A Multi-branch Linear Amplification with Nonlinear Components Technique

João Guerreiro^(1,2), Rui Dinis^(1,3), Paulo Montezuma^(1,3), João P. Oliveira⁽³⁾, Mário Silva^(1,2)

⁽¹⁾ IT, Instituto de Telecomunicações, Lisboa, Portugal.

⁽²⁾ UAL, Universidade Autónoma de Lisboa, Lisboa, Portugal.

⁽³⁾ FCT, Universidade Nova de Lisboa, Monte de Caparica, Portugal.

Abstract—Single carrier modulations with high order constellations or orthogonal frequency division multiplexing (OFDM) modulations can be employed to obtain spectrally efficient communications. However, such modulations lead to signals with large envelope variations, where the amplification is challenging and usually energy-inefficient. The linear amplification with nonlinear components (LINC) makes use of two efficient, nonlinear power amplifiers for an efficient amplification. In this work, we propose the multi-branch LINC (MB-LINC) technique, which allows the use of highly-efficient nonlinear amplifiers without increasing the nonlinear distortion effects, even if the amplifiers are not balanced¹.

Index Terms—OFDM, PAPR, amplification, energy efficiency.

I. INTRODUCTION

Two of the most important aspects of wireless communication systems are the energy and the spectral efficiency. Achieving these two goals simultaneously is a very challenging task, specially when considering spectrally efficient modulations such as single carrier modulations with high order constellations or orthogonal frequency division multiplexing (OFDM) modulations [1], which are known to exhibit large peak-to-average power ratio (PAPR). To avoid nonlinear distortion effects, which are more critical for larger constellations, and to avoid out-of-band emissions and maintain the high spectral efficiency of such modulations, one should consider quasi-linear power amplifiers (PA). However, these amplifiers have low amplification efficiency and require high output back-off to accommodate the high PAPR of the input signals in the linear amplification region, which reduces further more the energy efficiency. Thorough the last years, many techniques that aim to reduce the PAPR prior to the amplification and improve the amplification efficiency have been proposed [2], [3]. Besides these PAPR reducing techniques, another solution for the amplification issues associated with spectral efficient transmission schemes is the linear amplification with nonlinear components (LINC) technique [4], where two simple and

highly efficient, nonlinear amplifiers (such as the ones of class E, F [5] or D [6]) are combined to obtain an efficient amplification. The unavoidable losses associated with the combination process of the LINC can be compensated by the fact that we can employ efficient, grossly nonlinear power amplifiers, with each amplifier driven by a signal with a quarter of the required power that would be necessary if a single amplifier was employed.

In terms of monolithic implementation, namely using a low cost CMOS technology, the downscaling evolution of the technology nodes verified in the past decades has enabled the integration of an increasing number functionalities in the same integrated circuit (IC) die, in a system on chip (SoC) architecture [7]. However, this size downscaling is followed by a reduction of the maximum supply voltage (V_{supply}) that can be applied to the IC [8], thus imposing serious limitations on the maximum output power (related with V_{supply}^2) that can be extracted from one single traditional RF Power Amplifier (namely, using a class A biasing scheme). To overcome it, parallel circuit techniques [9] can be used to combine the power of a set of multiple RF power amplifier branches operating, each one, at a low supply voltage.

In this work, we propose the multi-branch LINC (MB-LINC) technique where P nonlinear amplifiers are considered to achieve an efficient amplification and at the same contributing to improve the total output RF power. Therefore, the power consumption of each amplifier can be made very low by increasing P . One of the major problems of conventional LINC is the nonlinear distortion effects caused by the imbalances of the amplifiers' gain. In this work, it is shown that the robustness of MB-LINC to nonlinear distortion effects regarding mismatches in the amplifiers' gain is similar to that of conventional, two-branch LINC.

This work is organized as follows: in Sec. II a brief review of conventional LINC schemes is presented. Sec. III is dedicated to the explanation of the proposed MB-LINC amplification scheme. In Sec. IV it is shown how the nonlinear distortion effects associated to the MB-LINC can be evaluated and a set of results regarding the power spectral density (PSD) and the self-to-interference (SIR) of the resultant MB-LINC signals are presented. Sec. V shows the conclusions of this paper.

¹This work is supported by the European Regional Development Fund (FEDER), through the Competitiveness and Internationalization Operational Program (COMPETE 2020) of the Portugal 2020 framework, Regional OP Centro (POCI-01-0145-FEDER-030588), Regional OP Lisboa (Lisboa-01-0145-FEDER-03058) and by FCT/MEC through national funds, under Project PES3N (SAICT-45-2017-02) and Instituto de Telecomunicações project UID/EEA/50008/2013.

II. CONVENTIONAL LINC

Let us start by considering a given bandpass signal written in the form

$$s^{(bp)}(t) = r(t) \cos(2\pi f_c t + \phi(t)), \quad (1)$$

where f_c represents the carrier frequency. Its complex-envelope, $s(t)$ can be written as

$$s(t) = r(t) \exp(j\phi(t)), \quad (2)$$

where $r(t)$ and $\phi(t)$ represent the absolute value and phase, respectively. In conventional LINC techniques, the signal $s(t)$ is separated into $N = 2$ constant-envelope and out-phasing signals, $s_1(t)$ and $s_2(t)$. Without loss of generalization, we will illustrate the LINC concept for the specific case where the envelope is real-valued i.e., $\phi(t) = 0$, which means that $s(t) = r(t)$ (the extension to other cases is straightforward).

Fig. 1 represents the conventional LINC decomposition under two components that lie in a circle with radius $R_M/2$. From the vectorial representation of the figure, we can define

Fig. 1. Vectorial representation of the conventional LINC decomposition.

the two LINC components as

$$s_1(t) = \frac{R_M}{2} \exp\left(-j\frac{\theta(t)}{2}\right), \quad (3)$$

and

$$s_2(t) = \frac{R_M}{2} \exp\left(j\frac{\theta(t)}{2}\right). \quad (4)$$

Since these two signals have constant-envelope, they can be efficiently amplified with nonlinear power amplifiers. When the amplifiers' gain are equal, the combination of the two LINC components produce a scaled replica of $s(t)$, eventually clipped to a level R_M . In fact, the combination of the two signals can be generally thought as a clipped, nonlinearly distorted version of $s(t)$, since the maximum amplitude of the combined signal is limited to R_M (which is a value obtained when $\theta/2 = 0$). Under these conditions, the LINC process can be modeled as a bandpass memoryless nonlinearity characterized by the corresponding amplitude modulation/amplitude

modulation (AM-AM) and amplitude modulation/phase modulation (AM-PM) conversion functions [10]. The complex envelope of the nonlinearly distorted signal can be written as

$$y(t) = A(r(t)) \exp(j(\phi(t) + \Theta(r(t)))), \quad (5)$$

where $A(r(t))$ and $\Theta(r(t))$ represent the AM-AM and the AM-PM conversion functions, respectively. For the specific case of an envelope clipping, we have $\Theta(r(t)) = 0$ and

$$A(r(t)) = \begin{cases} r(t), & r(t) \leq R_M \\ R_M, & r(t) > R_M, \end{cases} \quad (6)$$

where R_M denotes the clipping level. Still considering the geometry represented in Fig. 1, we can write, omitting the time dependence,

$$s = \frac{R_M}{2} \exp\left(-j\frac{\theta}{2}\right) (1 + \exp(j\theta)) = R_M \cos\left(\frac{\theta}{2}\right). \quad (7)$$

Note also that

$$\frac{\theta}{2} = \arccos\left(\frac{s}{R_M}\right). \quad (8)$$

Under these conditions, and since $\sin(\arccos(x)) = \sqrt{1 - \cos^2(\arccos(x))} \approx \sqrt{1 - x^2}$, we can write

$$s_1 = \frac{s}{2} + j\frac{R_M}{2} \sqrt{1 - \frac{s^2}{R_M^2}} \approx \frac{s}{2} + j\frac{1}{2} \sqrt{R_M^2 - s^2}, \quad (9)$$

and

$$s_2 = \frac{s}{2} - j\frac{1}{2} \sqrt{R_M^2 - s^2}. \quad (10)$$

Besides this “global nonlinearity” of the LINC process, which is an envelop clipping, we can define a bandpass memoryless nonlinearity for the p th branch (where p can be either 1 or 2, depending on the specific amplification branch), whose the AM/PM conversion function is defined as

$$\Theta_p(r) = (-1)^p \arccos\left(\frac{r}{R_M}\right), \quad (11)$$

and the AM/AM conversion function is $A_p(r) = \frac{R_M}{2}$.

III. MULTI-BRANCH LINC

In the following, we propose the MB-LINC transmitter, which generalizes the LINC concept by considering up to P amplification branches. In this technique, the signal $s(t)$ is divided into P components that are amplified separately through a set of nonlinear amplifiers, as is shown, in Fig. 2. As in conventional LINC, when there are P amplifying branches, the vectorial representation of the different LINC components is such that their sum produces the envelope at a given time instant.

Fig. 3 shows the vectorial representation of the MB-LINC components when P is even ($P = 4$) and when P is odd ($P = 5$). From this figure, it can be noted that, regardless of the value of P , the p th signal (i.e., the signal associated with the p th amplifying branch) of the MB-LINC decomposition is defined as

$$s_p = \frac{R_M}{P} \exp\left(j\frac{\theta}{2} (2p - P - 1)\right). \quad (12)$$

Fig. 2. Vectorial representation of the conventional LINC decomposition.

Fig. 3. Vectorial representation of the MB-LINC decomposition when $P = 5$ (a) and when $P = 4$ (b).

Therefore, the complex envelope of the MB-LINC output, given by the combination of the P signals, can be written as

$$\begin{aligned} s &= \sum_{p=1}^P s_p \\ &= \sum_{p=1}^P \frac{R_M}{P} \exp\left(j\frac{\theta}{2}(2p - P - 1)\right) \\ &= \frac{R_M}{P} \exp\left(-j\frac{\theta}{2}(P - 1)\right) \frac{1 - \exp(j\theta P)}{1 - \exp(j\theta)}. \end{aligned} \quad (13)$$

Under these conditions, we may write

$$s = \frac{R_M}{P} \frac{\sin\left(\frac{P\theta}{2}\right)}{\sin\left(\frac{\theta}{2}\right)} = R_M f_P(\theta), \quad (14)$$

with

$$f_P(\theta) = \frac{\sin\left(\frac{P\theta}{2}\right)}{P \sin\left(\frac{\theta}{2}\right)}. \quad (15)$$

It should be noted that, if $P = 2$, we have the conventional two branch LINC, i.e.,

$$s = \frac{R_M}{2} \frac{2 \sin\left(\frac{\theta}{2}\right) \cos\left(\frac{\theta}{2}\right)}{\sin\left(\frac{\theta}{2}\right)} = R_M \cos\left(\frac{\theta}{2}\right), \quad (16)$$

as obtained in (7).

Fig. 4 shows the envelope of a given OFDM signal before and after the LINC amplification scheme. The values of the envelope at the points A, B and C can be vectorial built by the LINC components represented in the Fig. 5.

Fig. 4. Envelope of a given OFDM signal before and after the LINC amplification scheme.

Fig. 5. Vectorial representation of the MB-LINC components with $P = 4$ for the three envelope values identified in Fig. 5: almost 0 (A), clipped to R_M (B) and lower than R_M (C).

Clearly, when the envelope is almost 0, the phase of the LINC components approaches $\pm\pi/2$. On the other hand, when the envelope is clipped, the LINC components are aligned with the direction of s , yielding the maximum output value. The case C illustrates a situation where $s < R_M$.

In general, for a given P , θ is defined as

$$\theta = f_P^{-1}\left(\frac{r}{R_M}\right). \quad (17)$$

Therefore, we can rewrite the output signal associated to the p th branch as

$$s_p = \frac{R_M}{P} \exp\left(j\left(f_P^{-1}\left(\frac{r}{R_M}\right) \frac{2p - P - 1}{2}\right)\right). \quad (18)$$

Thus, taking into account (12), one can define the AM/PM and AM/AM conversion functions associated to the p th branch as

$$\Theta_p(r) = f_P^{-1}\left(\frac{r}{R_M}\right) \frac{2p - P - 1}{2}, \quad (19)$$

and

$$A_p(r) = \frac{R_M}{P}. \quad (20)$$

It is well-known that imbalances in the complex gains of LINC's amplifiers lead to performance degradation [12], [13]. In that context, one could expect the situation to be more serious with MB-LINC, since there are more amplifying branches. To study the impact of these imbalances on the MB-LINC signal, let us consider the combined signal after amplification, which is given by

$$s = \sum_{p=1}^P g_p \frac{R_M}{P} \exp \left(j \left(f_P^{-1} \left(\frac{r}{R_M} \right) \frac{2p - P - 1}{2} \right) \right), \quad (21)$$

where g_p is the complex gain associated with the p th amplifier, defined as $g_p = 1 + \varepsilon_g$, with ε_g being a complex-valued deviation whose the real and imaginary parts have variance σ_ε^2 , i.e., $\varepsilon_g \sim \mathcal{CN}(0, 2\sigma_\varepsilon^2)$. Naturally, when $\sigma_\varepsilon = 0$, the amplifiers have equal gains and (21) turns into (16). On the other hand, when $\sigma_\varepsilon > 0$, the magnitude of nonlinear distortion effects associated with the MB-LINC signal will depend on the combination of the gains of the different amplifiers.

IV. NONLINEAR DISTORTION EFFECTS

By taking advantage of the Gaussian nature of OFDM signals with a large number of subcarriers N , the PSD of their nonlinearly distorted version can be obtained through their autocorrelation, $R_y(n)$, that is itself a function of the input signals' autocorrelation, $R_s(n)$, and of the power associated to the different intermodulation products (IMPs) at the nonlinearity output. The autocorrelation of a nonlinearly distorted OFDM signal can be computed as [14]

$$R_y(n) = \sum_{\gamma=0}^{+\infty} P_{2\gamma+1} \frac{R_s(n)^{\gamma+1} (R_s^*(n))^{\gamma}}{\sigma^{2(2\gamma+1)}}, \quad (22)$$

where σ is the standard deviation of the input signal and $P_{2\gamma+1}$ is the power associated with the IMP of order $2\gamma + 1$. The PSD of the nonlinearly distorted signal, $G_y(k)$, is given by $\text{DFT}(R_y(n))$. Moreover, from the Busgang's theorem [15], we can express the nonlinearly distorted signal as the sum of two components. Under these conditions, the PSD of MB-LINC signals can be written as

$$G_y(k) = |\alpha|^2 G_s(k) + G_d(k), \quad (23)$$

where α is a scale factor given by the scaled cross-correlation between the OFDM signal and its nonlinearly distorted version, and $G_s(k)$ and $G_d(k)$ are the PSDs of the input signal and nonlinear distortion term, respectively.

Fig. 6 shows the normalized PSD of the nonlinear distortion term, $g_d(k) = G_d(k) / \max(|\alpha|^2 G_s(k))$, considering different values of P , $\sigma_\varepsilon = 0$ and $R_M/\sigma = 1.5$ (due to the symmetry associated to the different amplification branches, only $(P-1)/2+1$ PSDs are shown). From the figure, it can be noted that although the PSD of the nonlinear distortion term associated to the different branches can vary, the PSD associated to the combined signal is the same, regardless of P . To have an

Fig. 6. Normalized PSD of the nonlinear distortion term associated to the different MB-LINC branches.

insight on the performance penalty associated to the nonlinear distortion, we can define the SIR for the k th subcarrier as

$$\text{SIR}_k = \frac{|\alpha|^2 G_s(k)}{G_d(k)}. \quad (24)$$

Fig. 7 shows the evolution of SIR_k considering an MB-LINC transmitter with $s_M/\sigma = 1.0$, different values of P and $\sigma_\varepsilon = 0.05$. From the figure, it can be noted that although

Fig. 7. Evolution of SIR_k considering an MB-LINC transmitter with different values of P and $\sigma_\varepsilon = 0.05$.

there are variations on the SIR for different gain combinations, the magnitude of these variations is similar. Therefore, the use $P > 2$ does not increase the nonlinear distortion effects comparatively to the conventional LINC, where $P = 2$.

V. CONCLUSIONS

In this work we propose the MB-LINC technique, which generalizes the conventional LINC. It is shown that, although

each amplifier can work with low power signals, the spectral characteristics of the combined signal does not change relatively to the conventional LINC techniques, even when there are imbalances in the amplifiers' gain.

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