Chapter 3

Distortions Reduction Techniques in Direct Conversion Receiver

3.1 General Considerations

Direct conversion receiver architecture is very attractive because of its simplicity and amiability to monolithic integration. But, as discussed in the previous section, it suffers from various distortions such as dc-offset, even-order distortion and I/Q mismatch problem. Efforts have been made to analyze various sources of distortion and various distortion reduction/removal techniques suitable for integrated RFIC receiver suggested by various researchers. Brief description with their strengths and weaknesses of various distortion cancellation or removal techniques are mentioned here.
3.2 DC-Offset Reduction Techniques

In the DCR, the desired signal is down-converted directly to a baseband, and the offset signals generated by the self-mixing problem can corrupt the baseband desired signal, as shown in figure 2.5. Therefore, it is necessary to mitigate these offset signals in order to successfully demodulate the desired signal at the DCR. The offset signals generated by the self-mixing problems can be classified into two types.

One is the DC offset generated by local oscillator (LO) leakages, and the other is the self-mixed interference generated by interference leakages [6]. The DC offset is a time-invariant offset signal while the self-mixed interference is a time varying offset signal. Numerous methods are proposed in literature to solve this problem. These methods can be classified in following categories:

3.2.1 AC-coupling

This is a simple approach to employ high-pass filtering in the down converted signal path to remove DC component present due to self-mixing. But this technique corrupt the down converted baseband signal if it exhibits a peak at DC, such as M-ary random data. In addition to cover full baseband, high pass filter must have the corner frequency of less than 0.1% of the data rate. For low data rate application, this frequency becomes so low that the high-pass filter implementation required large capacitors and resistors, which is not desirable. A low corner frequency in the high pass filter may also lead to temporary loss of data in the presence of wrong initial conditions [9]. If no data
is received for a relatively long time, the output DC voltage of the high pass filter (HPF) droops to zero. Now if data is applied, the time constant of the filter causes the first several bits to be greatly offset with respect to the detector threshold, thereby introducing errors. A possible solution to the above problems is to minimize the signal energy near DC by choosing “DC-free” modulation schemes. One possible solution to this problem is to select “DC-free” modulation schemes, such as binary frequency shift keying (BFSK) [5].

3.2.2 DC-Offset cancellation

In offset cancellation method, wireless device store the dc-offset on a capacitor during ideal mode and this stored offset will be subtracted from the received signal during active mode. In wireless standards that incorporate time-division multiple access (TDMA), each mobile station periodically enters an idle mode so as to allow other users to communicate with the base station. The offset in the receive path can be stored on a capacitor during this mode and subtracted from the signal during actual reception. Figure 3.1[6] shows a simple example, where capacitor stores the offset between consecutive TDMA bursts while introducing a virtually zero corner frequency during the reception of data.

For typical TDMA frame rate, offset cancellation is performed frequently enough to compensate variations due to moving objects. The major issue in the circuit of figure 3.1 is the thermal noise of $S_1$ (kT/C noise).
For example, if a 1μVrms signal received at the antenna experiences a gain of 30dB before offset cancellation, then $C_1$ must be at least 200 pF so that $kT/C$ remains 15 dB below the signal level. If the signal path is differential, then two capacitors, each equal to 400 pF, are required so that the overall noise is still 15 dB below the signal. Thus, with I and Q channels, the total capacitance reaches 1.6 nF. Note that since $C_1$ is a floating capacitor, it cannot be easily implemented with MOS transistors. Structures providing so much capacitance typically occupy a very large area.

A general difficulty with offset cancellation in a receiver is that interferers may be stored along with offsets. This occurs because reflections of the LO signal from nearby objects must be included in offset cancellation and hence the antenna cannot be disconnected (or “shorted”) during this period. While the timing of the actual signal (the TDMA burst) is well-defined, interferers can appear any time. A possible approach to alleviating this issue is to sample the offset (and the interferer) several times and average the results.
3.2.3 Adaptive DC-Offset cancellation

In adaptive methods, amount of dc-offset is estimated with the help of various techniques and then this estimated dc-offset is subtracted from the received signal. Such methods are discussed in [18,37-41]. They can be effective in mitigating time-invariant offsets while they cannot effectively mitigate time-varying offsets. Some digital signal processing methods for mitigating the time-varying offset have been introduced in [42]. However, the cancellation methods in [42] cancel the time-varying offsets only when the time-varying offsets are periodic. A novel receiver structure based on the adaptive interference cancellation (IC) method as shown in figure 3.2 is suggested in [43].

![Figure 3.2 Receiver structure based on self-mixed interference cancellation procedure](image)

As all the techniques are implemented mainly in DSP section, increase in the component count in RF section is not an issue, but they do not take care of other distortions. This may result into the requirement of other methods to
deal with remaining distortions, results in the increase in component count and computational load.

3.3 Even-order distortion reduction techniques

In direct conversion receiver, second order intermodulation distortion (IMD2) is a significant kind of interference because it falls in the baseband occupied by a down converted baseband signal. A large number of techniques have been suggested by various researchers. Brief description with their strengths and weaknesses of various IMD2 cancellation or removal techniques are mentioned here.

3.3.1 Layout Techniques

Layout techniques can be classified into device-matching improvement techniques and RF-LO coupling reduction techniques.

Careful, symmetrical layout of both active and passive devices as well as their interconnects improves matching between differential paths of the circuit. Parameter mismatches of active devices are significantly reduced by applying a so-called common centroid layout technique, which includes cross-coupling and inter-digitation patterns.

Another category of IMD2-cancelling layout techniques deals with minimization of the impact of RF self-mixing on the overall even order distortion performance. Three coupling mechanisms responsible for self-
mixing phenomenon need attention: EM crosstalk, substrate coupling as well as ground and power supply bounce. Shielding of LO transmission lines by using neighboring grounded metal layers [figure 3.3(a)] reduces EM crosstalk since metal ground planes confine electromagnetic fields [44].

![Diagram of shielding of LO transmission lines](image)

(a) *Shielding of LO transmission lines*

![Diagram of substrate coupling suppression](image)

(b) *Substrate coupling suppression*

**Figure 3.3** *LO-RF coupling reduction techniques*

The disadvantage is an increased power consumption of LO buffers driving such transmission lines required to maintain high-quality LO waveforms. Substrate coupling can be suppressed by placing highly resistive substrate rings around the sensitive downconversion mixer circuit [figure 3.3(b)]. Other substrate coupling reduction techniques are reported in [45].
Finally, ground and power bounce can be reduced by connecting circuit and substrate grounds together using densely placed substrate contacts as well as by using de-coupling capacitors between power and ground lines.

All of the presented layout techniques increase IIP2 on average i.e. shifts its histogram toward higher values. However, they cannot totally remove circuit and interconnection asymmetries as well as completely eliminate random parameter variations.

### 3.3.2 Circuit Architecture for Even-order distortion removal

Circuit techniques focus on decreasing second order nonlinearity of mixer devices rather than on reducing mismatches between them. Various categories of circuit techniques are as below.

**(a) RC Degenerated Input Stage**

In MOS transistor, negative feedback can be applied by connecting an impedance between the source of the active device and ground. In case of simple resistive feedback, this technique reduces the effective first and higher-order transconductances of the transistor at all frequencies. Since only second-order transconductance has to be reduced, it is advantageous to apply negative feedback with resistor and capacitor connected in parallel, as shown in figure 3.5 [45]. At low frequencies, the absolute value of the impedance of such network is determined mainly by the resistance $R$, implying a strong negative feedback and significant attenuation of low-frequency distortion products. At
high frequencies, the absolute value of the impedance is determined by the capacitance C, which shorts R to ground so that effectively no feedback exists. Consequently, the effective transconductance is larger, which improves mixer noise performance.

\[\text{Figure 3.4} \quad \text{IMD2 cancellation with RC degenerated input stage}\]

Advantage of this method is easy implementation. But the drawbacks are (i) Resistor will occupy larger area on chip, (ii) This negative feedback attenuate desired signal also (iii) This method does not alleviate the problem of second order distortion of the complete mixer because of the remaining distortion products due to RF-LO coupling.

(b) **Input Stage with IMD2-Cancelling Biasing Circuit**

Second order intermodulation distortion generated in the mixer input stage can be also reduced by using a special kind of biasing configuration. This biasing method is introduced in [46] and patented in [47], the biasing arrangement is shown in figure 3.5.
Advantage of this method is that, it is more suitable for mixers operating at low supply voltage since there are no linearizing feedback components, which consume voltage headroom. But drawback is inability to cope with coupling mechanisms and distortion generated by the switching stage.

**Figure 3.5 Mixer core with IMD2-cancelling biasing circuit**

**c) Reduction of Common Mode Distortion with CMFB**

IMD2 can be reduced by exploiting common mode feedback loops in order to eliminate the flow of common mode distortion current through certain elements, which could otherwise convert it to differential mode. Two such methods have been proposed.

The first technique suppresses common mode IMD2 current flowing through the mixer load impedances by injecting almost equal amplitude but out-of-phase common mode distortion current to the outputs of the switching stage, as shown in figure 3.6 [48-49]. In this manner, load impedance
mismatches contribute negligibly to the effective mixer mismatch. The amount of injected current is determined by the CMFB loop gain. Mismatches between the current sources affect the performance of this technique. Since only low-frequency distortion has to be suppressed in downconversion applications, relatively large (and thus better matching) current sources can be used so in practice performance degradation is small.

**Figure 3.6  Distortion cancellation with common mode feedback loop**

This method has advantage that it reduces the common mode distortion. But drawback is that in low voltage CMOS mixers, it does not solve all mismatch problems.

The second technique eliminates flow of common mode IMD2 current through the switching stage of the mixer by either controlling the biasing current of the mixer input stage [50] or by injecting common mode distortion current to the outputs of the input stage [51]. In both cases, the amount of current is controlled by the common mode feedback loop, which senses
common mode distortion voltage at the output of the mixer. This method effectively eliminates mismatch terms associated with switching stage leakage mechanisms and output stage imbalances. However, RF-LO coupling mechanisms and switching stage nonlinearities are not addressed.

(d) Other Techniques

In addition to the techniques described above, several other IMD2 mitigation methods based on certain circuit configurations have been published. Since they are less general in nature than the previous techniques, only a brief description is provided below.

In [52], a method for reducing even order distortion based on injection of out-of-phase distortion current has been presented. It is suitable only for mixers implemented in bipolar technologies. The idea is to utilize an additional bipolar differential pair. The bases of transistors forming the differential pair are connected to the outputs of the mixer and inject out-of-phase even-order distortion current, reducing the flow of distortion current through the load resistors. Good suppression of distortion is achieved provided that additional transistors are appropriately scaled to produce the amount of distortion close to that generated in the mixer core. Process variations as well as varying operating conditions limit the effectiveness of this technique.

Even-order distortion caused by RF-LO crosstalk can be significantly reduced by using harmonic mixers. In such mixers, the RF input signal is mixed with one of the higher order LO harmonics, usually the second
harmonic. The impact of the fundamental LO harmonic coupling on second-order distortion mechanisms becomes much smaller. However, second-order distortion products due to active device nonlinearities and mismatches still exist. Moreover, harmonic mixers offer smaller conversion gain for a given current consumption than traditional mixers.

An interesting concept, which reduces the impact of indirect leakage and distortion generated by the switching stage due to parasitic capacitances loading the switching pairs, has been presented in [53] and [54]. The idea is to attach an inductor to the common source node of the switches so as to cause a parallel resonance with the parasitic capacitor at the LO frequency. The effective impedance loading the common source node of the switching pair is thus high, which decreases the indirect leakage and linearizes the switching devices via negative feedback. Consequently, they contribute less to the effective mixer mismatch. However, the mismatch coefficients associated with coupling and output stage mismatches remain unaffected by the proposed technique.

Apart from its inability to deal with RF-LO coupling, the concept of tuning out parasitic capacitances by means of integrated inductors entails several practical limitations. First, such solution can function only for mixers operating in a single band. In case of multiband transceivers, this would require having separate mixers for each band, which increases the total chip area. As inductors are usually used in low-noise amplifiers and oscillators, adding another set of these passive devices to the whole receiver would make
the whole implementation quite costly. Additionally, multiple inductors introduce undesirable cross-talk between distinct signal lines. When using strong LO signals to drive the mixer switches quickly (for instance in order to decrease their noise contribution), distortion sidebands around higher order harmonics are still demodulated to baseband without significant attenuation, lowering the effectiveness of the proposed technique.

### 3.3.3 Dynamic Matching

The concept of dynamic matching for direct downconversion mixers has been introduced in [55] and patented in [56]. The idea, shown in figure 3.7(a) [55], is equivalent to chopper stabilization technique, which has been successfully used to combat DC offsets and low frequency noise in operational amplifiers. Prior to downconversion with the LO signal, the input signal is multiplied with a mitigating signal $\mu(t)$ in a dynamic matching block. The mitigating signal can be either a periodic waveform or a pseudorandom signal. After the main mixing process, the second dynamic matching block restores the desired signal. The low frequency IMD2 distortion as well as static DC offsets generated in the main mixer are either frequency translated or spreaded, depending on the choice of the $\mu(t)$ waveform. In addition to reducing IMD2 distortion and static DC offsets, dynamic matching simultaneously reduces flicker ($1/f$) noise generated in the main mixer, making it an appealing solution especially for CMOS mixers.
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Figure 3.7  Dynamic matching

An example of dynamically matched mixer implementation is shown in figure 3.7 (b) [55]. Differential input signal in the current domain is mixed with the mitigating signal $\mu(t)$ using a double balanced switching mixer structure. Another double balanced structure is placed after the main mixing block (driven with the differential LO signal) and it restores the desired signal in the voltage domain.
Despite its valuable properties, dynamic matching possesses a number of drawbacks. First of all, mitigating signals generate interference at the fundamental frequency of $\mu(t)$ and its harmonics. Despite the fact that differential signaling, transmission line shielding and decoupling capacitors in signal buffers reduce on-chip crosstalk, taking the mitigating signals into account makes the whole design more complicated and consuming more power, which is a disadvantage by itself. Besides, although the LO and mitigation signals can be derived in a simple way from a single reference source by means of odd and even integer dividers (e.g. odd dividers used to generate mitigation signals while even dividers used for LO signal generation), such approach may result in mixing of even division multiple frequency component with odd division multiple frequency component causing a spurious response which self quiets the receiver. This deficiency has been pointed out in [57], which also suggests a solution to the problem: the use of mitigation signals with frequency having a non-integer relationship to the LO frequency (or the reference oscillator signal frequency). This can be accomplished for example with the help of a direct digital synthesizer (DDS) unit. However, it further increases the complexity of the whole system. Another weakness of dynamic matching is poor noise performance of the whole block, since each mitigating stage introduces loss associated with the mixing process. This amplifies the noise contribution of the following stages when referred to the input. Finally, dynamic matching units are also nonlinear and prone to mismatches. Their contribution to the overall even-order distortion can be lowered by implementing them as highly linear passive mixers and driving them with
mitigation signals at low frequencies, where fast switching reduces the direct leakage while memory effects due to parasitic capacitances are less troublesome.

### 3.3.4 IMD2 Compensation

Instead of trying to eliminate sources of distortion, another approach is to try to eliminate the distortion itself [58]. This is where the idea of distortion compensation comes in. It can be loosely defined as addition of an out-of-phase, appropriately scaled distortion generated by a dedicated block featuring second order nonlinearity to the output signal of an RF block.

In case of downconversion mixers, input signals of the squaring block can be actually obtained either from the mixer input as shown in figure 3.8(a) as proposed in [59] or from the mixer output, before filtering of the interferers takes place, as shown in figure 3.8(b) and proposed in [60]. In the latter approach, a high pass filter removes the wanted signal while passing through the down-converted, out-of-channel interferers. Generally, separate gain settings are required for I and Q channels. The functions of the compensation system, including generation of reference distortion, scaling and subtraction can be carried out either in the analog or in the digital domain. When implemented in the analog domain, squaring operation can be performed by an instantaneous power detector. If implemented in the digital domain, squaring can be carried out by a dedicated multiplier. Since interfering signals are
relatively large, high resolution analog-to-digital converters in the reference path are not required. One more implementation is proposed in [61].

Figure 3.9 Methods for reference signal generation

Compensation techniques possess a number of drawbacks. First, they rely on accurate modeling of even-order nonlinearity of the RF block of interest. Since the modeling usually takes only second order distortion into account, compensation provides improvement for relatively small input
blocker levels while it fails at high input levels, where higher even order intermodulation terms become important. Next, compensation schemes require an additional path with its own set of filters, increasing the required area for on-chip integration. Moreover, an additional analog to digital converter is required if the compensation is to be carried out using digital techniques. In addition, extra equalization is required for matching of main path and auxiliary path, which in turn increase the complexity of the whole system.

3.3.5 Calibration Techniques

Instead of generating reference distortion by an additional block, it is possible to exploit common-mode distortion generated by existing circuit nonlinearities. Called calibration method, such self-compensating technique alters the RF block of interest by controlled tuning of component parameters in order to convert part of common mode distortion to differential mode. Since no additional active circuitry generating reference distortion is necessary, considerable savings in power consumption are possible. Moreover, the problem of modeling accuracy of the RF block nonlinearity is eliminated. For these reasons, calibration techniques are the most robust IMD2 cancellation techniques.

In [62], intentional mismatches were introduced between the biasing currents of the double-balanced harmonic mixer. In this way, operating points of the switching transistors were changed, affecting mismatches associated
with the indirect leakage and switching stage distortion. Although for such mixer the effective IMD2 mismatch is not exactly given, the idea is applicable also to standard Gilbert cell like mixers. In [63], a concept based on tuning the biasing voltages of both switching pairs separately was proposed. Therefore, values of mismatch coefficients associated with the first switching pair as well as the second switching pair could be decreased, cancelling simultaneously the impact of input stage transconductance mismatches. However, existence of other mismatches introduces undesirable ambiguity about which switching pair mismatch to tune in order to compensate for coupling or output stage imbalances. Similar tuning concepts were suggested in patents [64] and [65], suffering from the same drawbacks.

In [66], a remark was made that it is not necessary to change mismatches of both switching pairs separately. Accordingly, a technique based on introducing intentional mismatch in only one switching pair was proposed. Yet another approach to intentional mismatching of the switching stage by placing additional devices in parallel with the main switches was patented in [67]. The concept is more complicated than other IP2 tuning approaches described above. Moreover, it may degrade performance of the mixer because additional switches add their own parasitic capacitances loading the switching pairs.

In [68], calibration techniques for current mode output mixers were discussed, including input and switching stage mismatching. Although in the publication it was stated that introducing mismatch between linear
transconductances by tuning the biasing current flowing through each branch of the input stage has an impact on IIP2, it is strongly believed that in fact biasing current mismatching affects IIP2 much more through related switching stage mismatches than through input stage transconductance mismatches. Tuning circuits introducing mismatches in the output stage of the mixer can be divided into two groups: load resistance tuning circuits and common mode feedback loop mismatching circuits. In traditional voltage mode output mixers which don’t employ output common mode feedback loops, controlling mismatches between the load resistors has proven effective. A so-called load balancing technique, which can be implemented by connecting a bank of large resistors in parallel to the main loads, was presented in [69] and patented in [70]. The second output stage IP2 tuning method introduces mismatches in the CMFB loop, provided that the mixer is equipped with such block. Variants of this technique were patented in [71] and [72] and also shown in [73] and [74].

3.3.6 Automatic IMD2 Cancellation

An important issue associated with above methods is how to appropriately scale the reference distortion (in case of IMD2 compensation) or adjust the settings of tuners (in case of calibration) so that the required suppression of second order intermodulation distortion is achieved.

Performing necessary adjustments as a part of post-production testing with the aid of dedicated measurement setups is not desirable for cost and
time reasons. A much more preferable solution is to let every fabricated receiver chip adjust itself automatically, without the help of external equipment. And this leads to development of automatic IMD2 cancellation techniques.

One of the first publications dealing with the subject of automatic IMD2 cancellation was [75]. The calibration settings were set based on measurements of received signal strength indicator (RSSI) and frame error rate (FER) for a specific communication system (PHS). Such approach is not advantageous for two reasons. First, it is a system dependent concept, requiring interaction with receiver blocks performing higher level system tasks including FER and RSSI measurements. Although technically feasible, this concept may be difficult to implement for organizational reasons since RF transceivers are usually developed independently from the baseband units. Secondly, the method presented in [75] is rather slow as signal level measurements take time and it may happen that they have to be performed for the whole range of tuning codes.

In [76], a general automatic calibration scheme was patented based on exciting the receiver with an AM modulated interferer having a known AM content and correlating the response with that AM content in order to determine necessary adjustments of one of the differential branches at the output of the down-converter. In another patent, supplying test tones and measuring the resulting DC offsets was proposed to determine the scaling
factor of compensating currents injected to the differential outputs of the mixer [77].

A technique based on reusing existing transceiver building blocks for generation of out-of-band tones and AM modulating them using simple switches was proposed in [78]. The optimum IP2 tuner code was selected as the one corresponding to minimum interference level measured at the receiver output. Another method based on supplying an out-of-band test tone and detecting dynamic part of the output DC offset was presented in [79] and patented in [80] and [81].

3.3.7 Adaptive IMD2 Cancellation / Calibration

Adjustments of certain parameters in test signal based compensation and calibration schemes can be done only when the receiver is in idle mode, i.e. when it does not receive any desired signal. This happens for example just after power-up of the transceiver. However, it has been shown that the effective mixer mismatch varies with changes in operating conditions. For example, it depends on temperature, which changes rapidly after power-up. Thus, maintaining sufficient IMD2 suppression requires updating reference distortion gain coefficients (in case of IMD2 compensation schemes) or tuning codes (in case of calibration).

Such updates can be performed with test signals in a periodic manner. However, determination of when and how often to trigger the adjustment procedure requires careful planning on a system level. Moreover, in case of
wireless systems employing continuous transmission and reception, performing necessary adjustments periodically may be impossible. Thus, it is desirable to employ IMD2 cancellation techniques which are system independent and allow to keep satisfactory IMD2 suppression even with simultaneous reception of the desired signal. As a possible solution, statistical adaptive signal processing techniques can be utilized. Such methods have already proven useful in wireless receivers, for example in compensating static DC offsets and I/Q imbalances [37], [82], [83]. They are also an interesting option for IMD2 cancellation for a number of reasons.

The adaptive IMD2 compensation scheme was proposed in several publications. In [84], the reference signal was obtained by squaring the output signal of the mixer. For good cancellation of distortion, a multi-tap finite impulse response (FIR) equalizer was used, which combined the functions of scaling the amplitude of reference distortion and equalizing the main and reference signal paths. In [85], the reference second order distortion signal was obtained from the low-pass filtered output signal of the low-noise amplifier, thereby avoiding the use of a dedicated squaring block. However, the issue of nonlinear behavior of LNA and downconversion mixer at high input signal levels, where higher order distortion products start to play an important role, was not addressed. Publications [86] and [87] propose again the use of reference signal path, but apply it to cancel not only even order distortion but also odd order distortion. Other adaptive calibration techniques are proposed in [88-91].
3.3.8 Architectural Method

All the techniques discussed above can be classified in two categories. The first method uses an additional circuit for mixer non-linearity reduction. The second one uses an additional analogue or digital circuit for the IMD2 cancellation. Both solutions use dedicated software and hardware to reduce IMD2. The zero-IF receiver loses its advantages of high integration, simplicity and low cost by using these techniques. One possible solution is an architectural solution for distortion cancellation. One approach is to use five port architecture [92-94]. But, five port architecture consumes one more mixer and one more low-pass filter compared to conventional DCR. Therefore, use of five port receiver increases the component count and decreases the simplicity and compactness of the receiver. In this approach one more mixer will be added, which results in the addition of noise in the receiver. Other approach is to modified classical homodyne architecture [95].

3.4 I/Q mismatch reduction techniques

Gain and phase mismatch in the I and Q path will result in the I/Q mismatch problem. Gain error appears as a non unity scale factor in the amplitude, while phase imbalance corrupt one channel with a fraction of the data pulses in the other channel. This problem is a major obstacle in discrete designs, but it tends to decrease with higher levels of integration. Wireless devise for Cognitive – Radio (CR) application are very sensitive to this
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problem. Therefore, major concentration has been put on the removal of this problem nowadays.

Solutions based on digital signal processing offer better accuracy, a higher degree of freedom and the possibility to compensate for several impairments while consuming less chip area.

Several techniques are proposed in the literature to solve the I/Q mismatch problem. These approaches can be classified in two broad categories as below.

3.4.1  Data-aided approaches

Data-aided approaches [96]–[102] rely on known pilot or training sequences in the transmitted signal for compensation. Thus these approaches are strongly-standard dependent.

![Figure 3.10  Data-aided I/Q imbalance compensation in wideband DCR](image)

The received reference symbol contains the impairments of the entire signal processing chain in transmitter and receiver. Thus, such methods are suitable for the combined mitigation of several impairment sources. [96], [97]
compensate I/Q mismatch with least square approaches in combination with frequency offset as shown in figure 3., and [98], [99] together with channel estimation. [100] treats frequency dependent I/Q mismatch on the transmitter side and [101], [102] both on transmitter and receiver side in MIMO systems. With data-aided methods fast convergence and good performance are achieved, but often at the cost of high computational complexity, especially if considering compensation of frequency dependent I/Q mismatch.

### 3.4.2 Blind approaches

Blind methods [103-81] need longer convergence times but they are standard independent and thus are more flexible in its use. As changes of the I/Q mismatch due to temperature and aging appear slowly, convergence time is of less importance. Typically, blind methods rely on statistical properties of the influenced signal. As an example, [104] uses the statistical independence between the desired signal and its mirror image for frequency independent I/Q mismatch compensation in low-IF receiver by blind signal separation. Thus, such techniques can be used to selectively combat a single type of impairment. In OFDM-based systems the blind determination of the compensator coefficients in the receiver is possible before or after the FFT. In [105] a gradient descent search method in time domain and a frequency domain approach based on a single-tap matrix inversion for frequency dependent I/Q imbalance compensation is presented. [83] provides advanced blind source separation (BSS) techniques for frequency independent I/Q imbalance compensation in MIMO systems and [106] shows the same for the frequency
dependent case by using higher-order statistics (HOS) in an independent component analysis (ICA). A general feature of BSS techniques is their high implementation effort. A second-order statistical property which is destroyed by frequency dependent I/Q mismatch is properness, while in the frequency independent case it is called (second-order) circularity. It enables I/Q mismatch compensation algorithms in the time domain of the receiver in OFDM-based systems, which recover circularity [107-109] or properness [110-112] when the received signal is proper. The authors in [113-115] propose a Newton method based on a fourth-order moment for measuring frequency independent I/Q mismatch in communication signals used in LTE. This approach shows faster convergence speed compared with corresponding second order approaches. Compensation in the frequency domain [109], [113] can be realized only with significantly higher complexity, because the required statistics is subjected to a Fourier transformation. All of those blind I/Q mismatch compensation algorithms based on properness with complex signal models according to [110] are summarized in the following by the term complex-valued compensators (CVCs). In [116] such a CVC was introduced also on transmitter side for frequency dependent I/Q mismatch calibration. A realistic model for frequency dependent I/Q mismatch in currently used transceiver circuits is a finite impulse response (FIR)-filter with a few coefficients (typically 3-5) [117-118]. Less implementation effort can be achieved by using only real-valued filters for I/Q mismatch compensation according to [113], which we summarize under the term real-valued compensators (RVCs). The sensitivity of the objective function for I/Q
mismatch compensation algorithms with respect to other impairments needs to be taken into account for a practical implementation. [110],[119] examine the influence of several impairments on the properness. This properness property is efficiently utilized in [120]. Other I/Q imbalance estimation and removal methods are described in [121-123].

In conclusion, data-aided methods give fast convergence and good performance but at the cost of high computational complexity and dependency on standard utilized. Blind methods are standard independent and so give more flexible in its use, but need longer convergence times and sometimes high implementation effort.

### 3.5 Chapter Conclusion

All the above techniques are suffering from one or more of the below problems.

1. **Increase in the number of components in RF section.**
2. **Complex algorithm required to implement in the DSP section, results in the high computational loading.**
3. **Nullify only one distortion.**
4. **Required to modify the existing standard.**

As a solution to these problems, an architectural approach has been adopted here to propose simple architectural solutions which nullify multiple
distortions simultaneously. Here a novel method with self calibration strategy is suggested to nullify multiple distortions in DCR. Self calibration method makes this approach standard independent and simple algorithm makes it fast and less computationally complex. Comprehensive simulated and practically measured results are presented to indicate the effectiveness of the proposed method.