

# Error Vector Magnitude as a Figure of Merit for CDMA Receiver Design

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**Abstract:** Standards for modern digital CDMA communication systems specify limits on the error vector magnitude (*EVM*) of the transmitted signal at the transmitter, whereas performance at the receiver is characterized by the bit error rate (BER). Traditionally, RF receiver specification defines limits on amplitude- and phase ripple over a certain bandwidth, IP3, IQ-mismatch, DC-offset, etc. Unfortunately, these parameters lack an obvious relation to BER.

In this paper, we show that the *EVM* offers remarkable advantages when used as a conceptual figure of merit to characterize the distortions caused by the elements of the receiver's RF front-end. The use of the *EVM* as a parameter for the optimization of the RF receiver and for the description of the RF-BB (radio frequency-baseband) interface leads to a simple design process and avoids over-specification. For CDMA systems, we give an explicit relation between the *EVM* and the BER. Furthermore, the benefits of using the *EVM* in RF receiver design are illustrated by examples and verified by simulations.

## 1. Introduction

RF design engineers are faced with the need to provide specifications for the components or sub-components of their transmitter and receiver chains. This is a challenging task since modern digital wireless communication systems usually impose limits on allowed bit error rate and block error rate (BLER). Unfortunately, most of the traditionally used parameters like IP3, amplitude- and phase ripple, IQ-mismatch, DC-offset etc. lack an obvious relation to BER or BLER. Dealing with such a large amount of parameters often leads to over-specification of RF components.

Standards for modern communication systems define limits on the error vector magnitude of the *transmitted* signal at the air interface [1][2]. In [3], a classification of the mechanisms contributing to the *EVM* is presented. It is pointed out that in a communication link, the mechanisms contributing to *EVM* and BER are practically identical. However, the exact conditions of validity and an explicit relationship between the *EVM* and the BER are not given. Moreover, aspects of *receiver* design, like RF impairments that occur in the RX path only, such as noise or intermodulation, are also not addressed.

In this paper, we show that for a CDMA receiver a relatively simple relation between *EVM* on the one hand and BER or BLER on the other hand exists. The

reason is that demodulation of the received CDMA signal involves correlation with pseudo noise codes, which transforms, by the central limit theorem, any error of the received signal into a Gaussian random variable. Therefore, any RF impairment can be replaced by an additive white Gaussian noise source at the receiver input. The respective noise power, however, depends on the type of distortion and the test signal.

The benefits of using the *EVM* as a figure of merit (FoM) is that the set of parameters describing an RF component is reduced to one single parameter. This avoids over-specification and offers more degrees of freedom to the designer. Moreover, rules exist to describe the *EVM* of a cascade of RF components. The *EVM* also provides a simple means to optimize an RF receiver system and to describe the RF-baseband (RF-BB) interface. This allows to design and verify the RF front-end independently from the baseband part.

The paper is organized as follows. In Section 2 the benefits of *EVM* as a specification parameter will be highlighted by means of two examples. Section 3 describes the CDMA communication system considered. Section 4 introduces the mathematical definitions of the *EVM* and some related measures. In Section 5, semi-analytical analysis of the impact of different RF impairments on the system performance is developed. The relation between the *EVM* and the so-called code domain SNR on one side, and the BER or BLER degradation on the other side is pointed out. Section 6 provides simulation results which verify the analysis for W-CDMA standard test cases. The conclusions are given in Section 7.

## 2. Examples for Benefits of *EVM* Specification

### 2.1. Filter Design

As a first example, an analog surface acoustic wave (SAW) channel filter is considered. It is usually characterized by the parameters like stop-band attenuation, adjacent channel selectivity (ACS), insertion loss, and amplitude-, phase- or group-delay-ripple. The latter three describe the quality of the pass-band and are specified over a certain bandwidth. This bandwidth is a crucial factor for the optimum trade-off between ACS and the linear distortions. Unfortunately, the relation between the optimum bandwidth and the bit error rate is not obvious. Often the bandwidth of the pass-band is specified wider than really necessary. As a consequence, the suppression of the neighbor channels is diminished considerably while gaining almost nothing regarding BER-performance.

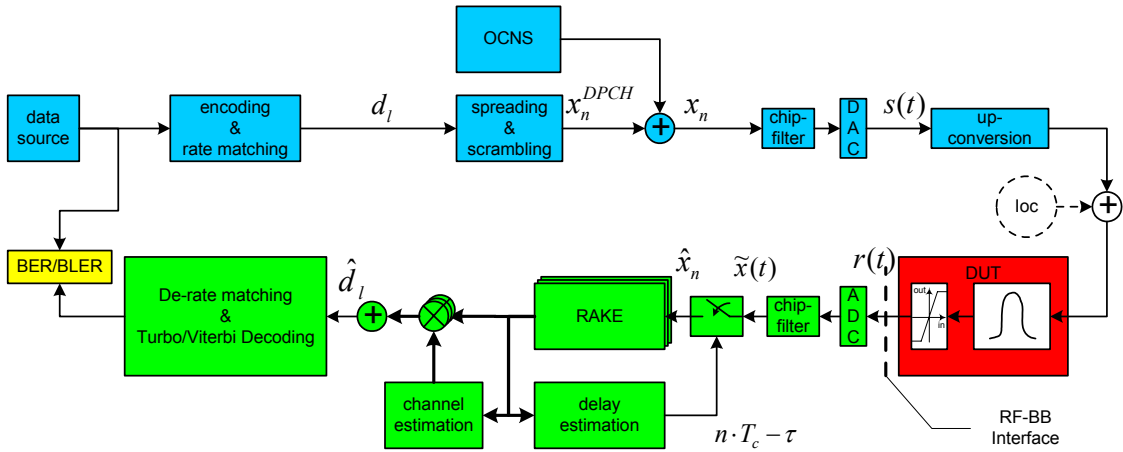


Fig. 1. CDMA system block diagram.

Using EVM and the noise gain (defined later) instead, the pass-band of a filter is completely characterized by parameters that are directly related to the resulting BER or BLER [5]. Specification work for the RF designer becomes much easier and the SAW filter designer has more degrees of freedom to find an optimum shape of the transfer function in terms of BER and ACS. This leads to cost- and size-optimized SAW filter solutions.

## 2.2. Simplification of RF system design

In a further step, the components are integrated into the RF receiver. System optimization is normally associated with a very time-consuming BER analysis using a test environment like the one depicted in Fig. 1. Let's assume a simplified RF front-end comprising a filter causing distortion  $filt\_dist$  and an additional DC-offset. A typical specification would look like

$$\begin{aligned} |filt\_dist| &\leq filt\_dist_{max} \\ |offset| &\leq offset_{max} \end{aligned} \quad (1)$$

As the interaction between filter distortion and the DC-offset is normally not known, the choice of the first limit, say  $filt\_dist_{max}$ , is quite arbitrary. The second limit can then be derived assuming worst-case conditions for the first parameter. The acceptable parameter space according to this specification is shown on the left hand side of Fig. 2.

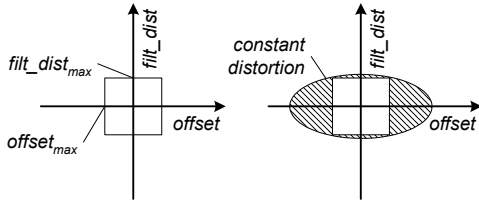


Fig. 2. Acceptable parameter space based on traditional (left) and EVM-based (right) specification.

Based on the EVM of the cascade, as a single parameter to describe the whole front-end, the device under test (DUT) is in fact characterized by an "allowed parameter subspace" which has ellipsoidal shape as depicted on the right hand side of Fig. 2 (see also Sec. 5). Obviously, exploitation of the EVM as a

single figure of merit for the RF-BB interface specification helps avoiding unnecessary over-specification (i.e. specification stricter than required).

It might furthermore even happen that substitution of a sub-component by an *improved* version results in an apparent BER *degradation* within the test- or simulation environment. In case of a channel filter this effect occurs e.g. if the basic filter delay does not fit to the discrete sampling grid of the RAKE receiver. In normal operating conditions with random propagation delays, however, the improvement would become evident. Even more simulations may be necessary to detect this phenomenon.

Using the EVM as an RF-BB interface description would allow to design and verify the RF front-end independently of the baseband module. Cascading rules permit the derivation of the EVM of the system by the EVM of its subcomponents. Perfect time synchronization is already included in the EVM definition. Thus, a clear separation between RF and BB impairments can be achieved.

## 3. System Description

The system description in this section is common to all CDMA systems, but it refers in particular to the UMTS/IMT2000 third generation radio standard. Fig. 1 shows the block diagram of the downlink of a CDMA system. The base-station transmitter is represented by a data source followed by encoding and spreading blocks. An "orthogonal channel noise simulation" (OCNS) block represents the signals of other interfering users with mutually orthogonal channelization codes. The resulting chip sequence is fed to a root-raised-cosine (RRC) chip filter for spectrum shaping and is finally digital-to-analog converted.

The RF part of the terminal receiver is the device under test in the context of this paper. The digital part of the receiver follows the RF-BB interface. Major components are the RAKE receiver and the channel decoder.

The mathematical relations behind the block diagram in Fig. 1 are listed below. Some simplifications are made to improve readability.

In order to whiten the transmit signal, the user symbols  $d_l$  of the desired user (DPCH) are spread by a factor  $SF$  and scrambled

$$x^{DPCH}_{k+l \cdot SF} = s_{k+l \cdot SF} \cdot c_k \cdot d_l, \quad k = 0, \dots, SF-1, \quad (2)$$

where  $s_n$  denotes a pseudo random scrambling sequence and  $c_n$  the channelization code. Both sequences are taken from the QPSK modulated alphabets  $\{1, j, -1, -j\}$ . Their product is abbreviated by  $s_{k+l \cdot SF} \cdot c_k = s_{c_{k+l \cdot SF}}$

Since the code domain is exploited for multiple accesses the multi-user chip sequence becomes

$$x_{k+l \cdot SF} = s_{k+l \cdot SF} \cdot \left( c_k \cdot d_l + \underbrace{\sum_{j=2}^J c_k^j \cdot d_l^j}_{OCNS} \right), \quad (3)$$

where  $c_k^j$  denote the mutually orthogonal channelization codes, which are used in the downlink to minimize multiple access interference. The interfering signals sum up to give the OCNS.

The chip sequence in (3) is sent through a pulse-shaping filter, which limits the spectrum to a certain system bandwidth. The continuous time baseband transmit signal becomes therefore

$$s(t) = \sum_l \sum_{k=0}^{SF-1} x_{k+l \cdot SF} \cdot RRC(t - (k+l \cdot SF) \cdot T_c), \quad (4)$$

where  $T_c$  is the chip duration. This complex baseband signal is up-converted to the carrier frequency and transmitted. Following [2], the total spectral density within the system bandwidth including OCNS at the input and output of the physical propagation channel<sup>1</sup> is denoted by  $I_{or}$  and  $\hat{I}_{or}$ , respectively. Depending on the actual test condition, white Gaussian noise with power spectral density  $I_{oc}$  is added.

The ideal receive processing starts with down conversion to  $r(t)$  and chip pulse matched filtering

$$\tilde{x}(t) = \int_{\tau} RRC(t - \tau) \cdot r(\tau) \cdot d\tau. \quad (5)$$

$\tilde{x}(t)$  is modeled as time continuous here because, despite of digital implementation, sufficient over-sampling is normally used such that reconstruction of the time continuous function based on the sampling theorem is still possible.

Next steps are down-sampling to the chip duration and de-spreading

$$\hat{x}_{k+l \cdot SF} = \tilde{x}((k+l \cdot SF) \cdot T_c), \quad (6)$$

$$\hat{d}_l = \sum_{k=1}^{SF} \hat{x}_{k+l \cdot SF} \cdot s_{c_{k+l \cdot SF}}^* \quad (7)$$

This describes the simplest operation mode of the RAKE receiver in Fig. 1 for a directly connected transmitter.

<sup>1</sup> The basic RF test cases do not employ a channel simulator. Therefore, the corresponding model is not shown in Fig. 1.

## 4. Definitions

This section gives the definitions for two measurement quantities to describe RF components as well as the RF-BB interface: The power signal-to-noise ratio, and the *EVM*. Some aspects of these measurements are discussed. In addition, the *code domain SNR* after the RAKE receiver and the transmit power ratio are introduced.

### 4.1. Power Signal-to-Noise Ratio

The power signal-to-noise ratio  $SNR_{RF-BB}^{power}$  is based on two power measurements at the RF-BB interface for different test conditions.

First, only the non-desired signal is measured. This is noise, OCNS or interference from other frequencies. The respective power measurement at the interface is denoted as  $P_N$ . Second, the desired signal is turned on and the new power measurement is denoted as  $P_{S+N}$ .

The power signal-to-noise ratio is then computed from these two measurements as follows:

$$SNR_{RF-BB}^{power} = \frac{P_{S+N} - P_N}{P_N}. \quad (8)$$

For narrow-band systems such as GSM, where the user separation is almost done on RF side and the phase error sensitivity is low, this is a reasonable interface specification parameter<sup>2</sup>.

Unfortunately,  $SNR_{RF-BB}^{power}$  is not suited as specification parameter for the CDMA system, which becomes immediately clear if we consider a DUT which introduces phase distortions like a SAW-filter or latch effects in non-linear components. Although the CDMA system is sensitive to such distortion, the power signal-to-noise ratio measurement is *not affected* by phase distortion. Additional restrictions on phase distortion are required therefore. This, however, leads again to over-specification as pointed out above.

### 4.2. Transmit Power Ratio

From system point of view, the *transmit power ratio* of a single user to the total transmit power at the base station is important because it is directly linked to (downlink) system capacity [6]. This parameter also specification [2], where it is called  $DPCH\_E_c/I_{or}$ . Based on the CDMA code properties, the *transmit power ratio* can be written as

$$\begin{aligned} \frac{DPCH\_E_c}{I_{or}} &= \frac{E\left\{ \left| x^{DPCH}_{k+l \cdot SF} \right|^2 \right\}}{E\left\{ \left| x_{k+l \cdot SF} \right|^2 \right\}} \\ &= \frac{E\left\{ \left| d_l \right|^2 \right\}}{E\left\{ \left| d_l \right|^2 \right\} + \sum_{j=2}^J E\left\{ \left| d_l^j \right|^2 \right\}} \end{aligned} \quad (9)$$

### 4.3. Error Vector Magnitude

The *EVM* of the observed signal  $r(t)$  is defined as the least squares fit of the reference sequence defined

<sup>2</sup> Conceptually, the power signal to noise ratio at a GSM RF-BB interface it is comparable to the code domain SNR of the CDMA system.

in (3) to the observed waveform after chip matched filtering and *synchronized* sampling (see also [1]).

$$EVM = \sqrt{\frac{\sum_{n=1}^N |\hat{x}_n^{opt} - \alpha^{opt} \cdot x_n|^2}{\sum_{n=1}^N |\alpha^{opt} \cdot x_n|^2}}, \quad (10)$$

$$\hat{x}_n^{opt} = \hat{x}_n(\tau^{opt})$$

$$[\alpha^{opt}, \tau^{opt}] = \arg \min_{\alpha, \tau} \sum_{n=1}^N |\hat{x}_n(\tau) - \alpha \cdot x_n|^2$$

$$\hat{x}_n(\tau) = \tilde{x}(n \cdot T_c - \tau)$$

where “synchronized sampling” refers to sampling optimized over  $\tau$  according to (10). The remaining error vector  $\mathbf{e} = [e_1 \dots e_N]^T$  with  $e_n = \hat{x}_n^{opt} - \alpha^{opt} \cdot x_n$  is orthogonal (uncorrelated) to the desired signal vector  $\mathbf{x} = [x_1 \dots x_N]^T$  due to the least squares optimization<sup>3</sup>. The power relation between desired signal and error can be written as follows:

$$EVM = \sqrt{\frac{\|\mathbf{e}\|^2}{\|\alpha^{opt} \cdot \mathbf{x}\|^2}} = \sqrt{\frac{P(e_n)}{P(\alpha^{opt} \cdot x_n)}}. \quad (11)$$

In Sec. 5 we show that any error (uncorrelated with the desired signal) is mapped to normally distributed noise at the RAKE output. Thus, reception of an observed signal with a certain *EVM* is equivalent to reception of distortion-free signal in additive white Gaussian noise with

$$SNR_{RF-BB}^{EVM} = \frac{1}{EVM^2}.$$

#### 4.4. Code Domain SNR

The *code domain SNR* is defined as the second order statistics at the RAKE receiver output, which is also the decoder input

$$SNR_{code} = \frac{|\mathbb{E}\{\hat{d}_i\}|^2}{\text{VAR}\{\hat{d}_i\}}. \quad (12)$$

### 5. Impact of RF Impairments on System Performance

RF impairments can in general be classified into *linear*, *non-linear* and *random* [3]. Their assessment is mostly done under static single path propagation (i.e. AWGN), for which the worst-case impact of RF impairments has to be expected. This is also reflected in the RF test conditions specified in [2].

In the sequel, we show that for all types of RF impairments the decision variables  $\hat{d}_i$  are Gaussian random variables. For CDMA systems, this arises due to the central limit theorem. A relationship between *EVM* and code domain SNR can always be established. The decoder performance is well-known for the AWGN channel and tabulated in typical “water fall” curves, which relate the BER and BLER to the code domain SNR and some coding parameters [6]. Thus, a

relationship between the *EVM* on the one side and the BER or BLER on the other side can be derived.

#### 5.1. Linear Distortion

Linear distortion of the DUT is described by

$$r(t) = \int_{\tau} \tilde{f}(t-\tau) s(\tau) \cdot d\tau + \tilde{o}_{ffset}, \quad (13)$$

where  $\tilde{f}(t)$  is the continuous time impulse response of the linear DUT and  $\tilde{o}_{ffset}$  its DC-offset. As usual in time discrete realizations, the time continuous relation can be modeled by a corresponding time discrete representation

$$\hat{x}_n^{opt} = \sum_m f_{n-m} x_m + o_{ffset}, \quad (14)$$

where  $f_m$  is the sampled total transfer function of the system, and  $RC(t)$  is the raised cosine impulse response

$$f_m = f(mT_c) = \int_{\tau} \tilde{f}(mT_c - \tau) \cdot RC(\tau) \cdot d\tau \quad (15)$$

$$RC(t) = \int_{\tau} RRC(t-\tau) \cdot RRC(\tau) \cdot d\tau$$

In the following, optimum synchronization as described in Sec. 4 is assumed and the noise source in Fig. 1 is the white Gaussian noise with power spectral density  $I_{oc}$ . It is filtered by the DUT and the receive RRC-filter with combined impulse response

$$h(t) = \int_{\tau} \tilde{f}(t-\tau) \cdot RRC(\tau) \cdot d\tau. \quad (16)$$

The (stationary) noise power of the filtered noise process becomes

$$P_n = I_{oc} \int_{-\infty}^{\infty} |h(t)|^2 dt = I_{oc} \int_{-\infty}^{\infty} |H(f)|^2 df. \quad (17)$$

The symbol estimates in (7) become

$$\hat{d}_i = \sum_{k=1}^{SF} \hat{x}_{k+l-SF}^{opt} \cdot s_{k+l-SF}^* =$$

$$d_i \cdot f_0 \cdot \sum_{k=1}^{SF} |s_{k+l-SF}|^2$$

$$+ f_0 \cdot \sum_{k=1}^{SF} OCNS \cdot c_k^* \cdot |s_{k+l-SF}|^2 \quad (18)$$

$$+ \sum_{m \neq 0} \sum_{k=1}^{SF} f_m \cdot (d \cdot c + OCNS) \cdot s_{k-m+l-SF} \cdot s_{k+l-SF}^*$$

$$+ \sum_{k=1}^{SF} o_{ffset} \cdot s_{k+l-SF}^*$$

$$+ \sum_{k=1}^{SF} n_{k+l-SF} \cdot s_{k+l-SF}^*$$

The first term in (18) contains the desired signal  $d_i$  weighted by the constant factor  $f_0 \cdot \sum_{k=1}^{SF} |s_{k+l-SF}|^2 = f_0 \cdot SF$ . The second term  $f_0 \cdot \sum_{k=1}^{SF} OCNS \cdot c_k^* \cdot |s_{k+l-SF}|^2$  is zero since the OCNS and the desired user's channelization code  $c_k$  are

<sup>3</sup> Superscript  $T$  denotes transposition.

orthogonal. The sequence  $sc_{k+SF-1}^*$  as well as the product sequences  $s_{k+m+1-SF} \cdot sc_{k+1-SF}^*$ ,  $m \neq 0$  have pseudo random property. The adds of the third and the fourth term are therefore binomially distributed and mutually independent<sup>4</sup>. For *sufficiently high spreading factor*, each term can be well approximated by a normal distribution. Finally, the fifth term  $\sum_{k=1}^{SF} n_{k+1-SF} sc_{k+1-SF}^*$  represents the contribution of the filtered noise to the symbol estimates. Due to scrambling with the pseudo random code the initially colored and therefore correlated noise samples  $n_{k+1-SF}$  become uncorrelated and normally distributed after passing the RAKE receiver. This shows that the estimated symbols are approximately Gaussian distributed and uncorrelated with adjacent symbols. Even the "deterministic DC-offset" is mapped to a random signal in the code domain. Thus, independently of a particular implementation, a simple AWGN channel between the encoder and decoder can replace the total CDMA transceiver. The AWGN channel is characterized by a *single* parameter, the *code domain SNR*. From (12) and (18) the *code domain SNR* can be computed as

$$SNR_{code} = \frac{DPCH\_E_c \cdot SF \cdot |f_0|^2}{I_{or} \sum_{m \neq 0} |f_m|^2 + |o_{ffset}|^2 + I_{oc} \int_{-\infty}^{\infty} |h(t)|^2 dt} = \frac{DPCH\_E_c \cdot SF}{I_{or} \cdot EVM_{total}^2} \quad (19)$$

where  $EVM_{total}$  is the cascaded  $EVM$  of linear distortion, offset and noise.

$$EVM_{total}^2 = \underbrace{\sum_{m \neq 0} |f_m|^2 / |f_0|^2}_{EVM_{filter}^2} + \underbrace{|o_{ffset}|^2 / (I_{or} |f_0|^2)}_{EVM_{offset}^2} + \underbrace{I_{oc} / I_{or} \cdot \int_{-\infty}^{\infty} |h(t)|^2 dt / |f_0|^2}_{EVM_{noise}^2} \quad (20)$$

The  $EVM_{filter}$  of linear filters according to [5] was inserted. Eq. (20) shows a typical cascading rule for  $EVM$ . If the processes causing the distortions are statistically independent, the cascaded  $EVM_{filter}^2$  is the quadratic sum of the partial  $EVMs$  (see also [4]).

The decoder performance is well-known for the AWGN channel and tabulated in curves, which relate the BER and BLER to the code domain SNR, and some coding parameters[6], respectively. Based on (19), the  $EVM$  of linear distortions is therefore directly linked to BLER. Note that the latter is not given for impairment parameters like amplitude or phase ripple.

Eq. (19) also verifies the relation in Fig. 2 where

$$filt\_dist = \sqrt{\sum_{m \neq 0} |f_m|^2} / |f_0| \text{ is the filter } EVM.$$

<sup>4</sup> This is not affected by the fact that two adjacent symbols contribute to each other for non-zero filter lag ( $f_m |_{m \neq 0}$ ).

## 5.2. Non-Linear Distortion

The effect of non-linear distortion depends on the signal level and amplitude statistics and can be memory-less or not. To simplify the discussion, we concentrate on a rather extreme non-linear effect

$$r(t) = 1 \cdot e^{j \cdot angle(s(t))} \quad (21)$$

This describes a limiter amplifier with constant output magnitude 1 independently of the input signal. It is well known that constant magnitude modulation schemes are robust against this kind of clipping. A multi-code CDMA signal, however, exhibits considerable amplitude variation as shown in Fig. 3.

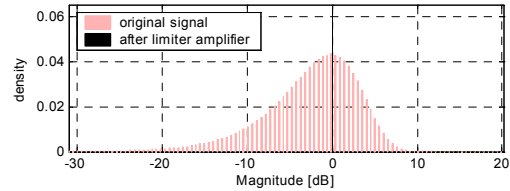


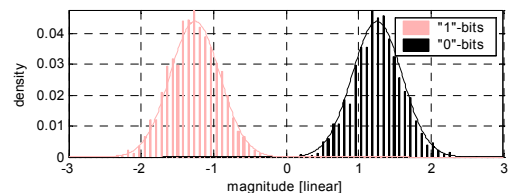
Fig. 3. Magnitude distribution of filtered multi-code CDMA signal before and after limiter amplifier.

The original signal in Fig. 3 consists of 128 orthogonal codes for spreading factor 128. They are used for multiplexing equal power users. The transmit signal constellation above is close to the "maximum input level" test case in [2]. No additional noise has to be taken into account since "maximum input level" occurs in the near vicinity of a base station with active connection to the terminal, where thermal noise or interference from other base stations can be neglected.

The more users are multiplexed, the more amplitude modulation and  $EVM$  arises. At the same time, the interference in the code domain due to clipping becomes more and more random. Fig. 4 shows the distribution of the estimated symbols  $\hat{d}_i$  after a few radio frames compared with a corresponding normal distribution<sup>5</sup>. The received symbols are marked with different colors depending on the transmitted data.

Fig. 4 shows that, as for linear distortion, the code domain symbols can be modeled as Gaussian distributed, and again  $EVM$  and *code domain SNR* provide a direct relation to the BER or BLER.

In summary, even extreme non-linearity allows replacement of the total CDMA receiver by a simple AWGN model. Although general equations for the  $EVM$  of some arbitrary non-linearity do not exist, the  $EVM$  and *code domain SNR* can still easily be measured, and finally the relation between  $DPCH\_E_c / I_{or}$  and the BLER can be derived.



<sup>5</sup> For the sake of simplicity, the sent sequence  $d_i$  was BPSK modulated.

Fig. 4. Magnitude distribution of the symbol estimates according to (7) for a fully loaded CDMA system received by a limiter amplifier.

### 5.3. Random Distortion

Random distortion may be additive or multiplicative [3]. Based on the analysis above (see (18) and (19)), it is obvious that the code domain symbols are again normally distributed and the  $EVM_{noise}$  measurement relates the *transmit power ratio* to the BLER.

Since the RF designer needs to find specifications for the sub-components of the receive chain, it is important to understand the interaction of noise with linear distortion. For that purpose we define a measure called *noise gain* that is related to the different transfer characteristics for signal and noise

$$noise\_gain = \frac{\int_{-\infty}^{\infty} |h(t)|^2 dt}{|f_0|^2}. \quad (22)$$

Using the *noise gain*, the *EVM* caused by noise simplifies to

$$EVM_{noise}^2 = \frac{I_{oc}}{I_{or}} \cdot noise\_gain. \quad (23)$$

Note that for a distortion-free DUT (e.g.  $\tilde{f}(t) = \delta(t)$ ) the *EVM* is 0 and the *noise gain* is equal to 1 (0 dB). Practical filters (e.g. SAW filters) usually have a *noise gain* greater than 1. This is because some energy of the filter's impulse response is not concentrated in the main lobe. Thus  $|f_0|^2$  reduces relative to the noise transfer characteristic  $\int_{-\infty}^{\infty} |h(t)|^2 dt$ .

It is important to note that the *noise gain* is derived from the temporal or spectral shape of the filter transfer functions. This has nothing to do with potential noise figure increase due to non-zero insertion loss of the passive device.

## 6. Simulation Results

This section evaluates the *EVM* as a FoM for the CDMA receiver design in a system simulation environment. Fig. 1 shows the simplified block diagram of the environment. All the receiver test cases specified in [2] can be configured. The transmitter is a standard compliant realization of the W-CDMA system. The receiver is a floating point implementation of various receive algorithms.

For the analysis of *linear distortion* we used an IF SAW filter and configured the environment for test case #1 for the "Demodulation of Dedicated Channel (DCH)" in section "Demodulation in static propagation conditions" as specified in [2]. This test case corresponds to speech transmission at 12.2 kBit/s with a spreading factor 128. Again white Gaussian noise with spectral density  $I_{oc}$  is added between the transmitter and the DUT, see also Fig. 1.

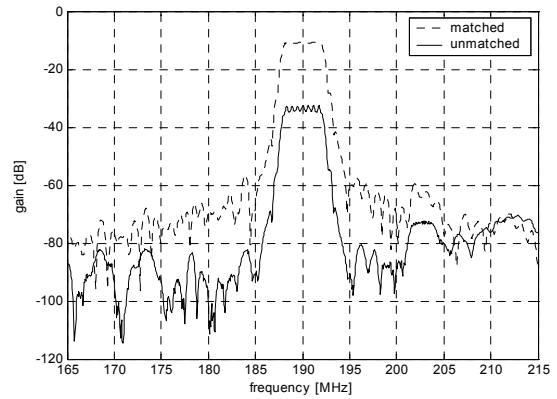


Fig. 5. Frequency domain transfer functions of matched and unmatched EPCOS SAW filter with 1.8% *EVM* (matched) and 12.3% *EVM* (unmatched).

Rather than using a direct measurement of the *EVM* at the RF-BB interface, we measured the *EVM* of the SAW filter in a separate test environment. An "EVM measurement module" performed the optimization with respect to  $\alpha$  and  $\tau$  according to (10). In order to illustrate the effect of strong distortion, we used an unmatched SAW filter that was directly measured in a 50 $\Omega$  environment without external matching elements. In this case  $EVM_{SAW}$  amounts to 12.3%. If the filter is properly matched the  $EVM_{SAW}$  can be reduced to 1.8%.

Fig. 5 depicts both frequency domain transfer functions. According to (19) and (22), additional interference proportional to  $\hat{I}_{or} \cdot EVM^2$  caused by the linear distortion and a noise gain of 0.086 dB must be taken into account. The ratio between the interference power with and without distortion characterizes the *SNR-degradation* as follows:

$$\begin{aligned} \frac{SNR_{ideal}}{SNR_{distorted}} &= \frac{I_{oc} \cdot noise\_gain + \hat{I}_{or} \cdot EVM_{SAW}^2}{I_{oc}} \\ &= noise\_gain + \frac{\hat{I}_{or}}{I_{oc}} \cdot EVM_{SAW}^2 \end{aligned} \quad (24)$$

Fig. 6 shows the respective BER and BLER curves (solid lines) in case of distortion-free DUT for different  $\hat{I}_{or}/I_{oc}$  ratios for the unmatched case. Moreover, the predicted BER and BLER curves in case of the DUT exhibiting linear distortions (dashed lines) are also shown. The latter are just transformations of the original water fall curves (solid lines) by (24). In this way the BER and BLER can be determined from tabulated ideal BER and BLER curves of respective decoders without simulations. The prediction and the actual discrete time system simulation (crosses in Fig. 6 including the SAW filter show a quite acceptable match. This verifies the concept for a (simulative) floating point baseband implementation.

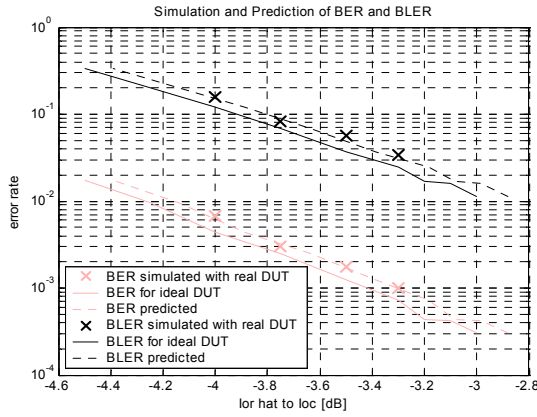


Fig. 6. Comparison of simulated and predicted BER and BLER for linear distortion with 12.3%  $EVM$ .

For an actual fixed point real time implementation we observe of course additional degradation of the measured BER or BLER. The additional degradation, however, must solely be attributed to the baseband part. Thus, the  $EVM$ -approach allows a clear separation between RF and BB impairments.

For non-linear distortions, we revisit the limiter amplifier of Sec. 5. After clipping, an  $EVM$  of about 39% can be “measured”.<sup>6</sup> For coherent BPSK – reception, (19) has to be modified to

$$SNR_{code,BPSK} = 2 \cdot \frac{DPCH\_E_c}{I_{or}} \cdot \frac{SF}{EVM^2}. \quad (25)$$

With the substitutions  $\frac{DPCH\_E_c}{I_{or}} = \frac{1}{128}$  and

$SF = 128$  the BPSK code domain SNR amounts to  $SNR_{code,BPSK} = 11.17dB$ . This matches again well to the measured SNR of about 11dB<sup>7</sup>. This value can guarantee a zero BLER after channel decoding. Thus, no BLER curves are shown here.

It is interesting to note that clipping of the whole “maximum input level” signal is surprisingly no problem for the required performance of the respective test case in [2]. A general assessment of non-linearity must of course also take into account adjacent channel and blocking conditions [2]. Note, however, that the (in-band) distortion of the respective inter-modulation is already included in the  $EVM$  measurement.

## 7. Conclusions

We showed that  $EVM$  can be used as a figure of merit for the CDMA receiver design. *Linear*, *non-linear* and *random* distortions are covered. A direct relation of the  $EVM$  to the BER and BLER by means of simple formulas and tabulated “water fall” curves, which characterize the decoder in AWGN conditions, is derived. The relations and formulas have been verified by W-CDMA system simulations for two

examples, a badly matched SAW filter and strong clipping behavior in the receiver. The test cases according to the 3GPP standard were used for the analysis.

Using the  $EVM$  as a design parameter for RF components, RF systems, and as an interface specification parameter between RF and baseband part of a CDMA receiver can greatly simplify the design and verification process, offer more degrees of freedom for the design, fasten the analysis, and avoid unnecessary over-specification.

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<sup>6</sup> This was simulated in a separate environment, where a “ $EVM$  measurement module” performs the optimization with respect to  $\alpha$  and  $\tau$  according to (10).

<sup>7</sup> The “measurement” is based on an estimation of mean and variance at the RAKE receiver output according to definition (12).