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# SOQPSK-TG STC : a new decoding scheme for a higher bit rate and a better link budget

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## Abstract

To protect telemetry from antenna masking during flight tests, two antennae must be used on-board. To avoid phase opposition of the two transmitted signals at reception, frequency diversity has been used for decades. However with spectrum auctions this technique is no longer compatible with the requirement of higher bit rate. To overcome this issue, well-known modulation format like COFDM may be used at the expense of power efficiency, with few constraints on antennae position. IRIG-106 standard proposed also SOQPSK-TG STC which is efficient both in power and in spectrum at the expense of the antennae position. To demodulate this format a dual-trellis solution has been proposed based on XTCQM representation [1] with results very close to SISO as far as distance between antennae does not exceed half bit delay. Therefore this decoding scheme is a hard decision scheme and consequently is less power efficient when a FEC such as LDPC is used. In this paper a new decoding scheme will be presented with soft decision outputs to increase power efficiency, and with a better tolerance to differential delay.

**Key words:** Flight test, spatial diversity, Space time coding, telemetry.

## Introduction

Transmitting real-time telemetry with low latency on aeronautical channel is very demanding. The data link has to be error free on at least four different channel models from high Doppler spread take-off fading channel to low SNR far flight channel [1]. Spectrum auctions constrain even more the data link pushing it to C band and requiring spectral efficiency to transmit up to 20Mbps to meet market demand. A first waveform based on COFDM meeting all these requirements has been proposed [1][2] to transmit telemetry from commercial aircraft. However for small aircraft or for carriers with little power available, the amplifier back-off required by COFDM could be incompatible with mission requirements. For such missions SOQPSK-TG combined with equalization techniques can meet power and spectral efficiency on channel with moderate frequency selectivity [3]. However under certain aircraft to ground station geometries, antenna masking occurs, leading to link outage. A MISO hard bit scheme Alamouti like has been proposed [4] to mitigate the shadowing and the associated differential delay based on XTCQM decomposition of STC-SOQPSK-TG waveform. In this paper a new low resource decoder with soft output and its performances will be presented.

## IRIG-106 STC-SOQPSK-TG

IRIG-106 has standardized constant envelope waveforms with ever increasing spectral efficiency enabling use of power amplifier at saturation

TABLE A-3. CHARACTERISTICS OF VARIOUS MODULATION METHODS.

| Characteristic | PCM/FM with single symbol detection | PCM/FM with multi-symbol detection | FQPSK-B, FQPSK-JR, SOQPSK-TG | ARTM CPM |
|---|---|---|---|---|
| Occupied Bandwidth | 1.16 bit rate | 1.16 bit rate | 0.78 bit rate | 0.56 bit rate |
| Sensitivity ($E_b/N_0$ for BEP=1e-5) | 11.8-15+ dB | 9.5 dB | 11.8-12.2 dB | 12.5 dB |
| Synchronization time | 100 to 10,000 bits | 250 bits | 5,000 to 30,000 bits | 30,000 to 150,000 bits |
| Synchronization threshold level ($E_b/N_0$) | 3 to 4 dB | 2 dB | 4.5 to 5 dB | 8.5 dB |
| Phase noise susceptibility* | 2 | 1 | 3 | 4 |
| Co-channel interference susceptibility* | 2 | 1 | 3 | 4 |

* 1=Best, 2=Second Best, 3=Third Best, 4=Worst

*Fig.1 IRIG-106 waveforms comparison*

However to prevent data link from on-board antenna masking outage, dual antennae (MISO) transmission is required to ensure error-free telemetry transmission, and SOQPSK-TG features an interesting trade-off between spectral efficiency with an occupied bandwidth of 0.78*bit rate and power efficiency thanks to its 4dB synchronization level and its performance when coupled to an ARJA LDPC FEC [5]. Combining SOQPSK-TG and MISO space diversity is however not straightforward as modulation is not linear and antenna spacing on-board induced a differential delay at reception which is not compatible with Alamouti scheme. In [6] an Alamouti like scheme has been proposed, and a SOQPSK-TG space time decoder based on XTCQM representation given. This scheme has been standardized by IRIG-106 group [7] enabling telemetry transmission with a high quality of service whatever the aircraft to ground station geometry.

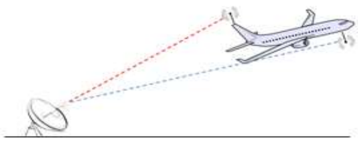

*Fig.2a Miso scheme with differential delay*

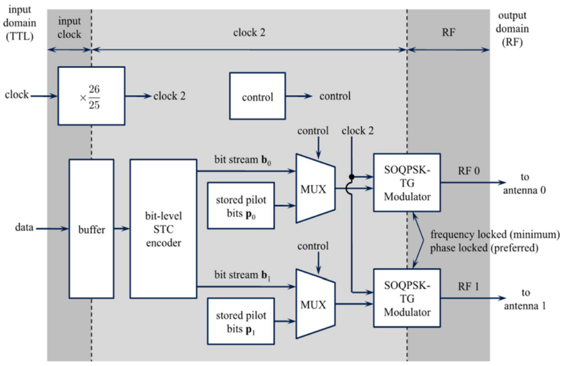


*Fig.2b STC on-board modulator*

| | Alamouti Code | | IRIG 106-15 | |
|---|---|---|---|---|
| | symbol index | | symbol index | |
| | $2k$ | $2k+1$ | $2k$ | $2k+1$ |
| TX 0 | $s_{2k}$ | $-s^*_{2k+1}$ | $b_{4k}, b_{4k+1}$ | $\bar{b}_{4k+2}, b_{4k+3}$ |
| TX 1 | $s_{2k+1}$ | $s^*_{2k}$ | $b_{4k+2}, b_{4k+3}$ | $b_{4k}, \bar{b}_{4k+1}$ |

*Fig.2c Bit level STC encoder*

As can been noticed on fig.3 the tolerated differential delay is strictly limited to one bit, and a noticeable BER degradation occurs at half a bit period. Taken into account plane geometry and constraints on antennae location, a direct consequence of this limitation is a limitation of the telemetry rate which cannot be higher than

$$Rmax < \frac{c}{2Dant}$$

(1)

Where Dant is the distance between antennae, considering differential delay due to on-board cables has been perfectly compensated.

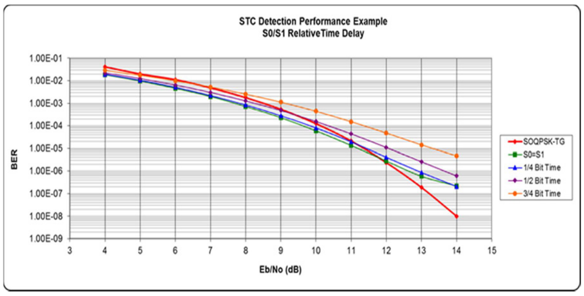


*Fig.3 STC detection performance vs differential time delay*

Consequently, the ability to decrease degradation related to differential delay could be a way to relax constraints on antennae position or to increase telemetry rate with the same position constraints. Another constraint of the differential delay on the XTCQM LS decoder [4][6] is it requires to implement two trellis, which increases its complexity.

## STC PAM LS decoder

In [9] a new PAM decomposition was proposed by Othman, STC signal can then be written as

$$s_p(t) \approx \sum_i \rho^p_{0,2i} w_0(t-2iT) - \rho^p_{1,2i+1} w_1(t-2iT-T) + \left(\sum_i \rho^p_{0,2i+1} w_0(t-2iT-T) - \rho^p_{0,2i} w_1(t-2iT)\right)$$

(2)

Where :

$p \in \{0,1\}$ is the on-board antenna number

$$\rho^p_{0,i} = \begin{cases} \left(2b_i^{(p)}-1\right) \text{ i even} \\ j\left(2b_i^{(p)}-1\right) \text{i odd} \end{cases}$$

$$\rho^p_{1,i} = \begin{cases} -j\left(2b_{i-2}^{(p)}-1\right)\left(2b_{i-1}^{(p)}-1\right)\left(2b_i^{(p)}-1\right) \text{ i even} \\ -\left(2b_{i-2}^{(p)}-1\right)\left(2b_{i-1}^{(p)}-1\right)\left(2b_i^{(p)}-1\right) \text{ i odd} \end{cases}$$

$b_i^{(0)}$ and $b_i^{(1)}$ are ith bits feeding respectively channel 0 and 1 encoded accordingly to fig2.C

$w_0(t)$ and $w_1(t)$are shaping pulse given in [9] and fig.4

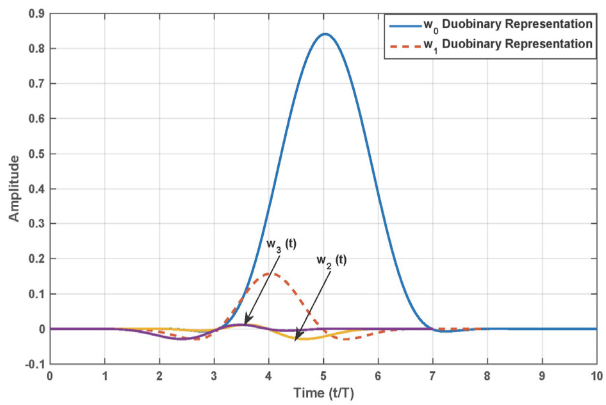


*Fig.4 DBD Pulse shape*

On ground antenna I the signal received is

$$r_I(t) = \left[h_{0,I} s_0\left(t - \Delta t_{0,I}\right) + h_{1,I} s_1\left(t - \Delta t_{1,I}\right)\right] e^{j2\pi \Delta f_I t} + z_I(t)$$

(3)

Where

$h_{p,I}$ is the LOS channel complex gain between antenna p and antenna I

$\Delta t_{p,I}$ is propagation delay between antenna p and antenna I and $\Delta\tau_I = \Delta t_{1,I} - \Delta t_{0,I}$

$\Delta f_I$ frequency offset, $z_I(t)$ additive noise

In Miso scheme I=0, considering perfect carrier synchronization and combining (2) and (3) received signal after ADC could be written

$$r_0(n) \approx h_0 \underbrace{\left[\sum_i \rho^0_{0,i} w_0(nT' - iT) + \sum_i \rho^0_{1,i} w_1(nT' - iT)\right]}_{s_0(nT')} + h_1 \underbrace{\left[\sum_i \rho^1_{0,i} w_0(nT' - iT - \Delta\tau) + \sum_i \rho^1_{1,i} w_1(nT' - iT - \Delta\tau)\right]}_{s_1(nT' - \Delta\tau)} + z(nT')$$

(4)

Where $T'$ is sampling time. As shown on fig.4 shaping pulse lasts more than T, consequently intersymbol interference remains and has to be filtered. Optimal detection filter is discussed in [10], after filtering and sampling of two TX channels samples can be written using (4) accordingly to Fig.5:

$$y(4k) \approx h_0 \sum_{i=-1}^{1} \rho^0_{0,4k-i} \widetilde{w}_0(iT) + h_0 \rho^0_{1,4k} \widetilde{w}_1(0) + h_1 \sum_{i=-1}^{1} \rho^1_{0,4k-i} \widetilde{w}_0(iT - \Delta\varepsilon T) + h_1 \rho^1_{1,4k} \widetilde{w}_1(-\Delta\varepsilon T) + \tilde{n}(4kT)$$

(5)

$$y_{\Delta\tau}(4k) \approx h_0 \sum_{i=-1}^{1} \rho^0_{0,4k-i} \widetilde{w}_0(iT + \Delta\varepsilon T) + h_0 \rho^0_{1,4k} \widetilde{w}_1(\Delta\varepsilon T) + h_1 \sum_{i=-1}^{1} \rho^1_{0,4k-i} \widetilde{w}_0(iT) + h_1 \rho^1_{1,4k} \widetilde{w}_1(0) + \tilde{n}(4kT + \Delta\varepsilon T)$$

(6)

Where $\widetilde{w}_i$ is filtered pulse shape, $\tilde{n}$ is z filtered, and $\Delta\varepsilon$ is nearest integer from $\Delta\tau / T$

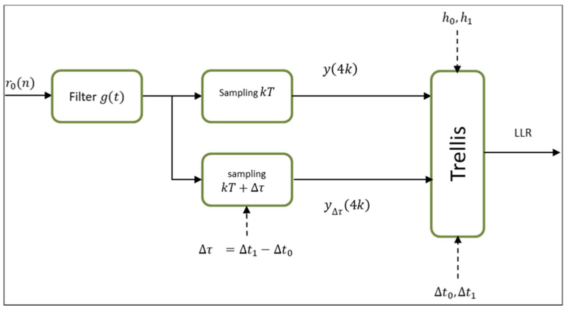


*Fig.5 STC SOQPSK-TG decoder*

Then the following likelihood function is considered

$$\Lambda(\underline{S}) = \sum_{n=0}^{(N-1)/4} \left( \sum_{m=-1}^{2} \left[ \left| B^{(0)}_{m,n} \right|^2 + \left| B^{(\Delta\tau)}_{m,n} \right|^2 \right] \right)$$

(7)

Where

$$B^{(0)}_{m,n} = y(4n+m) - h_0\left(\sum_{i=-1}^{1} \rho^0_{0,4n+m-i} \widetilde{w}_0(iT) + \rho^0_{1,4n+m} \widetilde{w}_1(0)\right) - h_1\left(\sum_{i=-1}^{1} \rho^1_{0,4n+m-i} \widetilde{w}_0(iT - \Delta\varepsilon T) + \rho^1_{1,4n+m} \widetilde{w}_1(-\Delta\varepsilon T)\right)$$

(8)

$$B^{(\Delta\tau)}_{m,n} = y_{\Delta\tau}(4n+m) - h_0\left(\sum_{i=-1}^{1} \rho^0_{0,4n+m-i} \widetilde{w}_0(iT + \Delta\varepsilon T) + \rho^0_{1,4n+m} \widetilde{w}_1(\Delta\varepsilon T)\right) - h_1\left(\sum_{i=-1}^{1} \rho^1_{0,4n+m-i} \widetilde{w}_0(iT) + \rho^1_{1,4n+m} \widetilde{w}_1(0)\right)$$

(9)

The ML estimate is then

$$\hat{\underline{S}} = \underset{S}{\operatorname{argmin}}\, \Lambda(\underline{S})$$

(10)

Where $S_n = [b_{4n} \quad b_{4n+1} \quad b_{4n+2} \quad b_{4n+3}]$

Thus the more likely underlying LLR sequence can be estimated with SOVA algorithm using a 16 states trellis (Fig.5 & 6)

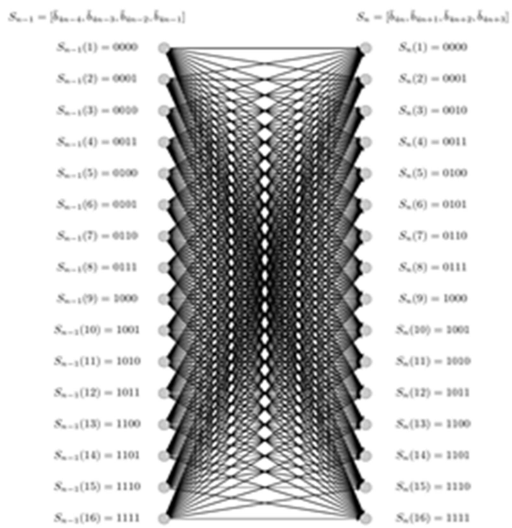

*Fig.6 STC-SOQPSK-TG DBD trellis*

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Instead of trellises derived from XTCQM decomposition [4], it should be noticed that the expressions (8) and (9) remains unchanged whatever the sign of the differential delay between channels, consequently a unique trellis has to be implemented in the decoder instead of two. Moreover considering (8) and (9) and taken into account that each STC symbol is a block of 4 bits, 2048 sub-metric should be computed to decode at bit level, however developing (8) and (9) as a function of the branch bits leads to consider only 320 sub-metric [10] against 480 in [4]. It could also be noticed that on AWGN channel at high SNR regime DBD PAM decoder outperforms XTCQM decoder of more than 1dB at BER 10E-6 for a 0.4 differential delay.

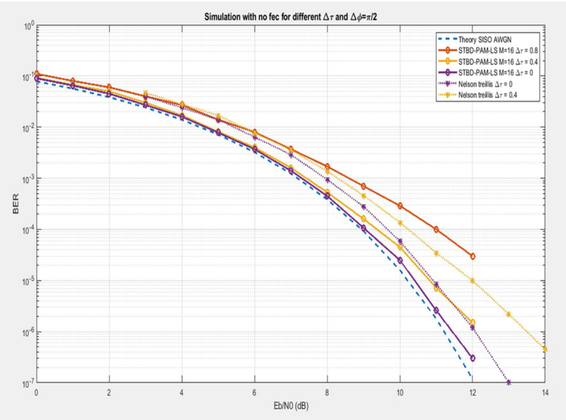


*Fig.6 BER proposed decoder vs XTCQM decoder*

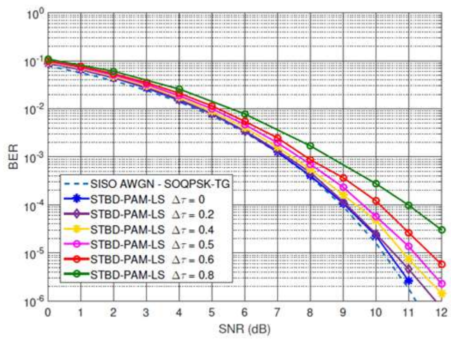


*Fig.7 Proposed decoder BER vs differential delay*

Instead of improving link budget of 1dB, looking at Fig6. and 7, as proposed decoder has the same BER performance at 10E-6 for a 0.6 differential delay than the XTCQM decoder at 0.4 differential delay, antennae position constraints could be relaxed of 50%, or bit rate could be increased of 50% with the same antennae position constraint .

### STC PAM LS soft decision decoder

Soft decisions may be extracted thanks to MAX-log-MAP equivalent SOVA for non binary codes [12]. The classical approach to estimate the underlying bit sequence and its soft information consists then in making a full trellis search up to a certain depth $\delta$ and then performing a trace-back loop, which means, for the proposed trellis, to calculate 256 metrics and to store 256 $\delta$ reliability values. Instead of extending all states from one symbol time to the next, a suboptimal technique in [11] consists in extending only the M best states, which is called M-algorithm. Othman in [8] proposed to adapt it to its non-binary STC 4 bits, which leads to calculate 16.M branch metrics.

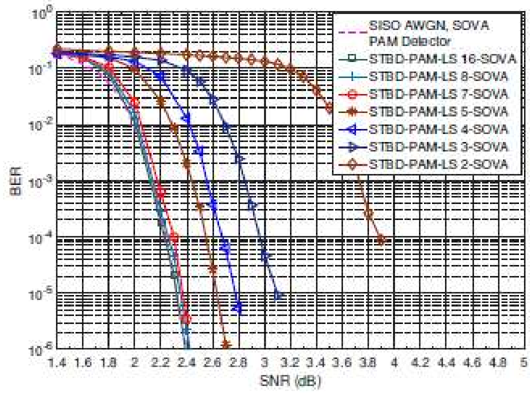


*Fig.8 STC M-algorithm AR4JA BER vs M parameter*

To estimate optimal M value decoder has then been simulated on AWGN channel in combination with IRIG-106 AR4JA LDPC code K=2/3 L=4096. As shown on Fig.8 for M > 4 BER degradation is below 0.2dB dividing by 3 metrics to be calculated, for M=8 metrics calculation is divided by 2 at no BER cost. Taken into account Fig.9 giving required resources, for M=8 we thus get a significant gain in gates between the XTCQM hard bit decoder and the proposed one with LLR output.

| | State of the art STBD-XTCQM-LS | Proposed solution STBD-PAM-LS with $M$-algorithm | Proposed solution STBD-PAM with $M$-SOVA | Proposed solution STBD-PAM-ML |
|---|---|---|---|---|
| Addition | 3968 | $640 + 128M$ | $644 + 176M$ | 576 |
| Multiplication | 960 | 640 | 640 | 64 |
| $\|.\|^2$ | 480 | 320 | 320 | – |
| Comparison | 240 | $15M$ | $46M + 40$ | 256 |
| Sorting operation of 16 elements | 0 | 1 | 1 | 0 |
| Number of stored values | 16 | 16 | $17M$ | 64 for hard decoding 128 for soft decoding |

*Fig.9 Operations required by STC decoder*

Both decoders have then been simulated using IRIG-106 AR4JA FEC on Fig. 10 exhibiting a 1.5dB gain. Once again this gain could be used to relax link budget or antenna position constraint or increase bit rate as the proposed decoder enables a higher differential delay margin at constant BER.

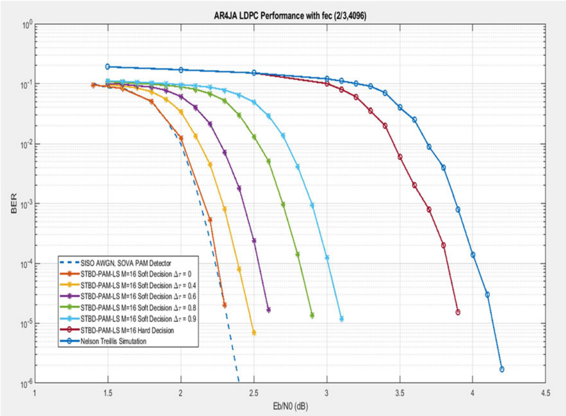


*Fig.10 Decoders performance with AR4JA 2/3,4096*

## Conclusion

Space time coding applied on SOQPSK-TG telemetry link, when constant envelope modulation is mandatory because of on-board power constraint, is a very performant solution to solve the antenna masking problem using a MISO scheme. However the telemetry bit rate is upper-bounded by a maximum bit rate that depends on differential delay (1). XTCQM decoder thus enables, with no significant loss, a differential delay of a quarter of bit, which means for a 20Mbps link a maximum distance of 3.75m between antennae with a perfect calibration of path from modulator outputs to antennae inputs. With proposed PAM-LS decoder it should be possible with the same antennae location to reach 32Mbps at equivalent Eb/N0 and 40Mbps at the cost of an extra 0.2dB. Moreover using IRIG-106 LDPC FEC PAM-LS decoder enables to gain an extra 1.5dB link budget thanks to soft decision output at the price of reduced gate consumption.

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