Bandwidth Requirements of Envelope Amplifiers for Polar Transmitters

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***Abstract* — This paper analyzes important bandwidth requirements of supply modulator envelope amplifiers for envelope elimination and restoration power amplifiers, EER PAs, or polar transmitters.**

Although desired spectral efficiency demands for highly confined spectra of modern wireless signals, their amplitude envelope is known to have a theoretically infinite bandwidth. So, since this specification must necessarily be relaxed from an engineering perspective, the present analysis will focus on the minimum linear processing bandwidth and slew rate required to meet the standard specs of total signal distortion and adjacent signal power ratio.

I. INTRODUCTION

The compromise between power efficiency and linearity has been one of the major design constrains of wireless communication systems. This is due to the fact that, while power efficiency determines the talk time of wireless handsets or the network grid consumption of base-stations (in this case not only it is important the waste of power as heat, as also the power needed to cool down the equipment must be accounted for), linearity specs determine the signal quality specifications, or the spectral usage efficiency. In fact, nonlinear impairments of wireless circuits play a key role on both co-channel and adjacent-channel distortion. With the additive noise, the former imposes a lower bound on the bit-error-rate of the transmission channel, while the latter sets a threshold on the attainable bit-error-rate of any potential adjacent-channel [1]. This is why two of the most important linearity specifications of wireless systems are the error vector magnitude, EVM, and the adjacent-channel power spectral density ratio. In this work, we will adopt the so-called raw-EVM, or the normalized mean square error, *NMSE*, defined as:

$$NMSE = \frac{\frac{1}{N}\sum_{n=1}^{N}\left|y(n)-\hat{y}(n)\right|^2}{\frac{1}{N}\sum_{n=1}^{N}\left|\hat{y}(n)\right|^2} \qquad (1)$$

- in which $y(n)$ is the actual output and $\hat{y}(n)$ stands for the desired output, i.e. a linear scaled version of the input - to quantify the co-channel distortion; and the adjacent-channel power ratio, *ACPR* defined as:

$$ACPR = \frac{\frac{1}{Bw}\sum_{k=-Bw/2}^{Bw/2}\left|Y(k)\right|^2}{\frac{1}{Bw}\sum_{k=Bw/2}^{3Bw/2}\left|Y(k)\right|^2} \qquad (2)$$

- in which $Y(k)$ is the output power spectral density - to measure the adjacent-channel distortion.

A simple, but surprisingly general, argument to support the widely known knowledge that power efficiency and spectral efficiency (linearity) are two conflicting requirements can be derived from the very basic energy conservation relations of any electronic transducer as a PA.

According to Fig. 1 we immediately conclude that the sum of the powers delivered to the amplifier – the signal input power, P_{in}, and the dc supplied power, P_{dc} – must equal the sum of the output powers – signal power delivered to the load, P_{out}, and dissipated power, P_{dis}:

$$P_{dc} + P_{in} = P_{out} + P_{dis} \qquad (3)$$

So, as the amplifier is supposed to be active, it must deliver a signal gain given by:

$$G \equiv \frac{P_{out}}{P_{in}} = 1 + \frac{P_{dc} - P_{dis}}{P_{in}} \qquad (4)$$

which leads us to the conclusion that, since the dissipated power must be non-negative, any PA dependent on a real power supply (i.e., one whose available power is finite) must evidence a compressing gain, and thus nonlinearity, after a certain level of input excitation power.

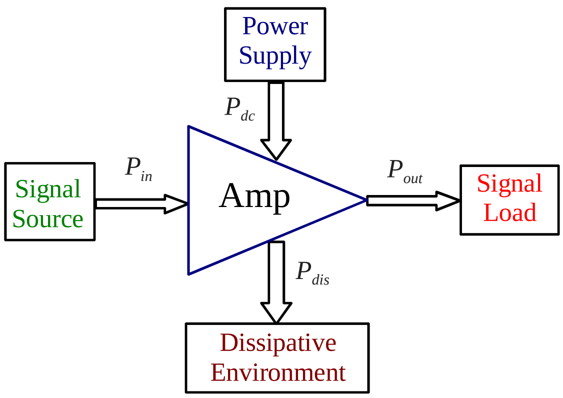

Fig. 1 Power relationships in a general electronic amplifier.

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Fortunately, recent advances in power amplifier design have shown that this is, by no means, a physical limit that must be expected from all systems. Indeed, in [2], [3] the authors have shown that, if one takes into account the power added efficiency definition and the power relations of the PA in Eq. (3):

$$PAE = \frac{P_{out} - P_{in}}{P_{dc}} \tag{5}$$

which can be rewritten as:

$$PAE = (G-1)\frac{P_{in}}{P_{dc}} \tag{6}$$

we immediately conclude that theoretical 100% *PAE*, i.e. 100% power efficiency, and perfect linearity, i.e., constant gain and thus ideal spectral confinement, can be obtained if the supplied power, P_{dc}, is made proportional according to the input power, P_{in}. This is the underlying theory behind the experiments that have recently being conducted with several types of supplied modulated PAs as the EER or even the polar transmitter.

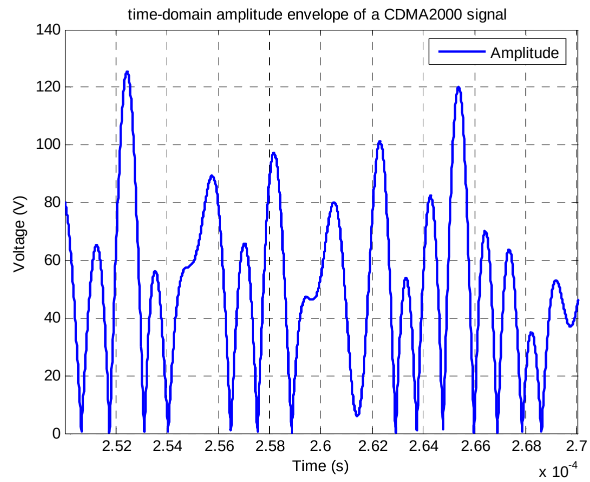
II. Supply Modulator Speed Requirements

A. General Background

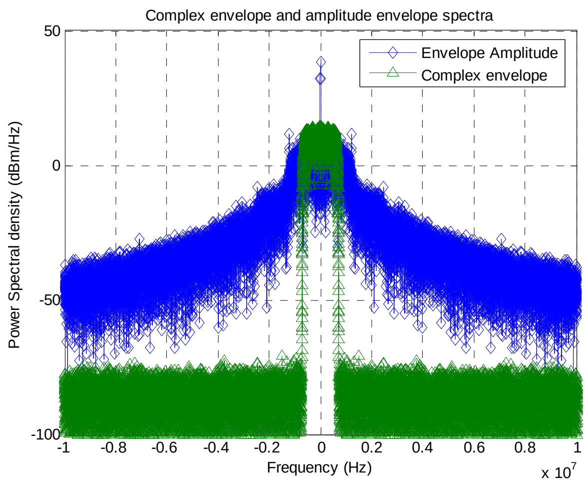
As happens in almost all situations in engineering, this theoretical possibility does not come without a price. Indeed, a closer look into Eq. (5) and (6) indicates that the high linearity requirements of modern wireless systems (with *NMSE* and *ACPR* on the order of -30 to -40dB and 40 to 45dBc, respectively) can only be achieved with almost ideally linear and highly efficient supply modulators. And, although easier to get than in a RF PA, because of the much lower frequency involved (please remember that we are no longer processing a modulated RF carrier of several hundreds of MHz or Ghz, but an amplitude envelope signal of several MHz), this still presents a tremendous challenge that has recently drawing the attention of an increasingly higher number of researchers.

First of all, one should keep in mind that a 100% *PAE* can only be achieved with a 100% efficient PA and a 100% efficient supply modulator, SM. Indeed, it can be easily shown that the overall efficiency of the system is given by the product of the individual efficiencies of the PA and the SM. Just to have a concrete idea of what this means, we need to have both a PA and a SM with (not easy to get) figures of slightly higher than 70% efficiency to reach the nowadays goal of 50% overall efficiency of wireless infrastructure base-stations. And this asks for switching mode operation of both the RF PA and the SM. Thus, classical class AB current mode RF PAs are being substituted by class F, class D or class E PAs, and the supply modulator must be a highly efficient switching mode power supply with variable reference [according to $a_x(t)$], something similar to the so-called digital audio amplifier technologies [4].

Furthermore, as the envelope input power, P_{in}, results from processing the original complex (amplitude and phase) envelope, $x(t) = a_x(t).e^{j\phi_x(t)}$, through a strong even-order nonlinearity, $a_x(t) = \text{abs}[x(t)]$, this amplitude signal will have an infinite bandwidth, even if $x(t)$ has a confined spectrum. Indeed, as illustrated in Fig. 2-(a) for a CDMA2000 signal, the sharp transitions near zero amplitude provide a much richer frequency content to $a_x(t)$, in comparison to the smooth $x(t)$.



(a)



(b)

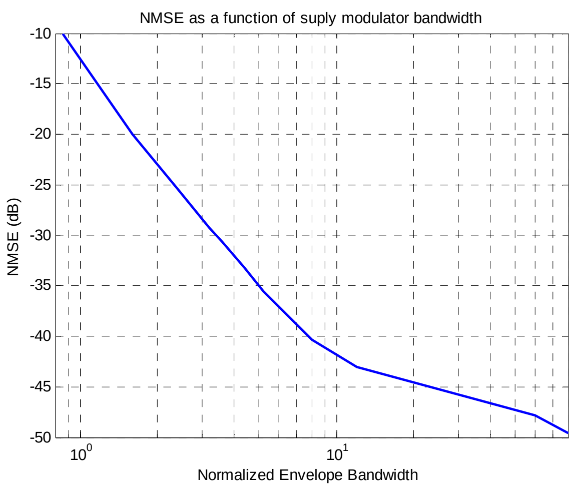
Fig. 2 Example of (a) the time-domain amplitude envelope of a CDMA2000 signal and (b) complex envelope and amplitude envelope spectra.

Fortunately, the amplitude signal spectrum depicted in Fig. 2-(b) shows that, not only there is a large amount of power concentrated at dc (consequence of the non-null average of $a_x(t)$, since $|x(t)|$ is assumed to be always positive), as the power spectrum density of $a_x(t)$ falls very rapidly with the increasing frequency. Therefore, it is important to devise the minimum speed requirements, both in terms of linear bandwidth, but also of nonlinear slew rate, the SM circuitry must obey, so that the supply modulated RF PA or the polar transmitter can meet the *NMSE* and *ACPR* specs. This is the main objective of the remainder of this paper.

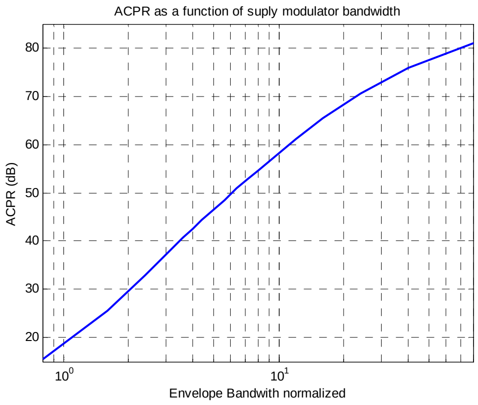
B. Linear Bandwidth Requirements

For studying the bandwidth requirements, we conceived an ideal linear supply modulator whose bandwidth limitation is imposed by a low-pass second-order L-C filter as the ones normally used in digital audio amplifiers, switching mode power supplies and in some of previously reported supply modulators [5].

We started by extracting the amplitude envelope with an ideal envelope detector, $a_x(t) = \text{abs}[x(t)]$, and then forced it to pass through the filter to obtain $a_y(t) = H[a_x(t)]$, where $H[.]$ stands for the linear dynamic filter operator. After this, we considered that the RF PA was ideal, so that the resulting complex envelope signal was rebuilt by: $y(t) = a_y(t).e^{j\phi_x(t)}$. Finally, the signal quality was quantified via the co-channel and adjacent-channel distortion with the *NMSE* and *ACPR* metrics defined in Eq. (1) and (2). These are the results shown in Fig. 3-(a) for *NMSE*, and in Fig. 3-(b) for the *ACPR*, as a function of the SM bandwidth normalized to the original complex envelope bandwidth, Bw = 1.25MHz. It is clear that a bandwidth of slightly less than 5Bw is enough for achieving both the desired 45 dBc of *ACPR* and -35 dB of *NMSE*. Although still high for the majority of switching mode supply modulators, this value is, at least for this signal format, considerably smaller than the 10Bw rule of thumb sometimes adopted.



(a)



(b)

Fig. 3 *NMSE* (a) and *ACPR* (b) as a function of supply modulator bandwidth normalized to the complex envelope bandwidth, Bw = 1.25MHz.

C. Nonlinear Slew Rate Requirements

Now, let us direct our attention to another bandwidth requirement, often overlooked: the slew rate. As is known from basic electronics [6], the slew rate, *SR*, of a voltage amplifier measures the maximum derivative with respect to time with which its output can vary. Above that speed limit, the output will not follow the amplified input, but will increase (or decrease) at a constant rate of *SR* (V/s), converting any waveform into a positive or negative slope ramp. Slew rate constitutes, therefore, a fundamental nonlinear process that, contrary to the linear bandwidth, only manifests itself when very high excitation amplitudes and frequencies are simultaneously involved. In high power devices, as the ones necessary for driving the Si LDMOS or GaN HEMT based drain-modulated PAs that are being designed for wireless base-stations, slew rates of tens or hundreds of V/μs can be the limiting factor in selecting appropriate SM amplifiers.

For studying the effect of SM slew rate, we again extracted the input amplitude envelope, $a_x(t)$, and processed it by the following algorithm accounting for the slew rate.

$a_y(1) = a_x(1)$
for n=2 to n
{
  erro= $A_v.a_x(n) - a_y(n\text{-}1)$
  if $|\text{erro}|/T_s > SR$
  then $a_y(n) = a_y(n\text{-}1) + sign(\text{erro}) \times SR/\, T_s$
  else $a_y(n) = A_v.x(n)$
}

Here, A_v is the amplifier voltage gain, $a_x(n)$ and $a_y(n)$ are the input and output amplitudes, respectively, sampled at the time instant n and with a sampling period of T_s. Then, we again used the resulting output amplitude to build the output complex envelope as $y(t) = a_y(t).e^{j\phi_x(t)}$.

Fig. 4 depicts a sample of the input and output time-domain waveforms of $a_x(t)$ and $a_y(t)$ when the SM is intended to provide a supply voltage of about a_{y_Max} = 30 V peak and is subject to a slew rate of 45 V/µs. Note that it is in the region of lowest values of $a_y(t)$ that the limited SR manifests its impact. In fact, it is there that the input amplitude has its highest time derivative, something on the order of $a_{y\_Max}.Bw/2$. So, we may anticipate that a signal format like EDGE, that is known for presenting a hole in the center of the constellation diagram, and thus having smoother transitions of $a_y(t)$ in the low voltage end, will be more immune to limited SM slew rate.

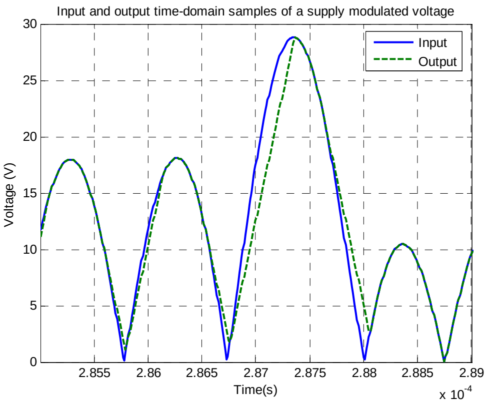
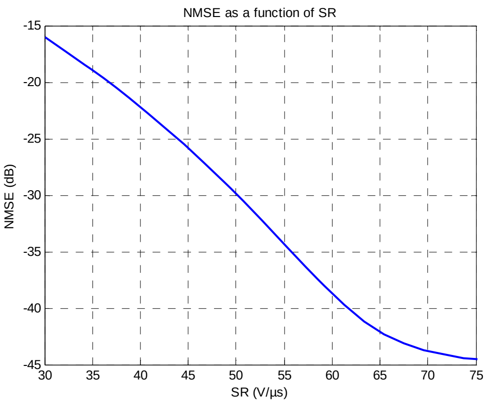


Fig. 4 Input and output time-domain samples of a supply modulated voltage of about 30V peak and subject to a moderate slew rate of 45 V/µs.

Finally, we tested several SR values and calculated the corresponding *NMSE* and *ACPR*. Fig. 5-(a) and Fig. 5-(b) depict these results. It is clear that, for the adopted CDMA2000 signal format, a slew rate of about 55 V/µs, i.e. of nearly one half the maximum output signal slope, is needed to guarantee the desired 45 dBc of *ACPR* and -35 dB of *NMSE*.



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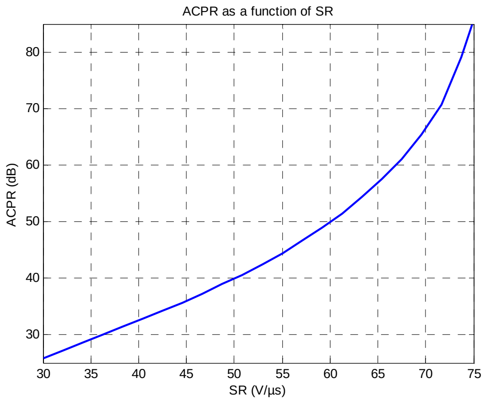

(b)

Fig. 5 *NMSE* (a) and *ACPR* (b) as a function of supply modulator slew rate for a CDMA2000 envelope amplitude signal with 30 V peak voltage.

III. Conclusions

In this work we investigated the frequency requirements, in terms of linear bandwidth, but also of slew rate, imposed by supply modulators of EER amplifiers or polar transmitters. Considering usually accepted specs of about -35 dB *NMSE* and 45 dBc of *ACPR*, we concluded that, for processing a 1.25 MHz bandwidth CDMA2000 signal, we would need about 5 times the original signal bandwidth and a slew rate of nearly one half the maximum output signal slope.

However, we also concluded that, since the maximum signal slope appears in the envelope zero crossings, other signal formats, as the EDGE, that show a hole in the center of the constellation diagram may be more immune for this type of supply modulator impairment.

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