Design Considerations for Supply Modulated EER Power Amplifiers

Jeffrey S. Walling
Electrical and Computer Engineering Dept.
University of Utah
Salt Lake City, UT
jeffrey.s.walling@utah.edu

David J. Allstot
Electrical Engineering Dept.
Stanford University
Palo Alto, CA
allstot@stanford.edu

Abstract—Envelope elimination and restoration (EER)-based transmitters provide increased spectral efficiency for modulation standards with large peak-to-average ratios. In contrast to an EER topology, increased spectral efficiency is traded off against decreased power-added-efficiency when a conventional linear power amplifier is used. Moreover, the advantage of scaled CMOS is exploited in the EER approach through the use of fast on-chip switching power amplifiers. Performance limitations associated with non-idealities in an EER system and its extensions (e.g., class-G) such as delay mismatches between the envelope (\(A\)) and phase (\(\phi\)) signal components, bandwidth limitations in the \(A\) and \(\phi\) paths, and glitches in the \(A\) signal path caused by switching between the dual-supply modulators, are examined.

Keywords—High Efficiency PAs, Linearization and Efficiency Enhancement Techniques, Supply Modulation

I. INTRODUCTION

Due to the ever increasing demands for bandwidth, the data modulation methods used in wireless services such as cellular, Wi-Fi and Bluetooth are made more complex in order to increase spectral efficiency. In order to amplify signals with increasingly complex modulation, the power amplifiers used must offer high linearity and efficiency at peak output power levels, while being able to operate over a signal with high peak-to-average power ratios (PAPR).

Examining the evolution of the Wi-Fi signal, shows modulation complexity has increased from Differential Quadrature Phase Shift Keying (D-QPSK) to Orthogonal Frequency Division Multiplexing (OFDM) with each subcarrier using up to 16 Quadrature Amplitude Modulation (16-QAM). These enhancements have enabled >5X improvements in data rates while increasing the PAPR of the signal.

Large values of PAPR are problematic because in conventional linear PAs, peak efficiency is only achieved for peak signals, such that the PA nearly saturates, a condition that occurs infrequently. Consider the OFDM modulated Wi-Fi signal, which has a PAPR≈13 dB with a Rayleigh distributed envelope; with such signal, the PA spends less than 1% of the time operating at peak amplitudes, and instead operates much closer to its average value much of the time. Because the power-added efficiency rolls off quadratically with a reduction in output voltage amplitude, this means that the average efficiency of the PA is greatly reduced.

Several techniques that improve PA efficiency while maintaining linearity are well known, including Doherty [1, 2], Chireix/Outphasing [3, 4], and Envelope Elimination and Restoration (EER) [5-7]. EER is attractive because it does not require passive components for matching and power combining which reduces cost and power losses.

EER is enabled using a mathematical transformation (e.g., CORDIC algorithm) of the signal from a Cartesian to polar coordinates. The resulting phase component, \(\phi\), is processed by a highly-efficient switching PA whereas the envelope signal, \(A\), is applied as the output from a power supply modulator as the power supply voltage of the PA.

A significant drawback of EER is that the transformations from Cartesian to polar coordinates are nonlinear [8, 9]:

\[
A(t) = \sqrt{I(t)^2 + Q(t)^2} \\
\phi(t) = \tan^{-1}(Q(t)/I(t))
\]

Thus, \(A\) and \(\phi\) occupy wider bandwidths than the original Cartesian signal with in-phase (\(I\)) and quadrature (\(Q\)) components, as shown in Fig. 2. While much of the energy of \(A\) resides at low frequencies, a much wider bandwidth is required to reconstruct the original signal with sufficient fidelity to meet system specifications.
paths are different; signals through the recombined signal. Moreover, the delays in the two bandwidths, information is lost, which reduces the fidelity between the tones used in [10, 11].

64 orthogonal tones occupy similar bandwidth as the two such as that used in the IEEE 802.11a standard, however, using only a two-tone excitation. For an OFDM signal system. Both of the above papers model non-linearity including the effects of EER bandwidth limitations on the paper also extends the work of Pedro, et al. [11] by dramatically from actual measured performance. This also examined herein, especially for EER transmitters that modulate and up-convert the \(\phi\) phase signal using a polar loop, minimizing the need for off-chip filters [12-14]. Finally conclusions are given in Section IV.

II. DELAY MISMATCH EFFECTS

Delay mismatches between the \(A\) and \(\phi\) phase paths occur in polar EER transmitters for two main reasons: (1) A \(\phi\) signal that is up-converted to RF experiences a greater group delay as it traverses interconnects, amplifiers, limiters, etc. (2) The \(A\) and \(\phi\) signal paths are both bandwidth limited; a power supply modulator limits the former and a phase-locked loop (PLL) the latter. Generally, the signals are low-pass filtered with different group delays. The effects of delay mismatches between the two paths are considered in this section, noting that in practice, the \(\phi\) signal is usually delayed relative to the \(A\) signal at the inputs of the PA.

Rudolph [9] has shown that delay mismatches affect the height of the shoulder of the complex signal (Fig. 2) but not the slope of the out-of-band roll-off. The roll-off characteristics are impacted, however, by the presence of a “vector hole” in the modulation, which is defined as a null in the envelope of the modulation. In other words, it is a region near the origin of the signal constellation diagram through which the modulation vector does not transition; i.e., large modulation phase shifts are avoided. The IEEE802.11a signal does not exhibit a vector hole.

Fig. 3 shows the power spectral density (PSD) of the recombined signals at the output of the PA relative to the mandated spectral mask and versus delay mismatch normalized to one symbol period; i.e., \(T_{\text{sym}} = 50\)ns for IEEE 802.11a. MATLAB® was programmed to generate 1 million random symbols for each of the two signals. From these symbols, a Welch spectral estimation with a resolution bandwidth of 30 kHz was performed and the maximum spectral components at all frequencies were selected. Note that the spectral masks for the IEEE 802.11a specification is violated for normalized delay mismatches \(>0.04T_{\text{sym}}\) (i.e., 2ns). It is interesting to note that as the delay mismatch is increased, the PSD of the recombined signals approach that of the \(\phi\) component. This follows because the absence of a vector hole means that the signal can undergo large phase jumps, which manifest as an increase in the out-of-band spectrum because a large change in phase corresponds to an instantaneous jump in frequency.
The margin-to-spectral mask and error vector magnitude (EVM) performance are plotted versus normalized delay mismatch in Fig. 4. Though both quantities degrade as the delay mismatch is increased, the margin-to-spectral mask is violated before the EVM. For relative delays >0.04\(T_{sym}\) (i.e., 2ns), the margin-to-spectral mask falls below 0 dB. The EVM specifications for the IEEE 802.11a (5.6%-rms) standard is violated for delay mismatches > 0.125\(T_{sym}\) (i.e., 6.25 ns). The key conclusion from this analysis is that delay mismatch has a greater impact on margin-to-mask performance so it is the parameter, rather than EVM, that should be optimized in such PA designs.

III. FINITE BANDWIDTH EFFECTS

Because of the finite bandwidths of the supply modulator and polar loop the \(A\) and \(\phi\) signals are low-pass filtered before recombination. Moreover, the bandwidths are different, which leads to unequal group delays for the two signals. In this section, an allpass filter is used to equalize the group delays so that finite bandwidth effects can be studied in isolation.

The \(A\) and \(\phi\) signals both occupy much larger bandwidths than the corresponding Cartesian representation as illustrated in Fig. 2. Because closed-form expressions do not exist for the exact bandwidth required in each path to reproduce the PA output signal with a given level of fidelity, MATLAB simulations again are used to determine the behavior.

A typical linear supply modulator is implemented using a low-dropout regulator, which comprises an operational amplifier enclosed in a negative feedback loop around a PMOS series pass transistor. These circuits typically display a low-pass characteristic dominated by the closed-loop frequency response of the amplifier. For this reason, the filter in the \(A\) envelope signal path is modeled as a first-order Butterworth low-pass filter with a variable 3 dB bandwidth.

A polar modulator typically comprises a type-I or type-II PLL which is more suited for high-bandwidth signals [13]. A type-I PLL loop has only one integrator so it exhibits a one-pole low-pass response; hence, the \(\phi\) phase path is also modeled using a first-order low-pass filter.

The PSDs for the IEEE 802.11a signals versus the \(A\) path bandwidth (\(\phi\) bandwidth held constant) are shown in Fig. 5, with the PSDs versus \(\phi\) path bandwidth (\(A\) bandwidth held constant) shown in Fig. 6. For delay mismatches (Fig. 3), the shoulder heights change but the slopes of the out-of-band spectra are unaffected (\(\approx 4\) dB/BWRF). Here, however, the bandwidth of the envelope path impacts the slope, shape and shoulder distance as shown. In fact, it must be significantly larger than the bandwidth of the original Cartesian signal (BW_RF) in order to meet the spectral mask requirements. Specifically, the \(A\) path for the IEEE 802.11a standards must have bandwidths greater than 2BW_RF (i.e., 40 MHz), respectively, while the \(\phi\) path must have bandwidth greater than 3BW_RF (i.e., 60 MHz). Achieving these bandwidths represents a significant design challenge; however, they are becoming more feasible with continued scaling of CMOS processes and ever improving design techniques.

Figs. 5 and 6 depict the bandwidth limitations in only the \(A\) path and \(\phi\) path, respectively. Margin-to-mask contours are plotted in Fig. 7 that depict the limitation of signals that are bandwidth limited in both the \(A\) and \(\phi\) paths simultaneously. Similar contours are plotted for EVM in Fig. 8. Many design choices meet both specifications, requiring different tradeoffs between the envelope and phase modulator bandwidths. As with delay mismatches, a system that satisfies the margin-to-mask specifications will generally satisfy the EVM specifications.

IV. CONCLUSIONS

A study of the effects of non-idealities/nonlinearities in EER/polar-based transmitters is presented. The effects of \(A\) and \(\phi\) path delay mismatches, \(A\) and \(\phi\) path band limiting and filtering are examined. EER/Polar systems are found to be tolerant of delay mismatches of approximately 4% of the symbol period, \(T_{sym}\), while still meeting the margin-to-mask and EVM specifications for the IEEE 802.11a standard. Moreover, the bandwidths of the \(A\) and \(\phi\) signal paths need to be approximately 2-3 times wider than the bandwidth of a Cartesian representation (BW_RF) of the original signal.
Fig. 5. PSD versus the $\phi$ path bandwidth ($A$ path bandwidth constant) for IEEE 802.11a.

Fig. 6. PSD versus the $A$ path bandwidth ($\phi$ path bandwidth constant) for IEEE 802.11a.

Fig. 7. Margin-to-mask contours (in dB) versus $A$ and $\phi$ path bandwidths for IEEE 802.11a.

Fig. 8. EVM contours (in %-rms) versus $A$ and $\phi$ path bandwidths for IEEE 802.11a.

REFERENCES


