Gain and Delay Mismatches Cancellation in LINC and Polar Transmitters

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Abstract— This article presents the application of a gradient algorithm on impairments correction for polar and LINC transmitters (LInear amplification with Non linear Component). The two aspects of the approach are presented: identification and correction. The large improvements obtained using such solutions is demonstrated for both transmitters.

I. INTRODUCTION

Current research in transceivers architecture is focused on the reduction of the overall power consumption of the system. For this reason, architectures such as polar [1] and LINC transmitters [2] are appealing alternatives to the traditional cartesian transmitter. The differences between these solutions can be simply viewed as different methods of writing the modulated signal. Indeed, the polar architecture is based on the polar representation of the signal, with its magnitude $\rho(t)$ and its phase $\phi(t)$, whereas the LINC solution consists in rewriting the signal as the sum of two constant envelope signals. These different ways of expressing the signal influence the architecture of the transmitter. In the case of polar solutions, the envelope signal and the phase signal are differently processed before being recombined at the output stage of power amplifier. For the LINC architecture, the two constant envelope signals are processed and amplified separately before being added.

The main drawback of these solutions is their sensitivity to mismatches between the two paths. In both cases, the output signal is significantly deteriorated, in terms of EVM (Error Vector Magnitude) and ACLR (Adjacent Channel Leakage Ratio) when a delay mismatch between paths and/or complex gain mismatch occur.

In this paper, we propose an adaptive algorithm for identifying and correcting delay and gain mismatches for these two transmitters.

We illustrate the issues and performances of proposed algorithms on a 16QAM modulation with a chip rate of 3.84 Mcps. This corresponds to the 3GPP standard. We mainly focus the analysis on the EVM (Error Vector Magnitude) and on the ACLR (Adjacent Leakage Channel Ratio) at ± 5 MHz from the assigned channel frequency. The specifications are an EVM lower than 17.5% and an ACLR higher than 33 dB [3].

In a first section, we introduce the general principle of identification and direct correction by an adaptive algorithm. These approaches are then applied to the polar transmitter in section III and to the LINC transmitter in section IV

II. ADAPTIVE ALGORITHM

A. Introduction to gradient algorithm

Let us consider a system, with input $x(t)$ and output $y(t)$. The output of the system is a function f of the input, characterized by k parameters $\sigma_1, \ldots, \sigma_k$. The definition of function f is such that $f(x(t), 0, \ldots, 0) = s(t)$. For instance, in case of a pure delay, we have $f(x, \tau_1) = x(t - \tau_1)$, and in case of a gain, we could have $f(x, g_1) = G_1 x(t) = e^{g_1} x(t)$. We will also denote by $d(t)$ a desired signal. The objective is to identify these parameters so that the output of the system becomes similar to the desired one. This is realized through the identification of parameters called σ_i such as we minimize the following mean square error:

$$J(\sigma_1, \ldots, \sigma_k) = \mathrm{E}\left[|f(x(t), \sigma_1, \ldots, \sigma_k) - d(t)|^2\right], \quad (1)$$

where $\mathrm{E}\left[\bullet\right]$ is the statistical expectation operator. The solution can be obtained with a descent algorithm. This algorithm consists in iterating for $i = 1 \ldots n$:

$$\sigma_i(n+1) = \sigma_i(n) - \gamma_i(n) \left.\frac{\partial J(\sigma_1, \ldots, \sigma_k)}{\partial \sigma_i}\right|_{\sigma_i = \sigma_i(n)} \quad (2)$$

where γ_i is the adaptation step. The gradient is computed according to

$$\frac{\partial J(\sigma_1, \ldots, \sigma_k)}{\partial \sigma_i} = \frac{\partial \mathrm{E}\left[|e(t)|^2\right]}{\partial \sigma_i} = 2\mathrm{E}\left[\Re\left(\frac{\partial e^*(t)}{\partial \sigma_i} e(t)\right)\right] \quad (3)$$

The theoretical algorithm is obtained with the derivatives in (3). In practice, we use a stochastic gradient algorithm that involves the instantaneous expression of the gradient rather than the statistical average. The formulation is finally

$$\sigma_i(n+1) = \sigma_i(n) - 2\gamma_i(n) \left.\Re\left(\frac{\partial e^*(n)}{\partial \sigma_i} e(n)\right)\right|_{\sigma_i = \sigma_i(n)}. \quad (4)$$

B. Identification procedure

The resolution of the minimization problem and the correction of the imperfections can be formulated in two ways. The first solution is to identify these defaults and then correct them. The identification process, Fig. 1, consists in modifying the ideal signal $x(t)$ with parameters $\mu_1, \ldots, \mu_k$ and then comparing it with the emitted one $z(t) = f(x(t), \tau_1, \ldots, \tau_k)$.
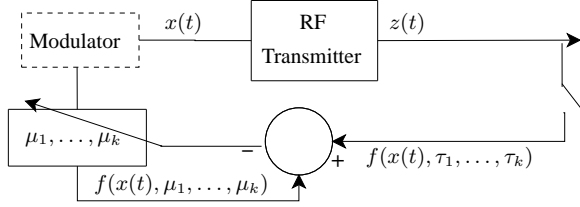

Fig. 1.   Identification implementation

The desired signal is here $z(t)$ and the parameters $\sigma_i = \mu_i$. The equation (2) becomes, for $i = 1 \ldots n$:

$$\mu_i(n+1) = \mu_i(n) - \gamma_i(n) \left. \frac{\partial J(\mu_1, \ldots, \mu_k)}{\partial \mu_i} \right|_{\mu_i = \mu_i(n)} \quad (5)$$

Once the τ_i values are identified, the correction can be applied, if necessary. Hence, the identification implementation allows both correction an monitoring.

C. Correction procedure

The other approach consists in directly correcting the signal before the emission, according to Fig. 2. The desired signal is here $x(t)$. This gives according to (2) the following iterative formulation:

$$\Delta_i(n+1) = \Delta_i(n) - \gamma_i(n) \left. \frac{\partial J(\Delta_1, \ldots, \Delta_k)}{\partial \Delta_i} \right|_{\Delta_i = \Delta_i(n)} . \quad (6)$$

This correction procedure avoids the intermediate step of the identification.

One has to pay attention to the fact that this algorithm needs, for the calculation of the instantaneous derivative of the gradient, to know perfectly the output signal. It implies that the correction algorithm cannot be applied when the transformation applied to the signal is non linear.
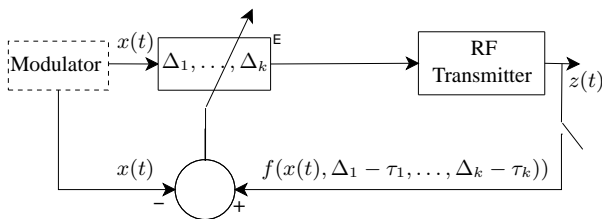

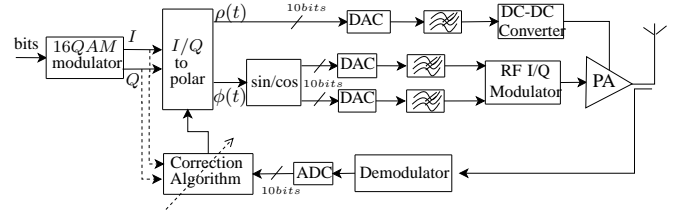Fig. 2.   Correction implementation

Fig. 3.   Simulation model of the transmitter

D. Application to desynchronisation

In the specific case of desynchronisation, the implementation of such algorithms requires a block that controls the advance and the delay of the signal. For this, the transmitter has to introduce a buffer between the emitted signal and the output of the modulator. Moreover, the solution for delaying or advancing signals by fractional amounts relies on the use of interpolation filters which allow to adjust the delay without recomputing the coefficients of the filter [4]. We choose to implement a fifth order Lagrange interpolator using the Farrow structure, as we did in [5].

III. APPLICATION TO THE POLAR TRANSMITTER

A. Polar architecture and simulation model

Polar architecture is a regarded as an efficient solution for SDR transmitter [6]. This architecture, based on the Kahn's Envelope Elimination and Restoration (EER) procedure, consists in amplifying the phase modulated signal with an efficient amplifier while the envelope is restored through the voltage supply of this PA (power amplifier). The simulation model used in this paper is shown Fig. 3, with a sampling period $T = T_s/12$, with T_s being the duration of an emitted symbol. With this model, the output EVM is on the order of 0.3% and the ACLR is 53 dB.

This architecture is sensitive to delay mismatches between the amplitude path and the phase path [7]. For example a delay mismatch of $0.1T_s$ between the two paths yields to an EVM of 5.5% and to an ACLR of 28 dB.

B. Delay cancellation

Let us consider the complex representation of the ideal signal: $x(t) = \rho(t) \exp(j\phi(t))$. This baseband signal is processed by the RF transmitter and results in the signal at the antenna denoted $z(t)$, Fig. 4.
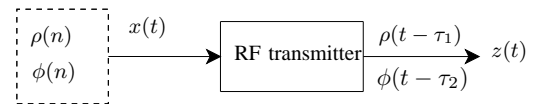

Fig. 4.   EER transmitter impairments

The envelope of the signal is delayed by τ_1 whereas the phase of the signal is delayed by τ_2. This gives for the emitted signal

$$z(t) = \rho(t - \tau_1) \exp(\phi(t - \tau_2)).$$

For this application, we can apply the correction algorithm without the preliminary identification step. This solution reduces the computation time. We also compute the mean square error only with the in-phase component of the complex signal (using both in phase and quadrature components implies higher computation load with little gain in performance). Following (1) and (6), we find that the mean square error is

$$J(\Delta_1, \Delta_2) = \mathrm{E}\big[|\rho(t)\cos(\phi(t)) - \rho(t_1)\cos(\phi(t_2))|^2\big], \quad (7)$$

and the updating rules for Δ_1 and Δ_2 are

$$\Delta_1(n+1) = \Delta_1(n) + \gamma_1(n)\frac{d\rho(u)}{du}\bigg|_{t_1}\cos\phi(t_2)\, e(n)$$
$$\Delta_2(n+1) = \Delta_2(n) + \gamma_2(n)\rho(t_1)\frac{d\cos\phi(u)}{du}\bigg|_{t_2} e(n)$$

where we noted $t = nT$, $t_1 = nT + \Delta_1 - \tau_1$ and $t_2 = nT + \Delta_2 - \tau_2$, T being the sampling period.

This algorithm is implemented and the simulation results are presented Table I.

τ_1	τ_2	tc Opt	EVM (%) Initial	After	ACLR @5 MHz Initial	After
$0.2T_s$	$1.1T_s$	$150T_s$	38%	0.3%	13 dB	51 dB
$0.45T_s$	$0.95T_s$	$130T_s$	25%	0.3%	13 dB	51 dB
$0.5T_s$	$0.12T_s$	$110T_s$	20%	0.3%	15 dB	51 dB
$0.1T_s$	$0.2T_s$	$80T_s$	5.5%	0.3%	28.5 dB	51 dB

These results show a huge improvement in the EVM which falls from 38% to 0.3%, and in the ACLR which rises from 13 dB to 51dB. The convergence time at 90% of the final value, t_c, is also evaluated and for our test cases is under $150T_s$. These results are detailed in [5].

C. Delay cancellation with gain identification

However, the results indicated above were obtained without considering the gain and phase offset introduced by the power amplifier. Indeed, from the transmitter point of view, the phase of the emitted signal is none of concern and the magnitude is controlled within a specified range (few dB). Since our algorithm relies on the comparison between the emitted signal and the ideal one, an uncertainty about the phase and the amplitude of the emitted signal impacts directly the performances. This is reported on Table II, on the (B) columns, for different scenarios for the gain and phase mismatches with $\tau_1 = 0.95\ T_s$ and $\tau_2 = 0.45\ T_s$. To enhance performances we add an identification loop dedicated to this complex gain according to the principle Fig. 1. The corresponding mean square error is then

$$J(K, \xi) = \mathrm{E}\big[|K\rho(t)cos(\phi(t) + \xi) - G\rho(t_1)\cos(\phi(t_2) + \Theta)|^2\big],$$

with $G\exp(j\theta)$ the complex gain representative of the incertitude in the return path and $K\exp(j\xi)$ the unknown gain. The implementation of the identification algorithm was elaborated

Gain mismatch	phase mismatch	tc A	EVM (%) A	B	ACLR @5 MHz B	A
3 dB	45°	$300\ T_s$	11%	1.5%	27 dB	49 dB
1 dB	30°	$250\ T_s$	9%	0.7%	30 dB	50 dB
0.5 dB	10°	$230\ T_s$	3.4%	0.6%	38 dB	50 dB
0.2 dB	2°	$100\ Ts$	1.2%	0.6%	47 dB	50.5 dB

from the cartesian expression of the gain $K\exp(j\xi) = K_r + jK_i$ and leads to the following equations:

$$\begin{cases} K_r(n+1) = K_r(n) + \gamma_3(n)\Re(X^*(t)e(n)) \\ K_i(n+1) = K_i(n) + \gamma_3(n)j\Re(X^*(t)e(n)) \end{cases}$$

The associated results are given in Table II, under the columns (A). As shown in Table II, we obtain a large improvement. The EVM falls to about 1.5% for the worst case compared to 11% without gain identification and the ACLR becomes higher than 49 dB. The only concern is a slower convergence which is the price to pay to account for these additionnal mismatches.

IV. APPLICATION TO THE LINC TRANSMITTER

A. LINC architecture and simulation model

The LINC solution is based on the separation of any modulated signal into two constant envelope (phase modulated) signals, Fig. 5. These two signals are processed along two parallel paths and are recombined after efficient power amplification. However, this architecture is sensitive to mismatches between the two transmit paths: gain impairment and differences in propagation delays. The gain imbalance is a well known drawback of this architecture and some solutions for solving this problem are presented in [8], [9]. In addition to this gain mismatch, the delay mismatch between the two paths has to be taken into account. It does not only degrade the output performances of the transmitter but also corrupts the gain correction by adding a time dependent perturbation on the phase. The simulation model implemented is similar to the one used for the polar model, Fig.3. It is constituted of the main path with two transmitters in parallel and a return path for correction. It takes into account quantization effects, with 8-bit DACs and 8-bit ADCs at a sample rate of $8F_s$ as detailed in [10]. With the simulation model, the EVM value is 0.3% and ACLR is 43 dB.

B. Delay and Gain identification and correction

The principle of the LINC transmitter consists in rewriting any modulated signal as the sum of two constant envelope modulated signals, denoted by $s_1(t)$ and $s_2(t)$:

$$s(t) = s_1(t) + s_2(t).$$

These two phase modulated signals, $s_1(t)$ and $s_2(t)$, are generated according to [2]:

$$s_1(t) = \exp j\left(\phi(t) - \theta(t)\right), \qquad s_2(t) = \exp j\left(\phi(t) + \theta(t)\right).$$
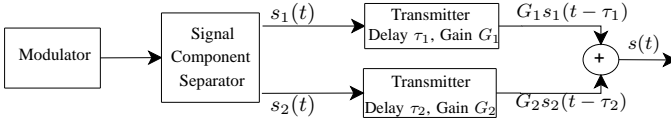
Fig. 5. LINC Transmitter impairments

Let us consider that $s_1(t)$ is delayed by τ_1 whereas $s_2(t)$ is delayed by τ_2, with $\tau_2 - \tau_1 = \Delta$. At the output of the transmitter, considering these delays, the original modulated signal $s(t)$ is modified into:

$$z(t) = s_1(t - \tau_1) + s_2(t - \tau_2). \tag{8}$$

This delay mismatch impacts both EVM and ACLR. For a differential delay of $\Delta = T_s/32$, the EVM reaches 3.2% and the ACLR is 32 dB, for $\Delta = T_s/10$ the EVM is near the limit with 10.7% and the ACLR is out of specification with 22 dB. The synchronisation has to be recovered.

Applying the correction algorithm to this architecture implies that we can extract the values of $s_1(t)$ and $s_2(t)$ from $z(t)$ with an additional SCS. But, as the operation is non linear, we cannot recover the expected signals. For this reason, only the identification approach is possible.

The equations for the identification of τ_1 and τ_2, according to Fig. 1 are, for $i = 1, 2$:

$$\mu_i(k+1) = \mu_i(k) + 2\gamma(k)\Re\left(e(n)\frac{\partial s_i^*(t - \mu_i)}{\partial \mu_i}\bigg|_{\mu_i = \mu_i(k)}\right),$$

with

$$e(n) = z(nT) - s_1(nT - \mu_1) - s_2(nT - \mu_2).$$

This algorithm allows the reduction of the EVM below 0.6% and the increase of the ACLR above 43 dB for differential delays up to $T_s/8$ after identification and then correction of the signals.

In addition to the delay correction, the gain impairments between the two paths have to be cancelled. The output signal for gain mismatches is:

$$z(t) = G_1 s_1(t) + G_2 s_2(t),$$

with G_1 and G_2 two complex gains. Using the identification procedure the error function is:

$$e(t) = K_1 s_1(t) + K_2 s_2(t) - z(t)$$

and the recursive equations are, for $i = 1, 2$:

$$\begin{cases} \Re(K_i(k+1)) &= \Re(K_i(k)) + 2\gamma_k(k)\Re(e(n)s_i^*(t)), \\ \Im(K_i(k+1)) &= \Im(K_i(k)) + 2\gamma_k(k)\Re(j.e(n)s_i^*(t)). \end{cases}$$

The gain correction cannot be correctly achieved without a preliminary synchronization between the reference and the emitted signal, even if there is no differential delays. The delays of the analog signal processing have to be identified otherwise the algorithm diverges. For this reason, the procedure adopted consists in: identifying the delay mismatch, correcting it, identifying the gain mismatch and finally correcting this gain mismatch.

Further simulations were realized including both delay mismatch and gain impairment. We introduced a differential delay mismatch of $T_s/32$ between the two paths. This corresponds to a mismatch of 10% of the propagation time between the two baseband analog filters. A comparison of the results was realized between the performances before the correction, after only a simple non differential delay identification and finally with the differential delay adjustment procedure. Results given in Table III demonstrate the efficiency of a differential delay correction as it allows a increase of about 15 dB on ACLR and lowers the EVM to less than 1%. It should be pointed out that for greater mimatches (above 1 dB and 15°) the algorithm exhibits, of course, poorer results in terms of ACLR and EVM.

TABLE III

ALGORITHM PERFORMANCES WITH AND WITHOUT DIFFERENTIAL DELAY ADJUSTMENT, WITH $\Delta = \tau_1 - \tau_2 = T_s/32$

		Before		W/o Delay Adj		W Delay Adj	
ΔG	$\Delta \phi$	EVM	ACLR	EVM	ACLR	EVM	ACLR
$0.1dB$	2°	4.2%	27dB	3.7%	28dB	0.7%	43dB
$0.2dB$	5°	6.9%	23dB	3.7%	28dB	0.8%	41dB
$0.5dB$	5°	6.9%	23dB	3.8%	28dB	0.7%	41dB
$0.6dB$	8°	10.8%	20dB	3.8%	28dB	0.8%	40dB
$0.8dB$	10°	14.1%	18dB	3.8%	28dB	0.9%	39dB

V. CONCLUSION

This paper addresses the correction of gain and delays mismatches in polar transmitter and LINC transmitter. A gradient algorithm is used in both cases, either through an identification procedure or a direct correction principle. The efficiency of the approaches is demonstrated for the two transmitters. The achieved performance exhibits dramatic improvements, with an EVM lower than 1% and an ACLR at 5MHz higher than 39 dB.

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