

Chapter 2

2. POWER AMPLIFIER LINEARIZATION

Traditionally, constant envelope modulation schemes have been used in radio telecommunications because of their simplicity and robustness to amplitude errors. This made it possible to use high efficiency power amplifiers (PA), which are intrinsically very nonlinear devices, near the saturation region where the amplifier efficiency is at its peak.

However these modulation schemes are spectrally inefficient and the current trend is to improve the spectral efficiency or the number of bits transmitted per bandwidth by using some linear modulation scheme such as quadrature amplitude modulation (QAM). Alas, when driven through a nonlinear device the fluctuating envelope of the linear modulation schemes cause intermodulation products to appear around the signal band. This spectral spillage is effectively impossible to filter away and so can cause the amplified signal to exceed its allowed adjacent channel interference (ACI) limits.

To compensate these unwanted effects, various amplifier linearization techniques have been presented. Table 2-1 shows a comparison between three basic linearization techniques, namely, Cartesian feedback, feedforward and predistortion. The cancellation performance of the Cartesian feedback is good, but the bandwidth is narrow, making the technique unsuitable for very wideband systems. The feedforward, on the other hand, can be employed for wideband linearization, but, unfortunately, the system is extremely complicated, resulting in great power waste and large physical size. The third method, predistortion, is an optimal solution in terms of power added efficiency and physical size.

2.1 Feedforward

In the 1920s, H. S. Black invented two schemes for reducing amplifier distortion, namely, feedforward [Bla28] and negative feedback [Bla37]. Feed-

forward became forgotten in favor of the feedback technique, even though feedforward predated the latter by several years. Today, we are well aware of the limitations of feedback owing to the work of Nyquist and Bode. Feedback is limited by conditional stability and finite inter-modulation distortion (IMD) suppression, whereas feedforward is unconditionally stable and can, in theory, completely eliminate the IMD. However, Black himself noted that the key problem with his feedforward prototype was the primary reason for feedforward to be relegated to the background.

The feedforward prototype required perfect gain match in the different signal paths and Black reported that the gain of the amplifier had to be constantly re-adjusted. Another reason, at that time, was the simple fact that the complexity of the feedforward system compared with negative feedback was considered as a major disadvantage. However, as applications with higher frequencies and bandwidths appeared, the disadvantages of negative feedback became more apparent. This has to some extent led to a renaissance of the feedforward technique and it is nowadays considered to be one of the most established and approved methods, especially for wideband and multi-carrier systems [Ken91a], [Ken91b], [Mye94]. Otherwise, it has been used in many areas spanning from low frequency audio applications [Van80] to high-frequency CATV [Pro80] and microwave [Se71b] applications.

A block diagram of the feedforward system is depicted in Figure 2-1. All blocks operate at RF. The main amplifier, a nonlinear power amplifier, is fed directly with the source signal. The distortion generated by the amplifier is isolated in the signal cancellation loop by subtracting the source signal from the amplifier output. This signal is often referred to as the error signal. In the distortion cancellation loop, the error signal is finally subtracted from the amplifier output. For perfect signal and distortion cancellation, an attenuator and auxiliary amplifier are required in the signal and distortion cancellation loops, respectively. For high frequency applications, it is evident that the performance of this scheme in terms of obtaining perfect signal and distortion cancellation is not only dependent on the amplitude match but also on the phase/delay match along the parallel signal arms. In practice, fixed delays can be inserted as noted in Figure 2-1 to balance the delays in the co-arms that are primarily dominated by the amplifiers.

The effects of delay, phase and amplitude imbalances have been treated

Table 2-1. Comparison of Three Basic Linearization Techniques [Mad99]

Technique	Cancel- lation Per- formance	Bandwidth	Power Added Efficiency	Size	Suitability to Multicarrier
Feedback	Good	Narrow	Medium	Medium	Low
Feedforward	Good	Wide	Low	Large	High
Predistortion	Medium	Medium	High	Small	Medium

in several papers [Ste88], [Wil92a], [Par94], and [Mye94]. As an example, to obtain 25dB suppression of the amplifier distortion, an amplitude error of better than 0.5dB or a phase error of better than 0.5 degrees is required [Wil92a]. For narrowband systems, the delay mismatch can be corrected by simply adjusting the phase. However, for wideband systems, a given delay corresponds to different phase shifts at different frequencies. It is shown in [Par94] that, if the delay mismatch corresponds to one wavelength of the carrier signal, then the distortion suppression will be limited to 30dB at both ends of the bandwidth, assuming 1% bandwidth with an otherwise ideal system. Thus, to take into account the delays when designing a feedforward system, the delaying elements should be quantified by means of measurements and/or simulations. A systematic approach based on simulation is presented in [Kon93]. The technique is based on harmonic balance simulation that takes into account the nonlinear effects in the amplifiers, provided that appropriate simulation models are available. The method was found to work quite well for a microwave feedforward amplifier. Simulation results gave 20dB suppression over 500MHz bandwidth and a 6GHz carrier frequency.

Fixed or manually controlled amplitude and phase matching networks are usually not sufficient to preserve reasonable distortion suppression. Component aging, temperature drift, change of operating frequency etc. cause variations that require automatic control of the amplitude and phase matching networks. Several solutions have appeared and three distinct techniques can be identified. One approach is based on measuring the power at some points and minimizing it [Obe91]. For example, the signal cancellation loop can be tuned by minimizing the power of the error signal and the distortion cancellation loop can be tuned by minimizing the out-of-band power at the transmitter output. Another method uses a pilot tone that is inserted at some point,

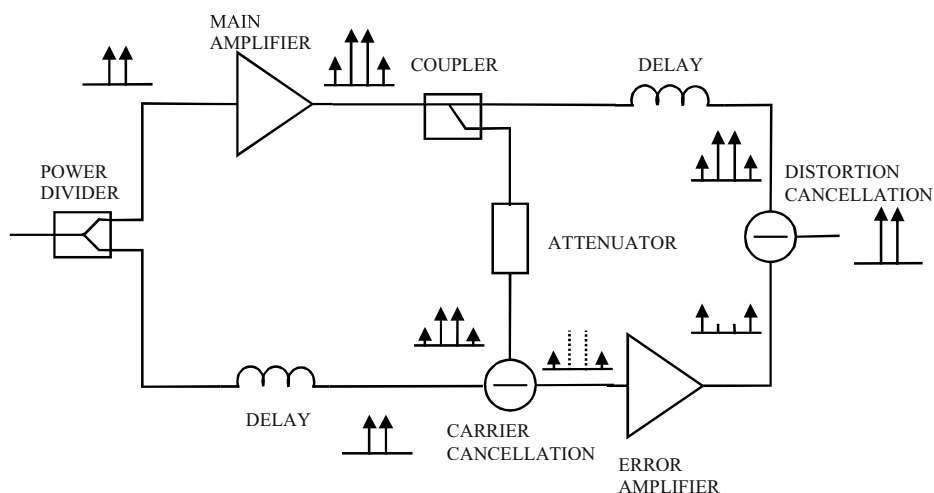


Figure 2-1. Block diagram of the feedforward system.

typically after the power amplifier, and a measure of the imbalance is obtained by detecting the pilot at another point [Nar91], [Cha91]. The level of the detected pilot tone guides a controller to adjust the amplitude and phase accordingly. The third method is based on the fact that the gradient for the function to be optimized can be calculated and used to guide the adjustment [Olv85], [Bau93], [Ken94], [Cav95], [Smi98]. Some of these solutions require a computer-based controller [Obe91], [Nar91], [Ken94], [Smi98]; others can be implemented using a continuous closed-loop system [Cha91], [Ken94]. Common for all the schemes is the fact that the total complexity of the feedforward scheme becomes quite large compared with the basic feedforward configuration shown in Figure 2-1.

Even though the main amplifier can be quite power efficient, the total efficiency of the feedforward scheme is drained due to losses in the main path delay, the couplers and the auxiliary amplifier. A high efficiency auxiliary amplifier should, of course, be used. But it must also be sufficiently linear so that no additional distortion is generated. Furthermore, the coupler that is used to subtract the error signal from the amplifier output should have a low coupling factor. With a low coupling factor, most of the power available from the main amplifier is fed to the antenna. On the other hand, the coupling factor should be high because the auxiliary amplifier must provide enough power to compensate for the losses in the coupler. Thus, an optimal coupling factor can be calculated based on the knowledge of the other components in terms of amplifier efficiency, intercept points and delay line losses [Ken92], [Dix86].

Several feedforward prototypes have been reported [Ken91b], [Se71b], [Ste88], [Mey74], [Nar91], [Dix86] with performance ranging from 20 to 40dB suppression both for narrowband and wideband systems with carrier frequencies from a couple of MHz to several GHz.

Typical applications are of narrowband type, i.e. the bandwidth is a couple of percentage points of the carrier frequency. Yet, it is interesting to note that feedforward has also been applied to systems with a bandwidth of one decade or more [Sei71a], [Mey74], even though the maximum frequency was rather low ($< 300\text{MHz}$). With such a large bandwidth, the designer is confronted with the problem of obtaining flat frequency responses from all components. In [Mey74], an interesting solution is presented because it combines the best sides of Black's two schemes. The main amplifier has a local negative feedback loop, while, for low frequencies, the loop-gain is large enough to make the amplifier itself sufficiently linear. However, for higher frequencies, the loop-gain is diminished and the feedforward technique, which is optimized for the higher frequencies, takes over.

2.2 Cartesian Modulation Feedback

Figure 2-2 shows the principle of the Cartesian feedback transmitter. The output of the amplifier is synchronously demodulated and compared with the source signal to obtain an error signal. The error signal is fed to the loop filter followed by upconversion in a quadrature modulator before it finally reaches the power amplifier. The Cartesian feedback was introduced by Petrovic [Pet83]. Several experimental systems have been reported operating with carrier frequencies ranging from a couple of MHz to 1.7GHz with modulation bandwidths of up to 500kHz [Joh91], [Wil92b], [Joh94], [Whi94]. Distortion suppression varies from 20dB up to 50dB with 35% to 65% amplifier power efficiency.

It is interesting to note that Cartesian feedback has been proven to work for wideband applications [Joh91]. Careful design and selection of components are required since the propagation delay will dominate the phase characteristics of the loop. The relations between loop delay, bandwidth and stability are investigated in [Joh95]. It is shown that for a 5MHz cross-over bandwidth the loop delay should not exceed 33ns (with 60° phase margin). Assuming that the loop gain is sufficiently high and that the loop contains a single pole, this corresponds to 20dB of IMD suppression at 500kHz.

A potential problem for Cartesian feedback transmitters is that the characteristics of the amplifier effectively degrade the gain and phase margins of the loop. Another problem is the phase shift that occurs in the loop when, for example, changing the carrier frequency. In practice, a phase adjuster is required to adjust the phase automatically to preserve the stability [Bro88], [Ohi92]. The phase adjuster can be placed in the loop or, as illustrated in

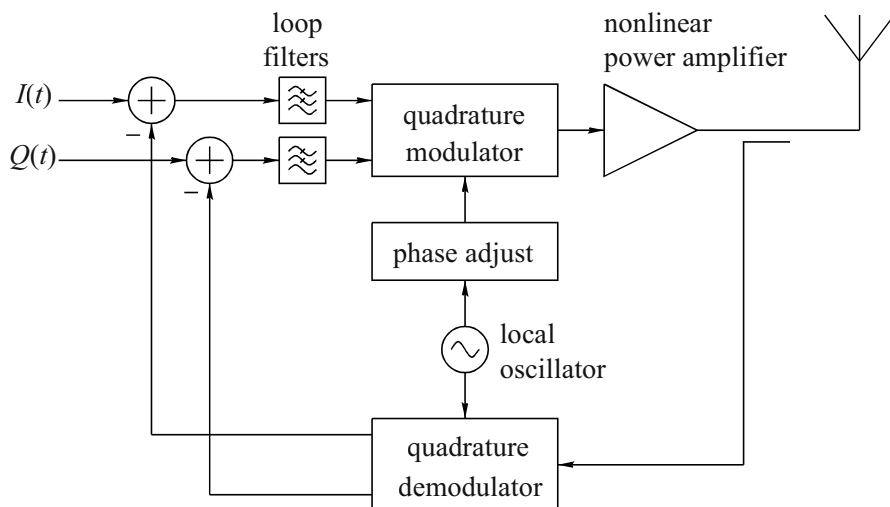


Figure 2-2. Cartesian feedback transmitter block diagram.

Figure 2-2, to differentiate the local oscillator phase between the modulator and the demodulator. This means that the complexity of a practical Cartesian feedback transmitter is much higher than the diagram in Figure 2-2 might imply.

One disadvantage of Cartesian feedback and other schemes where the amplifier is fed with a signal having a varying envelope is that the power efficiency is low for low input levels. An attempt to improve on efficiency is presented by Briffa et al. [Bri93]. Dynamic biasing is applied to the final amplifier stage. The biasing (of both base and collector) is controlled by the envelope of the input signal through mapping functions that apply predetermined biasing levels for maximum efficiency. Simulation results show that the efficiency can be increased from 50% to 60% near saturation and from 10% to 30% at low input levels. A small improvement in linearity was a welcome side effect.

To obtain low levels of ACI the dynamic range of the feedback path must be appropriately high. Upwards, the dynamic range is limited by intermodulation, especially in the quadrature demodulator, and, downwards, by the accumulated noise in the feedback path.

A derivative of Cartesian feedback was presented by Johansson et al. for multi-carrier applications [Joh93]. The principle is outlined in Figure 2-3. Several Cartesian feedback loops (Figure 2-4) operate in parallel on distinct frequency bands. Each loop can accommodate several carriers. An interesting property is that a loop can be assigned to a channel without an input sig-

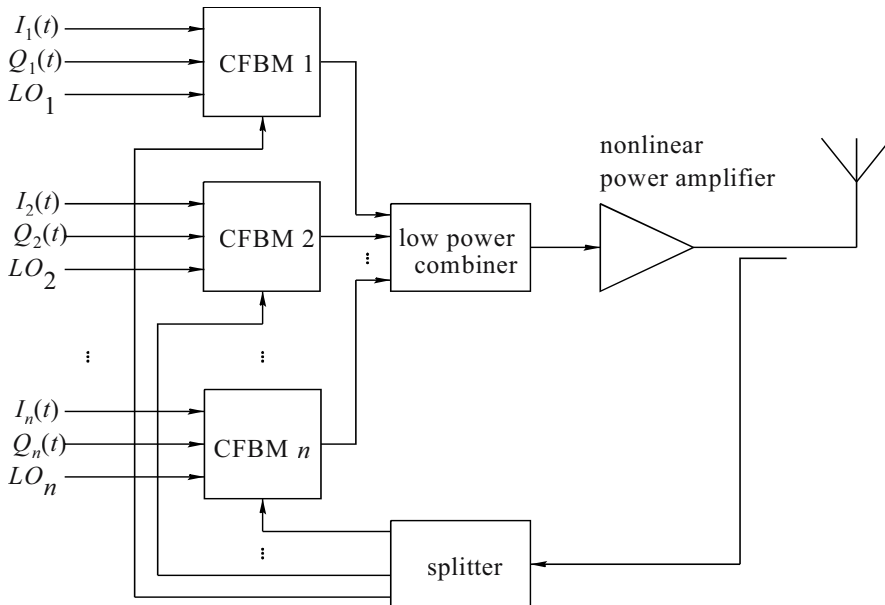


Figure 2-3. Block diagram of multi-loop Cartesian feedback system.

nal to reduce intermodulation products in that channel caused by the other channels in operation. Experimental results show that this is a viable method for broadband multi-carrier linearization with up to 30dB intermodulation suppression [Joh94]. The use of Cartesian feedback with a class-C PA amplifying an IS-136 (DAMPS) signal improves the first ACPR by 35 dB and allows the signal to be produced with an efficiency of 60% [Ken00].

2.3 Predistortion

From a mathematical point of view, predistortion is, next to feedforward, probably the most obvious technique for linearization. By preceding the nonlinear amplifier with its inverse counterpart, one-to-one mapping between the input and output can be obtained. As illustrated in Figure 2-5 predistortion in its most simple form is an open loop system. However, most solutions presented use some kind of feedback to enable adaptation of the predistorter. Several solutions have been developed to realize the predistorter, from digital baseband processing to processing the signal directly at RF using diodes as nonlinear devices.

Behavior models for PA have traditionally been developed on the basis of the AM-AM and AM-PM curves, and the PA gain is usually approximated as a complex polynomial function of instantaneous input power level. However, as the bandwidth of the signal increases, memory effects in the transmitter distort this simplified picture. Memory effects are attributed to filter group delays, the frequency response of matching networks, nonlinear capacitances of the transistors and the response of the bias networks. The performance of the predistortion algorithms that do not take these memory effects into account is severely degraded as the bandwidth of the input signal increases [Vuo01]. A nonlinear system with memory can be represented by Volterra series, which are characterized by Volterra kernels [Sch81]. How-

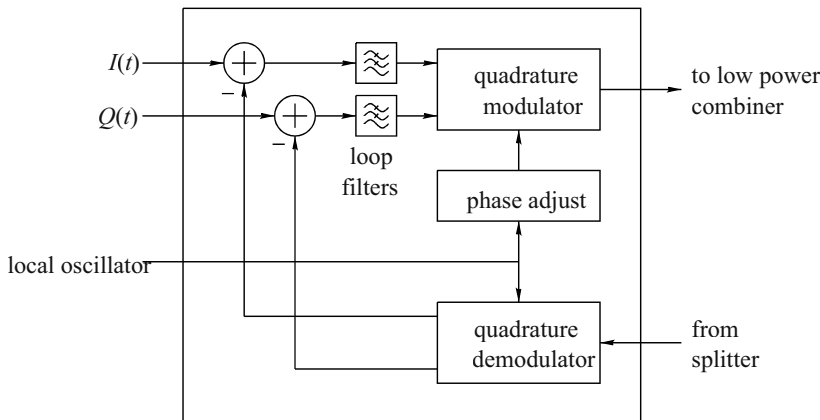


Figure 2-4. Cartesian feedback module (CFBM).

ever, the computation of the Volterra kernels for a nonlinear system is often difficult and time consuming for strongly nonlinear devices. In many applications that involve modeling of nonlinear systems, it is convenient to employ a simpler model. The Wiener model, which is a cascade connection of linear time invariant (LTI) system and memoryless nonlinear system, has been used to model nonlinear PAs with memory [Sal81]. A straightforward predistortion method is to add an adaptive filter in cascade with the memoryless predistorter [Kan98].

In applications with weak nonlinear amplifiers, or having moderate linearity requirements, coarse approximations of the wanted predistortion function can be applied using nonlinear analogue components. In fact, what this all really says is that the predistorter is essentially a technique, which can work usefully for well backed-off amplifiers showing only small amounts of compression. This is a very tough restriction for signals having high peak to average ratios.

The output of the power amplifier in Figure 2-5 is

$$v_o = v_p - a_3 v_p^3 \quad (a_1 = 1). \quad (2.1)$$

A predistorter generates output v_p

$$v_o = v_{in} = v_p - a_3 v_p^3, \quad (2.2)$$

so a predistorter has to "solve" the cubic

$$v_p^3 - \frac{1}{a_3} v_p + \frac{1}{a_3} v_{in} = 0. \quad (2.3)$$

This can be solved by

$$v_p = v_{in} + b_3 v_{in}^3 + b_5 v_{in}^5 + b_7 v_{in}^7 + \dots \text{etc.} \quad (2.4)$$

This shows that, in general, the predistorter contains an infinite series of terms of a higher order than the amplifier distortion itself. The output signal from any useful predistorter will have a spectral bandwidth significantly greater than the distorted PA output that it strives to linearize; this has important implications for the required bandwidth of any components used in predistorter design.

The output of the power amplifier is

$$v_o = a_1 v_p - a_3 v_p^3. \quad (2.5)$$

The transfer function of the predistorter is

$$v_p = b_1 v_{in} + b_3 v_{in}^3 \quad (b_1 = 1). \quad (2.6)$$

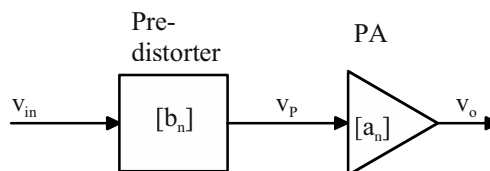


Figure 2-5. Open-loop predistortion block diagram.

Using (2.5) and (2.6), the power amplifier output is

$$\begin{aligned} v_o &= a_1(v_{in} + b_3 v_{in}^3) - a_3(v_{in} + b_3 v_{in}^3)^3 \\ &= a_1 v_{in} + (a_1 b_3 - a_3) v_{in}^3 - 3a_3 b_3 v_{in}^5 - 3a_3 b_3^2 v_{in}^7 - a_3 b_3^3 v_{in}^9. \end{aligned} \quad (2.7)$$

If $b_3 = a_3/a_1$, then the third harmonic will cancel. But we now have additional 5th, 7th and 9th order products. An important result is that a predistorter with a simple 3rd degree expansion characteristic can cancel the 3rd degree nonlinearity in an amplifier, but will create additional higher degree nonlinearities, that were absent in the basic PA itself.

2.3.1 Analog Predistortion

In [Noj84] Nojima et al. used diodes to build a third-order predistorter (see Figure 2-6) that operated at an intermediate frequency (130MHz) for a microwave system with 6GHz carrier frequency and a 30MHz bandwidth. More than 30dB suppression of the third-order IMD was obtained with a self-adjusting system that controlled the magnitude and the phase of the third-order predistorter. A continuation of this work was presented in [Noj85], [Noj85], [Nan85], where the predistorter worked at the carrier frequency instead of an intermediate frequency. One prototype was intended for mobile telephone systems operating at 800MHz and the other one for 6GHz microwave digital radio systems. For the 800MHz system, Nojima reported up to 20dB reduction of the third-order products over a bandwidth of 25MHz and, for the microwave system, about 6dB over a 500MHz bandwidth. A similar solution is presented in [Nam83], where the predistorter is realized with FET (field-effect transistors) amplifiers acting as nonlinear devices.

Third-order predistorters are not enough if we want to obtain higher accuracy or linearize amplifiers that are less linear. One solution is to use higher order polynomials. However, this requires more advanced adaptation algorithms since there will be more coefficients to adjust. A new technique for adaptation of this kind of predistortion linearizers is presented by Stapleton et al. in [Sta91], which is based on minimizing the out-of-band power. By describing both the predistorter and the amplifier with truncated 5th-order complex polynomials, the IMD power can be expressed as a function of the polynomial coefficients. From this analysis it was shown that for the dominant third order components the out-of-band power represented a quadratic surface. Although the analysis required that the input signal was a stationary Gaussian process, it was also demonstrated by means of simulations that quadratic-like surfaces were obtained for a 16-QAM signal. The fact that one global minimum exists suggests that we can choose from several powerful optimization methods. As a first step, Stapleton simulated a system using a modified version of the Hooke and Jeeves method, a direct search

scheme, to prove the viability of this approach. A prototype operating with an 850MHz carrier frequency is presented in [Sta92a]. The polynomial coefficients were adjusted manually to give a 15dB improvement in the third-order and 5dB improvement in the fifth-order IMD products. Simulations estimated that the adaptation time was rather long, in the order of minutes, in fact.

The method that was used to measure the out-of-band power included the use of a mixer and a separate local oscillator followed by a filter and a rectifier. An alternative solution is given in [Sta92b], [Sta92c] that uses convolution of the RF input signal and the transmitter output signal, thereby avoiding an additional oscillator for the down conversion of the output signal. Analysis showed that the resulting signal could be used as a measure of the out-of-band power, just as in the previous approach. Prototype results gave an 11dB improvement in linearity compared with the 15dB that was predicted by means of simulation. The adaptation time was decreased to approximately 10 seconds by using more advanced optimization schemes based on surface fit.

The analysis based on complex polynomials has been extended to include quadrature modulator errors in [Hil92], [Hil94]. It is shown that the out-of-band power is a quadratic function of both the amplifier nonlinearities and the quadrature modulator errors. A prototype was reported to work excellently with up to 20dB IMD suppression and with a convergence time of below 4 seconds.

Yet another system based on polynomial predistortion is presented in [Gha93], [Gha94]. This system has a significantly larger complexity compared with the previous solutions, especially in terms of digital signal processing. The adaptation process requires synchronous detection of the output signal instead of monitoring the out-of-band power. Simulation results promise up to 45dB distortion suppression. Still no prototype results have been reported for this scheme.

The use of polynomial functions and a simple adaptation scheme based

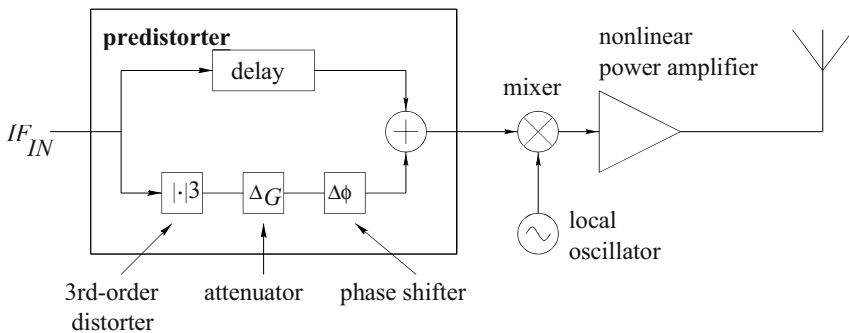


Figure 2-6. Principle of third-order predistorter.

on out-of-band power measurements offers low cost and complexity. However, a low-order polynomial is capable of canceling only weak nonlinearities. For more non-linear amplifiers, more general schemes based on DSP techniques and look-up tables have been developed.

2.3.2 Mapping Predistortion

In [Bat88], Bateman et al. suggested the use of DSP techniques and look-up tables with curve fitting to realize an adaptive predistorter. The approach required the transmission to be interrupted because of the special signal that had to be applied to characterizing the amplifier nonlinearities. However, the solutions presented below can adapt while transmitting and do not require any special signals.

A simple and "brute force" solution was presented by Nagata [Nag89], who used a huge two-dimensional table (see Figure 2-7). By using a two-dimensional table, any complex input signal represented by its Cartesian components can be mapped to a new constellation of Cartesian components. Thus, any distortion or error occurring in the conversion process can be cancelled. This even includes misalignments and nonlinearities in the quadrature modulator. For the purpose of adaptation, the amplifier output signal is synchronously demodulated and compared with the input signal. The suggested adaptation process is quite simple and is performed while transmitting. Actually, it is the time-discrete equivalent to the Cartesian feedback system described in Section 2.2. As such, the phase of the feedback signal has to be correct for stable operation. Nagata presented results from an experimental system with a 16kHz modulation bandwidth, 128kHz sampling rate and 145MHz carrier frequency. Up to 26dB distortion suppression was obtained,

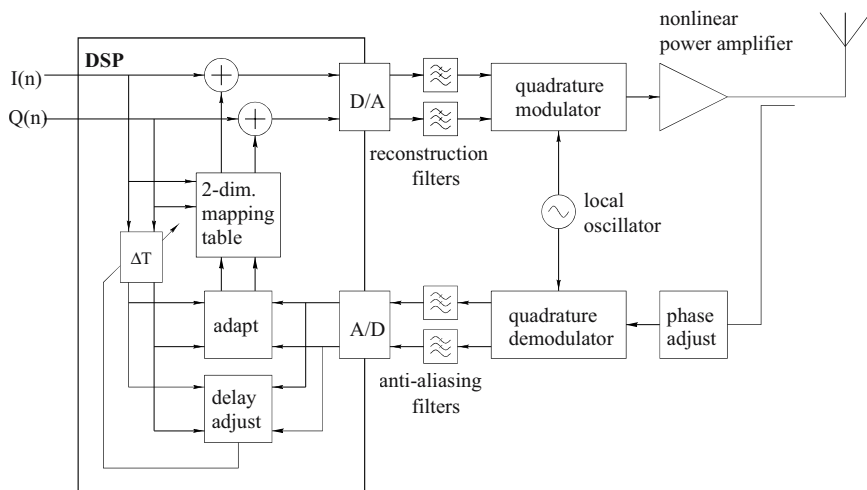


Figure 2-7. Mapping predistorter block diagram.

but the convergence time was quite long (10 seconds), which was primarily caused by the size of the table (2Mwords). The problem of correct sampling of the demodulated signal to be used in the adaptation process was addressed. It was shown by simulation that a sampling time deviation corresponding to 1% of the symbol time or more resulted in a significant degradation in performance. An automatic delay adjustment circuit that converged within 16 symbols was presented. Nagata reports that 2Mbits of memory were required in his design, along with 13000 gates in dedicated DSP hardware and up to 2W in the A/D and D/A converters. The total power consumption was therefore up to 4W, which is significantly more than the power output of the RF amplifier in most handportable equipment. It is therefore evident that the power-efficiency of an adaptive predistortion system will be poor until device technology is advanced sufficiently to enable the power consumption of the linearizer to become a small fraction of that of the RF power amplifier.

A similar system is presented by Minowa et al. [Min90], who investigated this technique in conjunction with amplifier back-off. Yet another prototype is presented in [Man94], where the computational burden was increased by using interpolation of the table entries. With this procedure, the table size could be reduced by a factor of 16 to 64 kwords. More memory efficient schemes have been developed (see Sections 2.3.3 and 2.3.4). In contrast to the mapping predistorter, these techniques can only compensate for phase-invariant non-linearities.

2.3.3 Complex Gain Predistortion

The major drawback of the mapping predistorter is the size of the two-dimensional table, which results in long adaptation times. However, if we restrict the predistorter to correct for nonlinearities in the amplifier alone, then a one-dimensional table will do, since the amplifier characteristic is a function of the input amplitude only. That is, such a table would approximate the inverse function of the amplifier nonlinearity with a finite number of table entries.

This approach has a table containing complex-valued gain factors given in Cartesian form, see Figure 2-8 [Cav90]. The address to the table is calculated as the squared magnitude of the input signal, which gives a uniform distribution of power in the table entries. Furthermore, the input signal is predistorted by a single complex multiplication. All in all, this leads to a substantially larger computational load compared with the mapping predistorter. The adaptation scheme of the complex gain predistorter is a multiplicative predistorter in contrast to the mapping predistorter, which is additive. This means that the complex gain predistorter is not sensitive to the phase of

the feedback signal, as is the case for the mapping predistorter. A phase adjustment circuit in the feedback path is therefore not necessary for stable operation.

Since this solution assumes a phase-invariant characteristic it depends heavily on the use of perfect quadrature modulators and demodulators. Such modulators and demodulators are difficult and expensive to build. Methods for automatic adjustment of these errors have therefore been suggested in Chapter 3.

Cavers analyzed the effect of table size on adjacent channel interference. The effect of adaptation jitter was also investigated and it was found that the table size should be increased by 20% to account for this effect. The adaptation process was formulated as a root finding problem; the secant method was found to perform significantly faster than the linear scheme described by Nagata in [Nag89]. The convergence time was estimated to be less than 4ms for a 25kHz system with 64 table entries, provided that every table entry was accessed exactly 10 times each.

The LUT size affects linearly the speed of adaptation, so one way to increase the adaptation speed is to reduce the number of entries in LUT. The number of LUT entries, however, determines how closely the predistorter is able to follow the inverse function of the amplifier distortion as well as the maximal signal to noise ratio (SNR) available in the output of the amplifier. Both the precision and the entry number requirement can be alleviated with a nonuniform organization of the LUT entries [Cav97]. By organizing the entries in such a way that the entries do not overlap [Has01], the required precision can be reduced, while by organizing the entries according to the probability density of the amplitude values, the required number of LUT entries can be reduced [Muh00].

An experimental system based on complex gain predistortion is presented in [Wri92]. This system gave an up to 25dB improvement in linearity in nar-

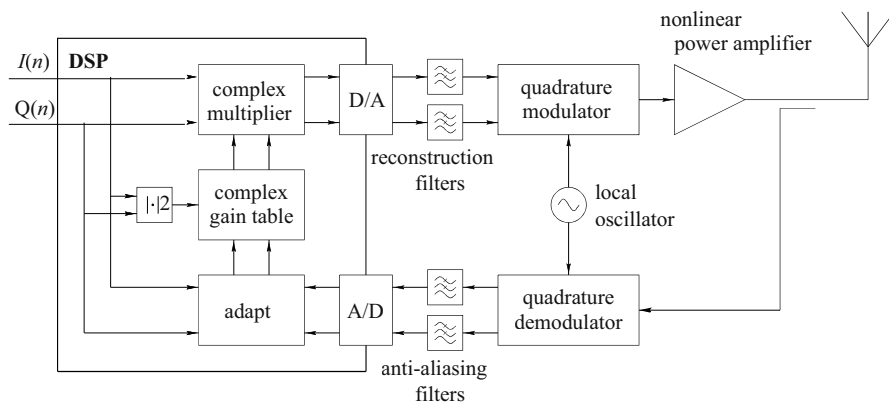


Figure 2-8. Complex gain table predistorter.

rowband operation (less than 1 kHz). A recent advance in this area has been described by Sundström et al. [Sun96], [And97], in which a dedicated predistorter ASIC is outlined. The performance of this device was shown to be very good over a broad range of channel bandwidths (up to 300 kHz) and the use of an ASIC helped to reduce the linearizer power consumption to sensible levels (roughly one-tenth of that of an equivalent clock-rate DSP device, whilst providing around seven times the channel bandwidth).

2.3.4 Polar Predistortion

The approach suggested by Faulkner et al. [Fau94] uses two one-dimensional tables containing magnitude gain and phase rotation, respectively. The principle is illustrated in Figure 2-9 [Sun95]. From the input signal, given in Cartesian components, the amplitude is calculated and used as an address to the look-up table containing amplitude gain factors. The input signal is multiplied by the gain factor obtained from the table. The magnitude of the input signal is multiplied by the same gain factor and the result is used to address the second table containing the phase. Finally, phase rotation is performed by the amount obtained from the second table. This final step includes two additional look-ups to get $\sin()$ and $\cos()$ values for the rotation matrix. This technique requires more operations to predistort the signal compared with the complex gain predistorter.

Since the adaptation process is based on polar coordinates each iteration involves rectangular-to-polar (R/P) conversion of the input signal and the detected output signal. All in all, this leads to a substantially larger computational load compared with the mapping predistorter. The adaptation scheme of the polar predistorter is a multiplicative predistorter in contrast to the mapping predistorter which is additive. This means that the polar predistorter is not sensitive to the phase of the feedback signal, as is the case for the mapping predistorter. A phase adjustment circuit in the feedback path is

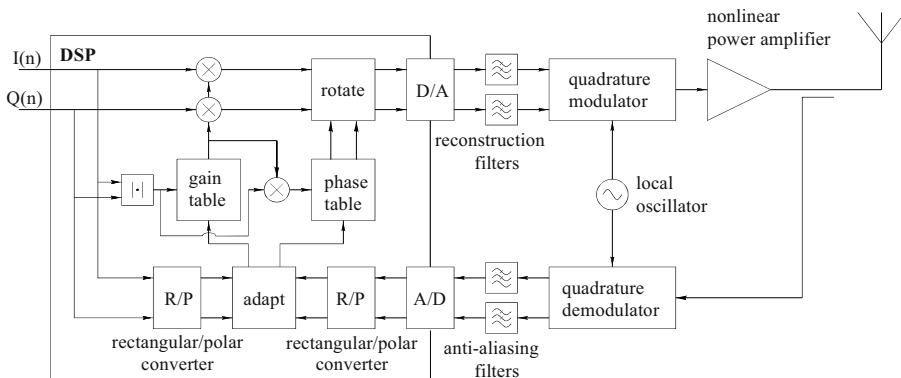


Figure 2-9. Polar predistorter block diagram.

therefore not necessary for stable operation.

This technique gives a considerable reduction in table size compared with the mapping predistorter [Fau94]. Faulkner reported that, as a compromise between convergence time and error it was found that 64 table entries gave the best result, provided that interpolation was used. This is four orders of magnitude less than the size of the mapping table. As a consequence, the adaptation time was estimated to be less than 10ms for a 25kHz system with a sampling rate eight times the symbol rate, assuming that all samples were used in the adaptation process. The experimental system gave up to 30dB IMD suppression when the system was exercised with a two-tone test [Fau94]. However, the modulation bandwidth was only 2 kHz.

The possibility of making the gain table a function of $|\cdot|^2$ (see Figure 2-8) instead of $|\cdot|$ (see Figure 2-9) was considered in [Fau94]. $|\cdot|^2$ is easy to calculate and causes the points of the table to be concentrated in the saturation region of the amplifier characteristics, while $|\cdot|$ concentrates more points into the turn-on region.

Since this solution assumes a phase-invariant characteristic, it depends heavily on the use of perfect quadrature modulators and demodulators. Therefore, methods of automatic adjustment of these errors have been suggested in Chapter 3.

2.3.5 RF-Predistortion Based on Vector Modulation

The RF-input/output predistortion operates completely independently of the circuit before the amplifier [Set00]. The RF-predistortion is implemented by transforming the analog PA input signal by analog means that are controlled by digital signals (Figure 2-10). The phase and envelope of the input and output signals are detected by analog means and A/D-converted. The result is fed to a DSP that retrieves phase and amplitude correction values corresponding to the input envelope value from a LUT. These values are D/A converted and used to control an analog AM/PM distorter that alters the RF-signal. The effect of RF-predistortion can be described with formula:

$$\begin{aligned} A_{PD}(|V_{IN}|) A_{PA}(|V_{PD}|) &= K \\ \phi_{PD}(|V_{IN}|) + \phi_{PA}(|V_{PD}|) &= \Delta\phi, \end{aligned} \quad (2.8)$$

where A and ϕ refer to the notation in Figure 2-10 and K and $\Delta\phi$ are constants.

The LUT is updated with a suitable algorithm on the basis of the phase and amplitude differences of the input and output signals. The signal can be detected with simple envelope and phase detectors [Ken01]. This configuration alleviates the problem of detector nonidealities that may limit the linearization ability of the baseband schemes, as both the output and input sig-

nals are detected with the same kind of detectors. Therefore, the possible error in the correction is only due to the mismatch between the detectors, which can be kept low with proper selection of the detectors. If the amplitude distortion is delayed compared to the phase distortion or vice versa, the upper and lower intermodulation sidebands become asymmetrical [Crip02]. This can be caused by, for example, different delays in the phase and amplitude feedback loops.

Another method used for the LUT update is the minimization of out of band distortion. This can be achieved by downconverting the signal to IF and bandpass filtering the signal to extract the out of band distortion. The power of the distortion is measured and the LUT is updated to minimize it. Another drawback is the slow convergence of the adaptation and complexity [Set00].

The limiting factors in the RF-predistortion mainly relate themselves to the analog parts of the circuit. The phase delay round the forward control loop reduces the effectiveness of the predistortion. The main sources for this delay are DSP latency, A/D and D/A converters latencies and the delays of the reconstruction and anti-aliasing filters. This can be alleviated to some extent by increasing the sampling rate of the digital parts and by adding an

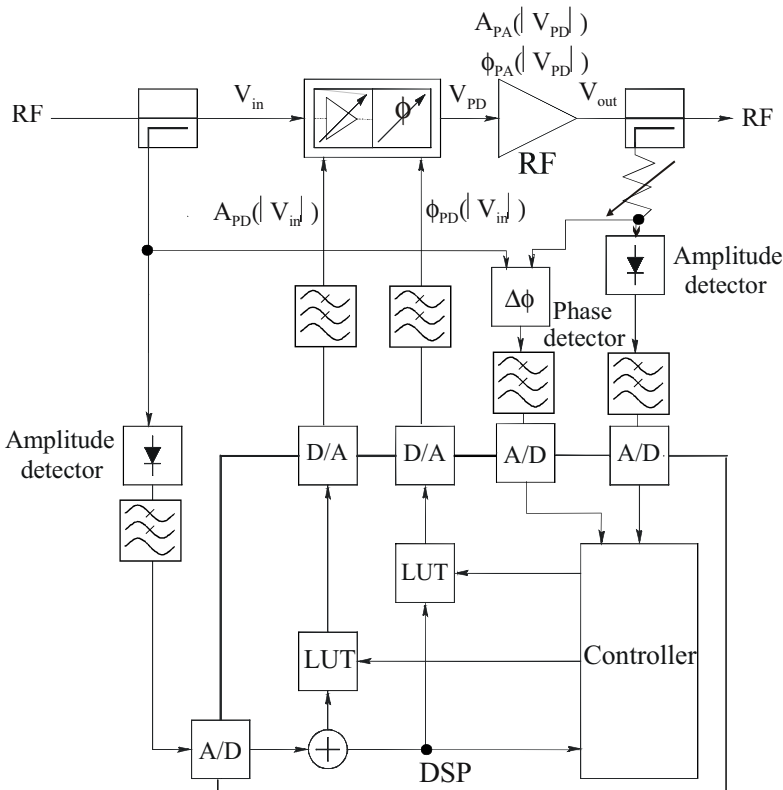


Figure 2-10. RF input/output adaptive digital linearizer.

analog delay element before the analog predistorter. The envelope detection process generates signals that have a higher bandwidth than the original modulation bandwidth of the RF signal. The RF predistortion based on vector modulation is therefore mainly suitable for narrowband systems. The wider bandwidth requires a higher sampling frequency in DSP. Furthermore, less delay for forward correction is tolerated. The operational bandwidths of the envelope detectors as well as the A/D and D/A converters limit the bandwidth of the correction signal. The use of the RF predistorter to amplify an EDGE signal improves the first ACPR by 20 dB and allows the signal to be produced with an efficiency of 30% [Ken01].

2.3.6 Data Predistorters

The technique operating at the transmitter is mainly related to distorting the data alphabet [Sal83], [Boy81]. Such predistorters compensate for the warping and clustering effects on the data constellation and therefore improve the eye openings at the maximum eye opening instants. They will thus improve the error magnitude of the amplifier output signal, but do not usually improve, intentionally, the adjacent channel performance or spectral purity of the transmitter output. They commonly employ look-up table based techniques to form the predistortion function and operate in a manner similar to their conventional equivalents. Biglieri proposed a data predistortion scheme with memory [Big88], which improves the system performance with respect to the memoryless predistorter. A key disadvantage with this form of the predistorter is that it is generally modulation format specific.

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