Filter Considerations in Polar Transmitters for Multi-mode Wireless Applications

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Abstract

Polar based transmitter architectures are attractive for future multi-mode wireless communication systems. This paper presents some considerations on the filters needed in polar transmitters to meet the spectrum mask requirements. A 1-bit $\Sigma\Delta$ modulator in the envelope path can push the quantization noise outside the channel bandwidth, but will unfiltered cause problems for fulfilling the spectrum mask. This noise can be removed by a low-pass filter in the envelope modulator. The paper shows that a 3^{rd} order Butterworth filter with a bandwidth three times the signal bandwidth can meet the system requirements. The delay caused by the filter must be compensated in the phase path to correctly align the phase and envelope in time. This compensation can be integrated into the base-band DSP block.

1. Introduction

Modern wireless communication systems use multi-level modulation schemes to achieve high spectral efficiency — examples are 8-PSK used in EDGE, and Hybrid-QPSK used in UMTS. These varying envelope modulations require a close to linear transmitter. Unfortunately, linear power amplification generally implies a poor power efficiency which obviously is a particularly bad situation for mobile terminals.

Some architectures rely on non-linear power amplifiers (PAs) combined with linearization techniques to achieve both high power efficiency and high linearity — one is Envelope Elimination and Restoration (EER) [2], and another is LInear amplification with Nonlinear Components (LINC) [5]. In principle, these transmitters can use switch-mode PAs, such as Class-D, -E, -F but each of them has some problems in practice. For the LINC architecture, the processing of two phase components expands the bandwidth, and the power combining of the two PA outputs suffers from low efficiency at low output power levels. Furthermore, the two power branches must be very well matched. The EER architecture suffers from the limited envelope bandwidth and the differential delay between the envelope and phase paths [4]. The transformation from I/Q to A/P (amplitude and phase) form expands the envelope bandwidth as seen in Figure 1. Figure 1 compares the relative power spectral density (PSD) versus the frequency of EDGE I/Q and A/P signals — here the relative PSD means PSD relative to the PSD value at 0 Hz. For wideband applications, the bandwidth may be a critical issue in the system design. The main source of the differential delay is the low-pass filter in the class-S envelope modulator.

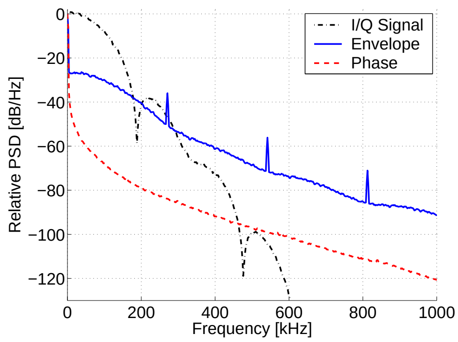

Figure 1. PSD of I/Q and A/P Signals.

The concept of EER is used in the polar transmitter shown in Figure 2. In order to minimize the degradation of the transmitter spectral performance due to the differential delay, the initial idea from Wang [7] is to $\Sigma\Delta$ modulate the envelope signal to control the switch-mode (class-S) power supply (SMPS) of the switch-mode PA. The $\Sigma\Delta$ modulator pushes the quantization noise to higher frequencies where it is removed by an RF bandpass filter after the PA. So, the main delay source, the low-pass filter, can be omitted. However, it is impractical for this RF filter to have a channel selection character due to size and loss. Its pass band must be the entire transmit band. The $\Sigma\Delta$ modulator should shape the quantization noise outside this band which is normally several tens of MHz.

On the other hand, in order to switch the SMPS the $\Sigma\Delta$ modulator must be 1-bit. This limits the modulator to 2^{nd} order for stability reasons. Then the 2^{nd} order, 1-bit $\Sigma\Delta$ modulator must have a very high over-sampling ratio[1] to shape the quantization noise to several tens of MHz away from the carrier frequency. This high sampling ratio results in a high switching frequency, such as several GHz. At this very high switching frequency the SMPS will have a low power efficiency and poor spectral performance.

2. Transmitter Architecture

In [1], a polar transmitter for multi-mode operation was introduced — see Figure 2. The envelope signal is $\Sigma\Delta$ modulated to switch the SMPS of the PA stage. The phase signal is up-converted to the RF band by a two-point-modulation PLL [3]. Finally, the envelope and phase information is recombined at the class-E PA stage [6]. Adaptive baseband digital pre-distortion is involved in this architecture to improve the system linearity and efficiency. It may also relax the requirement of the up-converter and power blocks.

Because it is difficult to design a channel-select post-PA band-pass filter to remove the quantization noise, a practical solution is to keep the low-pass filter in the class-S envelope modulator to reduce the quantization noise. The delay caused by this filter is compensated in the baseband DSP block to align the envelope and phase signals to meet the stringent spectral requirement. Furthermore, the class-S modulator and the PA stage will degrade the out-of-band noise performance. So, the output of this low-pass filter must meet the spectral specifications and have some margin.

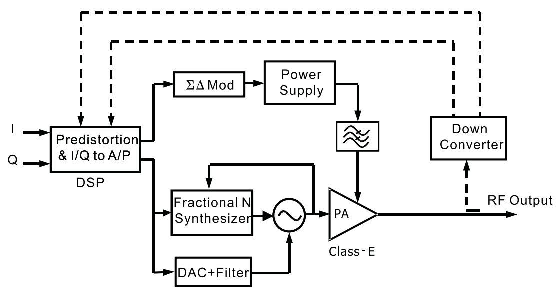

Figure 2. Polar transmitter architecture [1].

To design the low-pass filter in the envelope modulator, the noise shape of the $\Sigma\Delta$ modulation must be considered. Figure 3 depicts the PSD of the EDGE envelope signal before and after $\Sigma\Delta$ modulation (2^{nd} order, 1-bit, 256 times over-sampled). The quantization noise beyond the inflexion should be removed by filtering.

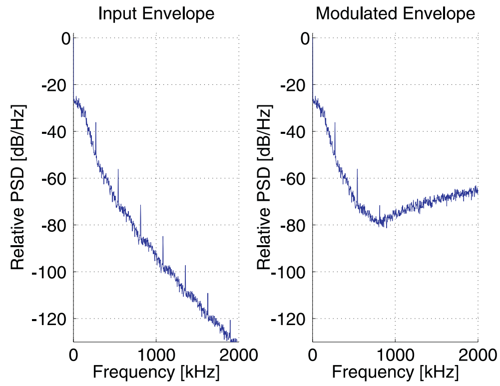

Figure 3. EDGE envelope PSD before and after the $\Sigma\Delta$ modulator.

When considering group delay, a FIR filter is a better choice than an IIR filter because the group delay of FIR filters is constant while it is frequency-dependent for IIRs. Constant group delay can easily be compensated by DSP delay components. But, the drawback of FIRs is that the transition from pass-band to stop-band is slower than IIRs, meaning that a higher order FIR is needed to remove the noise. If the frequency-dependent group delay can be compensated somehow, an IIR filter is a better choice.

In MATLAB, the function *iirgrpdelay* can be used to compensate for this frequency-dependent group delay. *iirgrpdelay* acts as an all-pass filter, only to compensate for the frequency-dependence of the group delay.

3. Simulations

This paper is mainly focused on the envelope modulation and filtering issues which are considered in the following. The phase path (up-converter) and PA (including the power supply) are ideally described as:

- **Up-converter;** The input phase of the PA is expressed as:

$$\phi_{PA,in}(t) = 2\pi f_0 t + \phi_{in}(t) \tag{1}$$

where f_0 is the carrier frequency, and $\phi_{in}(t)$ is the input phase.

- **Class-S Modulator;** The Class S power supply modulator is ideally described as:

$$V_{DD}(t) = A \cdot A_{SD}(t) \tag{2}$$

[1]Defined as sample rate divided by symbol rate.

where $V_{DD}(t)$ is the PA supply voltage, A is a constant to generate a proper switch-mode supply voltage, and $A_{SD}(t)$ is the output of the $\Sigma\Delta$ modulator in the envelope path.

- **PA;** A Class E PA is ideally described as:

$$y(t) = V_{DD}(t) \cdot \sin\left[\phi_{PA,in}(t)\right] \tag{3}$$

where $y(t)$ is the PA output waveform and $\phi_{PA,in}(t)$ is the phase information of the PA input.

3.1. Polar Imperfections

Delay Mismatch: This is simulated by delaying the envelope signal for a number of samples when having a high over-sampling factor to ensure a fine time resolution. Figures 4 and 5 show the simulation results for the spectral performance. The delay mismatch between the envelope and phase signals has a negative impact on both EVM and spectral performance. The simulations show that the impact on spectral performance is more serious than on EVM. The maximum delays are about 70 ns for EDGE (according to the spectrum mask) and 16 ns for UMTS (according to the ACLR [2] specification, 33 dB for the adjacent channel, and 43 dB for first alternate channel), and the RMS EVMs are less than 2% for EDGE and 7% for UMTS — lower than the specification requirements of 9% and 15%, respectively.

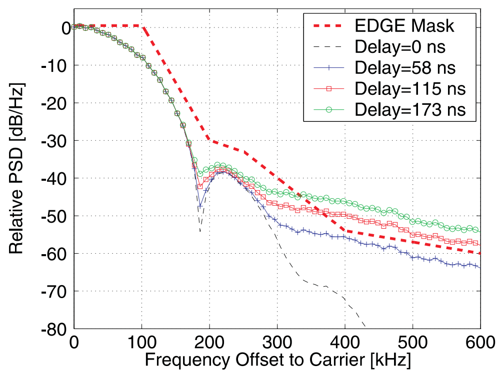

Figure 4. EDGE spectral performance due to delay difference.

Bandwidth limitation: This limitation is realized by low-pass filtering the envelope and phase signals (by the same filter). This limitation degrades the spectral performance of the polar transmitter further. Simulations show that when the envelope bandwidth is less than three times the signal bandwidth, the signal does not meet the spectrum

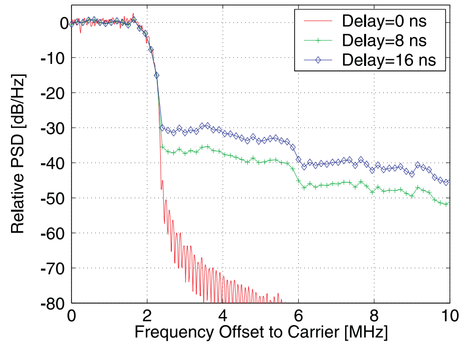

Figure 5. UMTS spectral performance due to delay difference.

mask. Figure 6 shows the simulation result when the envelope bandwidth is $3\,BW$ where BW is the signal bandwidth.

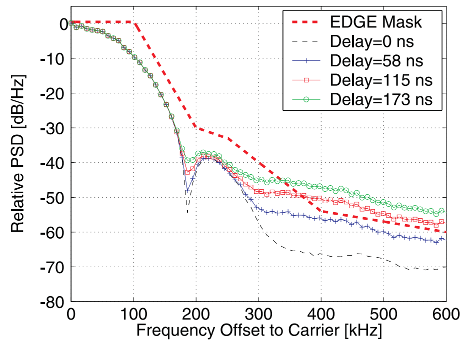

Figure 6. EDGE spectral performance due to delay difference and $3\,BW$ envelope bandwidth.

3.2. Proposed Architecture

In this part, the envelope modulation is included in the polar simulation. The $\Sigma\Delta$ modulator is 2^{nd} order, 1-bit.

The $\Sigma\Delta$ modulated envelope signal is filtered by a 3^{rd} order Butterworth low-pass filter to give the PA power supply signal. This filter causes a frequency-dependent group delay. The phase signal must be compensated to correctly align in time with the envelope signal. Two schemes are simulated: (i) The group delay is uniformed by an extra all-pass filter to align the phase signal, and (ii) to align the phase signal according to the group delay at 0 Hz. These two schemes have similar performance. It means that the all-pass filter for the group-delay unification is not necessary. In the following simulations, the second scheme is used to align the envelope and phase signals.

[2] Adjacent Channel Leakage power Ratio

Another important issue is the choice of over-sampling (OS) ratios of the signal generator and the $\Sigma\Delta$ modulator. For the signal generator, $OS = 8$ is sufficient to produce a baseband signal spectrum suitable for this polar transmitter. But the noise shaping character of the $\Sigma\Delta$ modulator with this over-sampling ratio can NOT meet the spectral requirements. Here, these two ratios are set to same value. Of course, the ratio of signal generator can be set to a lower value, then interpolate (up-sample) the data sequence before the $\Sigma\Delta$ modulator. In order to meet these requirements, the over-sampling ratio of the 2^{nd} order $\Sigma\Delta$ modulator should be 256 for EDGE and 64 for UMTS.

Figure 7 compares the PSD of the de-modulated EDGE signals with and without envelope and phase signal alignment. Here, the bandwidth of the post-PA bandpass filter is set to 20MHz and the de-modulator is ideal. It shows that the delay difference between the envelope and phase paths has a clear impact to the spectral performance.

For UMTS signals, Figure 8 compares the output PSD with and without A/P delay alignment. Here, the low-pass filter is 5^{th} order Butterworth and the bandwidth is twice the signal bandwidth, the baseband signal is 64 times over-sampled, and the bandwidth of the post-PA bandpass filter is 60 MHz. Figure 8 shows that the ACLR of the final RF signal is improved from 25 dB to 41 dB (35 dB to 43 dB for the first alternate channel ACLR) by using delay compensation. Furthermore, the EVM performance also is improved by the alignment from 13% to 3%.

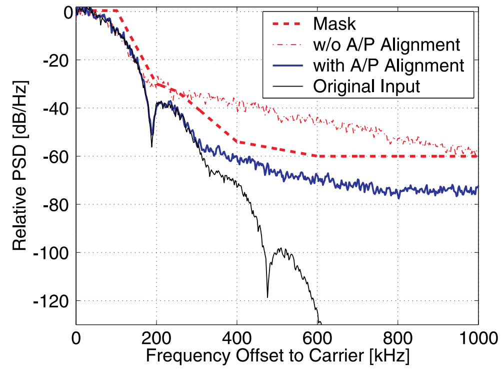

Figure 7. PSD of de-modulated EDGE signals.

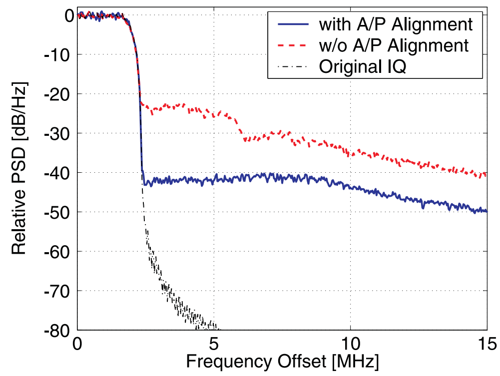

Figure 8. PSD of de-modulated UMTS signals.

4. Conclusion

In this paper, the filtering issue of the polar transmitter for multi-mode wireless applications is discussed based on the EDGE and UMTS systems. The envelope signal is $\Sigma\Delta$ modulated to control the switch-mode PA power supply. The signal over-sampling ratio must be sufficiently high (256 for EDGE) to push the quantization noise out of the signal band. A suitable filtering strategy is to apply a low-pass filter in the switch-mode power supply. A 3^{rd} order Butterworth filter can meet the system requirement with a bandwidth of three times the signal bandwidth. The group delay caused by this filtering 390 ns or 27 samples (with 256 over-sampled EDGE signal). This delay must be compensated in the phase path which can be implemented in the baseband DSP block.

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