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A Robust GSM/EDGE Transmitter Using Polar Modulation Techniques

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***Abstract* — A key property of polar modulator systems is the matching between amplitude and phase path. This paper shows how mismatches between these paths degrade performance and introduce difficulties in designing GSM EDGE systems. A transmitter overcoming these problems is presented. The chip has been implemented in a 0.13µm CMOS technology, with a nominal margin of 7 dB to the critical 400 kHz specification and a transmit EVM of 2.5%.**

I. Introduction

Increasing demands for high data rates in mobile communication resulted in specification of system standards superior to the 2nd generation ones (e.g. GSM, Global System for Mobile communication). 3rd generation standards like UMTS (Universal Mobile Telecommunications System) fulfill the requirements for high data rate, but need a huge amount of extra infrastructure and run in parallel to common GSM networks. EDGE (Enhanced Data rates for GSM Evolution) was designed for maximum compatibility to normal GSM networks, which enables the possibility of upgrading hard- and software with minimal costs.

The advantage of EDGE is its highly efficient 8PSK modulation scheme, which triples the transferable data rate just by modulation. The power spectrum is though very similar to the GMSK spectrum, allowing a high reuse of components – or at least very low effort for upgrade. The prize for this benefit is the non-constant envelope, which is one of the major challenges for transmitting EDGE signals, because this requires a linear power amplifier with reduced efficiency or at least intricate methods for linear transmission.

EDGE modulation can be done with conventional direct modulation architectures, which separate the complex base band signal into its real and imaginary part, or by polar modulators, which do the separation in amplitude and phase. The latter have the advantage of making the system invulnerable (because of the already modulated VCO) against the harmonics, which are folded back to in-band at the PA, especially those of the 3rd order (HD3). This eliminates the need of any isolator or extra filter components, which helps to keep the component count low. But, when using polar modulation, other challenges arise like mismatches in amplitude and phase path or increased bandwidth requirements.

This paper focuses on these mismatches and presents a transceiver overcoming those problems. Section II describes the polar modulation architecture and typical key requirements for polar modulator systems, section III shows the mismatch problems in detail and section IV shows a single chip CMOS EDGE transceiver coping with these critical requirements.

II. Polar Modulation Architecture

Conventional quadrature modulators using a direct conversion architecture, like in [1], separate the complex signal $s(t)$ into its real and imaginary part $I(t)$ and $Q(t)$, respectively:

$$s(t) = I(t)\cos(\omega_c t) + jQ(t)\sin(\omega_c t) \qquad (1)$$

The bandwidths of the separated paths are equal to the complex signal and both paths can be built up with the same signal processing blocks. Also, the combination of real and imaginary part is a linear transformation, which makes it a lot easier to predict the final output when knowing the properties of each path.

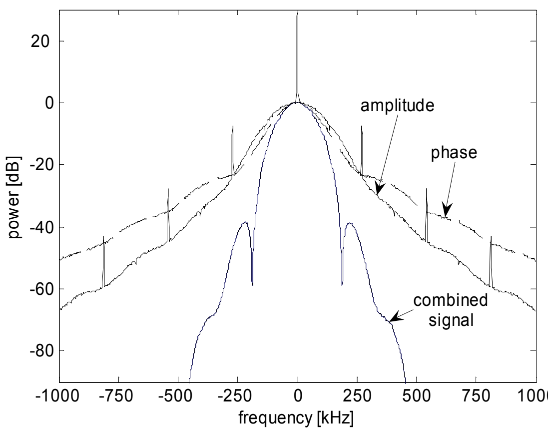

Fig. 1: Bandwidths of amplitude and phase part of an EDGE 8PSK signal

The principle of polar modulation is to separate a complex signal into its amplitude $a(t)$ and its phase part $\varphi(t)$:

$$s(t) = a(t)\cos(\omega_c t + \varphi(t)) \qquad (2)$$

Unlike GMSK signals, 8PSK signals do not have a constant amplitude part $a(t)$ for all t, introducing the linearity problems already mentioned in chapter I. Additionally, bandwidths of the amplitude and phase parts are much higher than that of the combined complex signal, as shown in Fig. 1. This requires e.g. a higher bandwidth of the PLL, which is likely to be used in the phase path, necessitating additional methods for bandwidth enlargement of conventional PLLs [2].

A block diagram of a polar modulator transmitter is depicted in Fig. 2.

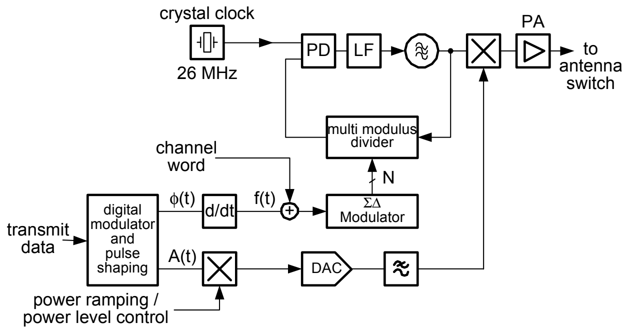

Fig. 2: Block diagram of a polar modulator transmitter

When looking at the amplitude and phase (or, for an even better view, the frequency) signals in the time domain, as plotted in Fig. 3, one can see, that amplitude and frequency signal both show large spikes. This reveals the source for the high bandwidths of these signals. When comparing the spikes in amplitude and frequency signal, it is obvious, that each peak in the amplitude signal has a corresponding one in the frequency signal. The combined signal does not have that high bandwidth, which means, that the mentioned spikes have to be compensated by each other. Clearly this must mean, if this compensation is worsened by e.g. mismatches in amplitude and phase path, a bandwidth increase of the restored signal is unavoidable.

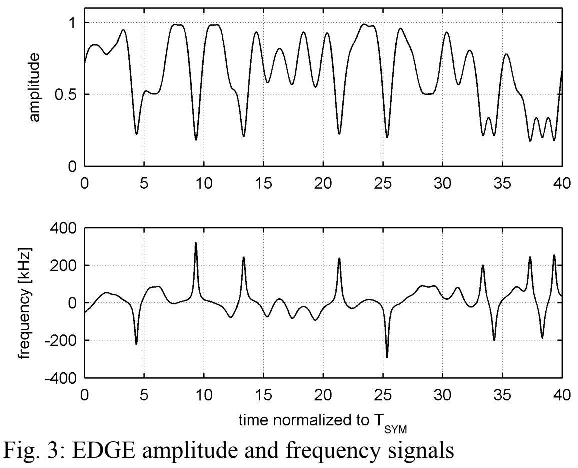

Fig. 3: EDGE amplitude and frequency signals

III. Degradation Effects

A. Delay between amplitude and phase path

The different signal processing stages in the separated amplitude and phase paths introduce different delays for these signal parts. Some of them, especially that from the digital domain, are static delays, which are constant for all frequencies. Others, e.g. filters, have a varying group delay with respect to frequency. In this subsection, static delays are investigated. If the delay difference is not corrected before combining the separated signals again, the unwanted spectral emissions and EVM increase dramatically.

The complex signal comprising static amplitude-phase delay with delay value τ is defined as:

$$s(t) = a(t-\tau)\mathrm{e}^{\mathrm{j}\varphi(t)} \tag{3}$$

Fig. 4 shows the degradation of spectral emissions and modulation accuracy as a function of delay mismatch. Although modulation accuracy (defined by EVM) is also affected, the spectrum mask is the more stringent specification for this type of mismatch. A value below approximately 3/96 symbol times (~0.12 µs) is required to keep spectral emissions below -54 dBc at the 400 kHz corner, as required by the GSM/EDGE specification. At this mismatch, RMS EVM is still well below the specification limit of 9% [3].

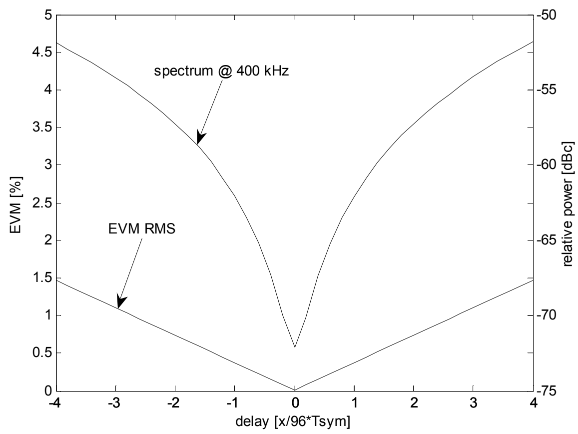

Fig. 4: effect of amplitude-phase delay on EVM and spectrum

B. Offset in the amplitude path

Analog components in the amplitude signal processing chain are often susceptible for offsets. Such offsets also result in additional spectral emissions, and detoriation of EVM. Amplitude offset is defined as a_{offs} in

$$a(t) = |s(t)| + a_{offs}\,. \tag{4}$$

Fig. 5 shows the influence of amplitude offset on spectrum and EVM. Again, the spectrum is influenced more than the modulation accuracy. A value of around 5% of the maximum signal amplitude would cause the spectrum to violate the specification limit of -54 dBc.

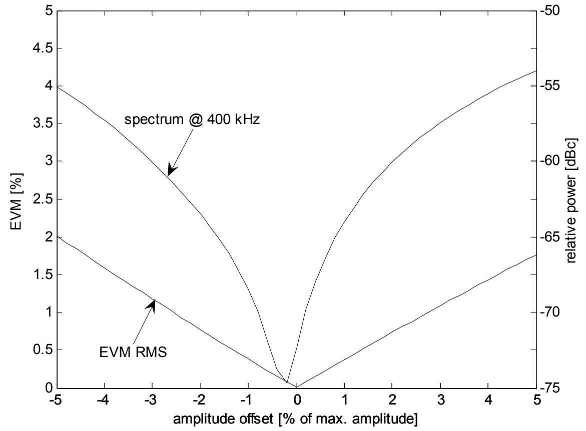

Fig. 5: Influence of amplitude offset on EVM and spectrum

C. Low pass filter in the amplitude path

To eliminate spectral images of the sampled signal after the DAC in the amplitude path an anti-imaging (or reconstruction) filter is required. This low-pass filter introduces two main issues: magnitude compression for higher frequencies and a non-ideal group delay response. For comparison, three filter types were investigated, each with varying cutoff frequency and order. The Butterworth filter is designed for a maximally flat magnitude response in the pass-band, but has some variation in group delay around the cutoff frequency. The Bessel filter has the flattest group delay response, but a very weak cutoff, which gives some attenuation in the pass-band. Finally, the Chebyshev filter has the worst group delay response, and a pass-band ripple in the magnitude response. The advantage of this filter type is its superior attenuation in the stop-band [4].

Results are shown in Fig. 6. The filters were placed in the amplitude path as digital filters at a sampling rate of 26 MHz. To avoid additional mismatch because of the introduced group delay, it was compensated statically, i.e. the group delay value of the filter for low frequencies was applied as static delay in the phase path. The Chebyshev filter was designed with 0.1 dB pass-band ripple.

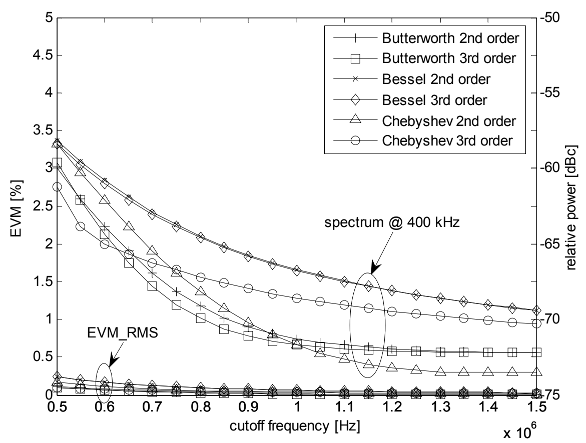

Fig. 6: influence of amplitude low pass filtering on EVM and spectrum

A first point which can be seen in this figure, is, that amplitude filtering does not affect EVM very much, but has much more influence on the modulation spectrum. Even the worst filter does not exceed 0.5% EVM RMS. Regarding filter order, only the Chebyshev filter shows spectral degradation when using the higher order. Generally, the comparison would lead to the conclusion, that magnitude response is more important than group delay response, because the Bessel filter, which is designed for best group delay response, has a worse spectral performance than the Butterworth or even Chebyshev filters.

D. Nonlinearities in the amplitude path

Analog components like DAC, filters, or amplifiers have nonlinear transfer characteristics, which affect the quality of the transferred signal. When placing such nonlinearities in the amplitude path, EVM and spectral performance are degraded, as shown in Fig. 7. A 3rd order polynomial model, described in equation (5), was used for this simulation. These models have been confirmed to be good models for real nonlinear building blocks.

$$y = \alpha_0 + \alpha_1 x + \alpha_2 x^2 + \alpha_3 x^3 \qquad (5)$$

While the 2nd order harmonics (HD2) can be kept within well defined limits by careful design, the 3rd order harmonics (HD3) can not be overcome so easily. Hence, this is the main critical linearity parameter for the polar modulator system design. For an explanation of the definition of HD2 and HD3, see Fig. 8.

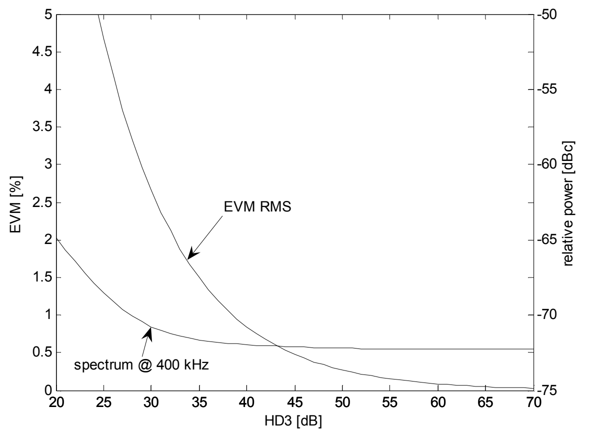

Fig. 7: Influence of 3rd order harmonics in the amplitude path on EVM and spectrum. HD3 is the distance of the third order harmonic to its fundamental.

As it can be seen in Fig. 7, the critical 400 kHz specification is not the key parameter for limiting the 3rd order harmonics. Here, EVM is much more critical – a value of 3% is reached already with a harmonic distortion of 30 dB, while the spectrum is nearly ideal at this point.

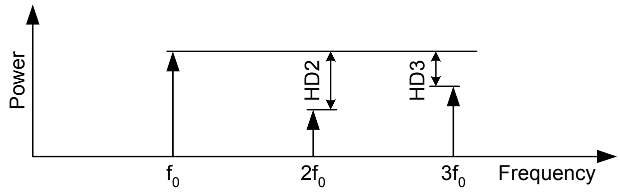

Fig. 8: Explanation of harmonic distortion terms HD2 and HD3

IV. Realization Of A Robust EDGE Transmitter

The EDGE transmitter was implemented in a complete transceiver IC in a 0.13 µm CMOS technology and operates at 2.8 V and 1.5 V power supply. In addition to the realized concept of a polar modulator for EDGE, the transmitter includes integrated PGA (Programmable Gain Amplifier) functionality. The transceiver is completely controllable through a common 3-wire bus interface and has a standard I/Q interface to the base-band processor. The chip is housed in a 40-pin green VQFN package (5.5 mm x 6.5 mm). A block diagram of the transceiver IC is depicted in Fig. 9.

The amplitude path is set up with mainly digital components. Some of the impairments described and investigated in the previous section, like amplitude-phase-delay or amplitude offset are measured on-chip and compensated digitally by special calibration algorithms via the amplitude path.

The phase processing path is realized as a ΣΔ-modulation loop as described in [5], using the technique of pre-emphasis to increase the available bandwidth. I.e., a pre-emphasis filter is implemented in the digital domain to compensate the low pass behavior of the PLL within the Nyquist frequency range.

The variation in PLL open loop gain is compensated by a digital adjustment algorithm. After the open loop gain is measured at VCO frequency, this algorithm controls the current of the charge pump.

To compensate for variations in the loop filter components, the characteristics are measured and subsequently corrected by another digital adjustment algorithm. The increased bandwidth of 8PSK phase signals with respect to GMSK signals mandates a high accuracy of the integrated PLL calibration loops.

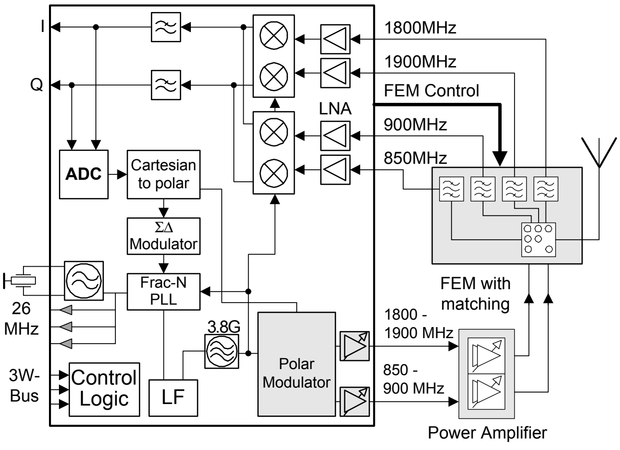

Fig. 9: Block diagram of an EDGE transceiver realized with a polar modulator transmitter

The high digital content of the presented transmitter reduces the amount of possible overall mismatch and simplifies the prediction of mentioned impairments, compared to mostly analog implementations, like presented in [6].

A measured output spectrum of the transceiver, compared to the EDGE mask is depicted in Fig. 10.

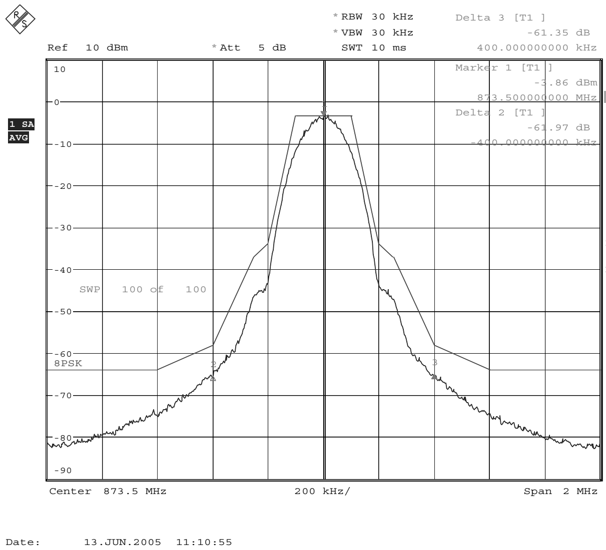

Fig. 10: Measured EDGE output spectrum

V. Conclusion

Degradation effects in polar modulator transmitters were investigated in this paper. It was shown, that transmission of EDGE signals with polar modulators requires narrow control of impairments in both amplitude and phase part, like delay mismatch, amplitude offset and filtering or nonlinearity. Good knowledge of AM and PM parameters and additional calibration loops are necessary for compensating environmental and process variations to receive a robust EDGE transmitter.

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