Evolution of Interplex Scheme with Variable Signal Constellation

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BIOGRAPHY

Mariano Vergara received the BSc. and the MSc. degrees in Telecommunications engineering from the University of Naples Federico II,Italy, in 2006, and from the Technische Universität Kaiserslautern (TU-KL), Germany, in 2009, respectively. He is currently pursuing a Ph.D. degree at the Universtat Autònoma de Barcelona (UAB), Spain, on signal design for global navigation satellite systems (GNSS). Since 2008, he has been an Associate Researcher with the Department of Navigation, Institute of Communications and Navigation of the German Aerospace Center (DLR).

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ABSTRACT

In this paper we present a modification of a multiplexing scheme for DS-CDMA signals, known as interplex scheme. The interplex allows to map several binary DS-CDMA signals onto a constant envelope signal, resulting in a signal that can be efficiently amplified. In order to obtain a constant envelope constellation, some additional power is transmitted that is not used for data transmission. This so called inter-modulation (IM) power can be too much compared to the useful power or the High Power Amplifier (HPA) non-linearities are not so severe to demand a perfectly constant envelope signal. The basic idea of this work is to adapt the interplex signal to the HPA at hand.

INTRODUCTION

The interplex scheme is a phase-shift-keyed/phase modulation (PSK/PM) that combines multiple signal components into a phase modulated composite signal [1]. The interplex offers a higher power efficiency than a conventional PSK/PM signal for a low number of signal components (less or equal than five [1]). In the following we assume that the signal components consist of direct sequence code division multiple-access (DS-CDMA) signals. Like the PSK/PM technique, the interplex mapping scheme is a constant envelope modulation, which means that the constellation points lie on a circle in the complex plain. This contributes to the reduction of the distortions due to the non-linearities of the high power amplifier (HPA). In order to establish a constant envelop modulation, some intermodulation (IM) product terms are introduced by the interplex mapping scheme. If the power of these terms is not used in the demodulation process, the transmit power efficiency is jeopardized [1, 2].

In former papers [1, 3] the attention was focused on the transmit power efficiency, that is to say on the share of transmitted power that is useable at receiver side. Nevertheless, what is ultimately important is the percentage of power that the receiver can use for the demodulation. Hence, we define the *receiver (Rx) power efficiency* as the ratio between the useful power at the output of the receiver's matched filters and the total transmit power.

In this paper we propose a two-step approach to adapt the signal to the characteristics of the HPA of the transmitter and thus to perform a signal mapping with high Rx power efficiency. In a first step we apply the so-called scalable interplex in order to achieve a shaping of the phase states of the signal constellation (constellation shaping) [4]. In a second step we apply the staggered interplex which introduces specific delays on the signal components of the interplex scheme, so that the sum of the Rx power at the correlator outputs at the receiver is maximized. This results to a non-linear optimization problem, which is solved

by an evolutionary algorithm [5]. This proposed two-step approach we call *scalable staggered interplex*. The signal is distorted by an HPA modeled after the well-known Saleh model [6].
We show that according to the degree of non-linearity of the HPA improvements of the Rx power efficiency of the order of 5-10% are possible. It is to be noted that this advantage does no require any hardware modification either at the transmitter or the receiver side.

SIGNAL MODEL

An N-channel interplex [1] signal is a PSK/PM signal

$$x_N(t) = \cos\left(2\pi f_c t + \Theta(t)\right), \tag{1}$$

in which the phase modulation is

$$\Theta(t) = \left[\theta_1 + \sum_{n=2}^{N} \theta_n s_n(t)\right] s_1(t), \tag{2}$$

where f_c denotes the carrier frequency and θ_n are the modulation (or interplex) angles, which are grouped into the vector

$$\boldsymbol{\theta} = [\theta_1, \ldots, \theta_N] \tag{3}$$

The signal components $s_n(t), \ \ n = 1, \ldots, N$ are DS-CDMA signals

$$s_n(t) = \sum_{m=1}^{M} s_m^{(n)}\, p_n(t - mT_n), \tag{4}$$

with the chip duration T_n, the pulse shape $p_n(t)$, and

$$\begin{aligned}
\mathbf{s}_n &= \left[s_1^{(n)}, \ldots, s_M^{(n)}]\right]^T \\
&= \mathbf{b}_n \otimes \mathbf{c}_n \ \in \{-1,1\}^{M\times 1}, && (5)\\
\mathbf{b}_n &= \left[b_1^{(n)}, \ldots, b_K^{(n)}\right]^T \in \{-1,1\}^{K\times 1}, && (6)\\
\mathbf{c}_n &= \left[c_1^{(n)}, \ldots, c_G^{(n)}\right]^T \in \{-1,1\}^{G\times 1}. && (7)
\end{aligned}$$

Here, $\mathbf{b}_n$ represents the sequence of K data symbols of the n-th signal component, $\mathbf{c}_n$ is the pseudo random binary sequence (spreading code) of the n-th signal component of length, and $\otimes$ denotes the Kronecker product.

HIGH POWER AMPLIFIER DISTORTION MODELING

The Saleh model [6] is an established model to describe the nonlinearities of a HPA. In this paper we use an extension of the Saleh model, known in the literature as modified Saleh model [7], [8, p. 113]. The AM-AM characteristic of the modified Saleh model that we use for our assessments is

$$\tilde{r}_{out} = \frac{\tilde{r}_{in}}{1 + \tilde{\beta}\tilde{r}_{in}^{\gamma}} \tag{8}$$

where $\tilde{r}_{in}$ and $\tilde{r}_{out}$ are the input and output signal envelopes respectively, expressed in $\sqrt{Watt}$. With respect to the original Saleh model[6], this extended model allows to characterize the degree of the nonlinearity for which the HPA is responsible, through the parameter γ. In the classical model $\gamma = 2$. In comparison to the formula in [7], we have ignored any scaling factor of the output as we are only interested in the distortion caused by the HPA. For the other exponents present in [7], we chose the values of the classical AM-AM Saleh model[6]. Moreover, since we are interested in a behavioral analysis, it is handy to write the input envelope as a function of the input saturation power:

$$P_{sat}^{in} = \sqrt[\frac{\gamma}{2}]{\frac{1}{(\gamma-1)\beta}} \tag{9}$$

The input envelope normalized to the saturation power is

$$r_{in} = \frac{\tilde{r}_{in}}{\sqrt{P_{sat}^{in}}} \quad , \tag{10}$$

with

$$\beta = \tilde{\beta}\,\sqrt[\gamma]{Watt} \quad . \tag{11}$$

Since we are interested only in the distortion and not in the gain brought about by the HPA, we normalize the HPA output to the square root of the power of the input. The power of the input determines the working point of the HPA, indicated by

$$P_{op} = E\left[\tilde{r}_{in}^2\right] \tag{12}$$

The AM-AM characteristic that we consider is such that the HPA does not alter the average signal power:

$$\bar{r}_{out} = \frac{r_{in}}{1 + \beta r_{in}^{\gamma}} \sqrt{\frac{E\left[\tilde{r}_{in}^2\right]}{E\left[r_{out}^2\right]}} \tag{13}$$

with

$$r_{out} = \frac{r_{in}}{1 + \beta r_{in}^{\gamma}} \tag{14}$$

where $\bar{r}_{out}$ is in $\sqrt{Watt}$ such that it always has the same power of the HPA input $\tilde{r}_{in}$. This formulation allows to highlight only the power loss caused by the distortion, independently from the HPA gain. We set the working point of the HPA, i.e. the average power of the input signal, at the input saturation power of the AM-AM characteristic:

$$P_{op} = E\left[\tilde{r}_{in}^2\right] = P_{sat}^{in} \quad \Longrightarrow \quad E\left[r_{in}^2\right] = 1 \quad . \tag{15}$$

This corresponds to an Input power Back-Off (IBO) equal to 0 dB. At this point of the AM-AM curve, both nonlinear distortions and HPA power efficiency are maximal. Furthermore, at this working point the PAPR (Peak-to-Average-Power Ratio) of the interplex constellation has the maximum impact on the power efficiency of the modulation.
For simplicity, in our study we consider an ideal AM-PM curve. The AM-PM curve describes the phase noise that

the HPA adds to the amplified signal. If the input signal has a constant envelope, the phase noise is a constant term and it creates no problem at receiver side. If the input signal has a high PAPR, the HPA output is affected by phase noise. A higher phase jitter reduces the power at the output of the receiver's correlator and thus it is also a power inefficiency. Thus, strictly speaking, it would be necessary to compute the increase of the phase jitter of the received signal, for which the HPA is responsible, and then to derive the consequent correlation loss. Nevertheless, in this study we make the approximation that at the working point at which we operate the AM-PM curve is almost constant. If this is the case, the output phase has a limited dependency on the dynamic range of the input envelope, and the phase jitter caused by the HPA can be neglected. This assumption is in agreement with the study of [9], in which the AM-PM curve is almost constant when the input power is equal to the saturation power of the AM-AM curve.

SCALABLE INTERPLEX

The idea of the scalable interplex [4] consists in adapting the interplex constellation to the HPA. Transmitting all IM power and obtaining a perfectly constant envelope signal might not be necessary and in this case the transmission would be power inefficient. On the other hand, not transmitting an IM product at all could cause non-linear distortion by the amplification of the signal through the HPA. It is thus logical that the optimal amount of IM power to transmit depends on the HPA at hand. The scalable interplex shapes the interplex constellation by scaling the IM terms of the standard interplex [1]. A scalable interplex (baseband) signal is of the kind:

$$\begin{aligned} x_N(t) = &-\sum_{n=2}^{N} g_n(\boldsymbol{\theta}) s_n(t) + \kappa_I\, v_I(t;\boldsymbol{\theta}) \\ &j\Big[g_1(\boldsymbol{\theta})\, s_1(t) + \kappa_Q v_Q(t;\boldsymbol{\theta})\Big] \end{aligned} \tag{16}$$

with $g_n(\boldsymbol{\theta})$ indicating the weighting factors and $v_I(t;\boldsymbol{\theta})$ and $v_Q(t;\boldsymbol{\theta})$ the in-phase and the quadrature intermodulation (IM) terms. The factors $\kappa_I \in [0,1]$ and $\kappa_Q \in [0,1]$ are the scaling factors of the IM terms. The values of the weighing factors $g_n(\boldsymbol{\theta})$ and of the IM terms $v_I(t;\boldsymbol{\theta})$ and $v_Q(t;\boldsymbol{\theta})$ for $N = 5$ are reported in the appendix.
Alongside the bandlimited scalable interplex (16), we define the constellation of the interplex as

$$\begin{aligned} \mathbf{x}_N = &-\sum_{n=2}^{N} g_n(\boldsymbol{\theta})\mathbf{s}_n + \kappa_I\, \mathbf{v}_I(\boldsymbol{\theta}) \\ &j\Big[g_1(\boldsymbol{\theta})\, \mathbf{s}_1 + \kappa_Q \mathbf{v}_Q(\boldsymbol{\theta})\Big] \end{aligned} \tag{17}$$

where $\mathbf{v}_I(\boldsymbol{\theta})$ and $\mathbf{v}_Q(\boldsymbol{\theta})$ are the vectors containing the products among the signal component as indicated in the appendix.

The interplex constellation is independent from the pulse shapes of the signal component. The constellation does not only describe the location of the states of the interplex signal but also the probability of each state, which is relevant for the determination of the PAPR of the constellation and as well as of the interplex signal(16). Note that even when the signal components are modulated by equiprobable symbols, the constellation states are not necessarily equiprobable. The scalable interplex concerns a modification of the signal constellation (17). The coefficients κ_I and κ_Q are varied in order to optimize the metric:

$$\eta = \frac{\sum_{n=1}^{N} z_n}{P_{Tx}} \tag{18}$$

with

$$z_n = \begin{cases} \left| \frac{1}{M}\mathrm{Im}\left\{\mathcal{T}\left[\mathbf{x}_N\right]\right\}^T \mathbf{s}_1 \right|^2, & \text{if } n = 1 \\ \left| \frac{1}{M}\mathrm{Re}\left\{\mathcal{T}\left[\mathbf{x}_N\right]\right\}^T \mathbf{s}_n \right|^2, & \text{if } n > 1 \end{cases} \tag{19}$$

where P_{Tx} is the total Tx power of transmitted signal and $\mathcal{T}[.]$ indicates the transfer function of the HPA described in (13). Hence, in order to derive the optimum constellation shaping we have to solve the problem

$$(\hat{\kappa}_I, \hat{\kappa}_Q) = \arg\max_{\kappa_I, \kappa_Q} \eta \tag{20}$$

The metric in (18) represents the Rx power efficiency without the effect of the pulse shapes of the signal components.

SCALABLE STAGGERED INTERPLEX

The staggered interplex [10] consists in introducing a relative delay among the signal components. The staggering does not affect the constellation and it impacts only the state transitions. The time offsets - which are smaller than a chip duration - can be seen as a particular form of pulse shaping, where the pulses are simply delayed. The scalable staggered interplex is described by

$$\begin{aligned} x_N^{stagg}(t) = &-\sum_{n=2}^{N} g_n(\boldsymbol{\theta}) s_n(t-\tau_n) + \hat{\kappa}_I\, v_I(t;\boldsymbol{\theta}) \\ &j\Big[g_1(\boldsymbol{\theta})\, s_1(t-\tau_1) + \hat{\kappa}_Q\, v_Q(t;\boldsymbol{\theta})\Big] \end{aligned} \tag{21}$$

For convention, the delay of the first component is taken as reference, thus $\tau_1 = 0$. The terms $\hat{\kappa}_I$ and $\hat{\kappa}_Q$ are derived in (20). The delays τ_n are to be chosen in order to maximize the metric:

$$\tilde{\eta} = \frac{\sum_{n=1}^{N} \tilde{z}_n}{P_{Tx}} \tag{22}$$

with:

$$\tilde{z}_n = \begin{cases} \left| \frac{\int_T \mathrm{Im}\{\mathcal{T}[x_N^{stagg}]\} s_n(t-\tau_n)\mathrm{d}t}{\int_T |s_n(t)|^2 \mathrm{d}t} \right|^2, & \text{if } n = 1 \\ \left| \frac{\int_T \mathrm{Re}\{\mathcal{T}[x_N^{stagg}]\} s_n(t-\tau_n)\mathrm{d}t}{\int_T |s_n(t)|^2 \mathrm{d}t} \right|^2, & \text{if } n > 1 \end{cases} \quad (23)$$

Thus, in order to derive the optimum staggering we have to solve the problem

$$(\hat{\tau}_2, \ldots, \hat{\tau}_N) = \arg \max_{\tau_2,\ldots,\tau_N} \tilde{\eta} \quad (24)$$

OPTIMIZATION

In this section we will show how both constellation shaping (scalable interplex) and staggered interplex can be applied in a two-step approach, the scalable staggered interplex, in order to optimize Rx power efficiency. By consecutively solving (20) and (24) this two-step approach yields a modified interplex signal that is adapted with respect to the characteristics of a given HPA. In the first step we optimize the metric (18) through the coefficients κ_I and κ_Q. This operation is called *constellation shaping*. In the second step we optimize the metric (22) through the time offsets $\tau_n,\ n = 2, \ldots, N$. This second operation is called *staggering* and it is a special case of pulse shaping, in which the pulse shapes are modified only by means of a time offset. The constellation shaping is performed by a line search and the optimization of the staggering is performed using a genetic algorithm as done in [10, 5]. In this paper we consider two HPAs with different degree of nonlinearity: one with $\gamma = 2$ and another with $\gamma = 5$. The HPA are always driven at saturation.

As an practical example we will consider the Galileo E1 signal which can be defined as a $N = 5$ signal interplex [2], where the signal components are defined as shown in Table (1). The interplex angles $\boldsymbol{\theta}$ are chosen in such a way that

Signal component	Service	$p_n(t)$
$s_1(t)$	PRS	BOC(15,2.5)
$s_2(t)$	OS pilot channel	BOC(1,1)
$s_3(t)$	OS data channel	BOC(1,1)
$s_4(t)$	OS pilot channel	BOC(6,1)
$s_5(t)$	OS data channel	BOC(6,1)

Table 1 Galileo E1 signal

the Public Regulated Service (PRS) contains twice as much as power as the Open Service (OS), and that the BOC(1,1) components contain 10 times the power of the BOC(6,1) components [2]. Moreover the signs of the weights of the BOC(6,1) components are different for the OS pilot channel and OS data channel [11]. This can be formulated as:

$$\begin{cases} g_1^2(\boldsymbol{\theta}) = 2\left(g_2^2(\boldsymbol{\theta}) + g_4^2(\boldsymbol{\theta})\right) = 2\left(g_3^2(\boldsymbol{\theta}) + g_5^2(\boldsymbol{\theta})\right) \\ g_4(\boldsymbol{\theta}) = -\frac{g_2(\boldsymbol{\theta})}{\sqrt{10}} \\ g_5(\boldsymbol{\theta}) = \frac{g_3(\boldsymbol{\theta})}{\sqrt{10}} \end{cases} \quad (25)$$

The one-sided bandwidth of the transmitted signal is $B_{Tx} = 70$ MHz for all signal components. The one-sided receiver bandwidth has been chosen differently according to the service: $B_{Rx,OS} = 10$ MHz for the OS and $B_{Rx,PRS} = 25$ MHz for the PRS.

Results

We report in Fig. 1 the constellation of the standard E1 interplex. The Rx power efficiency is slightly jeopardized (0.84) with $\gamma = 5$. Indeed Although the useful transmit power is 0.87, the Rx power efficiency is slightly reduced due to the effect of multiple access interference (MAI), inter-chip interference (ICI) and some marginal HPA impairments. Note that although the constellation is constant envelope, the bandlimited signal does not have a PAPR of exactly 0 dB. In Fig.2-2 the optimization results of the con-

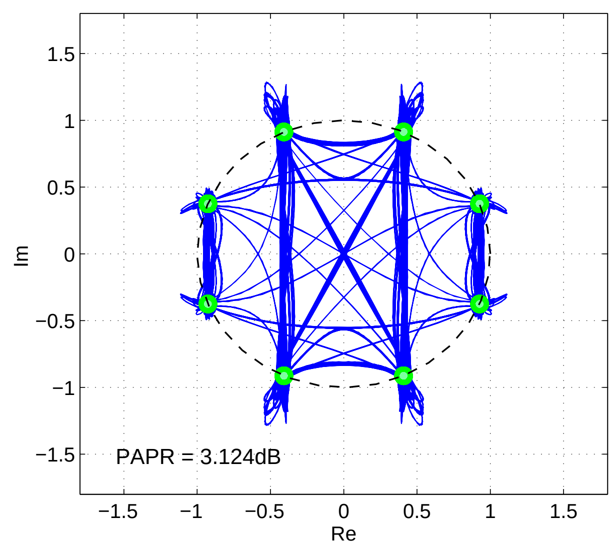

Fig. 1 Constellation diagram of the standard E1 interplex. For $\gamma = 2$ $\tilde{\eta} = 0.87$; for $\gamma = 5$ $\tilde{\eta} = 0.84$.

stellation shaping are depicted. Notice that the scaling factor κ_Q of the IM products on the Q branch has much more weight than the scaling factor κ_I on the I branch. In particular, the unscaled IM power on the Q branch is 10 times more than the IM power on the I branch. The standard interplex seems to be worth with highly non-linear HPA, but when the nonlinearities are less strong, the standard interplex is not the most power efficient solution. In Fig.4

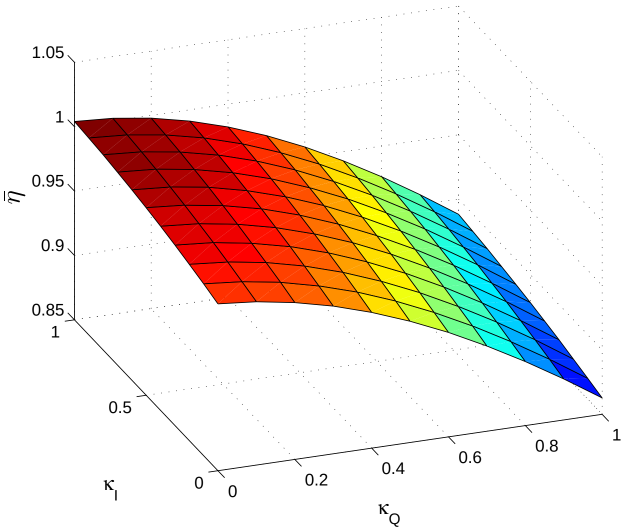

Fig. 2 Constellation shaping for $\gamma = 2$. The optimum is at $\kappa_I = 1, \quad \kappa_Q = 0$.

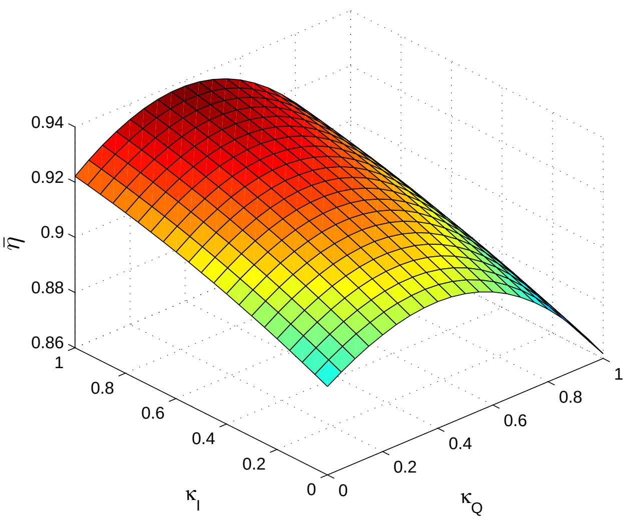

Fig. 3 Constellation shaping for $\gamma = 5$. The optimum is at $\kappa_I = 1, \quad \kappa_Q = 0.4$.

5 the results of the second optimization steps (staggering) are represented. In comparison with the standard interplex (Fig.1), the gain is more than 10% for the HPA with $\gamma = 2$ and roughly 5% with the HPA with $\gamma = 5$. When the HPA is less non-linear the standard interplex with full IM products results to be further from the optimum and thus the improvement margins are larger. To be noted that in comparison with the optimization of the first step, the staggering brings about very modest results. This can be explained as follows. The signal components are decorrelated mainly by the spreading codes, but due to the imperfect orthogonality of the codes, there is a residual cross-correlation. When the pulse shapes of the signal components are exactly the same, this cross-correlation is emphasized. The staggering decorrelates the signal components by minimizing the pulse cross-correlation. The staggering yields good results when the pulse shapes are equal for all signal components [10]. Nevertheless, in this example, the pulse shapes are not the same and the Open Service pulse shape is already spectrally separated from the PRS pulse shape and thus the optimization margins are small.

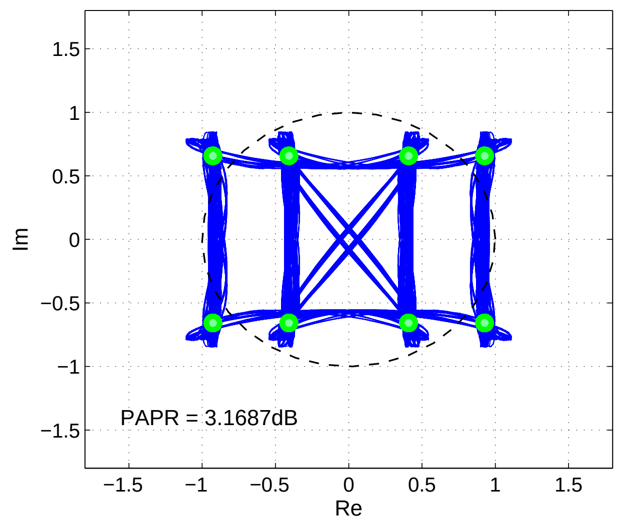

Fig. 4 Optimised (staggering) constellation diagram for $\gamma = 2$. The Rx power efficiency is $\tilde{\eta} = 1.01$

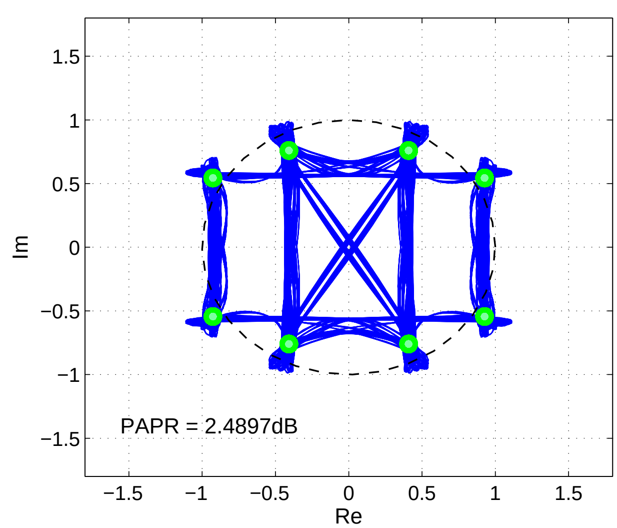

Fig. 5 Optimised (staggering) constellation diagram for $\gamma = 5$. The Rx efficiency is $\tilde{\eta} = 0.90$.

CONCLUSIONS

The interplex scheme is a method that is maximal efficient with highly non-linear HPA, but as the degree of nonlinearity of the HPA distortion diminishes, then the interplex is the most efficient solution to multiplex a stream of DS-CDMA signals. In this paper we proposed a modified interplex scheme that has the capability to adapt the interplex signal to the HPA at hand. As a result, this adaptive multiplexing technique, of which the standard interplex is a special case, allows higher Rx power efficiency. The adaptation to the HPA includes a shaping of the constellation and of the state transitions. This was also explored separately in [10] and [4]. We analyzed the example of a Galileo E1 signal. We found that the staggering does not bring significant improvements, because the pulse shapes are already decorrelated in frequency domain for the particular example of a Galileo E1. The staggering has much more impact on Rx power efficiency of the pulse shapes are all equal for all signal components [10].

APPENDIX

For the case $N = 5$ the weighting factors are

$$\begin{aligned} g_1(\boldsymbol{\theta}) &= \cos\theta_2\cos\theta_3\cos\theta_4\cos\theta_5 \\ g_2(\boldsymbol{\theta}) &= \sin\theta_2\cos\theta_3\cos\theta_4\cos\theta_5 \\ g_3(\boldsymbol{\theta}) &= \cos\theta_2\sin\theta_3\cos\theta_4\cos\theta_5 \\ g_4(\boldsymbol{\theta}) &= \cos\theta_2\cos\theta_3\sin\theta_4\cos\theta_5 \\ g_5(\boldsymbol{\theta}) &= \cos\theta_2\cos\theta_3\cos\theta_4\sin\theta_5\ , \end{aligned} \tag{26}$$

and the IM terms are

$$\begin{aligned} v_I\left(t;\boldsymbol{\theta}\right) &= s_2s_4s_5(t)\sin\theta_2\cos\theta_3\sin\theta_4\sin\theta_5 \\ &+ s_2s_3s_4(t)\sin\theta_2\sin\theta_3\sin\theta_4\cos\theta_5 \\ &+ s_2s_3s_5(t)\sin\theta_2\sin\theta_3\cos\theta_4\sin\theta_5 \\ &+ s_3s_4s_5(t)\cos\theta_2\sin\theta_3\sin\theta_4\sin\theta_5 \\ \\ v_Q\left(t;\boldsymbol{\theta}\right) &= s_1s_4s_5(t)\cos\theta_2\cos\theta_3\sin\theta_4\sin\theta_5 \\ &+ s_1s_3s_4(t)\cos\theta_2\sin\theta_3\sin\theta_4\cos\theta_5 \\ &+ s_1s_3s_5(t)\cos\theta_2\sin\theta_3\cos\theta_4\sin\theta_5 \\ &+ s_1s_2s_3(t)\sin\theta_2\sin\theta_3\cos\theta_4\cos\theta_5 \\ &- s_1s_2s_3s_4s_5(t)\sin\theta_2\sin\theta_3\sin\theta_4\sin\theta_5 \\ &+ s_1s_2s_4(t)\sin\theta_2\cos\theta_3\sin\theta_4\cos\theta_5 \\ &+ s_1s_2s_5(t)\sin\theta_2\cos\theta_3\cos\theta_4\sin\theta_5 \end{aligned} \tag{27}$$

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