dB
Output level resolution		0.01 dB
Output level repeatability	Typical values after 1 h warm-up time, always returning to same level and frequency, no temperature change, insignificant time change	
	Output level ≥ -80 dBm	< 0.01 dB
	Output level < –80 dBm	< 0.05 dB
VSWR		
RF1 COM, RF2 COM	70 MHz to 3300 MHz	< 1.2
	3300 MHz to 5000 MHz	< 1.5
	5000 MHz to 6000 MHz	< 1.6
RF1 OUT	70 MHz to 3300 MHz	< 1.5
	3300 MHz to 5000 MHz	< 1.5
	5000 MHz to 6000 MHz	< 1.6
Attenuation of 2nd harmonic		
RF1 COM, RF2 COM	70 MHz to 6000 MHz, $P < -10$ dBm	> 30 dB
RF1 OUT	70 MHz to 6000 MHz, $P < 0$ dBm	> 30 dB
Attenuation of 3rd harmonic		
RF1 COM, RF2 COM	70 MHz to 6000 MHz, $P < -10$ dBm	> 40 dB
RF1 OUT	70 MHz to 6000 MHz, $P < 0$ dBm	> 40 dB
Attenuation of nonharmonics	> 5 kHz offset from carrier, For output level > –40 dBm, For full scale CW signal	
	400 MHz to 3300 MHz, Except $f_{\text{nonharmonic}} = 3900 \text{ MHz} - f_{\text{carrier}}$, Except $f_{\text{nonharmonic}} = 3900 \text{ MHz}$ Except $f_{\text{carrier}} = (899 \text{ to } 901) \text{ MHz} + n \times 800 \text{ MHz}$ with $n = 1, 2, 3$	> 60 dB
	3300 MHz to 3600 MHz	> 25 dB
	3600 MHz to 6000 MHz, Except $f_{\text{nonharmonic}} = 2 \times f_{\text{carrier}} - 6400 \text{ MHz}$	> 40 dB
Phase noise	Single sideband, 70 MHz to 3300 MHz	
Carrier offset	$\geq 1 \text{ MHz}$	< –120 dBc, 1 Hz
Phase noise	Single sideband, 3300 MHz to 6000 MHz	
Carrier offset	$\geq 1 \text{ MHz}$	< –117 dBc, 1 Hz
Signal-to-noise ratio	70 MHz to 3300 MHz	
RF1 COM, RF2 COM	5 MHz offset from carrier, For output level > –30 dBm	> 95 dB, typ. > 101 dB, 1 kHz (> 125 dB, typ. > 131 dB, 1 Hz)
Signal-to-noise ratio	3300 MHz to 6000 MHz	
RF1 COM, RF2 COM	5 MHz offset from carrier, For output level > –30 dBm	> 92 dB, 1 kHz

(continued)

Table 1.12 (Continued)

Modulation source: arbitrary waveform generator (ARB) (R&S®CMW-B110A option)		
Memory size		1.024 Gbyte
Word length	I	16 bit
	Q	16 bit
	Marker	4 bit to 16 bit
Sample length	With 4-bit marker	Up to 227.55 Msample
Sample rate	Minimum	400 Hz
	Maximum	100 MHz
Maximum possible RF bandwidth	Depending on arbitrary waveform file	80 MHz
Trigger		
Trigger sources		BASE: external TRIG A, BASE: external TRIG B
RF analyzer		
VSWR		
RF1 COM, RF2 COM	70 MHz to 3300 MHz	< 1.2
	3300 MHz to 5000 MHz	< 1.5
	5000 MHz to 6000 MHz	< 1.6
Inherent spurious response	Without input signal, 70 MHz to 6000 MHz, Except 4000 MHz, 4800 MHz, 5162.5 MHz, 5600 MHz, 6000 MHz	
	Expected nominal power setting ≤ -10 dBm	< -100 dBm
	Expected nominal power setting > -10 dBm	< -90 dB below expected nominal power setting
Spurious response	For full scale single tone input signal	
	70 MHz to 3300 MHz	< -55 dB
	Except $f_{in} = 1962.5$ MHz and 3925 MHz	
	Except $f_{in} = 1962.5$ MHz + $f_{selected}$	
	3300 MHz to 3700 MHz, Except $f_{in} = 6400$ MHz - $f_{selected}$, Except $f_{in} = 6400$ MHz - $0.5 \times f_{selected}$	< -40 dB
	3700 MHz to 6000 MHz, Except $f_{in} = 6400$ MHz - $0.5 \times f_{selected}$	< -40 dB
Harmonic response	2nd harmonic	
RF1 COM, RF2 COM	$f_{in} = 70$ MHz to 1650 MHz, $f_{selected} = 140$ MHz to 3300 MHz	< -30 dB
	$f_{in} = 1650$ MHz to 3000 MHz, $f_{selected} = 3300$ MHz to 6000 MHz	< -30 dB
Harmonic response	3rd harmonic	
RF1 COM, RF2 COM	$f_{in} = 70$ MHz to 900 MHz, $f_{selected} = 210$ MHz to 2700 MHz	< -50 dB
	$f_{in} = 900$ MHz to 1100 MHz, $f_{selected} = 2700$ MHz to 3300 MHz	< -45 dB
	$f_{in} = 1100$ MHz to 2000 MHz, $f_{selected} = 3300$ MHz to 6000 MHz	< -50 dB
Phase noise	Single sideband, 70 MHz to 3300 MHz	
Carrier offset	≥ 1 MHz	< -120 dBc, 1 Hz
Phase noise	Single sideband, 3300 MHz to 6000 MHz	
Carrier offset	≥ 1 MHz	< -117 dBc, 1 Hz

Table 1.12 (Continued)

Trigger		
Trigger sources		BASE: external TRIG A, BASE: external TRIG B, GPRF: free run, GPRF: IF power, BB generators, BB signaling

Power meter

Frequency range		70 MHz to 3300 MHz Up to 6000 MHz with the R&S®CMW-KB036 option
Frequency resolution		0.1 Hz
Resolution bandwidths		Gaussian, 1 kHz to 10 MHz, in 1/3/5 steps, Bandpass, 1 kHz to 30 MHz, in 1/3/5 Steps, RRC, $\alpha = 0.1$, 3.84 MHz, RRC, $\alpha = 0.22$, WCDMA filter, 1.2288 MHz, CDMA filter
Expected nominal power setting range	For ADC full scale	
RF1 COM, RF2 COM	70 MHz to 100 MHz	–37 dBm to +42 dBm ^b
	100 MHz to 3300 MHz	–47 dBm to +42 dBm ^b
	3300 MHz to 6000 MHz	–37 dBm to +42 dBm ^b

Level range		
RF1 COM, RF2 COM	70 MHz to 100 MHz	
	Continuous power (CW)	–74 dBm ^c to +34 dBm
	Peak envelope power (PEP)	up to +42 dBm ^b
	100 MHz to 3300 MHz	
	Continuous power (CW)	–84 dBm ^c to +34 dBm
	Peak envelope power (PEP)	up to +42 dBm ^b
	3300 MHz to 6000 MHz	
	Continuous power (CW)	–74 dBm ^c to +34 dBm
	Peak envelope power (PEP)	up to +42 dBm ^b
	Maximum input DC level	0 V DC

Level uncertainty	In temperature range +20 °C to +35 °C	
RF1 COM, RF2 COM	70 MHz to 100 MHz	< 1.0 dB ^d
	100 MHz to 3300 MHz	< 0.5 dB ^d
	3300 MHz to 6000 MHz	< 1.0 dB ^d

Level uncertainty	In temperature range +5 °C to +45 °C	
RF1 COM, RF2 COM	70 MHz to 100 MHz	< 1.2 dB ^d
	100 MHz to 3300 MHz	< 0.7 dB ^d
	3300 MHz to 6000 MHz	< 1.2 dB ^d

Level linearity with fixed expected nominal power setting	In temperature range +20 °C to +35 °C	
RF1 COM, RF2 COM	Level range 0 dB to –40 dB	< 0.15 dB, typ. < 0.1 dB

Level resolution		0.01 dB
------------------	--	---------

Level repeatability	Typical values after 1 h warm-up time, always returning to same level and frequency, no temperature change, insignificant time change	
	Input level $\geq$ –40 dBm	< 0.01 dB
	Input level < –40 dBm	< 0.03 dB

(continued)

Table 1.12 (Continued)

Dynamic range	70 MHz to 3300 MHz, <i>RBW</i> → 1 kHz, With fixed expected nominal power setting	> 100 dB
Expected nominal power setting for full dynamic range		
RF1 COM, RF2 COM		−8 dBm to +42 dBm ^o
Dynamic range	3300 MHz to 6000 MHz, <i>RBW</i> → 1 kHz, With fixed expected nominal power setting	> 97 dB
Expected nominal power setting for full dynamic range		
RF1 COM, RF2 COM		+2 dBm to +42 dBm ^o

Spectrum measurements

FFT spectrum analyzer (R&S [®] CMW-KM010 option)		
Frequency range		70 MHz to 3300 MHz Up to 6000 MHz with the R&S [®] CMW-KB036 option
Frequency span		1.25 MHz, 2.5 MHz, 5 MHz, 10 MHz, 20 MHz, 40 MHz
FFT length		1k, 2k, 4k, 8k, 16k
Detector		peak, RMS
Level range		See general technical specifications
Level uncertainty	For center frequency and <i>detector</i> → <i>peak</i>	See general technical specifications

Dynamic range	70 MHz to 3300 MHz, For <i>FFT length</i> → 16k and <i>span</i> → 5 MHz (equivalent to <i>RBW</i> → 781 Hz)	> 100 dB
Expected nominal power setting for full dynamic range		
RF1 COM, RF2 COM		−8 dBm to +42 dBm ^o

Dynamic range	3300 MHz to 6000 MHz, For <i>FFT length</i> → 16k and <i>span</i> → 5 MHz (equivalent to <i>RBW</i> → 781 Hz)	> 97 dB
Expected nominal power setting for full dynamic range		
RF1 COM, RF2 COM		+2 dBm to +42 dBm Fehler! Textmarke nicht definiert.

Inherent spurious response	Without input signal	See general technical specifications
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Table 1.12 (Continued)**Possible configurations with two RF paths^f**

Necessary hardware (H570, H590X):

Selections: R&S®CMW-S590A RF frontend (BASIC) or R&S®CMW-S590D RF frontend (ADV.) and R&S®CMW-S570 RF TRX.

Options: R&S®CMW-B590A RF frontend (BASIC) or R&S®CMW-B590D RF frontend (ADV.) and R&S®CMW-B570 RF TRX.

Configuration with two H570 (RF TRX) and two H590A (RF frontend (BASIC))

The R&S®CMW-B570 and R&S®CMW-B590A options make the second RF path (RF path 2) available on the front of the instrument with three additional RF connectors, i.e. RF3 COM, RF4 COM and RF3 OUT.

RF3 COM	Equivalent to RF1 COM	See general technical specifications
RF4 COM	Equivalent to RF2 COM	See general technical specifications
RF3 OUT	Equivalent to RF1 OUT	See general technical specifications

Configuration with two H570 (RF TRX) and one H590D (RF frontend (ADV.))

The R&S®CMW-B570 option and R&S®CMW-S590D selection make the second RF path (RF path 2) available on the front of the instrument at connectors RF1 COM, RF2 COM and RF1 OUT.

RF path 1 and RF path 2 routed to separate connectors

RF generator 1 and RF generator 2	Switchable to RF1 COM, RF2 COM, RF1 OUT	See general technical specifications
RF analyzer 1 and RF analyzer 2	Switchable to RF1 COM, RF2 COM	See general technical specifications
Expected nominal power setting for full dynamic range	70 MHz to 3300 MHz	-5 dBm to +42 dBm ^g
	3300 MHz to 6000 MHz	+5 dBm to +42 dBm ^g

RF path 1 and RF path 2 routed to common connector

RF generator 1 and RF generator 2	Switchable to RF1 COM, RF2 COM, RF1 OUT	See general technical specifications
Output level range	Peak envelope power (PEP)	The specified value is valid for the total power of the two RF generators, see general technical specifications
Output level uncertainty	For each carrier	see general technical specifications + 0.2 dB
Signal-to-noise ratio	For the carrier with the highest output level (at least 3 dB higher than the other carrier)	see general technical specifications
RF analyzer 1 and RF analyzer 2	Switchable to RF1 COM, RF2 COM	See general technical specifications
Level uncertainty	70 MHz to 3300 MHz	See general technical specifications + 0.2 dB
	3300 MHz to 6000 MHz	See general technical specifications + 0.3 dB
Expected nominal power setting for full dynamic range	70 MHz to 3300 MHz	-5 dBm to +42 dBm ^g
	3300 MHz to 6000 MHz	+5 dBm to +42 dBm ^g

(continued)

Table 1.12 (Continued)**Possible configurations with four RF paths^h**

Necessary hardware (H570, H571B, H590D):

Selections: R&S®CMW-S590D RF frontend (ADV.) and R&S®CMW-S570 RF TRX.

Options: R&S®CMW-B590D RF frontend (ADV.) and R&S®CMW-B570 RF TRX and two R&S®CMW-B571B RF TX.

Configuration with two H570 (RF TRX), two H571 B (RF TX) and two H590D (RF frontend (ADV.))

The R&S®CMW-B570 option, two R&S®CMW-B571 options and R&S®CMW-B590D option make the four RF paths (RF path 1 RX and TX, RF path 2 RX and TX, RF path 3 TX only, RF path 4 TX only) available on the front of the instrument at connectors RF1 COM, RF2 COM, RF1 OUT and RF3 COM, RF4 COM, RF3 OUT.

RF path 1, 2, 3 and 4 routed to separate connectors

RF generator 1 and RF generator 3	Switchable to RF1 COM, RF2 COM, RF1 OUT	See general technical specifications
RF generator 2 and RF generator 4	Switchable to RF3 COM, RF4 COM, RF3 OUT	See general technical specifications
RF analyzer 1	Switchable to RF1 COM, RF2 COM	See general technical specifications
Expected nominal power setting for full dynamic range	70 MHz to 3300 MHz	–5 dBm to +42 dBm ⁱ
	3300 MHz to 6000 MHz	+5 dBm to +42 dBm ⁱ
RF analyzer 2	Switchable to RF3 COM, RF4 COM	see general technical specifications
Expected nominal power setting for full dynamic range	70 MHz to 3300 MHz	–5 dBm to +42 dBm ⁱ
	3300 MHz to 6000 MHz	+5 dBm to +42 dBm ⁱ

RF path 1, RF path 3 routed to common connector and RF path 2, RF path 4 routed to common connector

RF generator 1 and RF generator 3	Switchable to RF1 COM, RF2 COM, RF1 OUT	See general technical specifications
Output level range	Peak envelope power (PEP)	The specified value is valid for the total power of the two RF generators, see general technical specifications
Output level uncertainty	For each carrier	See general technical specifications + 0.2 dB
Signal-to-noise ratio	For the carrier with the highest output level (at least 3 dB higher than the other carrier)	See general technical specifications
RF generator 2 and RF generator 4	Switchable to RF3 COM, RF4 COM, RF3 OUT	See general technical specifications
Output level range	Peak envelope power (PEP)	The specified value is valid for the total power of the two RF generators, see general technical specifications
Output level uncertainty	For each carrier	See general technical specifications + 0.2 dB
Signal-to-noise ratio	For the carrier with the highest output level (at least 3 dB higher than the other carrier)	See general technical specifications
RF analyzer 1	Switchable to RF1 COM, RF2 COM	See general technical specifications
Level uncertainty	70 MHz to 3300 MHz	See general technical specifications + 0.2 dB
	3300 MHz to 6000 MHz	See general technical specifications + 0.3 dB
Expected nominal power setting for full dynamic range	70 MHz to 3300 MHz	–5 dBm to +42 dBm ⁱ
	3300 MHz to 6000 MHz	+5 dBm to +42 dBm ⁱ
RF analyzer 2	Switchable to RF3 COM, RF4 COM	See general technical specifications
Level uncertainty	70 MHz to 3300 MHz	See general technical specifications + 0.2 dB
	3300 MHz to 6000 MHz	See general technical specifications + 0.3 dB
Expected nominal power setting for full dynamic range	70 MHz to 3300 MHz	–5 dBm to +42 dBm ⁱ
	3300 MHz to 6000 MHz	+5 dBm to +42 dBm ⁱ

Table 1.12 (Continued)

<i>Timebase</i>		
Timebase TCXO		
Max. frequency drift	In temperature range +5 °C to +45 °C	$\pm 1 \times 10^{-6}$
Max. aging	At +25 °C, after 14 days of continuous operation	$\pm 1 \times 10^{-6}$ /year
Timebase basic OCXO (R&S®CMW-B690A option)		
Max. frequency drift	In temperature range +5 °C to +45 °C	$\pm 5 \times 10^{-8}$
Retrace	At +25 °C, after 24 hours power ON / 2 hours power OFF / 1 hour power ON	$\pm 2 \times 10^{-8}$
Max. aging	At +25 °C, after 10 days of continuous operation	$\pm 1 \times 10^{-7}$ /year $\pm 1 \times 10^{-9}$ /day
Warm-up time	At +25 °C, the frequency is in the range that is 10 times the frequency drift ($\pm 5 \times 10^{-7}$)	approx. 10 min
Timebase highly stable OCXO (R&S®CMW-B690B option)		
Max. frequency drift	In temperature range +5 °C to +45 °C, referenced to +25 °C	$\pm 5 \times 10^{-9}$
	With instrument orientation	$\pm 1 \times 10^{-9}$
Retrace	At +25 °C, after 24 hours power ON / 2 hours power OFF / 1 hour power ON	$\pm 5 \times 10^{-9}$
Max. aging	At +25 °C, after 10 days of continuous operation	$\pm 3 \times 10^{-8}$ /year $\pm 5 \times 10^{-10}$ /day
Warm-up time	At +25 °C, the frequency is in the range that is 10 times the frequency drift ($\pm 5 \times 10^{-8}$)	approx. 10 min
Reference frequency inputs/outputs		
Synchronization input		BNC connector REF IN, rear panel
Frequency	Sine wave	10 MHz to 80 MHz, step: 1 Hz
	square wave (TTL level)	1 MHz to 80 MHz, step: 1 Hz
Max. frequency variation		$\pm 10 \times 10^{-6}$
Input voltage range		0.5 V to 2 V, RMS
Impedance		50 Ω
Synchronization output 1		BNC connector REF OUT 1, rear panel
Frequency		10 MHz from internal reference or frequency at synchronization input
Output voltage		> 1.4 V, peak-to-peak
Impedance		50 Ω

^a Valid for a 12-month calibration interval.^b The maximum permissible continuous power is +34 dBm due to thermal limits.^c RBW $\rightarrow$ 1 kHz.^d Valid for a 12-month calibration interval.^e The maximum permissible continuous power is +34 dBm due to thermal limits.^f R&S®CMW500 only.^g The maximum permissible continuous power is +34 dBm due to thermal limits.^h R&S®CMW500 only.ⁱ The maximum permissible continuous power is +34 dBm due to thermal limits.

During the call, the tester must verify the handset's RF performance and checks the correct operation of the basic signaling features. The best testers do not rely on any special test modes in the mobile station, performing their measurements under conditions almost identical to those in a real network. A voice loop-back allows quick verification of the performance of a mobile as it is perceived by the user.

1.8.4 GSM

Measurement, test, and adjustment capabilities for GSM should include the following.

- Synchronization of mobile phone with base station (which is simulated by CTS)
- Location update
- Call setup (incoming/outgoing)
- Call release (incoming/outgoing)
- Control and measurement of transmitter power
- Handover (channel change)
- Sensitivity, including bit error rate (BER) and raw bit error rate (RBER), limit sensitivity via search routine, RxLev and RxQual
- Phase and frequency error
- Power ramp versus time
- Timing error
- AFC (automatic frequency correction) and RSSI (radio signal strength indication)
- *I/Q* modulator adjustment
- Echo test (voice test, includes also testing of loudspeaker and microphone)
- Functional test of mobile's keypad through display of dialed number
- Display of IMSI (international mobile subscriber identity), IMEI (international mobile equipment identity), power class, and revision level
- Short message service (SMS)

1.8.5 DECT

Measurement, test, and adjustment capabilities for DECT should include the following.

- Synchronization of DUT with the CTS
- Call setup
- Call release
- Echo test
- Detection and display of RFPI (FP)
- Normal transmit power (NTP)
- Power ramp versus time
- Modulation characteristics versus time
- Frequency offset
- Maximum modulation deviation
- Frequency drift

- Timing (jitter, packet delay)
- Bit error rate (BER), frame error rate (FER)

1.9 CONVERTING C/N OR SNR TO E_b/N_0

Figure 1.136 shows an application note for converting carrier-to-noise ratio (C/N) or SNR to energy per bit / normalized noise power (E_b/N_0).

Conversion of C/N or SNR to E_b/N_0

Often BER diagrams do not have C/N as abscissa (FIG) but E_b/N_0 , which is the energy per information bit E_b referred to the normalized noise power N_0 [1]. C/N describes the ratios in the transmission channel, SNR the signals after the $\sqrt{\text{cos}}$ receive filter. The following applies:

$$C/N = \text{SNR} + k_{\text{roll off}} [\text{dB}]$$

In converting the two quantities, some factors have to be taken into account as shown by equations 1 and 2 on the right.

To determine C/N [dB] or E_b/N_0 [dB], the logarithmic ratios have to be corrected using the following factors:

- Factor for Reed-Solomon FEC*

$$k_{\text{FEC}} = 10 \times \lg \frac{188}{204}$$

$$k_{\text{FEC}} = -0.3547 [\text{dB}]$$

- Factors for QPSK/QAM modulation

$$k_{\text{QPSK/QAM}} = 10 \times \lg(m)$$

Modulation	m	$k_{\text{QPSK/QAM}} [\text{dB}]$
QPSK	2	3.0103
16QAM	4	6.0206
64QAM	6	7.7815
256QAM	8	9.0309

- Factor for coding rate [P = 1 for QAM]*

$$k_p = 10 \times \lg(P)$$

Modulation	P	$k_p [\text{dB}]$
QPSK	1/2	-3.0103
	2/3	-1.7609
	3/4	-1.2494
	5/6	-0.7918
QAM	7/8	-0.5799
	1	0

- Factor for $\sqrt{\text{cos}}$ roll-off filtering in demodulator/receiver

$$k_{\text{roll off}} = 10 \times \lg \left(1 - \frac{\alpha}{4} \right)$$

Modulation	α	$k_{\text{roll off}} [\text{dB}]$
DVB-C	0.15	-0.1660
DVB-S	0.35	-0.3977

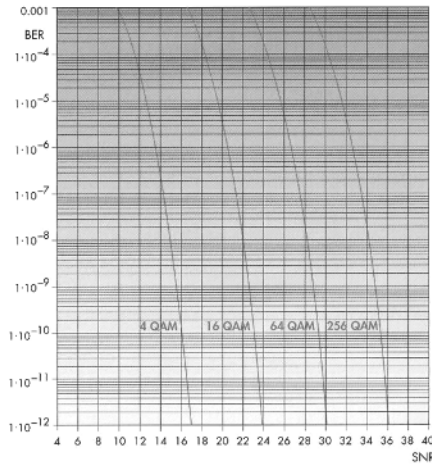
* To be taken into account if E_b only refers to the information bits; not to be taken into account if E_b refers to all bits transmitted (information plus error control bits)

$$C/N = E_b/N_0 + 10 \times \lg \frac{188}{204} + 10 \times \lg(m) + 10 \times \lg(P) - 10 \times \lg \left(1 - \frac{\alpha}{4} \right) [\text{dB}]$$

Equation 1

$$E_b/N_0 = C/N - 10 \times \lg \frac{188}{204} - 10 \times \lg(m) - 10 \times \lg(P) + 10 \times \lg \left(1 - \frac{\alpha}{4} \right) [\text{dB}]$$

Equation 2



BER referred to SNR for 4, 16, 64 and 256QAM

The types of correction factor required depend on whether measurement is made

- in the transmission channel,
- before or after Viterbi correction,
- with QAM or QPSK modulation.

Examples of conversion equations • For **in-channel** measurements with QAM transmission, the following applies:

$$E_b/N_0 = C/N - 10 \times \lg \frac{188}{204} - 10 \times \lg(m) [\text{dB}]$$

The factors for $\sqrt{\text{cos}}$ roll-off filtering and puncturing rate are not needed. For measurements in the **QAM demodulator**, $\sqrt{\text{cos}}$ roll-off filtering has to be taken into account.

$$E_b/N_0 = C/N - 10 \times \lg \frac{188}{204} - 10 \times \lg(m) + 10 \times \lg \left(1 - \frac{\alpha}{4} \right) [\text{dB}]$$

For measurements in the **satellite demodulator** with QPSK, the equation for determining the BER as a function of E_b/N_0 after Viterbi FEC is as follows:

$$E_b/N_0 = C/N - 10 \times \lg \frac{188}{204} - 10 \times \lg(m) - 10 \times \lg(P) + 10 \times \lg \left(1 - \frac{\alpha}{4} \right) [\text{dB}]$$

In the latter case all correction factors are included.

Reference:

[1] News From Rohde & Schwarz, Number 163 (1999/III)

Figure 1.136 Conversion of C/N or BER to E_b/N_0 .

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