On the Impact of Receiver RX filters on EDGE EVM Measurements

R. Mahmoudi and J.L. Tauritz

University Twente, P.O. Box 217, 7500 AE Enschede, The Netherlands, reza@mahmoudi.org
University Twente, P.O. Box 217, 7500 AE Enschede, The Netherlands, J.L.Tauritz@el.utwente.nl

ABSTRACT — Contemporary GSM networks use a modulation technique known as Gaussian Minimum Shift Keying, GMSK, while EDGE utilizes Eight Phase Shift Keying, 8-PSK. 8-PSK operates in the same 200 kHz GSM bandwidth supporting three times higher data rates while permitting the coexistence of EDGE and GSM services, [1]. The QPSK mapping technique converts the binary information into phase information, representing a specific location on the Fresnel plane. Noise and the nonlinear behavior of semiconductor devices can alter the location of recovered data on the Fresnel plane. To examine this dependency, one can employ the Error Vector Magnitude concept.

The European Telecommunications Standards Institute, ETSI, introduces five "Inter Symbol Interference, ISI," symbols to the processed signal, [2]. These complicate the hardware implementation as well as the interpretation of the recovered data when achieved by different methods.

This article considers the influence of noise and AM-AM and AM-PM distortion on EVM measurement by comparing three receiver RX-filter architectures, based on the zero forcing filter principal, a zero forcing filter combined with matched filtering principal and ETSI measurement protocol.

I. INTRODUCTION

In contrast to IS-95 CDMA, the EDGE GSM protocol does not employ spreading codes. The underlying idea behind EDGE is straightforward; improve the data rate capability while coexisting with contemporary GSM networks. This can be realized by employing eight phase shift keying, 8-PSK, in the same 200KHz bandwidth as GSM. Three times higher data rates for EDGE GSM are the result.

With analog as well as digital modulation, the characteristics of the sinusoidal carrier waveform are varied. Amplitude, phase, frequency or a combination of these may be used to provide the modulation. In the case of digital modulation schemes, the information is in digital format, a binary word composed of n bits. There are $M = 2^n$ states. M-PSK is a digital modulation scheme where the information to be transmitted is conveyed in the phase of the carrier. For 8-PSK, $M = 8 = 2^3$.

This technique can be presented vectorially using the Fresnel plane, better known as a constellation diagram. The 8-PSK signals' trajectories are linear transitions between from constellation point to the next. The processed trajectory path of the modulated data contains the origin of the Fresnel plane (see [1] and [5]).

This unfortunately enlarges the associated crest factor's span, intensifying the impact on system performance of the nonlinear behavior of the

semiconductor devices used. To confine the span of the power spectrum in the signal's envelope, it would be desirable to exclude the center of the Fresnel plane from the trajectory path.

Excluding the center of the constellation diagram can be achieved by continuously rotating the 8-PSK constellation planes by offsets of $3/8$ and swapping from one plane to the other for each successive symbol period. This condition can be concisely formulated as:

$$r[k] = \exp(2 \cdot \text{random}(0 \dots 7) + k \cdot \frac{3}{8}) \quad (1)$$

Where k symbolizes the consecutive symbol time.

II. GAUSSIAN-LIKE EDGE FILTER

Due to the random character of the sequenced phase, the generated pulse stream is stochastic with an infinite frequency spectrum. In order to bound the spectrum of the generated pulse sequences, a low pass filter is required. The European Telecommunications Standards Institute, ETSI [2], has proposed a Gaussian like filter for EDGE with the following impulse response:

$$\begin{aligned} Q(t) &= \frac{1}{\sqrt{2}} \exp\left(-\frac{x^2}{2}\right) dx = \frac{1}{2} \text{erfc} \frac{t}{\sqrt{2}} \\ (t) &= \frac{1}{2} Q \left(0.3 \frac{t - \frac{5T}{2}}{T\sqrt{\ln 2}} - Q \left(0.3 \frac{t - \frac{3T}{2}}{T\sqrt{\ln 2}} \right) \right. \\ &\quad \sin \frac{t}{2} (x) dx \quad \text{"for } 0 < t < 4T \\ &\quad 0 \\ (T) &= \sin \frac{t}{2} (x) dx \quad \text{"for } 4T < t < 8T \\ &\quad 0 \\ &\quad 0 \quad \text{Elsewhere} \\ h_{rx}(t) &= \sum_{i=0}^3 (t + iT) \quad \text{"for } 0 < t < 5T \\ &\quad 0 \quad \text{Elsewhere} \end{aligned} \quad (2)$$

The output of this filter is dependent on five consecutive symbols. This dependency, which is usually referred to as Inter Symbol Interference (ISI), complicates the recovery of the original data stream [3]. Employment of the filter leads to linear monotonic trajectories in the Fresnel plane as described in Eq. 3.

$$S[m] = \sum_k [k] h_{rx}[m - k] \quad (3)$$

III. EVM CALCULATION TECHNIQUES

Employing the QPSK mapping technique converts the binary information into phase information, associated with a specific location on the Fresnel plane. Noise and the nonlinear behavior of semiconductor devices can alter the location of recovered data on the Fresnel plane. The latter can lead to processing a different bit stream than the original one, leading to BER degradation. To examine this dependency, one can employ the Error Vector Magnitude concept. EVM simply monitors the deviation of the recovered data with respect to the reference signal.

Two major issues in modern digital mobile communication architecture are maximization of the signal-noise ratio at the detection plane and memory impact of the transmitter low-pass RX-filter.

1. Maximization of the signal-noise ratio can be achieved by carefully applying matched filter theory [4], which implies that:

the receiver and transmitter impulse response filters, $h_{rx}(t)$ and $h_{tx}(t)$, must satisfy the following criterion:

$$h_{rx}(t) = h_{tx}^*(-t) \quad (4)$$

An application of the above condition to the real and symmetrical properties of the ETSI transmitter filter (TX) reduces to the use of identical filters in the receiver (RX) and the transmitter path (TX).

2. Inter symbol interference due to memory effects in the transmitter filter (TX) can be avoided or reduced using the Nyquist filter principal which yields:

The overall filter is said to be a Nyquist filter if the effective transfer function of a pulse system is:

$$H(f) = \begin{cases} \frac{1}{2f_0} + Y(f) & |f| < 2f_0 \\ 0 & \text{elsewhere} \end{cases} \quad (5)$$

where $Y(f)$ is real function that is even symmetrical about $f = 0$, and odd symmetrical about $f = f_0$. There will be no ISI at the system output (TX-Filter) if the symbol rate is $f_s = 2f_0$.

The ETSI protocol, however, suggests a transmitter filter (TX) that doesn't meet the Nyquist criterion (see Fig 1).

Fig. 1: Impulse response of transmitter TX-filter, when it is twice oversampled.

Analog data passing through the ETSI transmitter filter (TX) will be interfered with by five preceding symbols. Based on the above, EVM can be predicted with respect to the original data or a reference signal.

IV. ZERO FORCING FILTER (RX)

Carrying out EVM calculations based on the original data stream requires appropriate data conversion from the analog domain to the Fresnel diagram, eliminating the inter-symbol-interference. In order to remove the added ISI, one can employ the zero-forcing algorithm yielding:

$$\begin{aligned} [m] &= h_{tx}[i]h_{rx}[m-i] && \text{Time Domain Representation} \\ 1 &= H_{tx}(f)H_{rx}(f) && \text{Frequency Domain Representation} \end{aligned} \quad (6)$$

where h_{tx} and h_{rx} are, respectively, the impulse response of the transmitter and receiver filters which can be realized using the linear transversal filter technique with a zero-forcing filter (see Fig 2).

Fig. 2: Impulse response of the zero-forcing RX-filter (left), Convolution of the transmitter TX-filter and zero-forcing RX-filter, satisfying the zero-forcing algorithm Eq. 6, (right).

V. COMBINED MATCHED AND ZERO FORCING FILTER

Although employing the zero-forcing algorithm removes the added ISI, it doesn't maximize the signal-noise ratio at the detection plane. To maximize the signal-noise ratio and eliminate ISI, a combined matched and zero-forcing RX-filter must be used. The time domain impulse response of the combined matched and zero-forcing RX-filter (see Fig 3) can be found in three steps:

- The impulse response of the transceiver and matched filter is obtained using the convolution theorem.
- Apply the zero-forcing algorithm to calculate the characteristic of the ISI removal filter for the combined transceiver and matched filter of the previous step.
- Employ the convolution theorem to obtain the impulse response of the ISI removal filter (step 2) and the matched filter.

$$[m] = \sum_i h_{ix}[i-m] (h_{Matched}[n] h_{ZeroForcing}[n-m])$$

$$1 = H_{ix}(f) H_{Matched}(f) H_{ZeroForcing}(f) \quad (7)$$

Simulation results in Fig. 7 emphasize the superiority of the combined matched and zero forcing RX-filter, influenced only by noise. However, simulation results also indicate that a combined matched and zero-forcing filter is more sensitive to the nonlinear behavior of semiconductor devices. To study the reason for this discrepancy, one can make use of the following expression:

$$[k] [k] h_{matched-zeroforcing}[m-k]$$

$$S[m] [k] \text{ Constant value}$$

$$[m] S[m] [k] \text{ Constant value} \quad (8)$$

where, $h_{matched-zeroforcing}[k]$ represents the time domain impulse response of the matched and zero-forcing RX-filter,

- $[k]$ is the time domain sample time of the analog signal,
- $[k]$ is the time random number representing the non-linearity added to the analog signal,
- $[k]$ is the deviation caused by the nonlinear behavior of the RF circuit blocks, symbolized by $[k]$.

Coefficients $[k]$ and the span of the impulse response filter determine the amount of this deviation. The sensitivity of the combined matched and zero-forcing RX-filter is apparently a result of its longer span-length.

From the above results one can conclude that the combined matched and zero-forcing RX-filter has a somewhat contradictory performance. This paradoxical behavior is further depicted in Fig. 7 and Fig. 8. The zero-forcing RX-filter performs better at high signal-noise ratios, while the combined matched and zero-forcing RX-filter has better performance for poor signal-noise ratios.

Fig. 3: Impulse response of matched and zero-forcing RX-filter (left), convolution of the transmitter TX-filter and combined matched and zero-forcing RX-filter, satisfying the zero-forcing algorithm, Eq. 8, (right).

VI. ETSI MEASUREMENT RX-FILTER

In order to calculate EVM, the ETSI protocol [2] suggests using a "raised-cosine filter with roll-off 0,25 and a single side-band 6 dB bandwidth of 90 kHz.". The measurement filter is windowed by multiplying its impulse response by the raised cosine window given by:

$$w(t) = \begin{cases} 1 & |t| < 1.5T \\ 0.5 + \cos\left(\frac{|t| - 1.5T}{2.25T}\right) & 1.5T \leq |t| \leq 3.75T \\ 0 & |t| > 3.75T \end{cases} \quad (9)$$

where T is the symbol interval. This filter processes a reference analog signal (see Fig. 4) containing the memory impact of the transceiver and ETSI measurement filter (ISI). The combination of the transmitter TX-filter proposed by the ETSI protocol and the receiver RX-filter fails to maximize the signal-noise ratio at the detection plane. Since the processed reference signal using this method doesn't represent the original signal, association of the calculated EVM of this reference signal to the BER measurement is contestable.

It is obvious that applying a different reference signal will lead to differences in the EVM results obtained.

Fig. 4: The EVM calculation technique based on the ETSI protocol with respect to a processed reference signal which differs from the original signal.

Fig. 5: Impulse response of ETSI measurement RX-filter (left), convolution of the transmitter TX-filter and the ETSI measurement RX-filter, containing ISI, hindering maximization of the signal-noise ratio, (right).

VII. SIMULATION RESULTS

To study EDGE signal sensitivity to AM-AM and AM-PM conversation, we have made use of a special AM2AM & AM2PM functional block, where the transducer's gain and phase can be specified as a linear and non linear function of the incident input power as depicted in Fig 6.

Simulation results indicate that a combined matched and zero-forcing RX-filter when compared with a stand alone zero-forcing RX-filter is less responsive to noise, due to maximization of the signal to noise ratio (Fig 7), and is more sensitive to component non-linearity due to its larger span, Eq. 8.

Fig . 6: The transfer function's gain (left) and phase (right) of the specially implemented AM-AM and AM-PM functional block.

Fig . 7: A zero-forcing RX-filter compared with a combined matched and zero-forcing RX-filter when influenced by noise.

Fig . 8: A zero-forcing RX-filter compared with a combined matched and zero-forcing RX-filter when influenced by noise and AM-AM and AM-PM distortion (AM-PM offset is equal to zero).

Fig 9 illustrates a significant difference between EVM measurement based on an ETSI RX-filter and a stand alone zero-forcing RX-filter.

Fig . 9: A zero-forcing RX-filter compared with an ETSI RX-filter when influenced by noise (left), AM-AM and AM-PM distortion (right).

V. CONCLUSIONS

The output of an ETSI transmitter filter is dependent on five consecutive symbols. This dependency complicates the recovery of the original data stream. The choice of the architecture of the receiver filter plays a significant role in the EVM measurement.

The ETSI RX-filter is neither a matched nor an ISI removal filter so that the EVM when calculated in accordance with the ETSI protocol, does not represent the BER behavior of the system under consideration.

ACKNOWLEDGMENT

The authors wish to acknowledge the assistance and support of Andre van Bezooijen, Philips Semiconductors for stimulating discussions and support.

REFERENCES

- [1] A. Mashhour, "Understanding Offset 8-PSK Modulation for GSM EDGE," Microwave Journal, vol. 42, pp. 78-92, Oct. 1999.
- [2] ETSI. "Mobile Station-Base Station Compatibility Standard for Dual-Mode Wideband Spread Spectrum Cellular Systems", 1995.
- [3] L.W. Couch, Digital and Analog Communications Systems, 4th ed. Upper Saddle River, NJ: Prentice Hall, 1993.
- [4] J.G. Gibson, The Mobile Communications Handbook. New York: IEEE Press, 1996.
- [5] R. Mahmoudi, A Multidisciplinary Design method For Second and Third Generation Mobile Communication System Component, Thesis, UTwente, November 2001, Enschede, The Netherlands