I

Functional Principle of Radio Receivers

I.1 Some History to Start

Around 1888 the physicist Heinrich Hertz experimentally verified the existence of electromagnetic waves and Maxwell's theory. At the time his transmitting system consisted of a spark oscillator serving as a high frequency generator to feed a dipole of metal plates. Hertz could recognize the energy emitted by the dipole in the form of sparks across a short spark gap connected to a circular receiving resonator that was located at some distance. However, this rather simple receiver system could not be used commercially.

I.1.1 Resonance Receivers, Fritters, Coherers, and Square-Law Detectors (Detector Receivers)

The road to commercial applications opened only after the Frenchman Branly was able to detect the received high-frequency signal by means of a coherer, also known as a *fritter*. His *coherer* consisted of a tube filled with iron filings and connected to two electrodes. The transfer resistance of this setup decreased with incoming high-frequency pulses, producing a crackling sound in the earphones. When this occurred the iron filings were rearranged in a low-resistance pattern and thus insensitive to further stimulation. To keep them active and maintain high resistance they needed to be subjected to a shaking movement. This mechanical shaking could be produced by a device called a Wagner hammer or knocker. A receiving system comprising of a dipole antenna, a coherer as a detector, a Wagner hammer with direct voltage source and a telephone handset formed the basis for Marconi to make radio technology successful world-wide in the 1890s.

The components of this receiver system had to be modified to meet the demands of wider transmission ranges and higher reliability. An increase in the range was achieved by replacing the simple resonator or dipole by the Marconi antenna. This featured a high vertical radiator as an isolated structure or an expanded fan- or basket-shaped antenna

Radio Receiver Technology: Principles, Architectures and Applications, First Edition. Ralf Rudersdorfer.

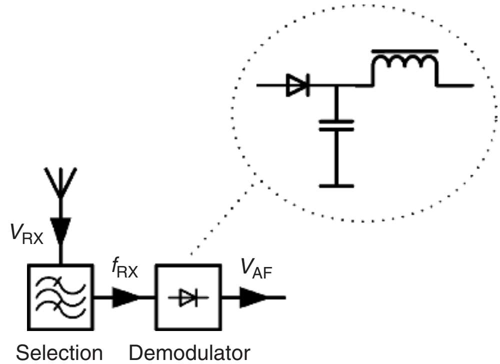

Figure I.1 Functional blocks of the detector receiver. The demodulator circuit shown separately represents the actual detector. With the usually weak signals received the kink in the characteristic curve of the demodulator diode is not very pronounced compared to the signal amplitude. The detector therefore has a nonlinear characteristic. It is also known as a square-law detector. (The choke blocks the remaining RF voltage. In the simplest versions it is omitted entirely.)

of individual wires with a ground connection. The connection to ground as a 'return conductor' had already been used in times of wire-based telegraphy.

The selectivity which, until then, was determined by the resonant length of the antenna, was optimized by oscillating circuits tuned by means of either variable coils or variable capacitors. At the beginning of the last century a discovery was made regarding the rectifying effect that occurs when scanning the surface of certain elements with a metal pin. This kind of detector often used a galena crystal and eventually replaced the coherer. For a long while it became an inherent part of the *detector* receiver used by our great-grandparents (Fig. I.1).

The rapid growth of wireless data transmission resulted in further development of receiving systems. Especially, the increase in number and in density of transmitting stations demanded efficient discriminatory power. This resulted in more sophisticated designs which determined the selectivity not only by low-attenuation matching of the circuitry to the antenna but also by including multi-circuit bandpass filters in the circuits which select the frequency. High circuit quality was achieved by the use of silk-braided wires wound on honeycomb-shaped bodies of suitable size or of rotary capacitors of suitable shape and adequate dielectric strength. This increased not only the selectivity but also the accuracy in frequency tuning for station selection.

I.1.2 Development of the Audion

Particularly in military use and in air and sea traffic, wireless telegraphy spread rapidly. With the invention of the electron tube and its first applications as a rectifier and RF amplifier came the discovery, in 1913, of the feedback principle, another milestone in the development of receiver technology. The use of a triode or multi-grid tube, known as the *audion*, allowed circuit designs that met all major demands for receiver characteristics.

For the first time it was possible to amplify the high-frequency voltage picked up by the antenna several hundred times and to rectify the RF signal simultaneously. The unique feature, however, was the additional use of the feedback principle, which allowed part of the amplified high frequency signal from the anode to be returned in the proper phase to the grid of the same tube. The feedback was made variable and, when adjusted correctly, resulted in a pronounced undamping of the frequency-determining grid circuit. This brought a substantial reduction of the receive bandwidth (Section III.6.1) and with it a considerable improvement of the selectivity. Increasing the feedback until the onset of oscillation offered the possibility of making the keyed RF voltage audible as a beat note. In 1926, when there were approximately one million receivers Germany, the majority of designs featured the audion principle, while others used simple detector circuits.

The nomenclature for audion circuits used ‘v’, derived from the term ‘valve’ for an electron tube. Thus, for example, 0-v-0 designates a receiver without RF amplifier and without AF amplifier; 1-v-2 is an audion with one RF amplifier and two AF amplifier stages. Improvements in the selective power and in frequency tuning as well as the introduction of direct-voltage supply or AC power adapters resulted in a vast number of circuit variations for industrially produced receiver models. The general interest in this new technology grew continuously and so did the number of amateur radio enthusiasts who built their devices themselves. All these various receivers had one characteristic in common: They always amplified, selected and demodulated the desired signal at the same frequency. For this reason they were called *tuned radio frequency* (TRF) *receivers* (Fig. I.2).

Due to its simplicity the TRF receiver enabled commercial production at a low price, which resulted in the wide distribution of radio broadcasting as a new medium (probably the best-known German implementation was the ‘Volksempfänger’ (public radio receiver)). Even self-built receivers were made simple, since the required components were readily available at low cost. However, the tuned radio frequency receiver had inherent technical deficiencies. High input voltages cause distortions with the audion, and circuits with several cascading RF stages of high amplification tend to self-excitation. For reasons of electrical synchronization, multiple-circuit tuning is very demanding with respect to mechanical precision and tuning accuracy, and the selectivity achievable with these circuits depends on the frequency (Fig. I.3). Especially the selectivity issue gave rise to the principle of superheterodyne receivers (superhet in short) from 1920 in the US

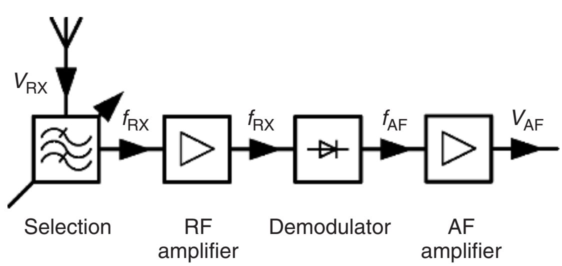

Figure I.2 Design of the tuned radio frequency receiver. Preamplification of the RF signal received has resulted in a linearization of the demodulation process. The amplified signal appears to be rather strong compared to the voltage threshold of the demodulator diode (compare with Figure I.1).

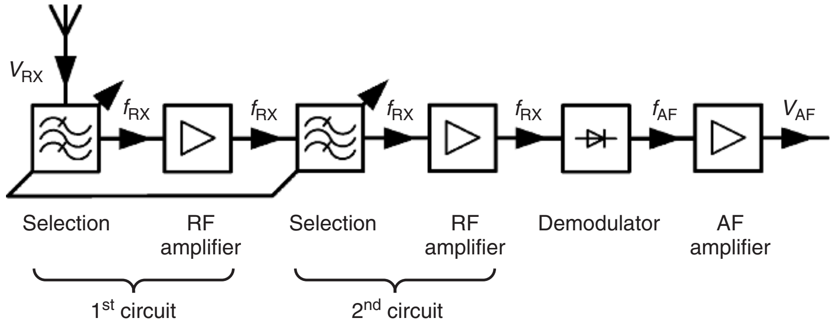

Figure I.3 Multi-tuned radio frequency receiver with synchronized tuning of the RF selectivity circuits. In the literature this circuit design may also be found under the name dual-circuit tuned radio frequency receiver.

and 10 years later in Europe. The superhet receiver solved the problem in the following way. The received signal was preselected, amplified and fed to a mixer, where it was combined with a variable, internally generated oscillator signal (the heterodyne signal). This signal originating from the local oscillator is also known as the LO injection signal. Mixing the two signals (Section V.4.1) produces (by subtraction) the so-called IF signal (intermediate frequency signal). It is a defined constant RF frequency which, at least in the beginning, for practical and RF-technological reasons was distinctly lower than the receiving frequency. By using this low frequency it was possible not only to amplify the converted signal nearly without self-excitation, but also to achieve a narrow bandwidth by using several high quality bandpass filters. After sufficient amplification the intermediate frequency (IF) signal was demodulated. Because of the advantages of the heterodyne principle the problem of synchronizing the tuning oscillator and RF circuits was willingly accepted. The already vast number of transmitter stations brought about increasing awareness of the problem of widely varying receive field strengths (Section III.18). The TRF receiver could cope with the differing signal levels only by using a variable antenna coupling or stage coupling, which made its operation more complicated. By contrast, the utilization of automatic gain control (Section III.14) in the superhet design made it comparatively easy to use.

I.2 Present-Day Concepts

I.2.1 Single-Conversion Superhet

The *superheterodyne receiver* essentially consists of RF amplifier, mixer stage, intermediate frequency amplifier (IF amp), demodulator with AF amplification, and tunable oscillator (Fig. I.4). The high-frequency signal obtained from the receiving antenna is increased in the preamplifier stage in order to ensure that the achieved signal-to-noise ratio does not deteriorate in the subsequent circuitry. In order to process a wide range from weak to strong received signals it is necessary to find a reasonable compromise between the maximum gain and the optimum signal-to-noise ratio (Section III.4.8). Most modern systems can do *without* an RF preamplifier, since they make use of low-loss selection and

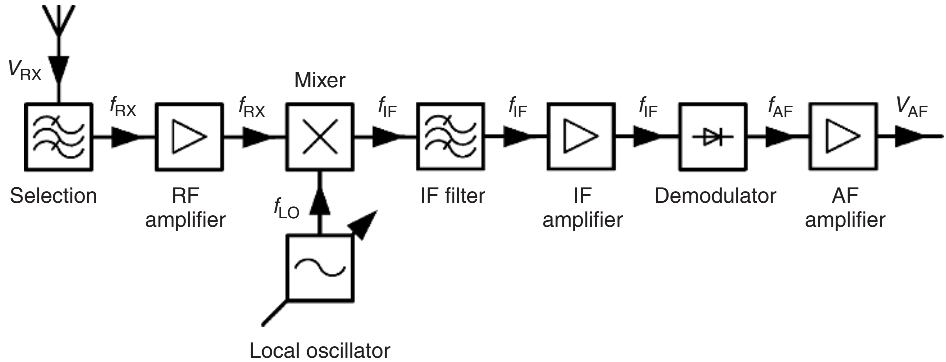

Figure I.4 Functional blocks of the simple superhet. Tuning the receiving frequency is done by varying the frequency of the LO injection signal. Only the part of the converted signal spectrum that passes the passband characteristic (Fig. III.42) of the (high-quality) IF filter is available for further processing.

mixer stages with low conversion loss. The required preselection is achieved by means of a tunable preselector or by using switchable bandpass filters. These are designs with either only a few coils or with a combination of high-pass and low-pass filters.

Previously, the mixer stage (Section V.4) was designed as an additive mixer using a triode tube. This was later replaced by a multiplicative mixer using a multi-grid tube like a hexode (in order to increase the signal stability some circuit designs made use of beam-reflection tubes as mixers). With the continued progress in the development of semiconductors, field-effect transistors were used as additive mixers. These feature a distinct square characteristic and are clearly superior to the earlier semiconductor mixers using bipolar transistors. Later developments led to the use of mixers with metal oxide field-effect transistors (FETs). The electric properties of such FETs with two control electrodes correspond to those of cascade systems and enable improved multiplicative mixing. High oscillator levels result in acceptable large-signal properties (Section III.12). Symmetrical circuit layouts suppressing the interfering signal at the RF or IF gate are still used today in both simple- and dual-balanced circuit designs with junction FETs. Only with the introduction of Schottky diodes for switches did it become possible to produce simple low-noise mixers with little conversion damping in large quantities as modules with defined interface impedances. Measures such as increasing the local oscillator power by a series arrangement of diodes in the respective branch circuit resulted in high-performance mixers with a very wide dynamic range, which are comparatively easy to produce. Today, they are surpassed only by switching mixers using MOSFETs as polarity switches and are controlled either by LO injection signals of very high amplitudes or by signals with extremely steep edges from fast switching drivers [1]. With modern switching mixers it becomes particularly important to terminate all gates with the correct impedance and to process the IF signal at high levels and with low distortion.

The first IF amplifiers used a frequency range between about 300 kHz and 2 MHz. This allowed cascading several amplifier stages without a significant risk of self-excitation, so that the signal voltage suitable for demodulation could be derived even from signals close to the sensitivity limit (Section III.4) of the receiver. Initially, the necessary selection was achieved by means of multi-circuit inductive filters. Later on the application of highly selective quartz resonators was discovered, which soon replaced the LC filters. The use of several quartz bridges in series allowed a bandwidth adapted to the restrictions of the band allocation and the type of modulation used. Since quartz crystals were costly, several bridge components with switchable or variable coupling were used instead. This enabled manual matching of the bandwidth according to the signal density, telegraphy utilization or radiotelephony. Sometime later, optimum operating comfort was obtained by the use of several quartz filters with bandwidths matched to the type of modulation used. Replacing the quartz crystals by ceramic resonators provided an inexpensive alternative. The characteristics of mechanical resonators were also optimized to suit high performance IF filters. Electro-mechanical transducers, multiple mechanical resonators and so-called reverse conversion coils could be integrated into smaller housings, making them fit for use in radio receivers. The high number of filter poles produced with utmost precision were expensive, but their filter properties were unsurpassed by any other analog electro-mechanical system.

Continued progress in the development of small-band quartz filters for near selection (Section III.6) allowed extending the range of intermediate frequencies up to about 45 MHz. Owing to the crystal characteristics, filters with the steepest edges operated at around 5 MHz. Lower frequencies required very large quartz wafers, while higher frequencies affected the slew rate of filters having the same number of poles. Modern receivers already digitize the RF signal at an intermediate frequency, so that it can be processed by means of a high-performance digital signal processor (DSP). The functionality of the processor depends only on the operating software. It not only performs the 'calculation' of the selection, but also the demodulation and other helpful tasks like that of notch-filtering or noise suppression.

The maximum gain, especially of the intermediate frequency amplifier, was adapted to the level of the weakest detectable signal. With strong incoming signals, however, the gain was too high by several orders of magnitude and, without counter measures, resulted in overloading the system. In order to match the amplifier to the level of the useful signal and to compensate for fading fluctuations, the automatic gain control (AGC) was introduced (Section III.14). By rectifying and filtering the IF signal before its demodulation, a direct voltage proportional to the incoming signal level is generated. This voltage was fed to amplifier stages in order to generate a still undistorted signal at the demodulator even from the highest input voltages, causing the lowest overall gain. When the input level decreased the AGC voltage also decreased, causing an increase in the gain until the control function is balanced again. However, the amplifier stages had to be dimensioned so that their gain is controlled by a direct voltage. Very low input signals produce no control voltage, so that the maximum IF gain is achieved. The first superhets for short-wave reception were designed with electron tubes having a noise figure (Section III.4.2) high enough that suitable receiver sensitivities could not be achieved without an RF preamplifier. In order to protect critical mixer stages from overloading, the RF preamplifier was usually integrated into the AGC circuit.

To ensure that signals of low receive field strength and noise were not audible at full intensity, some high-end receivers featured a combination of manual gain control (MGC) and automatic gain control (AGC), the so-called delayed control or delayed AGC (Fig. I.5).

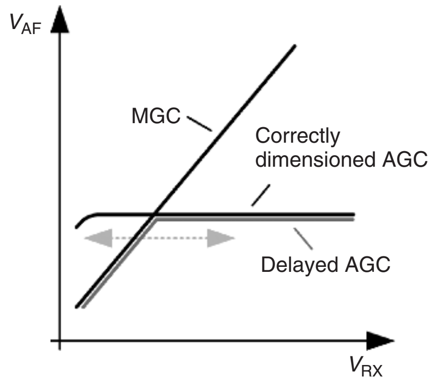

Figure I.5 Functional principle of different RX control methods. In the case of *manual control* the preset gain is kept constant, that is, the AF output voltage follows the RF input voltage proportionally. The characteristic curve can be shifted in parallel by changing the MGC voltage (the required control voltage is supplied from an adjustable constant voltage source). If dimensioned correctly, the automatic gain control (AGC) maintains a constant AF output voltage over a wide range of input voltages. The *delayed AGC* is not effective with weak input signals, but becomes active when the signal exceeds a certain preadjusted threshold and automatically maintains a constant AF output voltage – it is therefore called the 'delayed' gain control.

The automatic control of the gain cuts in only at a certain level, while with lower RF input signals the gain was kept constant. This means that up to an adjustable threshold both the input signal and the output signal increased proportionally. Thus, the audibility of both weak input signals and noise is attenuated to the same degree [2]. This makes the receiver sound clearer. In addition, the sometimes annoying response of the AGC to interfering signals of frequencies close to the receiving frequency (Section III.8.2) that may occur with weak useful signals, can be limited.

During the time when radio signals were transmitted in the form of audible telegraphy or amplitude-modulation signals a simple diode detector was entirely suitable as a demodulator. This was followed by a variable multi-stage AF amplifier for sound reproduction in headphones or loudspeakers. In order to make simple telegraphy signals audible an oscillator signal was fed to the last IF stage in such a way that a beat was generated in the demodulator as a result of this signal and the received signal. When the received signal frequency was in the centre of the IF passband (see Figure III.42) and the frequency of the beat-frequency oscillator deviated by, for example, 1 kHz, a keyed carrier became audible as a pulsating 1 kHz tone. This beat frequency oscillator (BFO) is therefore known as heterodyne oscillator (LO).

With strong input signals the generation of the beat no longer produces satisfactory results. The loose coupling was therefore soon replaced by a separate mixing stage, called the product detector since its output signal is generated by multiplicative mixing. With product detectors it then became possible to demodulate single-side-band (SSB) modulation that could not be processed with an AM detector.

Besides the task of developing a large-signal mixer, a symmetrical quartz filter with steep edges or a satisfactorily functioning AGC (that is well adapted to the modulation type used), especially the design of a variable local oscillator for the superhet presented an enormous challenge for the receiver developer.

The first heterodyning oscillators oscillated freely. Tuning was either capacitive by a rotatable capacitor or inductive after ferrites became available. The first generation of professional equipment used an oscillator resonance circuit that varied synchronously with the input circuits of the RF amplifier stages. For this the variable capacitors had the same number of plate packages as the number of circuits that needed tuning. In most amateur radio equipment, however, the input circuits were tuned separately from the oscillator for practical reasons. Any major detuning of the oscillator therefore required readjusting of the preselector. The frequency of the freely oscillating oscillators was lower than the received frequency. The higher the tuning frequency the lower was the stability with varying supply voltages and temperatures. Frequency stability could be achieved only by utmost mechanical precision in oscillator construction, the integration of cold thermostats, and the use of components having defined temperature coefficients. By combining these measures an optimum compensation was obtained over a wide temperature range. Manufacturing a frequency-stabilized tuning oscillator was difficult, even with industrial production methods, and required extra efforts of testing and measuring.

In order to prevent frequency fluctuations due to changing supply voltages and/or loads, oscillators are usually supplied with voltages from electronically regulated sources. Load variations originating from the mixing stage or subsequent amplifier or keying stages during data transmission are counteracted by incorporating at least one additional buffer stage. Its only task is the electrical isolation of the oscillator from the following circuits.

In the beginning, the receive frequency was indicated as an analog value by means of a dial mounted on the axis of the oscillator tuning element. The dial markings directly indicated the receive frequencies or wavelengths and, in the case of broadcast receivers, showed the stations that could be received. (A few units had a mechanical digital display of the frequency. Among them were the NCX-5 transceiver from National and the 51S-1 professional receiver from Collins. They allowed a tuning accuracy of 1 kHz.)

An accurate reproduction of the tuned-in frequency was possible only with a digital frequency counter used for determining and displaying the operating frequency. The display elements used were Nixie tubes, later the LED dot-matrix or seven-element displays, and recently mostly LC displays. To indicate the receive frequency, the frequency counted at the oscillator must be corrected when resetting the counter either by direct comparison of the BFO frequency counted in a similar manner or by preprogramming the complements.

I.2.2 Multiple-Conversion Superhet

The mixer stage of a superheterodyne receiver satisfies the mathematical condition for generating an intermediate frequency from the heterodyne signal with two different receive frequencies (III.5.3). Both the difference between the receive frequency (f_{RX}) and the LO frequency (f_{LO}) and the difference between the LO frequency and a second receive frequency generate the same intermediate frequency (f_{IF}). The two receive frequencies form a mirror image relative to the frequency of the oscillator, both separated by the IF. The unwanted receive frequency is therefore called the image frequency. The frequency of any such signal is equal to the IF and directly affects the wanted signal or, in extreme cases, covers it altogether. To avoid this, the image frequency must be suppressed. This is usually done by preselection, i.e. by means of the resonance circuits of the RF preamplifier or the preselector. At the beginning of the superhet era the near selection (Section III.6), responsible for the selectivity by filtering the useful signal from the adjacent signals, was possible only with high-quality multi-circuit bandpass filters having a low frequency. From the actual image frequency it is obvious that, for a low IF, it can be suppressed only with a considerable amount of filtering. Especially with receivers designed for several frequency ranges, the reception of high-frequency signals was strongly affected by an insufficiently suppressed image frequency (Section III.5.3). It was therefore necessary to find a compromise between image frequency suppression and selectivity, based on the intermediate frequency.

This problem was solved by twofold heterodyning. To reject the image frequency the first IF was made as high as possible; the higher the IF the lower the effort to suppress the image frequency (see Fig. III.36). A second mixer converted to a second IF so low that good near selection was possible at an acceptable cost (Fig. I.6). But the second mixer again produces both a useful frequency and an image frequency. The second image frequency must also be suppressed as far as possible by means of a filter operating on the first IF. In the era of coil filters this required very careful selection of the frequency.

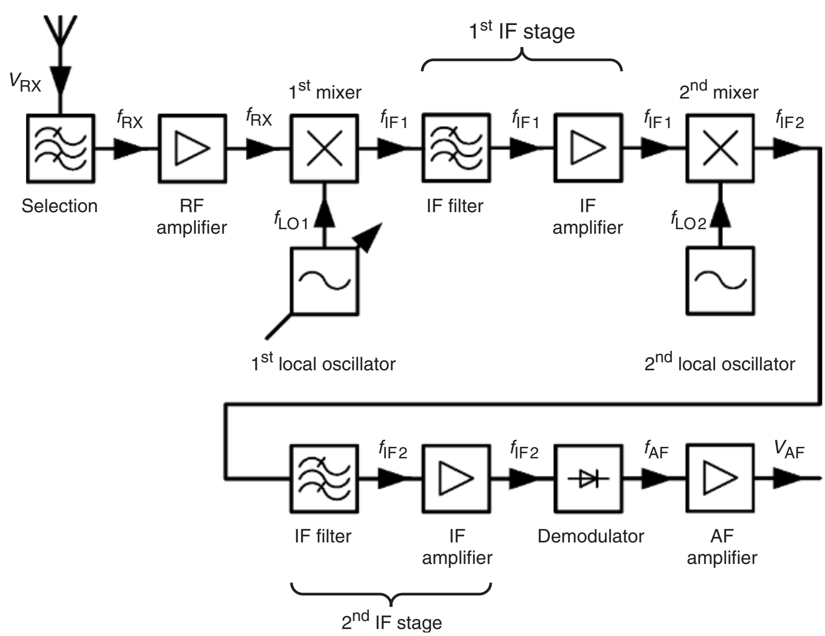

Figure I.6 Operating principle of multiple-conversion superheterodyne receivers. The design shown here is called a dual-conversion superhet. The first IF is a high frequency and serves mainly to prevent receiving image frequencies. The second mixer changes to a lower IF in order to perform the main selection.

The higher the first IF was chosen in the *dual-conversion superhet*, the more difficult it became to manufacture a variable freely-oscillating first local oscillator with a frequency low enough to cause sufficient frequency drifts (Section III.15), for example, for stable telegraphy reception at narrow bandwidths. If the LO frequency was above the receive frequency in one frequency range and below it in the other, the analog frequency scales had to be marked in opposing directions, making operating the equipment cumbersome. Attempts were therefore made to stabilize the first oscillator as well as possible. Initially, this utilized the converter method – the first oscillator remained untuned and was stabilized by a quartz element, while tuning was achieved with the second local oscillator. However, this required that the filter of the first IF be as wide as the entire tuning range. This design was used in almost all early equipment generations for semi-professional use (including amateur radio service) like those produced by Heathkit or Collins. In order to minimize overloading due to the high number of receiving stations within one band, the tuning range was limited to only a few hundred kHz. In the Collins unit, featuring electron tubes, the first IF was merely 200 kHz wide. With a tunable second local oscillator at a lower frequency the conversion to a lower, narrower second IF was simple and stable.

Nevertheless, the problem of large-signal immunity (Section III.12) remained. By using a first tunable local oscillator at a high frequency it was attempted to again reduce the bandwidth of the first IF to the strictly necessary maximum bandwidth, depending on the widest modulation type to be demodulated. At first, the premix system was used. This consisted of a low-frequency tuned oscillator of sufficient frequency stability and a mixer for converting the signal to the required frequency by means of switchable signals from the quartz oscillators. Since the mixing process produced spurious emissions, subsequent filtering with switchable bandwidths was necessary. This is a complex method, but free of the deficiencies described above. It established itself with Drake and TenTec in the semi-professional sector (Fig. I.7). With a tunable first local oscillator it is sufficient for the second LO to use a simple quartz oscillator with a fixed frequency.

As long as the required frequency bands were restricted to a reasonable number (like the short-wave broadcasting bands or the classical five bands of amateur radio services) this principle left nothing to be desired. However, the need for receivers covering all frequency ranges from <1 MHz to 30 MHz inevitably increased the number of expensive quartz elements and increased the demands on near selection of the premixer. This changed only with the availability of low-cost digital integrated semiconductor circuits, which simplified frequency dividing. When dividing the output frequency of an oscillator to a low frequency and comparing it with the divided frequency of a reference signal stabilized by quartz elements, the oscillator can be synchronized by means of a voltage-dependent component (like a varactor diode) using a direct voltage derived from the phase difference between the two signals for retuning the oscillator. This was the beginning of phase-locked loops (PLL) and voltage-controlled oscillators (VCO) (Fig. I.8). Particularly the PLL circuits gave an enormous boost to the advancement of frequency tuning in receivers. Today, highly integrated circuits enable the design of complex and powerful tuned oscillator systems for all frequency ranges. Using several control loops they achieve very high resolution with very small frequency tuning increments [3], short settling times (Section III.15) even with wide frequency variations, and little sideband noise (Section III.7.1). Those circuits used for generating heterodyne signals are called synthesizers.

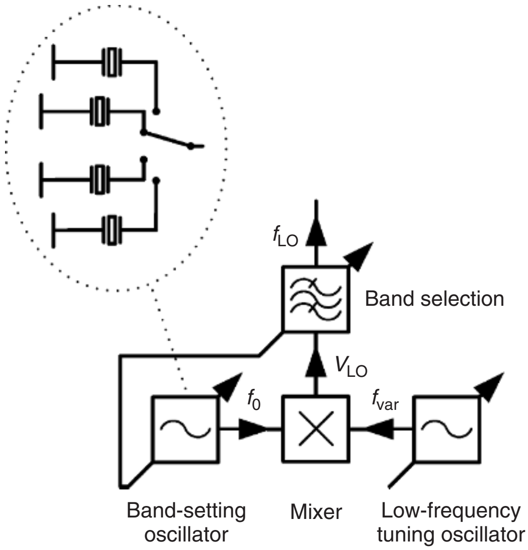

Figure I.7 Architecture of a premixer assembly which feeds an LO injection signal of a stable frequency to the first mixer of a multiple-conversion superhet receiver. The separately depicted circuit design of switchable quartz elements is of course part of an oscillator in actual equipment.

But it is necessary to use processors to make such circuits more ergonomic and the many functions easier to use. With processors the operating frequency can be tuned almost continuously by means of an optical encoder or be activated directly by a number entered via the keyboard. It is possible to store many frequencies in a memory. In the latest developments the loop for fine-tuning is replaced by direct digital synthesis (DDS) (Fig. I.9). This generates an artificial sinusoidal from the digital input information and the signal is tunable in increments of $\ll$1 Hz. It is controlled by the operating processor,

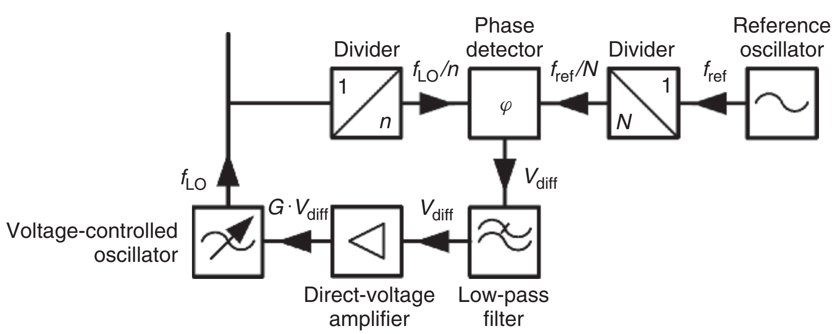

Figure I.8 VCO with phase-locked loop. The direct voltage V_{diff} for automatic frequency tracking is smoothed in the so-called loop filter to prevent spurious signals and sideband noise. V_{diff} is adapted to the required voltage range of the voltage-controlled oscillator via subsequent amplification by the factor G. This results in the constant output frequency $f_{LO} = n \cdot f_{ref}/N$.

Figure I.9 Complete DDS capable of producing output signals up to 400 MHz with a resolution of 14 bits. Only a reference clock and a low-pass filter must be provided externally. (Company photograph of Analog Devices.)

which is required in any case. Depending on the resolution of the D/A converter in the DDS module the output signal generated has very little phase noise (Fig. III.50) and unwanted spurious components (Fig. III.51). Owing to the rapid progress made in this technology DDS generators are currently used in almost every radio receiver. Fully integrated circuits that can generate output signals up to 500 MHz are available. (An example of this technology is AD9912 from Analog Devices, featuring a phase noise as low as −131 dBc/Hz at 10 kHz separation distance with an output frequency of 150 MHz. The output frequency can be varied by increments as small as 3.6 μHz [4]. The spurious emissions actually occurring depend to a large extent on the type of programming.)

It was quickly realized that large-signal problems can be eliminated only if the first narrow-band selection takes place in an early stage of the receive path. In multiple-conversion systems quartz filters with a frequency in the range of about 5 MHz to 130 MHz were therefore included already in the first IF. The first IF is amplified just enough so that the subsequent stages do not noticeably affect the overall noise factor (Section V.1). In high-linearity RF frontends there is no amplification at all upstream of the first mixer. The narrower the bandwidth in the first IF the higher is its relieving effect for the second mixer. Usually the second mixer stage is much simpler than the first mixer. Nowadays, the latest high-end radio receivers match the selected bandwidth already in the first IF stage to the respective transmission method by switching roofing filters (Fig. I.10). (Quartz filters are used in most cases. The commonly used term 'roofing' filter indicates its protective effect on all subsequent stages, just as the roof of a house protects all rooms underneath from the weather.) This satisfies the need for matching the selection to the modulation in order to achieve optimum large-signal immunity or for processing the useful signal with low frequency spacing to strong interferences.

For the second IF, almost all professional receivers used a frequency for which selection filters were readily available on the market, usually the frequency of 455 kHz. Telefunken developed their own mechanical filters of 200 kHz and 500 kHz, while Japanese developers chose to use their own frequencies, probably for competitive reasons. In professional systems amplification was made so high that the AGC cut in even with the weakest signals. This made such signals strong enough to be displayed (Section III.14) and to

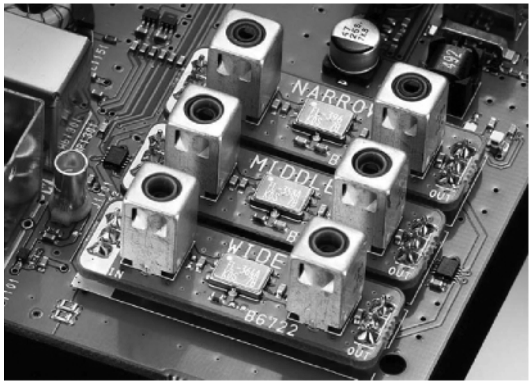

Figure I.10 Switchable filters with a bandwidth of 15 kHz/6 kHz/3 kHz matched to the requirements of the transmission methods F3E/A3E/J3E. In modern HF radio transceivers they are placed in the first IF stage (here at a frequency of 64.455 MHz) of the receiving section. Visible are the matching networks arranged close to the actual filters. (Company photograph of ICOM.)

produce a constant AF output level. With these high IF amplifications a control range of 110 dB was no rarity. (For amateur equipment this philosophy never gained ground. Many older-generation radio amateurs were accustomed to the low noise background from their use of low-gain electron tube units which, for weaker signals, needed a ‘boost’ from the AF amplifier. In order to reproduce such a low background modern amateur receivers also have a low noise level and thus sufficient sensitivity, but the IF amplification is so ‘narrow’ that only signals with an input voltage of several microvolts produce a signal indication, i.e. a constant output voltage. The control element marked MGC (manual gain control) is often used to shift the threshold value of the delayed AGC (Fig. I.5).)

The demodulation and the AF circuits of a dual-conversion receiver are not much different from those of a single-conversion superhet.

Unlike commercial radio services (Section II.3) that usually work with only a few permanently assigned and sparsely occupied frequencies, search receivers (Section II.4.2) used in radio monitoring, radio reconnaissance and amateur radio services, are dedicated to the reception of weak signals in an interference-prone environment. Very early, those units were therefore equipped with auxiliary devices for interference suppression. Notch filters are used to blank out constant whistling sounds or telegraphy signals from the voice band, while interferences at the periphery of the basic channel can be eliminated by parallel shifting of the filter passband without altering the receiving frequency. The latter method is called passband tuning (Fig. I.11). Eventually, IF systems were developed that allowed independent variation of each of the filter edges (Fig. III.42) of the selection filter in order to respond individually to interferences. A simple passband tuning system can be realized in a single-conversion superhet, while the so-called IF shift for independent edge adjustment always requires a dual-conversion superhet design. When adding the capability to receive signals with frequency modulation (F3E) by means of a dedicated low-frequency limiting IF circuit, all receive functions can be realized in a multiple-conversion superhet receiver. Some units generate a low-frequency IF simply to

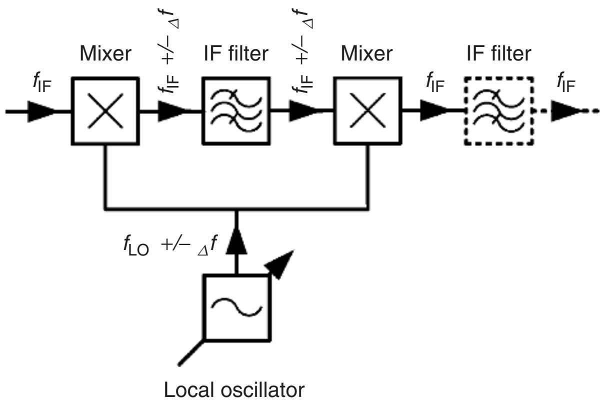

Figure I.11 Passband tuning principle, also known as IF centre frequency shifting. This allows shifting the passband of the IF stage without changing the receive frequency. By including another IF filter behind this stage the IF bandwidth can be varied continuously, that is, it can be matched to the input signal [5]. With this simple method of continuous IF bandwidth adjustment only one filter edge is actually shifted. This makes the passband asymmetrical to the centre frequency. With narrower passbands, however, the shape factor (Section III.6.1) of the IF passband characteristic deteriorates due to the fixed edge steepness of the two IF filters.

enable the use of an efficient notch filter. In order to prevent the mutual interference of the oscillator signals necessary for the multiple-conversion superhet and the resulting mixer products, it is essential not only to plan the frequencies very carefully but also to exercise great care to ensure electronic decoupling and shielding in the mechanical construction.

I.2.3 Direct Mixer

If the oscillator frequency of a superhet receiver is allowed to drift ever closer to the receive frequency the intermediate frequency becomes lower and lower until it reaches zero. The modulation contents of the useful signal are then converted directly to the low frequency range. A receiver working on this principle is called a *direct mixer*, direct-conversion receiver or zero-IF receiver. It avoids the use of an intermediate frequency and thus allows relocating the circuits for amplification, selection and AGC to the AF section (Fig. I.12). This is easily done by using operational amplifiers, such as active filters, amplifiers and control units.

Receivers based on the direct mixer principle remained in the shadows for a long time. The inverse mixing of the signal emitted by the oscillator (Section III.17) with the desired signal leads to hum noise, especially in units operated from a power line. This is why battery-powered units are preferred. With ‘simple’ heterodyning the image frequency adjacent to the received frequency is also within the baseband. (The baseband is that frequency range that normally contains the useful information, the news contents. In radio technology the transmitted news contents are ‘within the baseband’ prior to modulation

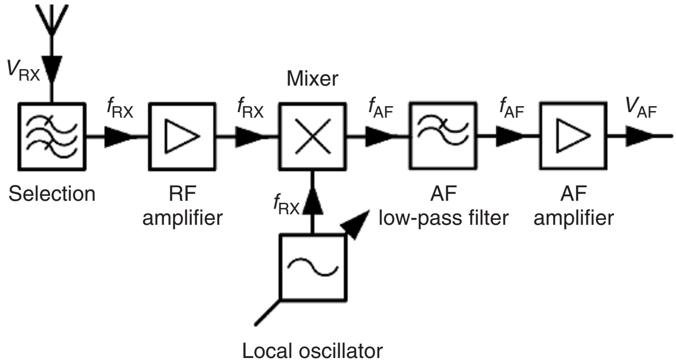

Figure I.12 Basic design of a direct mixer. The image frequency (Section III.5.3) is identical with the incoming frequency f_{RX}. Image frequency reception provides the same tuned frequency, but the demodulated signal spectrum appears inverted, indicating an interference signal.

and after demodulation.) Directly adjacent signals at the image frequency can, therefore, not be suppressed. For a long time this was regarded as such a serious disadvantage that there appeared to be no promise of developing this design to a high-performance 'station receiver'. But systematic implementation of RF/AF engineering enables the direct mixer to provide good receiving performance.

By in-phase splitting of the received signal behind the RF preamplifier and by feeding the two resulting signals to two mixers, where they are converted with the same oscillator signal into two basebands, the two basebands are vectorially orthogonal as AC voltage indicators, provided that the split oscillator signal is also fed to one of the two 90° out of phase (Fig. I.13). *Using these two orthogonal basebands allows the demodulation of signals of all modulation types!* One baseband represents the real component and the other the imaginary component of the complex signal (see also Section I.3.3). Other commonly used terms for these so-called quadrature signals are:

- For the real component: I component or in-phase component.
- For the imaginary component: Q component or quadrature-phase component.

Owing to the fact that RF amplifiers, mixer stages and both baseband branches can be integrated and that after digitization the baseband signals can not only be selected but also demodulated by a highly integrated digital signal processor (DSP), this principle was soon adopted for use in GSM technology. Today, it forms the basis of RF receivers in almost any mobile phone. In mobile radio technology (Section II.3.5) the system is usually referred to as a *homodyne receiver*. Another name for this version of a direct mixer is *quadrature receiver*. Due to the lack of synchronization between the received frequency and the frequency of the LO injection signal, a frequency error occurs because of the limited accuracy even when tuning to nominally the same frequency. For proper functioning [2] this error must be kept small compared with the receive bandwidth (Section III.6.1), since

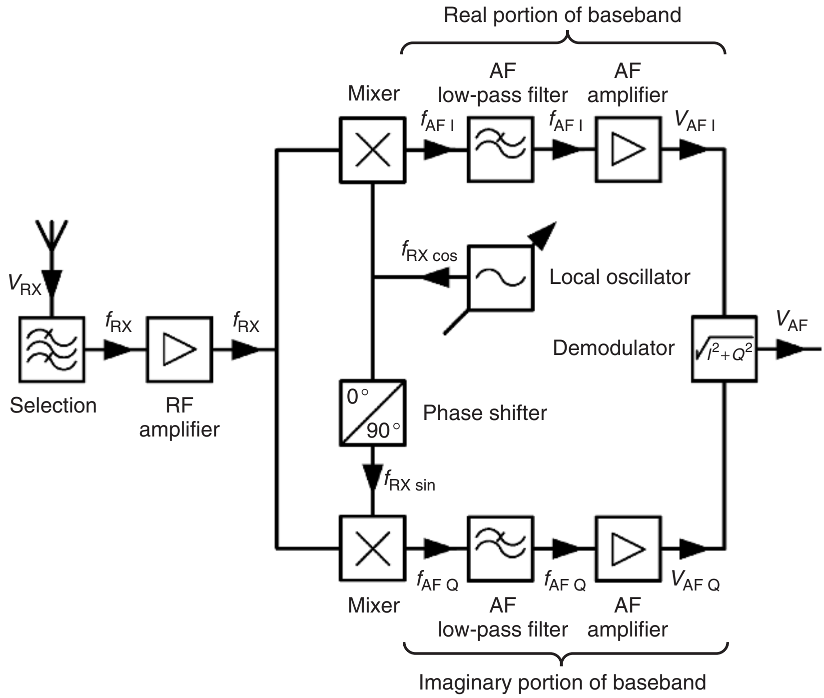

Figure I.13 With the quadrature receiver the main selection is achieved by AF low-pass filters in the I path and the Q path. High performance data can be achieved with fully digitized receiver designs (Section I.2.4) thanks to the very accurate signal processing which this principle makes possible.

slight deviations do not cause any interference, as can be demonstrated mathematically for AM reception:

$$\begin{aligned} S(t) &= \sqrt{(A(t)\cdot\sin(\omega\cdot t))^2 + (A(t)\cdot\cos(\omega\cdot t))^2} \\ &= A(t)\cdot\sqrt{\sin^2(\omega\cdot t)+\cos^2(\omega\cdot t)} \\ &= A(t) \end{aligned} \tag{I.1}$$

where

$S(t)$ = demodulated AF signal at time (t), in V
$A(t)$ = AM signal at time (t), in V
$\omega \cdot t$ = difference between carrier frequency and LO frequency, in rad
t = considered time, in sec

The term ω is not contained in the result, which proves that the frequency deviation from the LO injection signal is insignificant. This presumes, however, that the two mixed spectra are symmetrical to the LO frequency. This is not the case with selective fading. In this respect this demodulator is inferior to the synchronous receiver. For SSB reception the quadrature receiver requires another 90° phase shifter to enable suppressing the

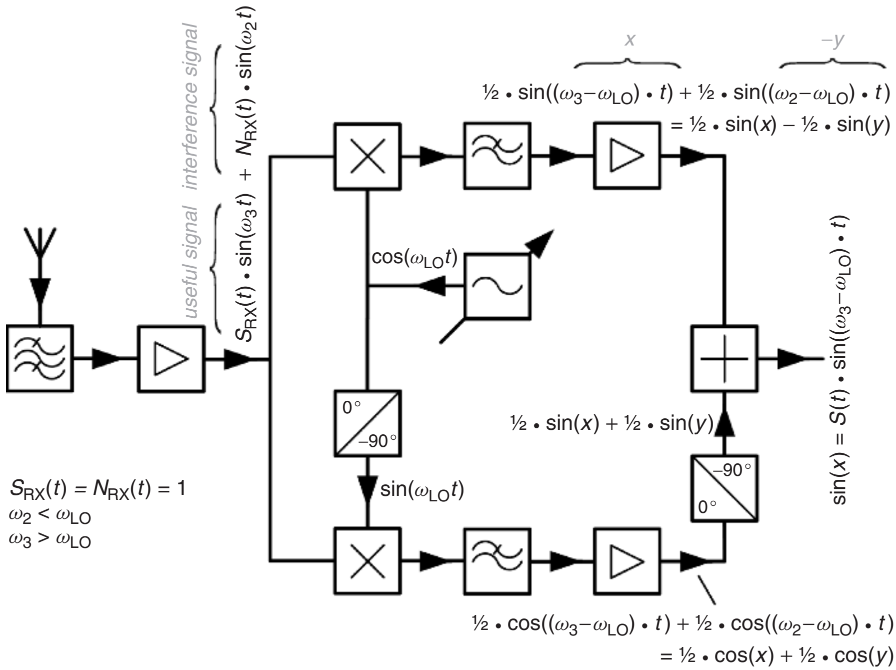

Figure I.14 The quadrature receiver with sideband suppression requires an additional 90° phase shifter. With the fully digitized unit (Fig. I.24) a sideband suppression of more than 100 dB can be obtained without problems.

unwanted sideband (Fig. I.14). The constant phase shift over several octaves in the AF range presented a major challenge in analog technology. This may be another reason why this type of receiver was rarely seen in earlier times.

Synchronizing the LO injection signal with the receive frequency by means of a phase control loop, can accomplish demodulation of FM/PM and AM signals without a demodulator. Such a design is called a *synchronous receiver* (Fig. I.15) which, apart from the omission of the demodulator, is identical to the quadrature receiver. Because of the strictly identical carrier frequencies of the signal and image behind the mixing stage, the even AM sidebands are the same in phase and shape. The same is true for the uneven FM/PM sidebands, assuming 90° out-of-phase mixing in the second branch. In each case, the other component is canceled out. Thus, demodulation takes place during the mixing process [2].

I.2.4 Digital Receiver

All functional blocks of the receiver designs discussed so far can be described mathematically (with regard to the time domain and frequency domain of the transfer characteristics). This means that basically all stages can be reproduced by algorithms in a fast digital processor, provided that A/D conversion (Section I.3.2) is sufficiently fast to convert the signals to a form (bit sequence) suitable for processing. The same considerations

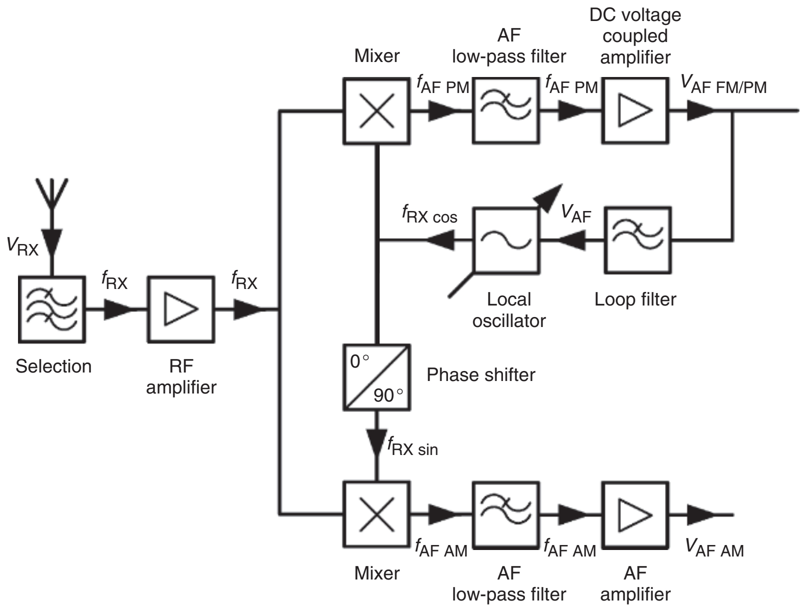

Figure I.15 The synchronous receiver is the second design of the direct mixer that receives the signal without image. If the signal received is phase modulated with a modulation frequency above the limit frequency of the PLL loop filter, the modulation contents can be extracted from the upper branch. Demodulated AM signals are available at the end of the lower branch.

apply as for conventional circuit designs. The in-principle ideal digital architecture has its deficiencies in quantization effects.

In the units marketed from around 1980, digital components were used only for control functions and audio signal processing. These were *first generation digital receivers*.

Using digital signal processors at low intermediate frequencies for 'computationally' processing the useful signal received has been standard in high-end equipment for several years. Modern receivers select the desired signal by means of a DSP from the signal spectrum of the input bandpass converted by the mixer to the intermediate frequency. The DSP performs arithmetic demodulation and keeps the useful signal free of interferences like continuous carrier whistling, noise or crackle. It then evaluates the signal and provides an AGC criterion for controlling the overall gain [6]. A modern DSP is capable of performing the required computing in 'real-time', i.e. with a time delay that is no longer subjectively detectable. Today, these units are called *second generation digital receivers* (Figs. I.16 and I.17). Despite arithmetic processing of the signal, considered unusual from analog perspectives, the significant advantages of this technology are cost savings, particularly for expensive quartz filters, and the enormous flexibility of the characteristics as a result of the software. The analog components, including the RF frontend, must meet high RF demands since these essentially determine the overall receiver properties (III). However, well-functioning digital signal processing alone is by no means sufficient for the manufacture of a radio receiver suitable for practical applications.

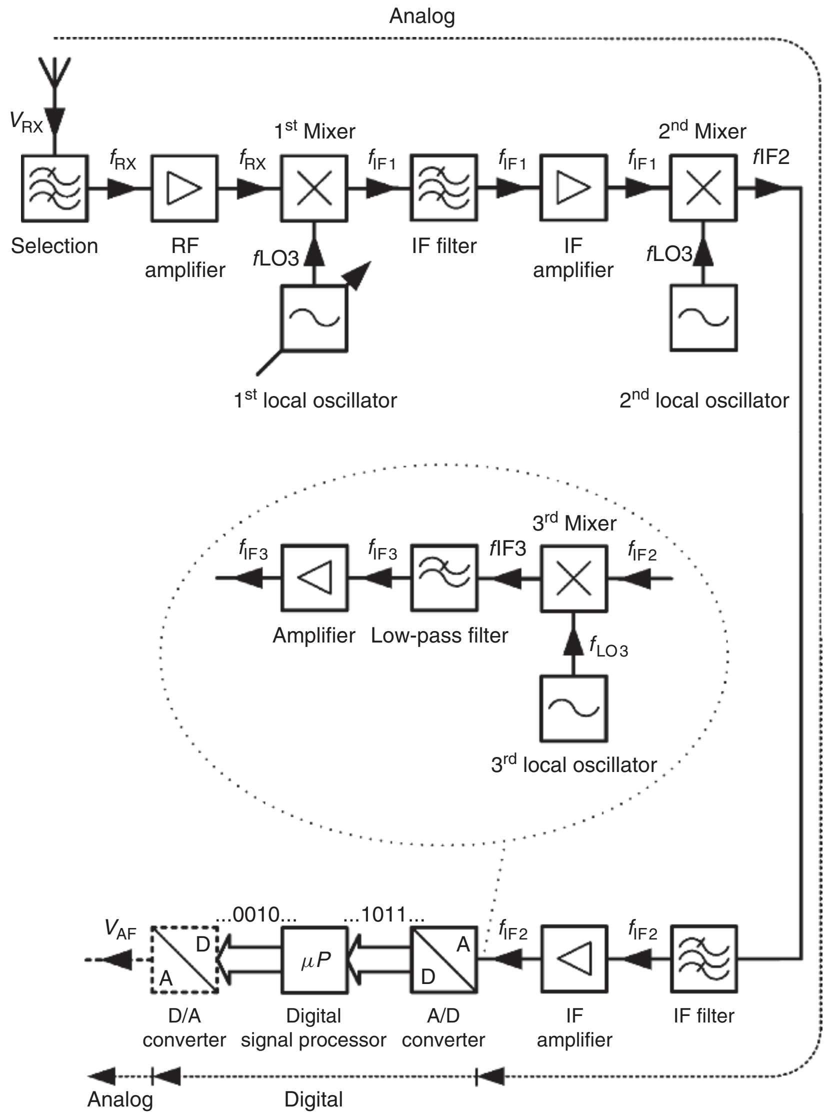

Figure I.16 Second generation digital RX using the superhet principle. Depending on the design concept, A/D conversion is achieved either by subsampling (Section I.3.7) or by the circuitry inside the dotted oval. The fast IF filter, having a bandwidth equal to the widest signal type to be demodulated, guarantees a limitation of the signal frequencies reaching the A/D converter, thus preventing phantom signals (such as those caused by aliasing).
The circuit shown separately depicts the components used for the additional conversion to a lower 3rd IF (usually with a frequency between 12 kHz and 48 kHz). The A/D conversion takes place behind the low-pass filter, having a limit frequency slightly below half the sampling rate. The signal has then passed three mixers and some filters (often too wide for narrow emission classes). (This principle is used in many radio receivers for semi-professional use, as well as in equipment like the VLF/HF receiver EK896 from Rohde&Schwarz.)

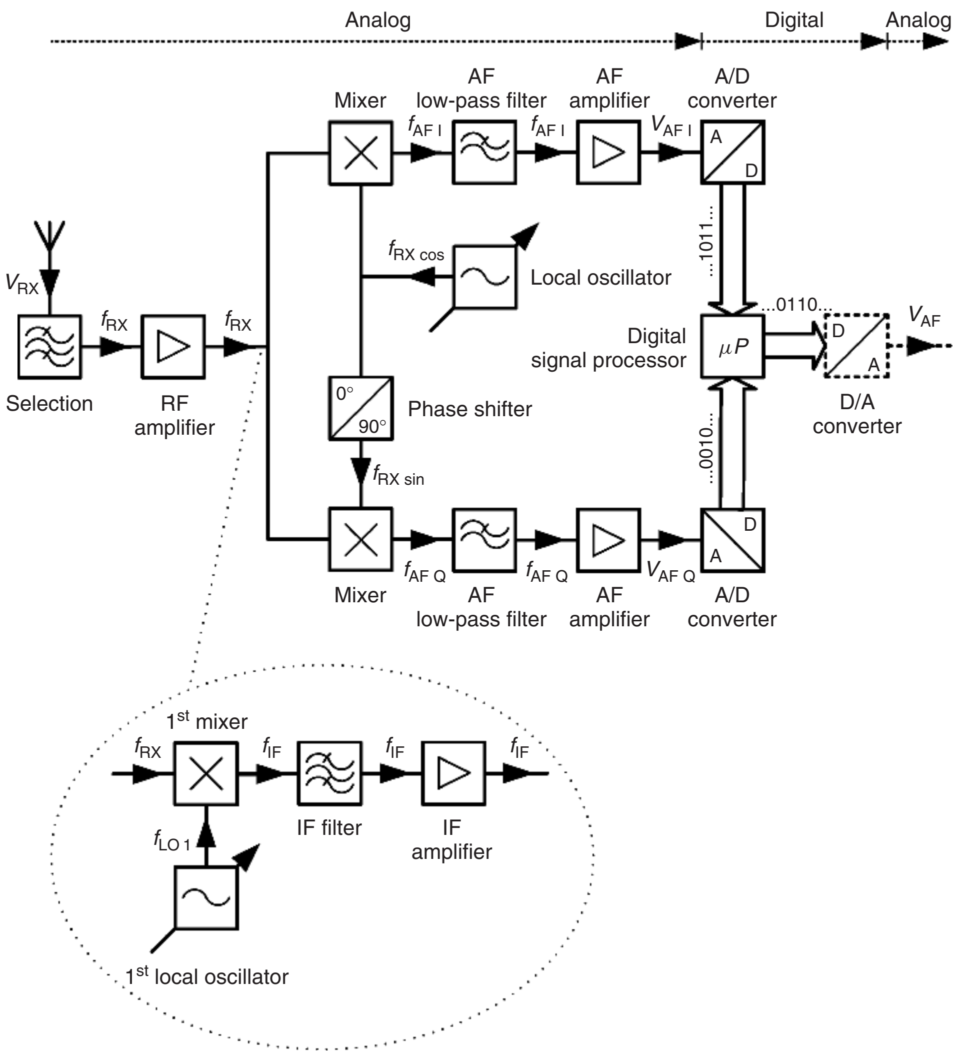

Figure I.17 Second generation digital RX using the quadrature principle. Direct mixers (Section I.2.3) of this design perform a separate A/D conversion of the basebands (as well as of the real and imaginary signal components), which are then combined for subsequent demodulation.
In a different version, shown as a separate circuit, the receive spectrum is converted to a first IF in a highly linear mixer and is then selected by a narrow-band IF filter. This frees the subsequent IQ mixer from sum signals. (The principle was used in the mid 1990s in model 95S-1A from Rockwell-Collins. It covers a receive frequency range from 500 kHz to 2 GHz.)

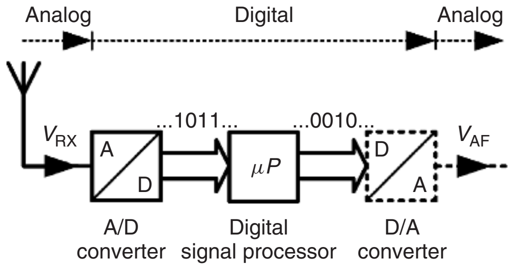

Figure I.18 In an ideal all-digital receiver the A/D conversion takes place close to the antenna socket. The entire signal processing is done by the DSP using mathematical algorithms. However, due to the limited sampling speed of A/D converters, at least one additional low-pass filter is required between the antenna and the A/D converter to prevent exceeding the Nyquist frequency and to avoid aliasing.

Almost all well-known manufacturers of radio equipment [7] now make use of this advanced technology. (The diagram in Figure I.20 shows the classification of the various digital receiver designs according to the location of the A/D converter within the receiver layout.)

In recent years, this technology has made significant progress in digital resolution and clock speed. It seems reasonable therefore to design receivers using only digital signal processing (Fig. I.18). After band selection and analog RF amplification, which is still necessary to achieve a sufficient signal-to-noise ratio, the RF signal is fed directly to a fast A/D converter with high signal dynamics. The subsequent digital signal processor performs all functions previously executed in analog mode, like amplification, selection, interference elimination, and demodulation. The processed signal can now be subjected to digital/analog (D/A) conversion, so that only the resulting AF signal has to be amplified for feeding, for example, a headset or loudspeaker (Fig. I.19). For further signal processing

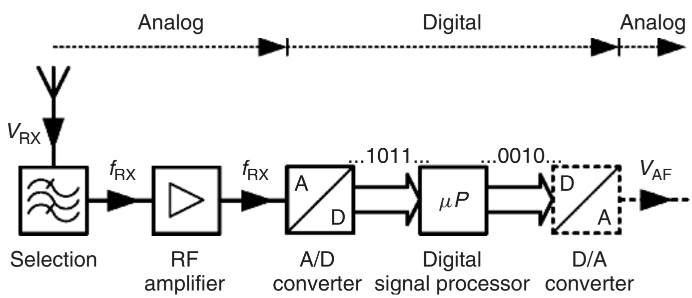

Figure I.19 Digital receiver of generation 2.5. The first units of this type are currently available covering a receive frequency range up to approximately 50 MHz. Depending on the required quality level, it is possible to produce models using only a low-pass filter behind the antenna input instead of a circuit for the specific selection of the desired receiving band.
The final D/A converter is of importance only if the demodulated signal must be available in analog mode, for example, for loudspeakers.

in digital mode, like in decoders or for screen displays, the last D/A converter stage is no longer necessary. Such receiver designs offer a number of advantages [8]:

- Digital signal processing is free of any distortion. Only initial signal conditioning requires special care.
- Problems experienced in analog circuits, like unwanted coupling effects, whistling sounds, and oscillating tendencies, do not exist.
- All modulation modes from AM to complex modes, like quadrature amplitude modulation (QAM) or code-division multiple access (CDMA, Section II.4.1), are supported by one and the same hardware. By using suitable software it is possible to design a multitude of receiver versions up to multi-standard platform models.
- New functions, extensions and modifications of radio standards, like conceptual improvements, can be added by simply installing an improved operating software version (firmware).
- Hardware expenditures based on the effective component costs are much lower than those for analog versions.
- The accuracy is scalable. With suitable software the display of, for example, the relative receive signal strength (Section III.18) can reach an accuracy of better than ± 1 dB over a range of 120 dB.
- Reproducibility is unrestricted. A filter trimmed to a certain shape factor (Section III.6.1) has exactly the same properties in every unit.
- Filter characteristics are freely definable over a wide range of values. This was also desirable with analog filters, but for physical reasons, could not be achieved.

However, with these concepts the technical data of high-end analog receivers can only be partly achieved despite the realization of some still extremely costly professional solutions and first interesting research results (Section I.3) as well as a few experimental models produced by the amateur radio services. For professional use there are already some solutions, however these are still very costly. But owing to new and continuously improved components the feasible range of receiving frequencies is being constantly extended to higher frequencies.

Presently, especially the interference-free dynamic range (Section I.3.2) of A/D converters is still inferior to that of high-performance mixers in combination with narrow-band analog signal processing. The demands on A/D converters regarding a wider bandwidth and a larger dynamic range (to do away with extensive analog prefiltering) are diametrically opposed to each other [7]. It is almost impossible to achieve both goals simultaneously. The best performance is therefore obtained with hybrid concepts (Figs. I.16 and I.17), using analog circuits to generate the IF and digital processing after the respective preselection by quartz filters.

I.2.4.1 Software Radio and Software-Defined Radio

Professional terminology sometimes differentiates between software radio and software-defined radio (SDR) [9]. The first term refers to the ideal *software radio*, i.e. a fully digitized receiver (Fig. I.18). (As already indicated, the software runs on generally available hardware. Since it is primarily the software which defines the functionality of the unit, this is also known as the ideal software radio.)

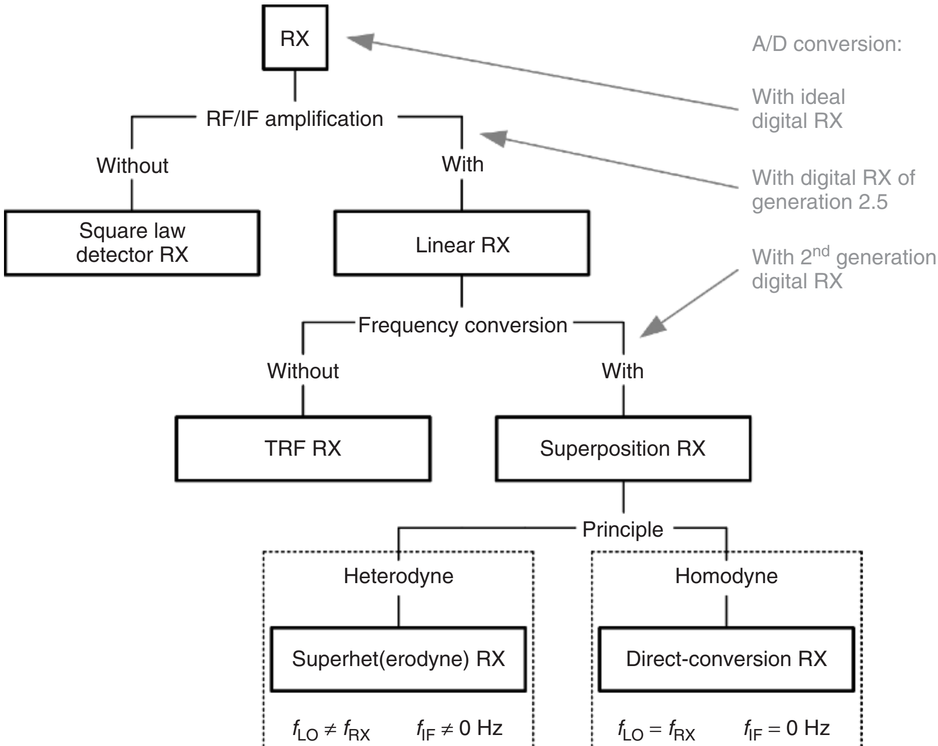

Figure I.20 Survey of possible receiver designs. The various concepts differ fundamentally in their complexity and achievable properties (Part III).

The collective term *software-defined radio* includes all solutions having deficiencies in one or several aspects but which pursue the basic ideas and advantages of a software radio, while considering its technical and economic feasibility on the basis of the hardware available [10]. A/D conversion takes place as close to the antenna as possible (Fig. I.20).

Table I.1 reviews the advantages and disadvantages of current receivers.

I.3 Practical Example of an (All-)Digital Radio Receiver

Already in 1988 reference was made to the technology of fully digitized receivers [11]. The prognosis was made that ‘despite all optimism digital receivers of satisfactory quality will hardly appear on the market before the middle of the 1990s . . . ’. In fact it took even longer, since really usable chipsets have only been developed in the laboratories of various renowned semi-conductor manufacturers within the last few years. The concept (Section I.2.4) of an all-digital receiver (ADR) outlined above will now be described in more detail. State-of-the-art high-quality professional receivers covering a frequency range up to 30 MHz are represented by the first commercially available units, like the ADT-200A from the Swiss engineering consultants Hans Zahnd (Figs. I.21 and I.22) or the MSN-8100-H from Thales Communications, developed for tactical marine communication (both

Table I.1 Principle-related advantages and disadvantages of today's receiver concepts according to [12]

Advantages	Disadvantages
Single-conversion superhet	
+ Most common receiver architecture	– Image signals inherent to the operating principle
+ Good selectivity	– Spurious signal reception
+ Least distortion and highest dynamics achievable with single heterodyning	– IF filter usually not integratable
Low-IF superhet	
+ Can be integrated in monolith	– Image frequency rejection is very sensitive to tolerances
	– Emission of LO injection signal
Multiple-conversion superhet	
+ Excellent receiving characteristics	– Very demanding in design, energy consumption, and number of components
+ Best receiver concept, since a partly digital system	– IF filter can usually not be integrated
Direct mixer	
+ Requires relatively few components	– Emission of LO injection signal
+ Can be entirely integrated in monolith	– Limitations due to inherent parasitic coupling and non-ideal components
+ Potentially low energy consumption	– Very low dynamics
Digital receiver >2nd generation	
+ Can also be integrated in monolith	– Very high energy consumption
+ Flexibly adaptable to changing receiver requirements	– Requires extremely fast linear A/D conversion
	– Requires significant computing power for receiving algorithm
	– Limited dynamics

systems contain an additional digital transmit path, so that the complete unit may correctly be called an all-digital transceiver (ADT)). Especially tailored to the needs of modern radio monitoring (Section II.4) is the EM510 model of Rohde&Schwarz. Controlling the DRM emissions (Section II.6) in full conformity is a feature of model DT700 from the Fraunhofer Institute for Integrated Circuits (Fig. II.47). In these units, an A/D converter samples the sum signal over the receiving range using a sample frequency of more than double the highest possibly frequency received and forwards the information as a parallel bitstream to the signal processing circuitry. Signal processing is all digital and software-controlled. For this the architecture of the homodyne receiver (Section I.2.3) is particularly advantageous [2], since it performs the main selection with comparatively few arithmetic operations. Equipment of this type is casually dubbed *direct receiver* by analogy to direct mixer receivers with their conventional circuitry and because they sample the RF receiving band directly without any conversion.

Figure I.21 ADT-200A is a first prototype of an almost fully digitized radio receiver (Digital RX of generation 2.5), designed as a stand-alone unit. Using a 14 bit A/D converter with a signal-to-noise ratio of 74 dB above half the Nyquist bandwidth of 36.86 MHz, in combination with the subsequent decimation it achieves a dynamic range (Section III.9.7) obtained so far only in high-end multiple-conversion superhets. A high-performance signal processor of the latest generation from Analog Devices having a processing power of up to 2 billion instructions per second performs the actual signal processing. (Company photograph of Hans Zahnd engineering consultants.)

I.3.1 Functional Blocks for Digital Signal Processing

The entire frequency range from DC to 30 MHz is fed to an A/D converter via a steep low-pass filter with a limit frequency of 30 MHz (Fig. I.19). The task of the low-pass filter is to prevent frequencies above half the sampling rate (32.5 MHz in this example) from reaching the A/D converter. The A/D converter is the link between analog and digital signal processing. This block essentially determines the receiver properties and should therefore be given special care! To achieve the high performance of a multiple-conversion receiver (like, for example, those of the 95S-1A from Rockwell-Collins or the TMR6100 from Thales Communications) in analog design up to the last IF stage would require a converter of at least 17 bits. The first 12 bit converters having a suitable speed became available in the year 2000. With the ADS852, Burr Brown was one of the first manufacturers to offer a 12 bit converter with a sampling rate of 65 mega-samples per second (MS/s) that was of high quality and still affordable [13]. With the AD6645, Analog Devices offered an improved component featuring 80 MS/s or 105 MS/s [14] and also includes a 16 bit converter, the AD9446 [15], in its sales program. The Linear Technology model LTC2208 is available in versions with 14 bit or 16 bit resolution and 130 MS/s [16]. What are the receiver characteristics that can be expected from such components?

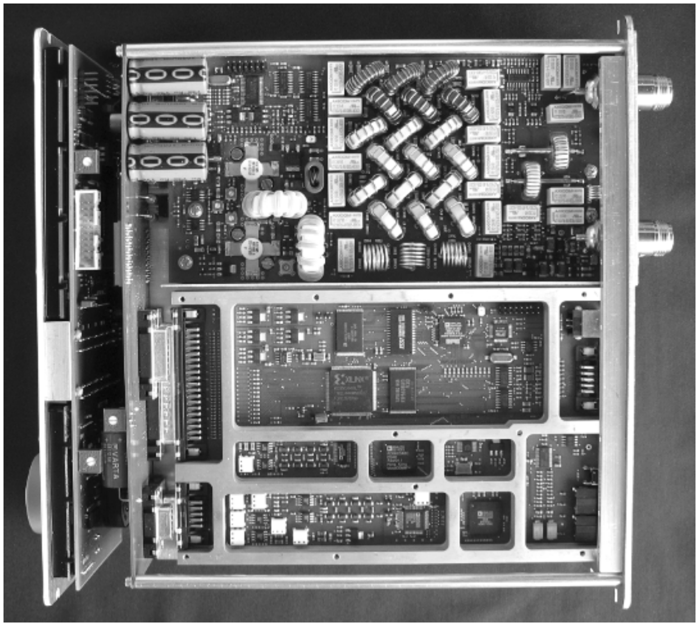

Figure I.22 Hardware layout of the communications receiver ADT-200A shown in Figure I.21. The lower section illustrates the module performing the digital signal processing (Fig. I.29). The upper section shows the 50 W HF transmit output stage. (Company photograph of Hans Zahnd engineering consultants.)

I.3.2 The A/D Converter as a Key Component

An ideal A/D converter is capable of splitting the input signal into 2^n equal voltage components, thus a 14 bit converter with an input voltage range of, for example, 1 V can convert this range into 2^{14} portions of 61 µV each. Any values in between are rounded off. Rounding errors cause noise, the so-called quantization noise. The theoretically possible signal-to-noise ratio (Section III.4.8) for sinusoidal signals is

$$SNR = Bit_{\text{spec}} \cdot 6.02 \text{ dB} + 1.76 \text{ dB} \tag{I.2}$$

where

SNR = signal-to-noise ratio of an ideal A/D converter, in dB
Bit_{spec} = specified resolution of the A/D converter, in bits

According to this equation the signal-to-noise ratio of the 14 bit converter considered would be

$$SNR = 14 \text{ bit} \cdot 6.02 \text{ dB} + 1.76 \text{ dB} = 86 \text{ dB}$$

In reality it is not possible to approach this ideal value. The data sheet for this converter specifies $SNR = 75$ dB. This first impression is not very encouraging, since the noise level of 75 dB below 1 V corresponds to a voltage of 178 µV or S9 + 11 dB (Section III.18.1).

However, this noise level is in relation to the entire bandwidth of 32.5 MHz (the Nyquist bandwidth). The A/D converter generates an enormous bitstream of

$$14\ \text{bit} \cdot 65\ \text{MS/s} = 0.91\ \text{GBit/s}$$

This includes the entire receive signal contents, from DC to 30 MHz! In fact, however, only a very narrow portion of it, the receive bandwidth (Section III.6.1), is of interest for, for example, the demodulation of an SSB signal. The huge amount of data needs to be reduced. In signal processing the reduction of the sampling rate is called decimation (see Fig. I.24) and is performed by a special digital filter [17] that averages the signal values of a certain number of sample values and forwards them with a reduced number of (combined) sample values to the subsequent decimation stage. Averaging reduces the quantization noise, resulting in a process gain:

$$G_{\text{dB p}} = 10 \cdot \lg\left(\frac{f_s}{2 \times B_{-6\,\text{dB}}}\right) \tag{I.3}$$

where

$G_{\text{dB p}}$ = process gain figure by decimation, in dB
f_s = sampling rate of the A/D converter, in S/s
$B_{-6\,\text{dB}}$ = receive bandwidth (−6 dB bandwidth) of the receive path, in Hz

With a receive bandwidth of 2.4 kHz, as is common for demodulating class J3E emissions, and a sampling rate of 65 MS/s the resulting process gain figure is

$$G_{\text{dB p}} = 10 \cdot \lg\left(\frac{65\ \text{MS/s}}{2 \cdot 2.4\ \text{kHz}}\right) = 41.3\ \text{dB}$$

This reduces the initially calculated noise floor from 178 µV to 1.53 µV. To achieve the usual value of the input noise voltage (Section III.4.7) of 0.2 µV EMF, it is necessary to include an upstream preamplifier. In fact, high intermodulation immunity (Section III.9.6) is possible only without or at most with low RF preamplification. This remains the weakest point of this concept.

The 14 bit A/D converter LTC2208 features a third-order intercept point (Section III.9.8) as high as 47 dBm, which can only be achieved with a high input noise level (Section III.4.2) of almost 30 dB noise figure (Fig. I.23). To obtain a receiver noise figure of $F_{\text{dB}} = 12$ dB by using a preamplifier ($F_{\text{dB}} = 6$ dB) would require a high amplification of 19 dB. Despite this high amplification a total intercept point ($IP3_{\text{tot}}$) of more than 25 dBm is possible, provided that the preamplifier alone has an output $IP3$ of more than 50 dBm. These high demands can be somewhat relaxed if the A/D converter is accessed by an impedance transformer. The input impedances of the various blocks are between 100 Ω and 800 Ω. The amplification can be reduced by 12 dB, which under favourable conditions may result in an $IP3$ of more than 30 dBm. However, the intermodulation response of an A/D converter cannot be compared to that of analog non-linear circuits. Section III.9.5 gives a more detailed description.

The theoretically achievable properties of various A/D converters as determined by the methods of calculation described are shown in Table I.2. The values stated have been confirmed to a large extent by practical tests.

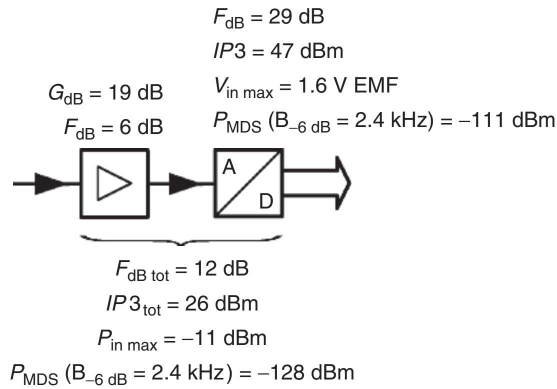

Figure I.23 Characteristic properties of the components for the calculation described and the resulting overall parameters.

Owing to the relatively high amplification the maximum signal level at the receiver input is reduced to the problematic value of less than −11 dBm. For LW/MW/SW reception even higher (sum) receive levels may be available from high-performance antennas [18]. The use of an attenuator from 0 to 25 dB controlled by the AGC (Section III.14) is an effective counter-measure. This shifts the dynamic range (Section III.9.7), which causes a response to the different signal strengths of the receiving bands. (Such a measure for signal conditioning has already been described in [19]. It stipulates that a switchable level attenuator be inserted before the analog receive section (the RF frontend) and a control amplifier be introduced between the analog receive section and the converter. The concept was initially designed for a second-generation digital receiver. The attenuator is controlled at the input of the control amplifier by the signal level. This arrangement is said to provide optimum utilization of the dynamic range of the A/D converter and a high immunity against overloading.) In the unit shown in Figure I.21 the automatic inclusion of such an attenuator complies with the following principle:

- The sum signal from the A/D converter is monitored at the input of the digital down converter (Fig. I.24). If the peak value exceeds the value 1 dB below the overload point several times within 1 second, the warning 'Intermodulation!' is displayed.
- One second later, the attenuation is incremented by 5 dB.
- If the sum signal remains 8 dB over the overload point for at least 5 seconds the attenuator switches back one increment (5 dB) until the originally preset value is restored.

Since the indication of the relative strength of the receive signal compensates the values of the attenuator and preamplifier, the process remains unnoticed by the operator. The sensitivity (Section III.4) of the receiver of course decreases with an increase in the attenuator damping, but in most cases remains below the external noise received [18].

A significant improvement of the large-signal properties (Section III.12) can be obtained by means of sub-octave filters, such as used in conventional receivers. Such filters reduce

Table I.2 Calculated parameters (Part III) of digital receivers using the components described

	AD6645	AD9446	LTC2288
Sensitivity			
Max. input voltage (= 0 dBc)	2.2 V_{pp}	3.2 V_{pp}	2.25 V_{pp}
Recommended source impedance (Z)	800 Ω	800 Ω	100 Ω
Max. input level ($P_{\text{in max}}$) with matched Z	−1.2 dBm	2.0 dBm	8.0 dBm
Signal-to-noise ratio***	75 dB	81 dB	78 dB
Noise level in 1st Nyquist band	−76.2 dBm	−79 dBm	−70 dBm
Process gain figure ($G_{\text{dB p}}$)	41.3 dB	41.3 dB	41.3 dB
Minimum discernible signal ($P_{\text{MDS A/D}}$) at A/D converter	−117.5 dBm	−120.3 dBm	−111.3 dBm
Noise figure ($F_{\text{dB A/D}}$) of the A/D converter	22.7 dB	19.9 dB	28.9 dB
Preamplification figure ($G_{\text{dB preamp}}$) for $F_{\text{dB tot}} = 12$ dB**	11.9 dB	9.1 dB	18 dB
Minimum discernible signal (P_{MDS}) of overall RX	−128 dBm	−128 dBm	−128 dBm
Operational sensitivity at 50 Ω, 10 dB $(S+N)/N$	0.28 µV	0.28 µV	0.28 µV
Dynamic range of preamplifier at a receive bandwidth $B_{-6\text{ dB}} = 2.4$ kHz	114.9 dB	120.9 dB	118 dB
Third-order intercept point ($IP3$)			
Intermodulation ratio ($IMR3_{\text{A/D}}$) of the A/D converter at −7 dBc***	−90 dBc	−96 dBc	−93 dBc
$IP3_{\text{A/D}}$ of the A/D converter*	36.8 dBm	43.0 dBm	47.5 dBm
Required output $IP3_{\text{preamp}}$ of the preamplifier	45 dBm	45 dBm	50 dBm
$IP3$ of the overall RX	24.3 dBm	31.8 dBm	24.1 dBm
Maximum intermodulation-limited dynamic range (ILDR) of the overall RX			
ILDR = $2/3 \cdot (IP3 - P_{\text{MDS}})$	101.6 dB	106.6 dB	101.5 dB

*$IP3_{\text{A/D}} = (P_{\text{in max}} - 7\text{ dB}) + IMR3_{\text{A/D}}/2$.
**$G_{\text{preamp}} = (F_{\text{A/D}} - 1)/(F_{\text{tot}} - F_{\text{V}})$; $F_{\text{dB V}} = 6$ dB; (F: numerical value, not in dB).
***according to datasheet [14], [15], [16].

Table I.2 contains the theoretically achievable properties of a digital RX of generation 2.5 with commercially available A/D converters with 65 MS/s sample rate. Not taken into consideration are the attenuation through upstream filters and the influence of sideband noise (Section III.7.1) of the reference clock oscillator on the noise factor of the converter. It is particularly important that the preamplifier is correctly matched to the converter input, which is usually of high resistance. Since the manufacturer of the LTC2208 recommends a source impedance of 100 Ω, this will perform rather poorly.

the process dynamics, since the difference between the weakest and the strongest signals (or the sum voltage) within the passband of the respective bandpass filter is lower than the entire short-wave range.

An important and, at the same time, critical parameter of any A/D converter is the spurious-free dynamic range (SFDR). This defines the ratio of the (unwanted) signal mix of higher order to the maximum input signal. These mixing products are caused by interference between the input signal and the sampling frequency f_s, whereby products

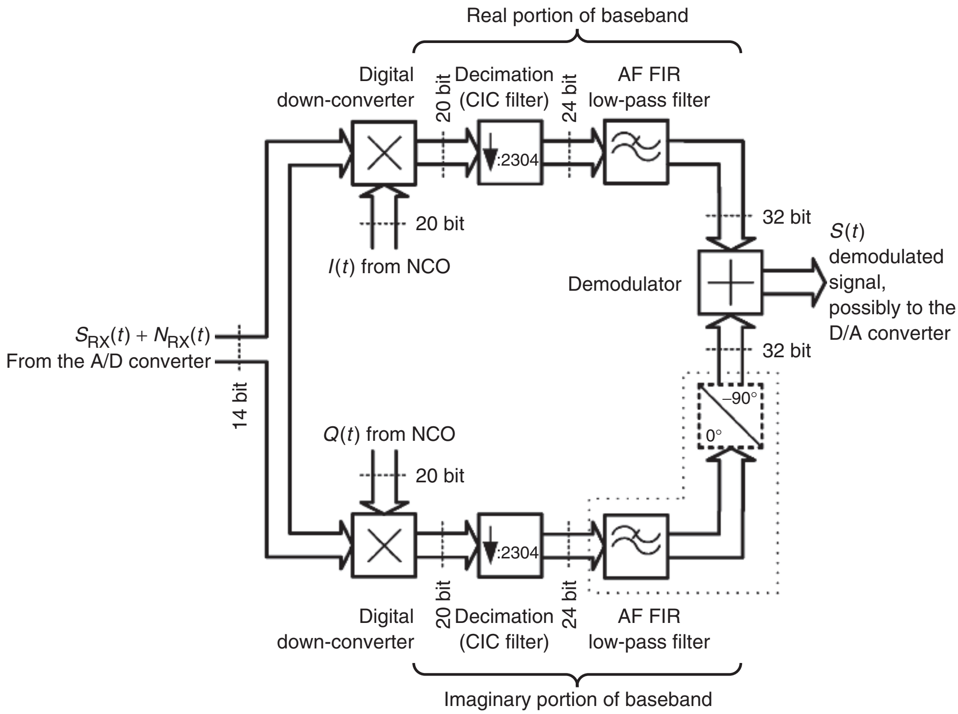

Figure I.24 Design of IF zero mixing with demodulation in a fully digitized mode. Decimation for reducing the sampling rate, without which meaningful post-processing by means of DSP would not be possible, is done by a CIC (cascaded integrator comb) filter. SSB demodulation requires an additional phase shifter in the Q path.

larger than $f_s/2$ are the result of so-called aliasing in the range between DC and $f_s/2$ (see also Fig. I.32). The datasheets for modern 14 bit A/D converters specify an SFDR of more than 100 dB.

Even better results can be achieved by using specific sigma delta A/D converters [9], which perform noise shaping at a sufficiently high oversampling rate [20]. The higher the order of the sigma delta A/D converter the more quantization noise is shifted out of the desired frequency range without the need to increase oversampling. Their use in digital receivers has been investigated for some time [21, 22].

I.3.3 Conversion to Zero Frequency

Selecting the desired signal is similar to selecting with conventional receivers, that is by mixing (Section V.4) with the signal from a local oscillator, utilizing the principle of direct-conversion receivers. The local oscillator is tuned exactly to the carrier of the signal to be received. This principle corresponds to that of a synchronous receiver (Fig. I.15) for

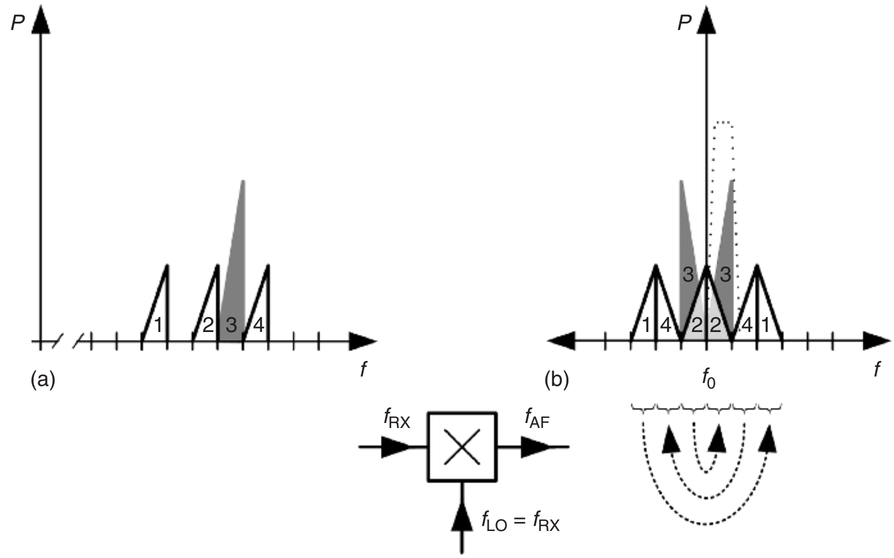

Figure I.25 Overlapping with direct mixer caused by mixing at zero position. Graph a) shows a possible signal scenario in the RF frequency band, and graph b) shows the same after shifting to zero position by mixing for the subsequent demodulation. Interference signals 1 and 2 appear mirrored about f_0. This presents no problem for AM, since both sidebands are symmetrical to the carrier and have identical information contents. However, with SSB this means that interference signal 2, directly adjacent to the desired signal 3, is displayed in the reverse position.

AM, mixing both sidebands of the AM signal in the audio frequency region. The carrier is therefore exactly at zero frequency. This causes the lower sideband of the AM signal to be in a negative frequency range. It can be demonstrated mathematically that the signal on the negative frequency axis is folded around the zero frequency to the positive frequency axis (Fig. I.25). For AM this presents no problem, since the two sidebands are symmetric to the carrier and have identical information contents. With SSB this situation is different; this process causes a reversal of the interference signal immediately adjacent to the useful signal in the basic channel. This problem must also be solved with the conventional analog direct-conversion receivers. The DC receiver uses two mixers controlled by a quadrature LO signal of 0° and 90° phases, as described in Section I.2.3. This produces a real component and an imaginary component (also called I channel and Q channel), which are fed separately to a low-pass filter (Fig. I.13). In order to suppress the unwanted sideband, it is necessary to shift the phase of the Q path again by −90° after demodulating the SSB (Fig. I.14).

In SSB modulators this method is also known as the phase method. This principle is easily explained mathematically on the basis of Figure I.24. It can be shown that the interference signal 2, which in Figure I.25 overlaps the useful signal 3 (upper sideband), is in fact suppressed. Quadrature mixing of the useful signal $S_{RX}(t) = \sin(\omega_3 \cdot t)$ and the interference signal $N_{RX}(t) = \sin(\omega_2 \cdot t)$ with the LO frequency ω_{LO} produces the following products behind the low-pass filter in the I path or behind the phase shifter in the *Q* path:

$$\begin{aligned} S_I(t) &= \sin((\omega_3 - \omega_{LO}) \cdot t) \\ S_Q(t) &= \cos((\omega_3 - \omega_{LO}) \cdot t - 90^\circ) \end{aligned} \tag{I.4}$$

where
$S_I(t)$ = real component of the useful signal at time (t), in V
$S_Q(t)$ = imaginary component of the useful signal at time (t), in V
ω_3 = angular frequency of the useful signal, in rad/s
ω_{LO} = angular frequency of the LO injection signal, in rad/s
t = considered time, in s

$$\begin{aligned} N_I(t) &= \sin((\omega_2 - \omega_{LO}) \cdot t) \\ N_Q(t) &= \cos((\omega_2 - \omega_{LO}) \cdot t + 90^\circ) \end{aligned} \tag{I.5}$$

where
$N_I(t)$ = real component of the interference signal at time (t), in V
$N_Q(t)$ = imaginary component of the interference signal at time (t), in V
ω_2 = angular frequency of the interference signal, in rad/s
ω_{LO} = angular frequency of the LO injection signal, in rad/s
t = considered time, in s

Substituting $(\omega_3 - \omega_{LO}) \cdot t = x$ and $(\omega_2 - \omega_{LO}) \cdot t = -y$ ($-y$ is negative when the time is positive, because $f_2 < f_0$), it follows that

$$\begin{aligned} S_I(t) &= \sin(x) \\ S_Q(t) &= \cos(x - 90^\circ) = \sin(x) \\ N_I(t) &= \sin(-y) = -\sin(y) \\ N_Q(t) &= \cos(-y + 90^\circ) \ = \sin(y) \end{aligned}$$

Finally, the summation of *I* path and *Q* path results in the AF signals:

$$S(t) = S_I(t) + S_Q(t) = 2 \cdot \sin((\omega_3 - \omega_{LO}) \cdot t) \quad (\rightarrow \text{useful signal with double amplitude})$$

$$N(t) = N_I(t) + N_Q(t) = 0 \quad (\rightarrow \text{interference is suppressed})$$

It can be seen that the condition $N(t) = 0$ is met only if the two components $N_I(t)$ and $N_Q(t)$ have exactly the same amplitude and a phase of 180°. A deviation of only 0.1 dB in the amplitude or 1° in the phase decreases the suppression of the interfering sideband to only

45 dB. With conventional analog signal processing in a direct mixer, sufficiently small tolerances can be obtained only with complex circuitry and arduous tuning, especially when covering a wide frequency band.

I.3.4 Accuracy and Reproducibility

The strong points of digital processing are high accuracy and reproducibility. When using digital circuits for the functional blocks of local oscillator, mixer, filter, and phase shifter only the resolution (number of bits) affects the accuracy. With a 24 bit DSP an attenuation figure of the unwanted sideband of more than 100 dB can be achieved. In fact, floating point processors for real-time processing are available with 32 bits and more [23]. Figure I.26 shows the passband characteristic of such an SSB filter. Without the complex signal processing described above, there would virtually be a second receive channel. The filter was designed in the finite impulse response (FIR) structure with 256 taps according to [24]. It shows a very good shape factor (see Fig. III.42) of better than 1.2 (−6 dB/−60 dB) and has a constant phase of 0° for the I channel and 90° for the Q channel (Fig. I.24) in its passband. For this reason, the 90° phase shifter of the Q channel can be omitted.

The frequency response in Figure I.27 corresponds to a 7 kHz filter designed and optimized for AM reception. With little frequency separation from the limit frequencies the attenuation figure for the cutoff region is already approximately 105 dB.

Another advantage of the digital solution is the fact that such filters can be produced in large quantities with high precision, while there is no aging or drift with temperature variations.

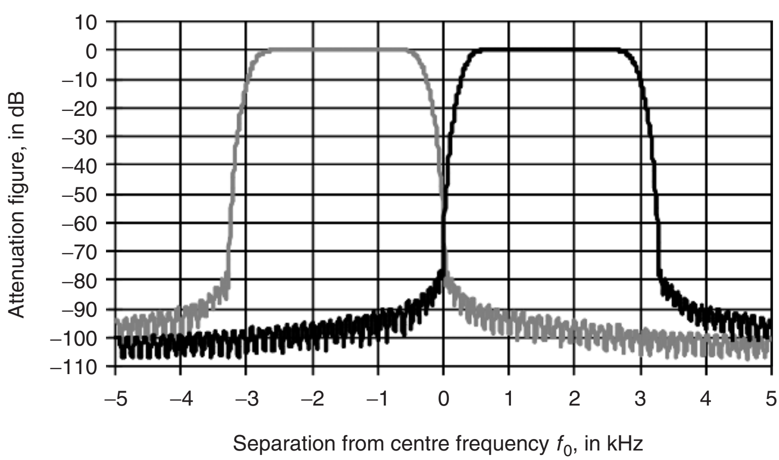

Figure I.26 Measured frequency response of a 2.7 kHz filter for receiving class J3E emission for demodulating the upper or lower sideband. The passband characteristics are fully symmetrical and provide a close-in selectivity (Section III.6) that is clearly above 90 dB already in low separation to the limit frequency.

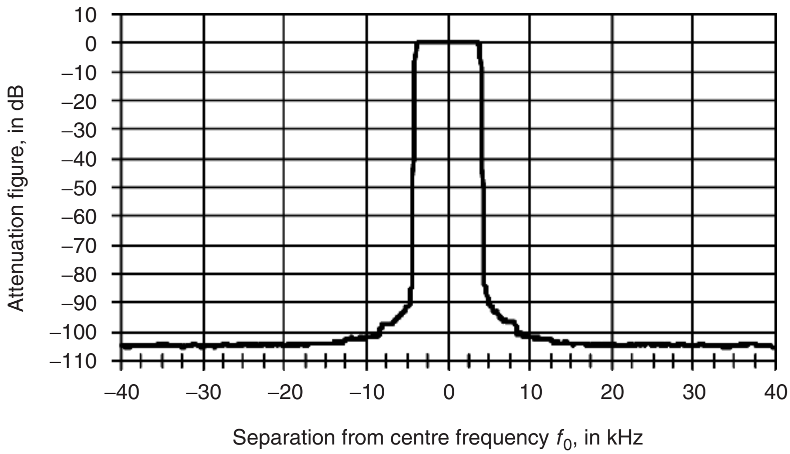

Figure I.27 Measured frequency response of a 7 kHz filter for receiving class A3E emission. No selection gaps were found. The shape factor is below 1.2 and the effective passband ripple (Fig. III.42) is in the range of about 0.3 dB.

I.3.5 VFO for Frequency Tuning

Another important component is the numerically controlled oscillator (NCO). Its construction follows the design of DDS generators (Section I.2.2) used in newer transceivers, but in contrast to these does not require a D/A converter. DDS components with moderate resolution generate a high amount of spurious signals (Section III.7.2). They are not suitable for use as variable frequency oscillators (VFOs), even though the frequency range would be suitable. This is because the low resolution of the D/A converter, ranging from only 8 to 12 bits while signal processing, is carried out with 32 bits. In an all-digital receiver there is no need to change to analog signals. The mixer, which is actually a digital down-converter (DDC), can be controlled with a resolution of 20 bits without problem. A DDS generator with a 10 bit D/A converter has a spurious signal ratio of about 55 dB. Owing to its doubling to 20 bits, a ratio of 110 dB is to be expected. Spurious signals are therefore negligible.

Figure I.28 illustrates the operating principle of the NCO. In the accumulator an increment is added to the value of the 32 bit wide sum register, and the resulting value is stored with every clock cycle in the sum register. The value increases linearly with every clock period until the register overflows at $\geq 2^{32}$. The result is a sawtooth signal having the frequency

$$f_{\mathrm{NCO}} = f_{\mathrm{cl}} \cdot \frac{s}{2^N} \tag{I.6}$$

where

f_{NCO} = NCO output frequency, in Hz
f_{cl} = clock frequency, in Hz
s = decimal value of the $(N-1)$ bit wide control increment, dimensionless
N = word length of the phase accumulator, in bits

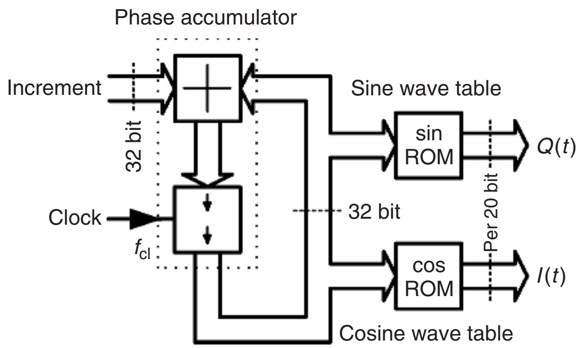

Figure I.28 Principle of the digital oscillator (NCO) for shifting the received signal by 0 Hz. The adder functions together with the sum register as a phase accumulator to which a fixed increment (the control word s) is added with a clock frequency of f_{cl}.

and an achievable frequency resolution [25] of

$$\Delta f_{NCO} = \frac{f_{cl}}{2^N} \tag{I.7}$$

where
Δf_{NCO} = achievable frequency resolution of the NCO, in Hz
f_{cl} = clock frequency, in Hz
N = word length of the phase accumulator, in bits

A table stored in a non-volatile memory (ROM – read only memory) is used for converting the sawtooth wave to a sine or cosine signal pattern in 2^{20} steps, which determine the possible phase resolution of the output signal:

$$\Delta \phi_{NCO} = \frac{2 \cdot \pi}{2^M} \tag{I.8}$$

where
$\Delta \phi_{NCO}$ = achievable phase resolution of the NCO, in rad
M = number of address bits of the ROMs, in bits

This essentially influences the number and spectral separation of the spurious signals from the useful signal [25].

With a clock frequency of 65 MHz and, for example, a decimal value of 231,267,470 as increment in the sum register the NCO generates an output frequency of

$$f_{NCO} = 65\text{ MHz} \cdot \frac{231{,}267{,}470}{2^{32\text{ bit}}} = 3.5\text{ MHz}$$

A deviation of 1 to 231,267,471 causes a frequency alteration of 0.015 Hz. This shows that it is possible to tune the frequency with a resolution as high as

$$\Delta f_{NCO} = \frac{65\text{ MHz}}{2^{32\text{ bit}}} = 15\text{ mHz}$$

and a phase resolution of

$$_{\Delta}\phi_{\text{NCO}} = \frac{2 \cdot \pi}{2^{20\ \text{bit}}} = 5.99\ \mu\text{rad} \mathrel{\widehat{=}} \frac{360^\circ}{2^{20\ \text{bit}}} = 0.000{,}343^\circ$$

Compared with the conventional analog design the digital design has the following additional advantages:

- The frequency achieved is as stable as the quartz crystal.
- Large frequency jumps can be made within microseconds and with extremely short transient periods (Section III.15) (for e.g. in spread-band technology applications).
- With well-considered dimensioning and the corresponding width of the ROMs the sideband noise, and therefore reciprocal mixing (Section III.7) is very low.
- There is not any response of the NCO to the receiver input, so that there is no stray radiation from the receiver (Section III.17).

I.3.6 Other Required Hardware

A possible circuit design for realizing the functional blocks described will be described based on the example of the unit shown in Figures I.21, I.22 and I.29. Analog Devices' IC AD6624 down-converter has been chosen for performing the functions of mixer, NCO, decimation and filtering [26]. In addition a digital signal processor is required. The same manufacturer's model ADSP-21362 already described is suitable for this purpose [23]. This DSP will perform the following functions:

- Filtering and near selection of the receive signal (several receive bandwidths from 50 Hz to 25 kHz, see Figs. I.26 and I.27).
- Measuring the receive signal voltage for the S meter (Section III.18.1) and the automatic gain control (AGC).
- IQ demodulation.
- Communication with the audio CODEC for the connection of loudspeakers, headphones or media for analog AF recording.
- Controlling the digital down-converter (configuration, frequency tuning, AGC).
- Communication with the operating unit or control device (PC or keyboard, display, or incremental encoder for frequency tuning).
- In addition, other functions, like adaptive noise suppression, modem functions, demodulation of coded modulation modes, and the functions of CW decoder or terminal node controller (TNC).

Figure I.30 shows a block diagram of the resulting all-digital VLF/HF communications receiver. The chipset consists of only four blocks. Almost unbelievable is the drastic reduction in the amount of hardware compared with a unit of conventional circuit design having the same characteristics.

The spurious-free dynamic range (Section I.3.2) of the AD6645 A/D converter used can be increased up to 100 dB by dithering (Section III.9.5). In the course of development the use of a noise source has been investigated. However, tests have shown that in

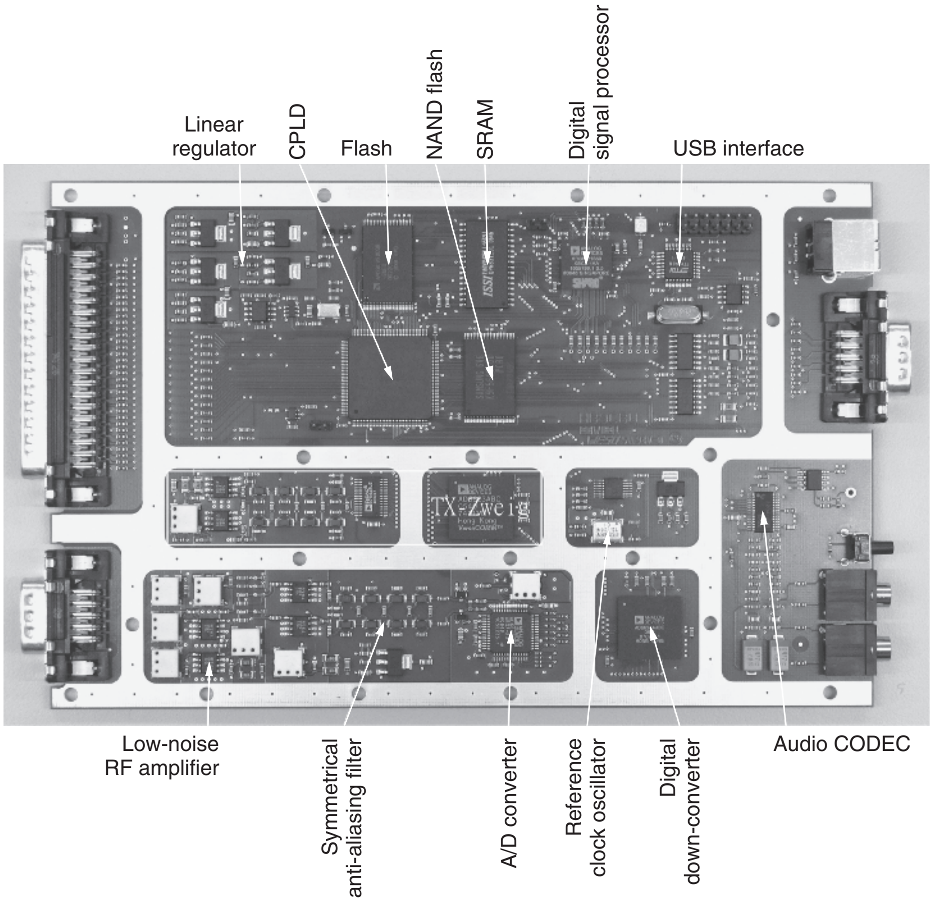

Figure I.29 The signal processor module of the ADT-200A (shown in Fig. I.21) in detail. (Company photograph of Hans Zahnd engineering consultants.)

practical applications there are always a sufficiently high number of stochastic interference signals, so that the additionally generated noise brings no further improvement except in intermodulation measurements (Section III.9.10) using the established two-tone measuring method (Fig. I.31).

I.3.7 Receive Frequency Expansion by Subsampling

The Nyquist sampling theorem specifies that a signal can be correctly reconstructed only if the sampling frequency (f_s) is at least twice as high as the highest frequency component of the sampled signal. If this condition is not met, aliasing of the frequencies above $f_s/2$ takes place in the region below $f_s/2$. For example, with $f_s/2 = 32.5$ MHz a receive frequency of 35 MHz will be superimposed on a receive frequency of 30 MHz. When increasing the frequency the process is repeated in segments of $f_s/2$. These segments are

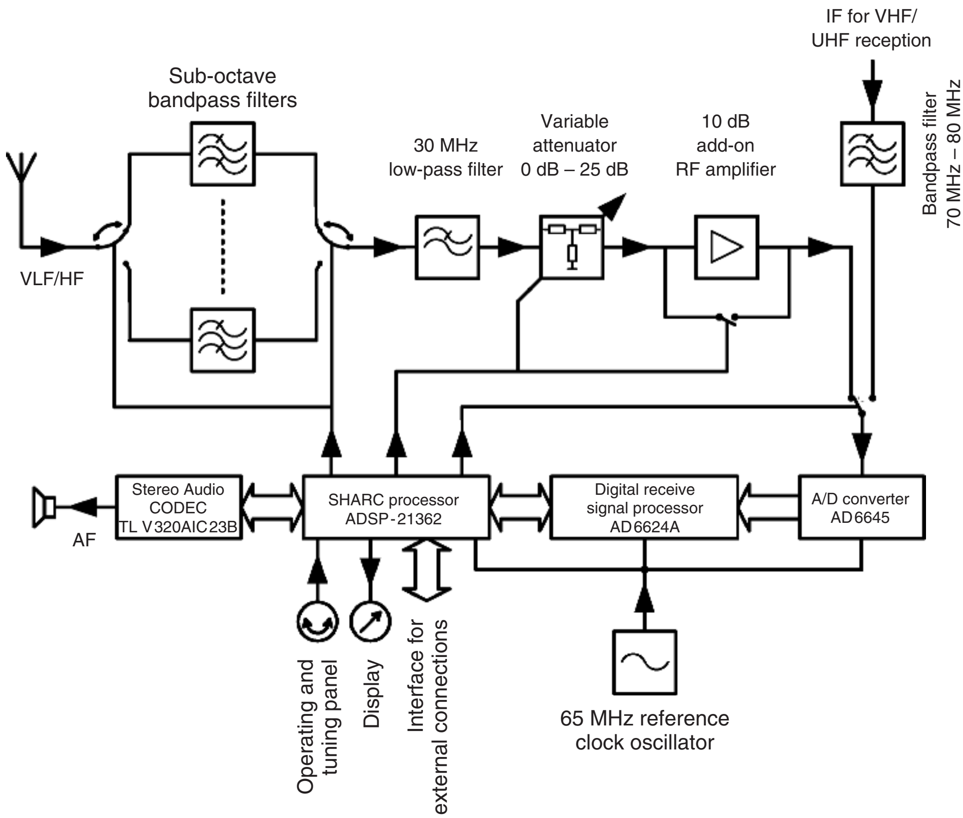

Figure I.30 Block diagram of the (fully) digitized radio receiver in the ADT-200A (Fig. I.21). The frontend, consisting of the bandpass filters and the 30 MHz low-pass filter, is constructed in keeping with present day technology. Signal processing is done by a chipset comprising four highly integrated components of the latest generation. In the VLF/HF receiving range the unit functions as a digital RX of generation 2.5 and for receiving frequencies in the VHF/UHF range as a second-generation digital RX by including subsampling (Section I.3.7).

called Nyquist windows (Fig. I.32). The phenomenon can be useful for receive frequency ranges above the sampling frequency, provided that they do not exceed the segment limits. A suitable A/D converter can use a low sampling frequency and still cover a range of several hundred MHz. This is called *subsampling*. The upper frequency limit is determined by the uncertainty in the sampling circuit of the A/D converter, which is called aperture jitter. Also, the phase noise (Fig. III.49) of the sampling frequency becomes more relevant with an increase in the signal frequency. (The implementation of sigma delta A/D converters for sampling such bandpass-filtered frequency ranges is advantageous [9]. How the bandpass subsampling [27] can be performed with a sigma delta A/D converter is investigated in [28].)

Generation of the intermediate frequency as shown in the block diagram in Figure I.30 for VHF/UHF reception is according to this principle. The IF is in the rather unusual range

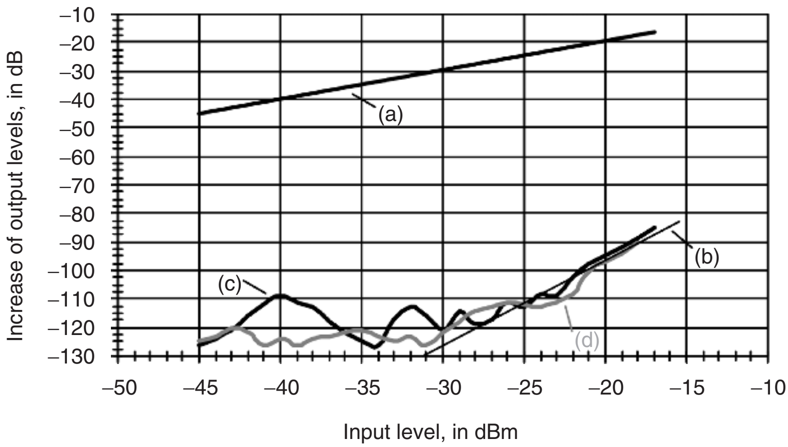

Figure I.31 Actual intermodulation response of the third order (Section III.9.2) as used in the all-digital radio receiver described. Curve a) represents the power of one of the excitation signals fed in. The level increase of an IM3 product without dithering is indicated by curve c), and with active dithering by curve d) (Section III.9.5). Especially with the received levels under normal operating conditions the dithering function brings a significant improvement in the intermodulation immunity (Section III.9.6). The IM3 increase expected by definition is shown in curve b) for comparison. (The increase in the input level of over −25 dBm is caused by intermodulation in the analog frontend of the unit. This is the reason for the level increase of 3 dB per 1 dB increase in the excitation signal.)

between 70 MHz and 80 MHz, which allows the use of a simple transverter with a fixed heterodyne frequency, thus providing high image frequency rejection (Section III.5.3) even with moderate input selection properties. If an IF signal is sampled by means of bandpass subsampling, the desired spectrum is mirrored at $f_s/4$ [9]. With subsampling the range 70 MHz to 80 MHz is shifted downward to 5 MHz to 15 MHz (Fig. I.32).

I.4 Practical Example of a Portable Wideband Radio Receiver

Possible implementations of modern wideband receivers covering a wide(r) receive frequency range differ from previous designs in several respects, both in regard to the basic parameters covered (frequency range, demodulated class(es) of emission, intended use, etc.) and in the specific circuit layout. Most of these are designed as multiple-conversion superhets (Section I.2.2) with a high first intermediate frequency. Depending on the required technical properties and, particularly their flexibility in terms of equipment configuration, they often operate with digital signal processing from the IF stage further downstream with a lower frequency.

To discuss this present state-of-the-art design, the unit shown in Figure I.33 can be taken as an example [29], and [30]. This unit covers the receive frequency range continuously from 9 kHz to 7.5 GHz. Despite its small dimensions it also offers a wide range of functions while being highly mobile at the same time. The moderate power consumption allows prolonged line-independent operation from a rechargeable lithium ion battery pack.

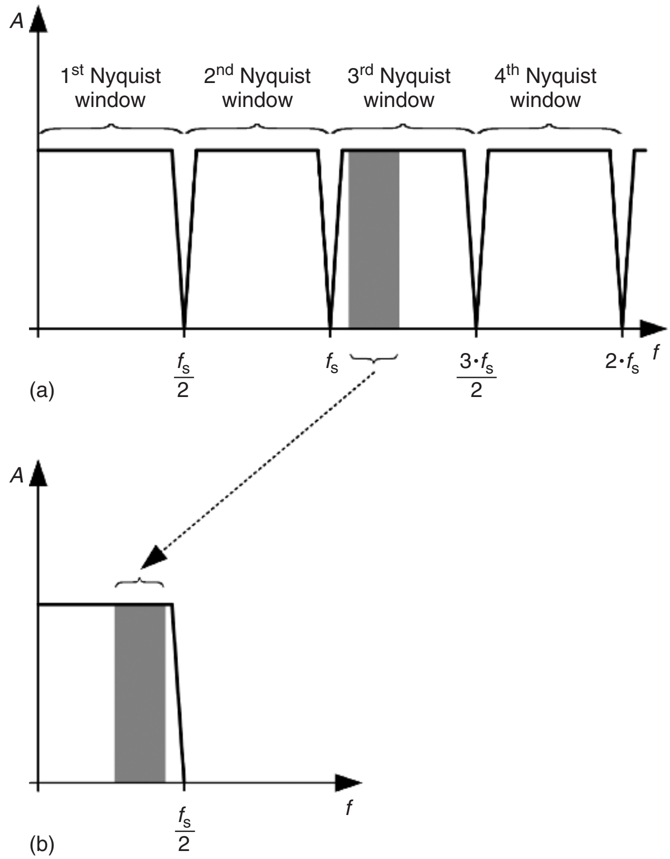

Figure I.32 Expanding the input frequency range by subsampling. Graph (a) shows the Nyquist windows and graph (b) the shifted frequency segment that was initially above the receive frequency range. The subsampling frequency range fed to the A/D converter must have a bandwidth of $<f_s/2$.

Its graphic display of frequency occupancy and analysis of receive signals extends the functionality compared with units designed for demodulation alone.

I.4.1 Analog RF Frontend for a Wide Receive Frequency Range

The signals received at the antenna port pass a low-pass filter, limiting the frequency spectrum for further processing to the filter's limit frequency of 8 GHz. Subsequent signal processing is carried out in three different paths, depending on the receive frequency selected (Figs. I.34 and I.35):

- Signals in the frequency range between 9 kHz and 30 MHz are fed via a 30 MHz low-pass filter and a HF preamplifier directly to the A/D converter. In a multi-functional portable unit of the given dimensions (Fig. II.41) a selective frontend selection of this

Figure I.33 PR100 allows continuous tuning for receiving in a frequency range from 9 kHz to 7.5 GHz with a noise figure (Section III.4.2) of less than 20 dB across the entire receive frequency range. The portable unit has a weight of 3.5 kg, including the battery pack (see Fig. II.41). Analysis of the received signals can also be carried out on the 6.5 inch colour display. Information can be stored on a built-in SD memory card without external accessories. For special applications all receiver functions can be remotely controlled via a LAN interface. The unit can be upgraded optionally for use as a single-channel direction finder in the range from 20 MHz to 6 GHz (Company photograph of Rohde&Schwarz).

frequency spectrum with its high levels is not possible. With low receive frequencies below 30 MHz the unit therefore operates as a direct receiver (Section I.3).

- In the frequency range between 20 MHz and 3.5 GHz the signal passes several automatically activated bandpass filters of moderate quality under operating conditions (Section III.11) or a high-pass filter and a subsequent RF preamplifier. For high-level input signals an attenuator allows bypassing of the preselector and RF preamplifier to prevent the generation of high sum signals and to ensure that the first IF stage operates in the linear region of its dynamic range. It forms the front block of the IF-generating circuit to which the filtered or attenuated input spectrum is fed via a 3.5 GHz low-pass filter.
- Signals in the frequency range above 3.5 GHz to 8 GHz are fed to an IF generating circuit via a high-pass filter of 3.5 GHz limit frequency and an RF preamplifier followed by another 8 GHz low-pass filter.

The above-mentioned IF-generating circuitry converts the respective receive frequency band to three intermediate frequencies, of which the last analog third IF is 21.4 MHz (the analog unregulated frequency of 21.4 MHz is available at a BNC socket for external processing). Following this signal preparation the receive frequency range from 20 MHz to nearly 8 GHz can now be fed to the same A/D converter as the lower-frequency receive signals. At higher frequencies, the concept of the multiple-heterodyne receiver with A/D conversion after the third IF stage is therefore used (see Fig. I.16). To make these specific equipment parameters available [30], the subsequent stages effectively process signals only up to 7.5 GHz.

I.4.2 Subsequent Digital Signal Processing

From the A/D converter on, the signals conditioned as shown in Figure I.34 can be processed in the downstream functional blocks according to the principles outlined in Sections I.3.1 to I.3.6.

In order to use the portable wideband radio receiver shown in Figure I.33 for graphic evaluations and analyses in addition to demodulation, the signal path is divided after the A/D converter into two parallel branches (Figs. I.36 and I.37). This allows simultaneous demodulation, receive signal level measurement (Section III.18), and display of a spectral panorama [29]. These two branches are described in detail in the next two sections.

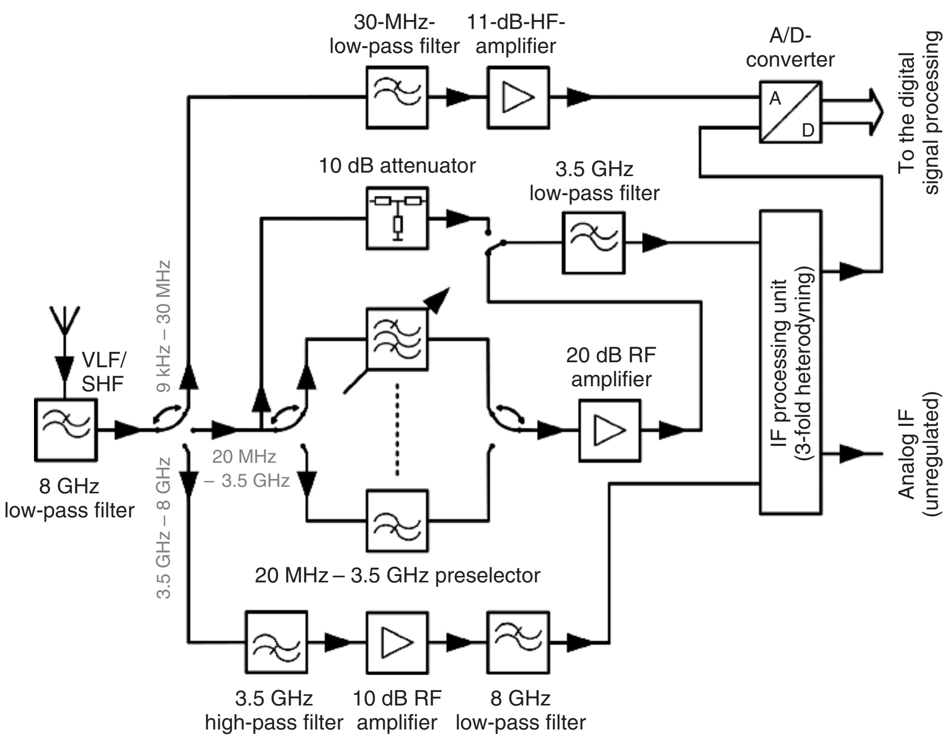

Figure I.34 Block diagram of the frontend up to the A/D converter of the PR100 (Fig. I.33). The receive frequency range is extended by a combination of second-generation digital RX (for receiving frequencies above 30 MHz) and of generation 2.5 (for frequencies below 30 MHz). For input signals >20 MHz the analog unregulated signal is available externally via the 41.4 MHz IF (subsequent digital processing paths are illustrated in Figures I.36 and I.37).

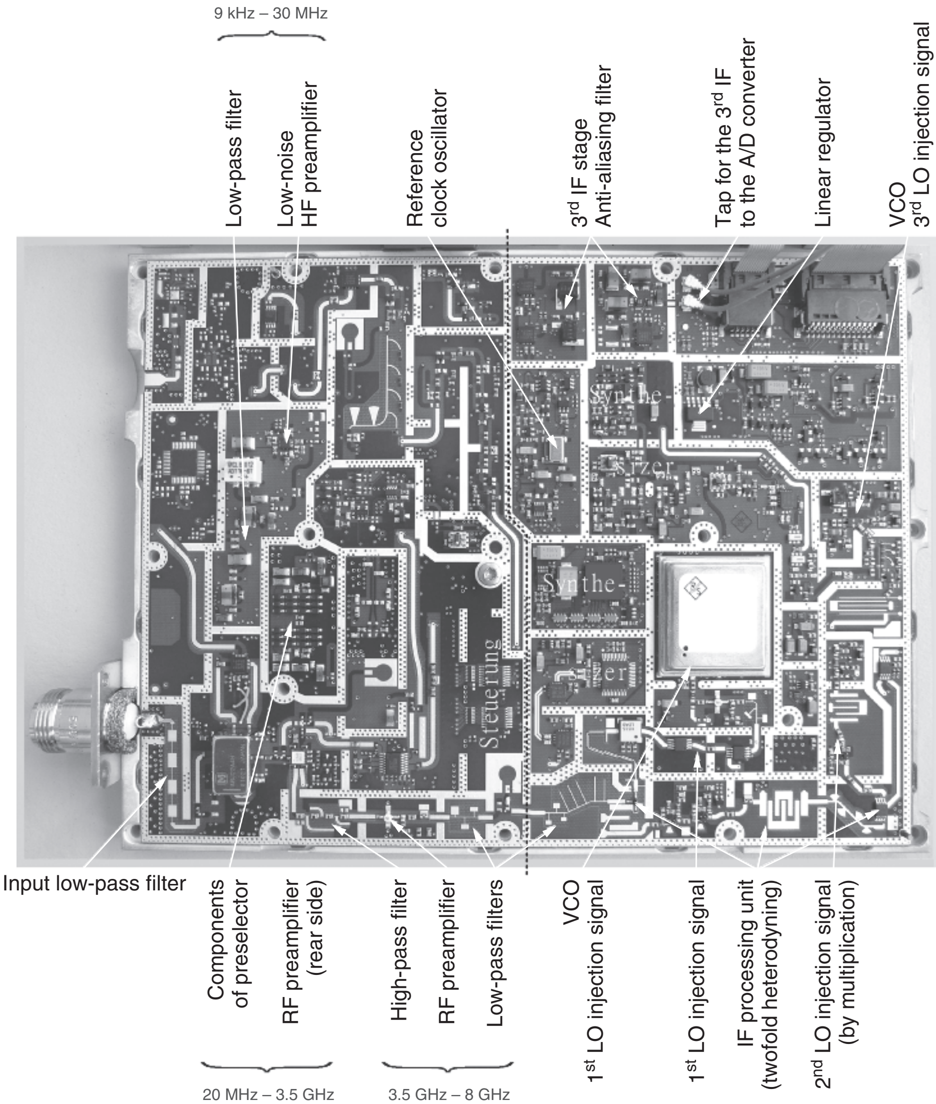

Figure I.35 Front detail of the analog RF frontend module of the PR100 (shown in Fig. I.33; see also Fig. I.34). The entire half to the left of the dotted line contains the preselection of the three receive paths, while the right half includes the frequency processing and the IF paths. (Company photograph of Rohde&Schwarz.)

I.4.3 Demodulation with Received Signal Level Measurement

The signal is prepared for demodulation or level measurement (Fig. I.36) by a digital down-converter (DDC) (Section I.3.5) and a digital bandpass filter. For the matched reception of different classes of emission and for an optimized signal-to-interference ratio in different receiving situations, the receiver offers the possibility of choosing from 15 digitally realized IF filter bandwidths from 150 Hz up to 500 kHz (in part depending on

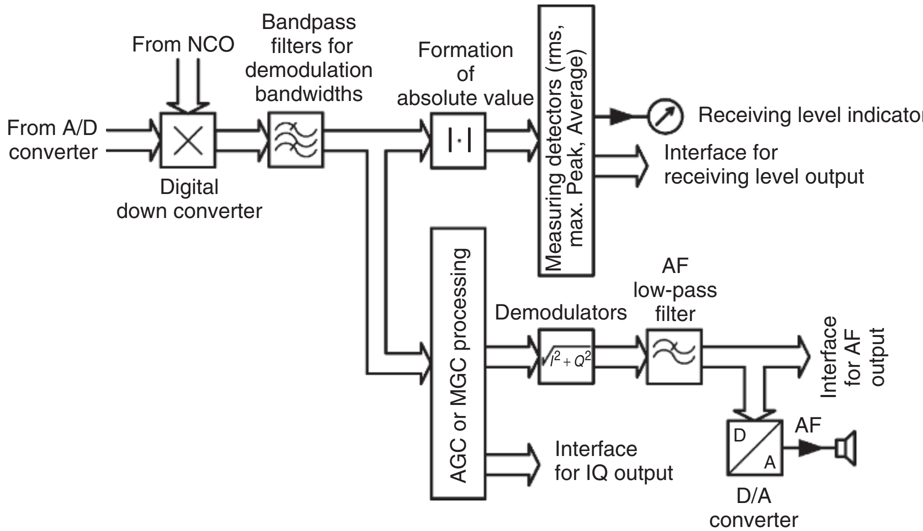

Figure I.36 Execution of the digitally based demodulation stage (lower signal path) and measuring the level of the receive signal (upper signal path) (the frontend up to the A/D converter is shown in Figure I.34).

the emission class selected). These can be selected independently of the display range and the resolution bandwidth of the spectral display described in Section I.4.4.

For demodulating analog signals the complex baseband data (Section I.2.3) are fed via the bandpass filter to the AGC or MGC stage (Fig. I.5). They are then subjected to the selected demodulation algorithm for A1A (CW), A3E (AM), B8E (ISB), F3E (FM), J3E (SSB, upper and lower sideband), or pulse. The results are in digital form and made available via the LAN interface. For loudspeaker output the digital audio data stream must be converted back to an analog signal.

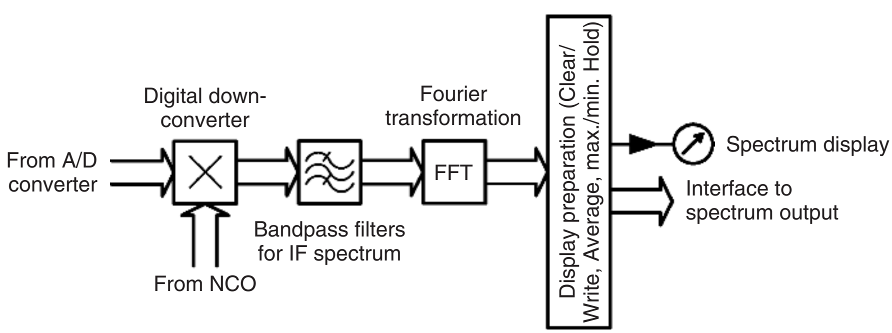

Figure I.37 Structure of the digital path for displaying the receive signal spectrum from the intermediate frequency by FFT analysis. Displaying the signal levels requires comprehensive logarithmic calculations of all the bins (the frontend up to the A/D converter is shown in Figure I.34).

After the AGC or MGC stage the complex IQ data (Section I.2.3) of the digital signals are directly available for further processing.

For the purpose of measuring the receive signal level, the signal strength is determined and the value assessed according to the measuring detector selected (rms, maximum peak, sample, average, as known from spectrum analyzers). The measured and evaluated levels are then available on the display and at the LAN interface. In addition, the unit is able to refer to a set of internally stored correction factors that enable measurement of the field strengths for known antenna factors (Section III.18).

I.4.4 Spectral Resolution of the Frequency Occupancy

The second signal path with DDC and digital bandpass filter is used for the calculation of the signal spectrum around the receive frequency in the FFT block (Fig. I.37) from the intermediate frequency. The bandwidth of the bandpass filter and, with it, the associated spectral span on the display can be selected by the user in the range from 1 kHz up to a maximum of 10 MHz.

The calculations based on the fast Fourier transformation (FFT) of the IF-filtered data stream (Section II.4.2) have a considerable advantage: the receiver sensitivity and signal resolution are clearly superior to those of a conventional analog receiver with the same spectral display span.

When selecting the setting of, for example, 10 kHz for sensitive signal reception, the following steps are performed in the course of the FFT calculation. Based on the finite steepness (Fig. III.42) of the IF filter, the decimated sampling rate (Section I.3.2) must be higher than the selected display width. This means that the quotient of the decimated sampling rate and bandwidth is >1 and represents a measure of the steepness of the IF filter (this may be seen as similar to the shape factor described in Section III.6.1). Its numerical value depends on the display range selected and may vary. For the 10 kHz display span, the constant is 1.28 and results in the necessarily decimated sampling rate of $10\,\text{kHz} \cdot 1.28 = 12.8\,\text{kHz}$. The FFT standard length n of the unit described in Figure I.33 is 2,048. The calculation (with a Blackman window) divides the frequency band of 12.8 kHz into 2,048 equidistant FFT lines (also called frequency lines or bin widths). Each of these FFT lines represents a quasi receive channel with a resolution bandwidth of

$$B_{\text{res}}(n) = \frac{f_{\text{s}}}{n} \tag{I.9}$$

where
B_{res} = resolution bandwidth of an FFT line, in Hz
f_{s} = decimated sampling rate prior to FFT analysis, in Hz
n = number of FFT lines, dimensionless

For the display range considered, this results in a resolution bandwidth of

$$B_{\text{res}}(n = 2{,}048) = \frac{12{,}800\ \text{Hz}}{2{,}048} = 6.25\ \text{Hz}$$

per frequency line, roughly effective as an equivalent noise bandwidth (Section III.4.4), and corresponds to

$$B_{\mathrm{dB\,N}} = 10 \cdot \lg\left(\frac{6.25\ \mathrm{Hz}}{1\ \mathrm{Hz}}\right) = 8\ \mathrm{dBHz}$$

This allows the determination of the minimum discernible signal (Section III.4.5) (the noise floor) of the spectral display using Equation (III.10):

$$P_{\mathrm{MDS}}(B_{-6\ \mathrm{dB}} \approx 6.25\ \mathrm{Hz}) = -174\ \mathrm{dBm/Hz} + 20\ \mathrm{dB} + 8\ \mathrm{dBHz} = -146\ \mathrm{dBm}$$

According to specification [30], with some receive frequencies the noise figure (Section III.4.2) of the receiver is below the value of 20 dB used for the calculation (in some cases below 10 dB), which suggests an even better sensitivity within this frequency ranges.

The minimum display range above 1 kHz results in the maximum sensitivity, while the widest range of 10 MHz produces the lowest sensitivity. The high spectral resolution of the FFT calculation shows that closely adjacent signals appear well separated in the spectrum displayed.

Prior to feeding the IF spectrum to the display or LAN interface, the type of display is prepared according to the user's specification (normal or clear/write, average, max. hold, min. hold) (as is also known from spectral analyzers).

For a survey over a wider spectral panorama several of the up to 10 MHz wide FFT display ranges can be combined on the frequency axis to form wide display ranges (so-called panorama scans, Figs. II.23 and II.24). In this operating mode the user can choose from 12 bin widths between 120 Hz and 100 kHz. Based on the selected bin width and the start and stop frequency settings, the required FFT length and the width of the frequency window for each individual viewing increment are determined automatically. However, the panorama scan must be stopped when operating the receiver in the listening mode [30].

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