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A New Approach On Delay Coding: The Receiver

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Abstract - This paper represents the second part of an issue from the„Communications 2004". Now, herewith we are presenting the receiving part following the idea to see the Delay Coded Signal DCS as a QAM signal, namely a Vestigial SideBand VSB one. One uses the fact that the DCS is a four phase signal, rotating $0^0/90^0$ clockwise. The demodulated signal is ternary too and alternating on the two branches, elsewhere identical. The carriers have half the signaling rate frequency. One expose three demodulation methods and two probability density functions together with the simulation results.

Key words: Miller code, floating threshold, integral criterion.

INTRODUCTION

1. *General considerations*

The electrical representation of the data must have few qualities: restricted bandwidth, lack of energy at very low frequencies, good noise protection. For example, the Alternated Mark Inversion (AMI) code is ternary and so noise sensitive. but allows an easy synchronization. The differential biphasic (DBPh) signal is binary and hence robust to noise. But its bandwidth is double than necessary. A special role is played by the Miller coded signal (MCS). also called Delay Coded Signal (DCS, DC). Its bandwidth is comparable with that of a baseband NRZ signal, but does it does not possess very low frequency components. Some drawbacks result from its intrinsic phase modulation. Unfortunately, a catastrophic carrier and bit timing ambiguity appears. But the Miller coding can however be improved. We provide in this paper some methods to perfect its efficiency.

2. *The contents*

The writing has three sections.

In Section I one remember the obtaining of a VSB-QAM signal using three level quadrature carriers and two data flows: the first, *divided data* (DD) is obtained from the (DCD) by dividing it in a T type flip-flop; the second is the modulo-2 sum of this two flows. The two operations result in a flow of Gray-coded dibits. They possess a structure of Hilbert pairs.

In Section II we present three demodulation methods. The first method uses a QAM procedure, but the decision rule is with floating threshold.

The second method exploits Gray coded data and obtain two decision branches, two unknown data flows and the final DCD. The third method use an integrating criterion. Both are using an original decision proceeding device. Comparisons are presented with the classic Miller detection and the binary baseband model. The AWG noises are inserted both in the front and the end of the channel.

In Section III we find the probability density function PDF $d(h)$ and $d(f)$ for the difference h and sum f of the absolute values of the inphase and quadature signals. The angle between them is $\psi=\mathrm{acos}(2/\pi)$ for the difference h and $(\pi-\psi)$ for the sum f.

The timing problem is mooted too. We proved that it is not necessary a bit timing of $2f_s$ frequency, the only possible till now, but an fs/2 one. It eliminates the T/2 ambiguity regarding the sampling points.

Section I Previous results

1.1 *Main Baseband data signals*

In Fig. 1 are illustrated baseband (b.b.) data. The DCS can be constructed by simply dividing a DBPh code in a T-flip flop. We will denote v_s as the signaling rate. The periodic signal P, with $v_s/2$ frequency, is one of two quadrature carriers. We can see the Miller code as a 4-PSK signal.

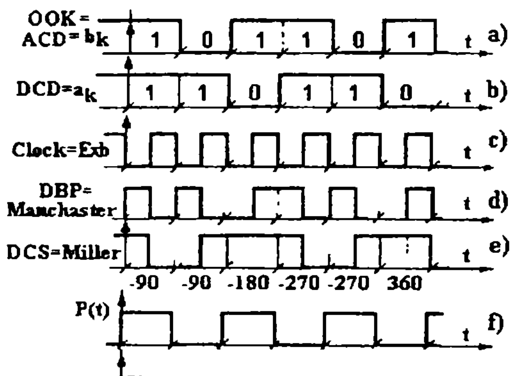

Fig.1. Data representation: a) On-Off Keying (OOK); b) DCD; c) Bit timing, is the carrier for the DBP code. d) Differential biphase code, with binary 0⁰ 180⁰ jumps; e) DC Signal: a transition at half-interval represents a "1" in DCD data a_k. f) P is the "In phase" carrier, with 0^0 180^0 phases. g) AMI (bipolar) code with "0 Volt" for logic zero and alternating "1 V" for logic 1;

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The OOK data will be denoted b_k and called absolute coded data (ACD). DCS is a quaternary DPSK signal. Data b_k must be differentially coded, leading to the data sequence a_k (DCD). The ACD produce for P a phase change of -90 when b_k=1 and no change when b_k=0. We will have $0^0/180^0$ when a_k=0 and $-90^0/-270^0$ when a_k=1.

1.2. *Plain Old Demodulation Method of DC data*

Summing up the values from 2 sub-intervals, we obtain Mimod(ified) data. If the phase φ_k on that interval is 0/180 (Fig.3), the sum is 0 (Fig.2) and will be sampled on the second sub-interval, resulting a_k=0.

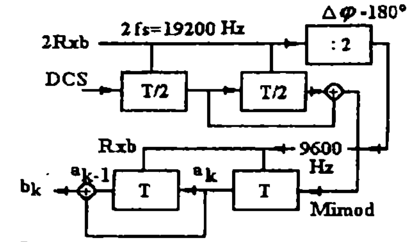

fig.6a

Fig2 - Standard Miller detector. The detector works in two steps. It decides Mimod: a. the DCD a_k and b. the bit differences ACD b_k.

If the phase is e.g. 90/270, it changes during T and the sum will be a_k=1 (Fig.4, Mimod signal). Then one obtain $b_k = a_k \oplus a_{k-1}$. The drawback is that it needs a clock with double frequency, $2v_s$. By division it is possible equally to obtain an inversed clock, $\Delta\varphi=180^0$, sampling at the points denoted by xx in Fig.4. The so decided data are erroneous and the recovery is difficult, after few hundred of erroneus bits.. This is the phase ambiguity we mentioned previously

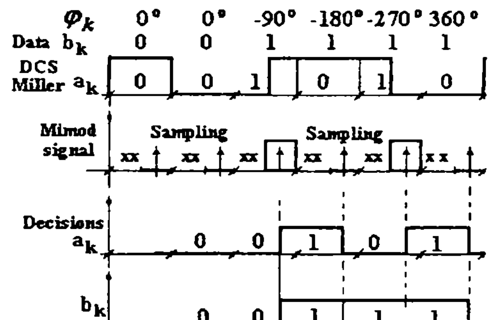

fig.6b

Fig.3 Signal diagram. DCS is the received signal. Mimod is to be evaluated in the second sub-interval; x are no-matter data

1.3 *DC modulation is a VSB -QAM with ternary carrier (VSB-QAM-TC)*

DCS can be built by modulating two quadrature carriers by two data bit streams at the v_s rate, which are then summed up. Summing displaced binary carriers result in a ternary one. So it is necessary these carriers to be ternary to resulting in a binary signal. In Fig.4 the constellation 4QAM is rotated by 45^0 w.r.t. P and Q. In Fig.5 we have the ternary carriers Π and K; their sum and difference are P and Q signals.

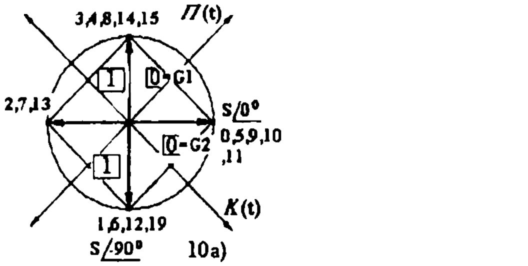

Fig.4 - Signal constellation with rotated carriers. The figures 0,1,3, ... denote the intervals kT

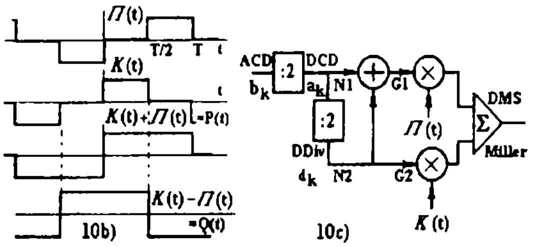

Fig.5 - QAM ternary carrier carriers with phases -180^0 and -270^0 are obtained by inverting P and Q. Fig 10c. The resulting schematic, N1, N2 are BCD numbers. G1, G2 are Gray coded bits

The synthesis of the bit streams G2.G1 is obtained based on the constellation from fig.5 and the modulation steps kT. The phases $l \cdot 90^0$ with l=0,1,2,3 written in BCD provide the N2N1 bits. By writing G2G1 as being the coordinates of the projections of the signal phasors on the axes, one can obtain Gray coded data (GC), as illustrated by the QAM scheme in Fig.6, right (see10c). Projections of the phasors φ_k on the axes Π and K take the values on the constellation; then G2 G1 also takes the necessary values. Fig.6 illustrates the PSD of the VSB signal.

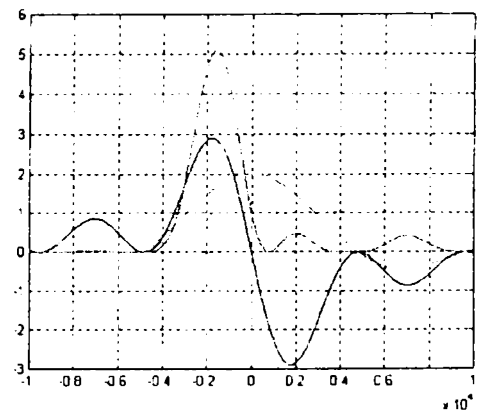

Fig 6- The VSB spectrum; spectra of the Gray quadrature data flows are visible as odd and even functions

Section II New demodulation methods

II.1 *Coherent quadrature detection – floating threshold decision*

a. Receiver model

The channel is simulating a cooper pair in a city network. The noise is placed at the channel entry or output. The carriers P, Q have $f_s/2$ frequency and are rectangular. The two low pass post detection filters have a cut-off frequency of $f_s/2$ too, at the Nyquist limit. The idea of the method is becomes clear by observing the three level demodulated signals, amazingly the same, in fact complementary. If one of them is "big" (1 or −1) the other is 0. The succesion is 1,0.-1.0,1,-1..

Table I

a_k differ.data	0 0 1 0 1 1 0 0 0 1 0 1
φ_k sent phases	0 0 −90 180 90 90 0 0 0 −90 180 270
p=inphase sign.	1 1 0 -1 0 0 1 1 1 0 -1 0
q=quadr. signal	0 0 1 0 -1 -1 0 0 0 1 0 -1
$\lvert u\rvert - \lvert v\rvert = -a_{k-}{}^{dc}$	1 1 -1 1 -1 -1 1 1 1 -1 1 -1
$\lvert u\rvert - \lvert v\rvert^{sc} = -a_k$	1 1 0 1 0 0 1 1 1 0 1 0

In the example from Table 1 a_k are arbitrary. The modules difference is finally DCD a_k. Let +/-A be the amplitude of p and q. The decision may be effected by simple rectifying one of them and comparing it with a threshold of A/2 value. The noise can be at most A/2.

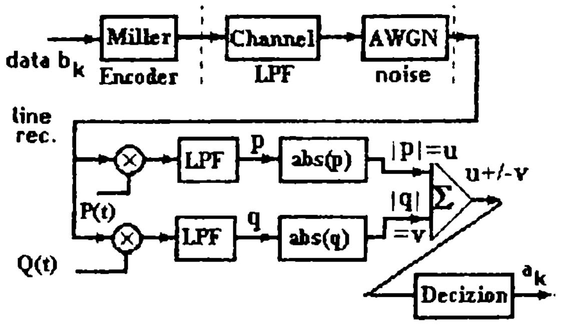

Fig 7 - The transmission system: Miller encoder sends DCS to a low pass filter as a channel with additive white Gaussian noise. Two multipliers translate the line band in the base band, aided by two l.p.f. abs(p,q) represent modules, Σ is an adder/substractor. The name of this scheme is Misnd. The signals are thresholds each other and are noisy. Σ is used as an adder only for computations purposes.

In our solution (fig.7) a threshold circuit decide between a voltage A and an electric 0 on the threshold entry, usually set at 0 voltages. But herein this entry is "hot" because always there exists the noise. In fact it happens on both entries. The sum will be of double power (3dB) wilethe permitted noise can be at most A. These get the advantage of 6 dB for the signal. The noise increases with only 3 dB. It can be expected a large advantage, of 3 dB.

Our simulations implied a "plain old" Miller receiver, denoted Milvec(no) for old or new structure concern ing the placement of the noise source in the transmi ssion chain. In fig.8 it is put at the receiver entry, at the end of the channel. The eye diagrams are plotted on the two branches in fig.8. The three levels are visible and the sampling points are too. The transitions in the above part figure are producing only between the neighbouring levels. The intersymbol interference ISI is not present. In the figure below there exist transitions between extreme levels. But the ISI is not them imputable. At a careful insight one observe on the two figures two adjacent and not identical eyes in every bit interval. They are displaced each other and the second is sampled in a wrong moment. So the two branches are not equally noise protected. This method will be referred as Msnd. The error rate is presented in the tables below together with standard Miller, denoted Mold and a base band, NRZ code, Mnyq. In the first hypo thesis the noise is placed at the channel issue, in the second is at the entry. All upper mentioned names receive a "2" at the end (e.g. Misnd2).

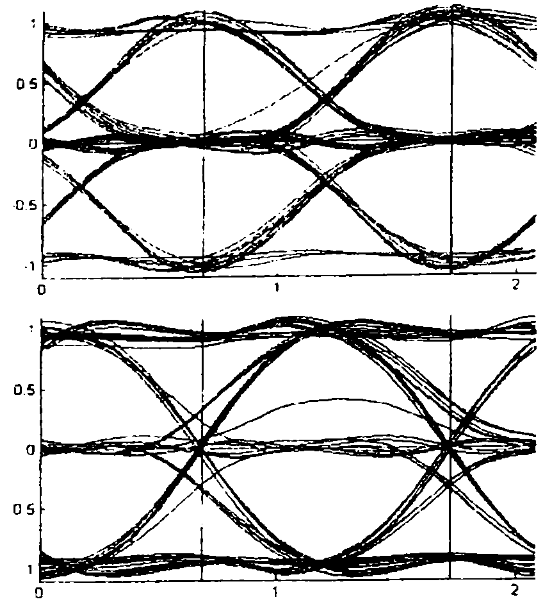

Fig. 8 Eye diagrams for Miller code demodulation on the branches; the ternary signal at the upper image is without ISI while this below has a big amount of ISI.

Table 2 - BER for noise at the issue

Snr/ber	4 dB	6 dB	8 dB	9 dB
Msnd	5e-3	.9e-4	1.6e-4	8.6e-6
Mold	2.8e-1	1.3e-1	3.5e-2	1.3e-2
Mnyq	1.8e-1	8.6e-2	2.7e-2	1.1e-2
snr/ber	10 dB	11 dB	12 dB	13 dB
Mold	5.2e-3	1.4e-3	2.6e-4	1.2 e-5
Mnyq	4.9 e-3	1.5 e-3	4.2 e -4	7.5e-5

Mnyq has 12dB (11.4) for BER=10^{-4}. Figure 4.2 is the effect of the descrambler. Miller old is equivalent with a 4 PSK. Remember the penalty of double BER for DPSK. So, our method is 5 dB superior to the classic Miller decoder. See for all that the detection has included two LPF's.

Table 3 - BER for noise at the entry

snr/ber	4 dB	6 dB	8 dB	9 dB
Msnd	2.2ᵉ-3	5.8ᵉ-5	6.4ᵉ-6	-
Mold	2.5e-2	3.4ᵉ-3	3.3ᵉ-4	2.3ᵉ-5
Mnyq	4ᵉ-3	6.6ᵉ-4	8.1e-6	-

When the noise is inserted at the entry (table3) it undergoes a loss of ≈9.5DB. The hierarchy remains but the differences are smaller. However, an effective consistent advantage of 2.5dB is real. 1dB is due to the post detection filters. A 1.5 value results from the peculiar probability density function of the module difference (see later).

II.2 *Coherent detection – double flow decision*

Remember that the DCS can be view as a VSB-QAM signal, with two Gray coded data flows. In table4 one observes the detected u, v signals and their sum and difference. Dc is double current (polar) and sc is simple current (biased). The decisions d_{k+} and d_{k-} of u+/-v are the data flows obtained from DCD with a T flip-flop on the positive and negative edge respectively! Their sum is desired a_k data. With the price of two decision devices and a sophisticated timing procedure one obtain a significant gain in SNR and BER

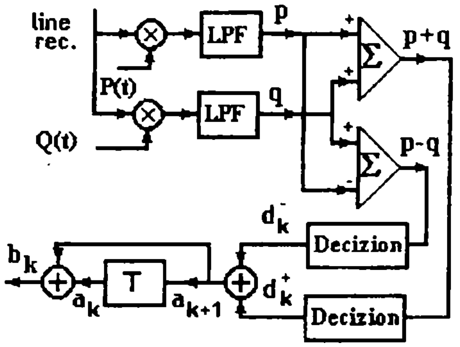

Figure9 Miller detector with double flow data decision; the sum and difference between branches are two level signals; the thresholds are set to 0 and are noiseless.

Table.4 – Relationship between data, phases and divided data. DCD a_k result as a modulo 2 sum of two data flows.

b_k absolute data	0 0 1 1 1 0 1 0 0 1 1 1
a_k DCD	0 0 1 0 1 1 0 0 0 1 0 1
φ_k sent phases	0 0 –90 180 90 90 0 0 0 –90 180 270
u inphase sign.	1 1 0 -1 0 0 1 1 1 0 -1 0
v quadr. signal	0 0 1 0 -1 -1 0 0 0 1 0 -1
$(u-v)^{dc}$	- - - - - - - - - - - -
$(u-v)^{sc}=d_{k+}$	1 1 0 0 1 1 1 1 1 0 0 1
$(u+v)^{sc}=d_{k-}$	0 0 0 1 1 1 0 0 0 0 1 1
$d_{k+}+d_{k-}=a_k$	0 0 1 0 1 1 0 0 0 1 0 1

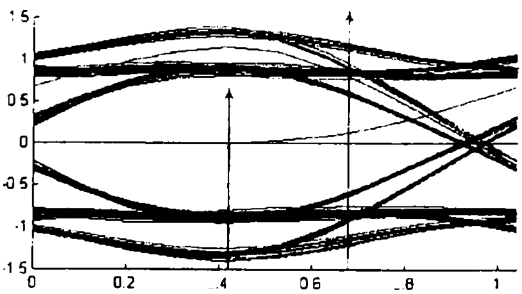

Figure 10 a - The in phase signal u+v

We have 3dB gain. The reason is the binary eye diagram and the possibility to sample both in their maximum opening point. By summing and subtracting the signal is doubled but the noise is only increased by a sqrt2 factor.

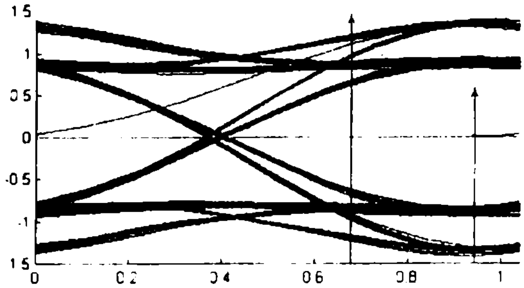

Figure 10b - The quqdrature signal u-v Eye pattern with two sampling locations single point for both branches in a compromise style (up) an' two points, left an' right in a perfect soluti n

Table 5 Double detection, decision and sampling method, the noise is applied at the entry of the channel

Snr/ber	5 dB	6 dB	7 dB	8dB
Mddds	8^e-5	3.6^e-5	3.3^e-6	<1^e-6
Mold	7e-3	3.4^e-3	7^e-4	3.3^e-4

II.3 *Coherent detection – integratig criterion*

Nyqui '' hird cri'eri n ^'T^) ---'u -- ' -- ---di'i-- for the lack of interference as:

$$\int_{kT-T/2}^{kT+T/2} h(t)dt = \delta_{0,k} = \begin{cases} 1, \ for \ k = 0 \\ 0, \ for \ k \neq 0 \end{cases} \quad (2)$$

The solution of replace a LPF with an integrator circuit is attractive by reason of the smaller number of operations. A certain filtering effect also does exist. The DCS is of the form: $s=p\cos\omega_c t+q\sin\omega_c t$, $p=\cos\varphi_k$; $q=\sin\varphi_k$. The proof of the NTC validity is immediate. The scheme in fig.11 uses the same method of as in fig.7. The carriers P, Q realize the zero-turning of the integrals at end of the interval T.

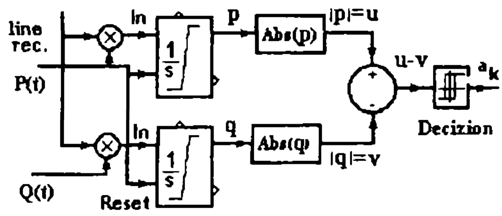

Figure 11 - DCS detector using an integrating proceeding (Mint)

Table 6 - BER vs. SNR for the integrating criterion

Snr/ber	4 dB	6 dB	7dB	8 dB	9 dB
Mint	4.4^e-3	2.3^e-4	5.3e-5	3.3^e-6	<1^e-6
Mold	2.5e-2	3.4^e-3	7.0^e-4	3.3^e-4	2.3^e-5

Mint is 1.5 dB better than plain old Miller, due to the filtering quality of a LPF with 1/s transfer function.

II.4 *Carrier phase and bit timing recovery*

Is a matter of fact that the carrier frequency of P, Q is half the signaling rate. Let a phase ε appear in the line signal $s=\cos(\omega_c t+\varphi_k+\varepsilon)$. p and q become:

$p = \cos(\varphi_k + \varepsilon)$, $q = -\sin(\varphi_k + \varepsilon)$;

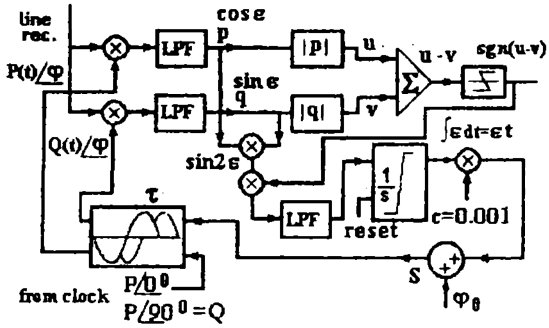

Figure 12 – PLL for data aided carrier recovery

The product is: $pq = \cos 2\varphi_k \sin 2\varepsilon = a_k \sin 2\varepsilon$; so, it is necessary to use a decizion directed algorythm. Data a_k are taken from the trigger that delivers sgn(u-v). The control signal in the loop is superposed to an initial value j0 with the aim to avoid negative delays.

SECTION III Theoretical Results

III.1. Distributions

In order to evaluate the error rate for our detection method we need to find the probability density function for the sum and difference of absolute values of the noise on the two channels, f=u+v and h=u-v. Let denote by d(f) and d(h). Their values are:

$$d(f) = \frac{2}{\sigma\sqrt{2\pi}} e^{-\frac{f^2}{4\sigma^2}} erf\,(f/2\sigma);\ \text{for } f \in \Re \quad (3)$$

$$d(h) = \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{h^2}{4\sigma^2}} erfc\,(h/2\sigma);\ for\ h \geq 0$$

(4)

Both u and v has normal, Gaussian distribution:

$$g(u) = d(u) = \begin{cases} \frac{2}{\sigma\sqrt{2\pi}} e^{-\frac{u^2}{2\sigma^2}} & \text{for } u \geq 0 \\ 0 & \text{for } u < 0 \end{cases} \quad (5)$$

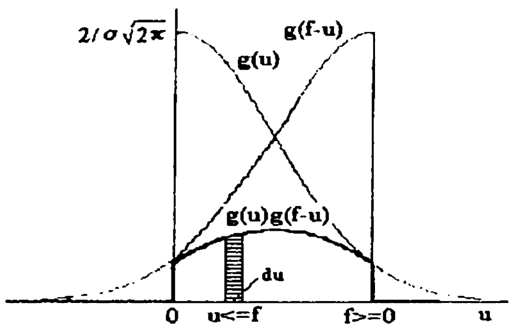

Figure13 – Half-Gaussian distributions are convoluted from 0 to f.

$$f = u + v > 0, u \leq f; d(f) = \int_{-\infty}^{f} g(u) g(f-u) du \quad (6)$$

$$d(f) = \frac{4}{2\pi\sigma^2} \int_0^f e^{-\frac{u^2}{2\sigma^2}} e^{-\frac{(f-u)^2}{2\sigma^2}} du \quad (7)$$

$$d(f) = \frac{2}{\pi\sigma} e^{-\frac{f^2}{4\sigma^2}} \int_{-f/2\sigma}^{f/2\sigma} e^{-\frac{x^2}{\sigma^2}} dx; \quad (8)$$

$$d(f) = \frac{2}{\pi\sigma} e^{-\frac{f^2}{4\sigma^2}} erf\,(f/2\sigma); \quad (9)$$

$$h = u - v \in \Re, v > 0, u > h, d(h) = \int_0^\infty g(v) g(h+v) dv \quad (10)$$

$$d(h) = \frac{2}{\pi\sigma^2} e^{-\frac{h^2}{4\sigma^2}} \int_0^\infty e^{-\frac{(v+h/2)^2}{\sigma^2}} dv \quad (11)$$

$$d(h) = \frac{2}{\sigma\sqrt{\pi}} e^{-\frac{h^2}{4\sigma^2}} erfc\,(h/2\sigma) \quad (12)$$

$$erfc\,(x) = \frac{2}{\sqrt{\pi}} \int_x^\infty e^{-y^2} dy = F(\infty) - F(x); \quad (13)$$

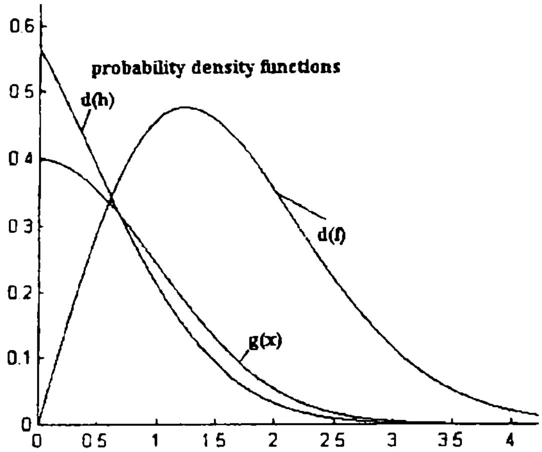

Figure 14 – Statistic distributions for the nonlinear processed noises

The error probability is to be approximated by:

$$P_{e,new} = \int_y^\infty d(h) dh \cong \int_y^\infty \frac{2}{\pi x} e^{-x^2/2} dx = \frac{1}{\pi y^2} e^{-y^2/2} \quad (14)$$

$$erfc\,'(x) = -F'(x) = -\frac{2}{\sqrt{\pi}} e^{-x^2} \quad (15)$$

We retain the first term from the Laurent series:

$$erfc\,(x) = \frac{e^{-x^2}}{\sqrt{\pi}} \left(\frac{a1}{x} - \frac{a2}{x^3} + \frac{a3}{x^5} - \ldots \right) \quad (16)$$

where a_k are obtained from derivative of erfc and identifying:

$$erfc\,(x) = \frac{e^{-x^2}}{x\sqrt{\pi}} \left(1 - \frac{1}{2x^2} + \frac{1 \cdot 3}{4x^4} - \frac{1 \cdot 3 \cdot 5}{8x^6} + \ldots \right) \quad (17)$$

So d(h) decreases more rapidly than Gauss's curve with a 1/x term.

$$erfc(x) \cong \frac{e^{-x^2}}{x\sqrt{\pi}}; d(h) \cong \frac{2}{\pi x} e^{-h^2/2}; d(x)/g(x) \cong \frac{2}{x}\sqrt{\frac{2}{\pi}}; \quad (18)$$

An approximation for the standard error rate is:

$$p_{e,gauss} = \int_y^\infty g(x) dx \cong 1/2.erfc(y/\sqrt{2}) \cong \frac{1}{y\sqrt{2\pi}} e^{-y^2/2}; \quad (19)$$

The ratio η between classical and our error rate is:

$$\eta = \frac{p_{e,gauss}}{p_{e,new}} = \sqrt{\frac{\pi}{2}} y = \sqrt{\frac{\pi}{2} \frac{P_{signal}}{P_{noise}}} \quad (20)$$

$$\lg \eta = \lg p_{e.gauss} - \lg p_{e.new} = 0.196 + \frac{1}{10} n_{signal/noise}; \quad (21)$$

III.2 – *How to explain some numbers*

Now, let's expose a peculiar result. If we denote by:

$$\Sigma^2 = \langle f^2 \rangle = \int_0^\infty f^2 d(f) df = 3.2732 \rightarrow 5.15 dB; \Sigma = 1.809 \quad (22)$$

$$s^2 = \langle h^2 \rangle = \int_{-\infty} h^2 d(h) dh = 0.727 \rightarrow -1.39 dB; s = 0.8525$$

(23)

$$\sigma^2 = \langle g^2 \rangle = \int_{-\infty}^{\infty} x^2 g(x) dx = 1 \rightarrow 0 \text{ dB}; \sigma = 1; \quad (24)$$

the ms values of u+v, u-v and p,q respectively, then (see fig.15) the effective values for this noises are summing as sinusoidal signals and the angle between them is obtained from: $\cos \phi = 2/\pi = 0.6366$.

$$\Sigma^2 = \sigma^2 + \sigma^2 + 2\sigma^2 \cos \phi = 2(1 + \cos \phi) = 3.2732, \quad (25)$$
$$s^2 = 2(1 - \cos \phi) = 0.7268;$$

$$\cos \phi = 2/\pi = 0.6366!; \phi = 50^0 40' \cong a\cos \frac{\sqrt{5}-1}{2}, \quad (26)$$

The value for it is: $\phi = 50^0 40'$, very close near to:

$$\kappa = a\cos \frac{\sqrt{5}-1}{2} = 51^0 50' \quad (27)$$

This is right the angle with the basement for the faces of Keops's great pyramid. And is in visible relation with a root of Fibonaci's series characteristic equation: x^2-x-1=0.

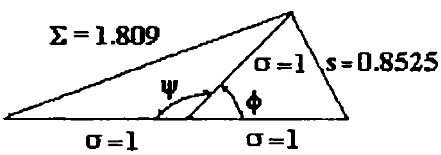

Figure 15 – A new gold triangle: the vectorial sum and difference of rectified noises is realized whet their angle is a golden one.

The powers σ^2, s^2 and Σ^2 were measured as temporal quadratic means and the logarithmic values confirm the calculus.

$$\Delta n_{calcul} = 6.565 \text{ dB}; \Delta n_{masur} = 6.8 \text{ dB} \quad (28)$$

II. 3. More about the DCS reception

1. It is possible to demodulate DCS as a QAM even though the result is ternary and the SNR is deteriorating.
2. Here exists a large amount of correlation between branches.
3. Other methods can be used to improve the noise protection

Conclusions

a) Three new methods were proposed, with good and very good results: decision with floating threshold, detection following an integrating criterion and decoding by two data flows decision and double sampling

b) Formulas for two new found distributions in concordance with good experimental results.

c) A graphic image for sums and difference of powers of the rectified noises as a.c. phasors; there virtual angle has peculiar proprieties.

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