A Digital Amplitude-to-phase Conversion for High Efficiency Linear Outphase Power Amplifiers

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Abstract – This paper introduces a digital amplitude-to-phase conversion scheme to facilitate the outphase amplifying concept, enabling the use of high efficiency, non-linear power amplifiers in linear systems. The digital implementation enables a fast and accurate amplitude-to-phase conversion by exploiting the computational capability that can be easily incorporated in today's integrated circuits (ICs). The proposed scheme minimizes amplitude variation of the outphase signals, which reduces the gain requirement of the power amplifiers that can be traded for higher efficiency. Our analysis demonstrates that outphase amplifying can be twice as efficient as the conventional class-A power amplifiers, and is suitable for wireless systems with large peak-to-average ratios.
Index terms – **Outphase, LINC, power amplifiers, power combiner, peak-to-average power ratios.**

I. MOTIVATION

A tradeoff exists between efficiency and linearity in conventional power amplifiers (PAs). Conducting class PAs such as class-A, and -AB offer excellent linearity but are inefficient [1-3]. On the other hand, switching class PAs such as class-E, and -F are highly nonlinear, but very efficient [4-6]. Modern wireless communication systems often employ intricate modulation schemes such as Orthogonal Frequency Division Multiplexing (OFDM) with multi-channel Quadrature Amplitude Modulation (QAM) in order to maximize bandwidth efficiency. Such modulation usually results in amplitude-modulated signals with large peak-to-average power ratios (PAPR) that require power amplifiers with extremely good linearity. Enabling the use of high efficiency, non-linear power amplifiers for linear systems is essential to significantly improve PAs' efficiency. This general technique is referred to as LInear amplification using Nonlinear Components (LINC) [7].

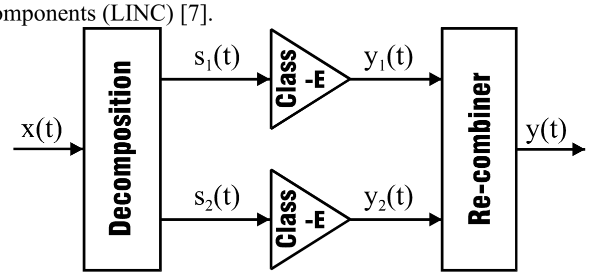

Figure 1: LINC Principle Diagram

Figure 1 describes the general principle of LINC. An amplitude-modulated signal $x(t)$ is first decomposed into two signals $s_1(t)$ and $s_2(t)$ that can be amplified using two highly efficient, non-linear power amplifiers. The PAs' outputs, $y_1(t)$ and $y_2(t)$, are then recombined to yield $y(t)$ for transmitting.

With this architecture, the goal is to design the decomposition and recombination such that the overall input-output characteristic is close to

$$y(t) = G \cdot x(t) \tag{1}$$

where G is the gain of the constituent power amplifiers.

For a practical implementation, the signal decomposition has to satisfy two conditions. First, in order to employ highly efficient, non-linear switching power amplifiers, the decomposed signals $s_1(t)$ and $s_2(t)$ can not have amplitude modulation. Furthermore, their amplitude envelope variations have to be limited to achieve the best efficiency. This is because switching power amplifier efficiency depends largely on its input drive threshold, which is the input level required to sufficiently drive the transistors into switching mode. Theoretically, switching power amplifiers can be designed for infinitely small input drive threshold. However, smaller drive thresholds correspond to larger gain. Unreasonably large power amplifier gain tends to be both unstable and inefficient. Second, the decomposition has to be such that its inverse function, which is performed by the re-combiner block, can be efficiently implemented using analog circuitry.

II. OUTPHASE AMPLIFYING

One method of decomposition is outphase amplifying, which was originally proposed by Chireix in 1935 [8,9]. An amplitude-modulated signal can be represented as a sum of two constant-amplitude, phase-modulated signals according to the simple identity,

$$a(t)\cos(\omega t + \theta) = \frac{A_{\max}}{2}\cos(\omega t + \theta + \phi) + \frac{A_{\max}}{2}\cos(\omega t + \theta - \phi)$$
$$\phi(t) = \cos^{-1}\left(\frac{a(t)}{A_{\max}}\right) \quad with \quad A_{\max} = \max|a(t)| \tag{2}$$

Essentially, the decomposition is the amplitude-to-phase mapping that results in two constant-amplitude signals. The re-combiner is simply the addition of the two power amplifiers' outputs. The amplitude is normalized to the maximum amplitude, $A_{\max}$, to ensure the inverse cosine conversion.

While the basic concept of outphase amplifying is simple, challenges exist in implementation. First, the practical implementation of the outphase amplifying technique critically depends on the accuracy and efficiency of the amplitude-to-phase conversion. Second, the sensitivity of bit-error-rate (BER) to gain and phase mismatch between the outphase paths determines the applicability of outphase amplifying in modern wireless communication systems.

III. OUTPHASE DECOMPOSITION

Traditionally, implementing the inverse cosine and other trigonometric functions using analog circuits are non trivial and inevitably complex. That is the main reason for the lack of popularity for the outphase amplifying concept since its invention in 1935. Instead, we propose a digital implementation to obtain the required accuracy and efficiency. The principle diagram in Figure 1 is re-drawn to explicitly reflect the digital implementation of the amplitude-to-phase conversion in Figure 2.

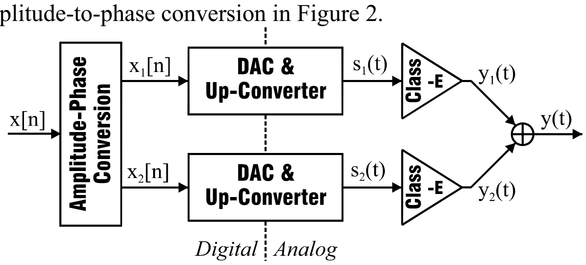

Figure 2: Outphase Amplifier with Digital Amplitude-to-Phase Conversion

Here, the amplitude-to-phase conversion is done in the digital domain, on the complex series $x[n]$ representing the quadrature time sample instances of the input signal. The outputs of the conversion box are two series $x_1[n]$, and $x_2[n]$ that correspond to the two constant-amplitude, phase-modulated signals $s_1(t)$ and $s_2(t)$ after the digital-to-analog converter (DAC) and Up-Converter. The re-combiner box is also replaced by the addition symbol to represent the power combiner functionality at the output.

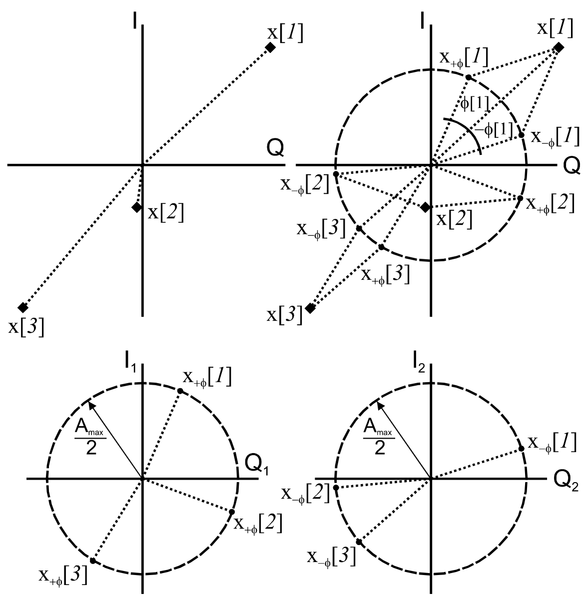

Figure 3: Amplitude-Modulated Constellations, and Their Corresponding Outphases

The digital amplitude-to-phase conversion is best illustrated using the constellation diagram in Figure 3. This figure shows an example of an amplitude-modulated signal with its three time sample instances on the top left I-Q quadrature constellation diagram. Each constellation $x[k]$ can be represented as the sum of two outphase vectors $x_{+\phi}[k]$ and $x_{-\phi}[k]$, such that all the outphase vectors have the same amplitude as shown in the top right. The angle between the two outphase vectors depends on the amplitude of the original constellation. A constellation with small amplitude (e.g. $x[2]$) corresponds to a large angle between the two outphases ($\angle x_{+\phi}[2]x_{-\phi}[2]$), while the constellation with large amplitude (e.g. $x[3]$) corresponds to a much smaller outphase angle ($\angle x_{+\phi}[3]x_{-\phi}[3]$). The first outphase channel contains all the $x_{+\phi}[k]$, while the second contains all the $x_{-\phi}[k]$. The two outphase channels are plotted in the bottom I-Q diagrams, showing constant-amplitude constellations.

The relation between the two digital outphases $x_1[n]$, and $x_2[n]$, and the original signal $x[n]$ is as follows.

- Input: $x[k]=\|x[k]\|\exp(j\theta[k])$
- Decomposition: $x[k]=x_1[k]+x_2[k]$, where

$$\begin{cases} x_1[k]=\dfrac{A_{\max}}{2}\exp(j(\theta[k]+\phi[k])) \\ x_2[k]=\dfrac{A_{\max}}{2}\exp(j(\theta[k]-\phi[k])) \end{cases}$$

with $\phi[k]=\cos^{-1}\left(\dfrac{\|x[k]\|}{A_{\max}}\right)$ and $A_{\max}=\max\|x[n]\|$

Although each of the outphase series $x_1[n]$, and $x_2[n]$ contains only constant amplitude constellations, their corresponding analog signals $s_1(t)$ and $s_2(t)$ do not have constant amplitude due to transition since

$$s_i(t)=\sum x_i[n]p(t-nT) \text{ where } p(t)\times p(-t) \text{ satisfies the Nyquist condition} \quad (3)$$

Figure 4 illustrates the nature of the outphase analog signal envelopes. The solid traces, connecting the outphase constellations ($x_{+\phi}[1], x_{+\phi}[2], x_{+\phi}[3]$) and ($x_{-\phi}[1], x_{-\phi}[2], x_{-\phi}[3]$), represent the continuous-time envelope amplitude of the two outphase analog signals $s_1(t)$ and $s_2(t)$.

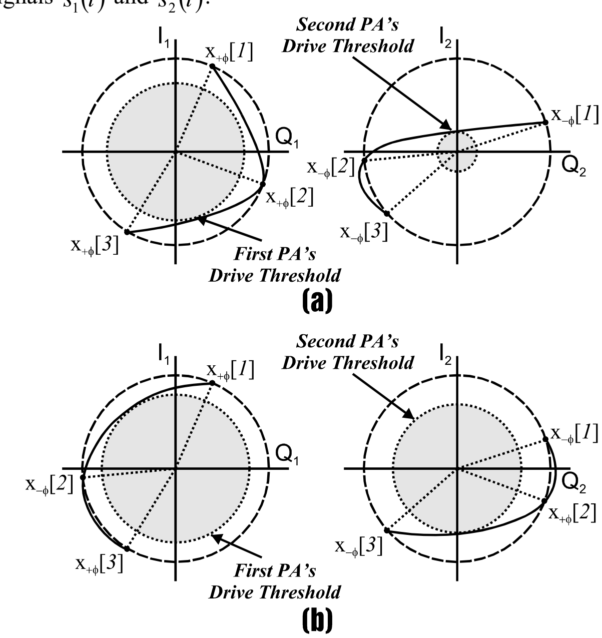

Figure 4: (a) Outphase Envelope Variation Due to Large Transition Angles (b) Envelope Variation Reduced by Interchanging $x_{+\phi}[2]$ and $x_{-\phi}[2]$

The non-constant envelope is similar to the fact that the analog waveform of M-ary PSK (Phase Shift Keying) modulation exhibits significant envelope variation despite all M-ary PSK constellations are on the constant amplitude circle. The envelope variation is most severe when there is a transition through zero. In Offset-QPSK and π/4-QPSK, the envelope variation is reduced by eliminating transition through zero, and limiting the transition angles.

As mentioned earlier, non-constant outphase envelope is not a problem as long as the envelope is larger than the input drive threshold required to sufficiently operate the pair of switching PAs. When the input signals are less than the drive thresholds, the switching PAs shutdown and their outputs collapse. In order to accommodate the full range of the outphase varying envelope, smaller input drive can be achieved by severely sacrificing efficiency and stability. Therefore, additional envelope varying control is needed to make the outphase technique practical.

To control envelope variation, we exploit the fact that the outphase channel assignment of the decomposition is not unique. For every symbol index k, the "forward-angle" outphase $x_{+\phi}[k]$ can be assigned to the first channel $x_1[n]$, and the "reverse-angle" $x_{-\phi}[k]$ to the second channel $x_2[n]$, or vice versa. This additional degree of freedom can be used to limit the transition angles and control the envelope variation. Following is the outphase assignment procedure.

1. Let $x_1[k-1]$ and $x_2[k-1]$ be the previous symbol of the first and second outphase channels, respectively.
2. For the current symbols, calculate the two transition angles when $x_{+\phi}[k]$ is assigned to the first outphase, and $x_{-\phi}[k]$ to the second outphase.

   $a_1 = \angle x_1[k-1]x_{+\phi}[k]$ and $a_2 = \angle x_2[k-1]x_{-\phi}[k]$

   (transition angle is defined as the smallest positive angle between the two constellation vectors)
3. Calculate the two alternate transition angles if the assignment is reversed, $x_{+\phi}[k]$ to the second outphase, and $x_{-\phi}[k]$ to the first outphase.

   $b_1 = \angle x_1[k-1]x_{-\phi}[k]$ and $b_2 = \angle x_2[k-1]x_{+\phi}[k]$
4. Eliminate the assignment that results in the largest transition angle among the four possible angles.

   If $\max(a_1, a_2) > \max(b_1, b_2)$ then

   $x_1[k] = x_{-\phi}[k]$ and $x_2[k] = x_{+\phi}[k]$

   Otherwise

   $x_1[k] = x_{+\phi}[k]$ and $x_2[k] = x_{-\phi}[k]$

This technique is demonstrated by comparing Figure 4(a) to Figure 4(b). In Figure 4(a), the transition angle between $x_{-\phi}[1]$, and $x_{-\phi}[2]$ is large, causing significant amplitude variation. Realizing that the transition angle between $x_{-\phi}[1]$, and $x_{+\phi}[2]$ is much smaller, we can simply swap $x_{+\phi}[2]$, and $x_{-\phi}[2]$. The result is plotted in Figure 4(b), showing great reduction in amplitude variation for both outphases.

IV. EFFECT OF MISMATCH

This section investigates the effect of mismatch among the two outphase paths on the outphase power amplifier performance.

Let α be the total gain mismatch, and δ° be the total phase mismatch between the two outphases as shown in Figure 5. The mismatch causes the outphase constellation x_2 to shift by an error vector $\vec{e}$ to x_{2err}. After recombining, the original constellation x is also shifted by the same error vector $\vec{e}$. The error vector is a function of the mismatch and the nominal gain of the outphase channels,

$$\|e\| = \frac{G \cdot A_{\max}}{2}\sqrt{1+(1+\alpha)^2 - 2(1+\alpha)\cos(\delta)} \qquad (4)$$

where G is the nominal gain of each channel

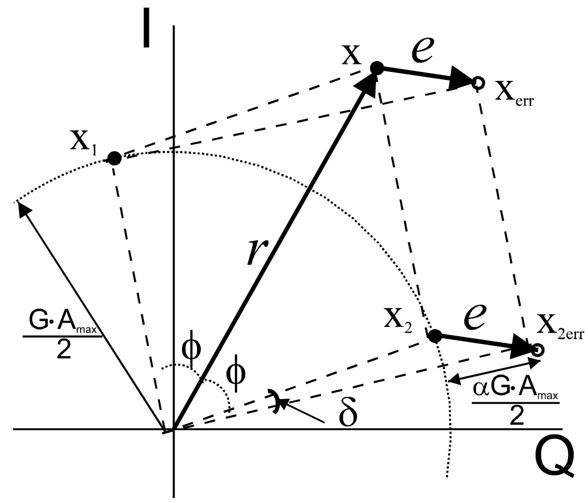

Figure 5: Error Vector Dues to Gain and Phase Mismatch Between the Two Outphase Paths

The error vector can be treated as the additional noise caused by the mismatch in the outphase power amplifier. The effect of mismatch can be quantified by signal-to-noise ratio (SNR) degradation as follows,

- Let N be the mismatch-free noise power, the corresponding mismatch-free SNR is defined as

$$SNR_{orig} = \frac{\|r\|^2}{N} \text{ where } \|r\| = G \cdot A_{\max}\cos(\phi) \text{ is the signal amplitude} \qquad (5)$$

- Given the mismatch vector $\vec{e}$, the total SNR can be calculated as,

$$SNR_{total} = \frac{\|r\|^2}{N + \|e\|^2} \qquad (6)$$

Or

$$SNR_{total} = \frac{1}{\dfrac{1}{SNR_{orig}} + \dfrac{\|e\|^2}{\|r\|^2}} \qquad (7)$$

where $\dfrac{\|e\|^2}{\|r\|^2} = \dfrac{1+(1+\alpha)^2 - 2(1+\alpha)\cos(\delta)}{4\cos^2(\phi)}$

SNR degradation due to mismatch is plotted Figure 6. A 1% gain mismatch causes unnoticeable change in SNR. A 2% gain and 2° phase mismatch, which is achievable with current IC technology, causes a 2dB SNR degradation.

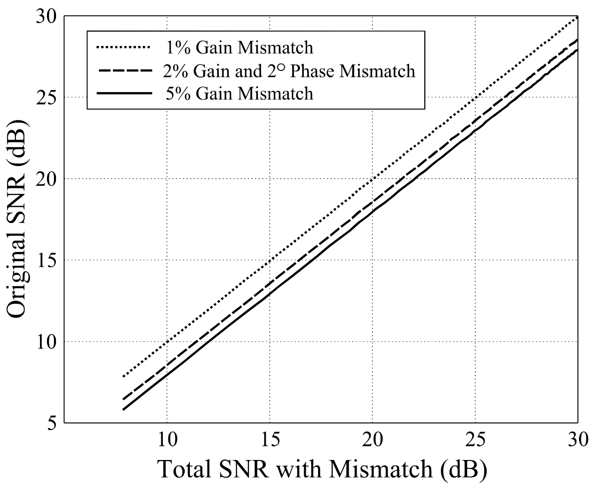

Figure 6: SNR Degradation Due to Mismatch

V. POWER CONSUMPTION COMPARISION

Previous sections have demonstrated that the outphase amplifying concept can be digitally implemented, and is suitable for linear amplification. This section compares the DC power consumption of the outphase PA to that of a conventional class-A PA, which offers similar linearity. For a fair comparison, both the class-A PA, and the outphase PA have the same maximum transmitting power of $P_{\max}$. Referring to Figure 2, let P_{conv} be the cost of power to carry out the amplitude-to-phase conversion, P_{DAC} be power consumption of one DAC and up-converter, $P_{Class-E}$ be the power consumption of one class-E PA, then the total power consumption of the outphase PA is given in (8). The DAC and class-E PA power are doubled because of the two outphase channels.

$$P_{outphase} = P_{conv} + 2P_{DAC} + 2P_{Class-E} \tag{8}$$

Since the PA usually consumes much more power than any other blocks in the transmitter [10], the total power consumption can be approximated to,

$$P_{outphase} \approx 2P_{Class-E} \tag{9}$$

The efficiency of a class-E PA is defined as the ratio of its output power to its DC power consumption,

$$\eta_{Class-E} = \frac{P_{out-E}}{P_{Class-E}} \tag{10}$$

The two class-E PAs always generate the same amount of output power P_{out-E}. Depending on the outphase angle, the in-phase components add up to the total output transmitting power, while the quadrature components are dissipated in the power combiner. Therefore, the maximum transmitting power is twice the output power of the individual class-E PA, when the two outphases are perfectly in-phase ($P_{\max} = 2P_{out-E}$). The outphase PA power consumption in (9) can be rewritten in terms of the maximum transmitting power P_{max}, and individual class-E efficiency $\eta_{Class-E}$.

$$P_{outphase} \approx P_{Class-E} = \frac{P_{\max}}{\eta_{Class-E}} \tag{11}$$

The outphase PA can be compared to a conventional class-A PA with the same maximum transmitting power $P_{\max}$. The same approximation, namely $P_{DAC} << \frac{P_{\max}}{\eta_{class-A}}$ where $\eta_{Class-A}$ is the efficiency of a class-A PA, is made for the total DC power consumption given in (12).

$$P_{linear} = P_{DAC} + \frac{P_{\max}}{\eta_{class-A}} \approx \frac{P_{\max}}{\eta_{class-A}} \tag{12}$$

Then,

$$\frac{P_{outphase}}{P_{linear}} = \frac{\eta_{class-A}}{\eta_{class-E}} \tag{13}$$

A typical integrated class-A PA has an efficiency of 30%, while the class-E efficiency is typically around 50-70%. The outphase PA reduces the power consumption by approximately a factor of 2 compared to a class-A PA.

VI. CONCLUSION

Practical implementation of the outphase amplifying technique are realized digitally to take advantage of the vast computational capability of advanced IC technology. In addition, an optimal outphase assignment scheme has been proposed to drastically reduce the outphase amplitude variation and relax the switching PA gain requirement. The outphase PA's mismatch tolerance makes it suitable for modulation systems with demanding linearity while providing significantly higher efficiency than a traditional class-A linear power amplifier.

VII. ACKNOWLEDGEMENT

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