A BLIND PARTIALLY COHERENT
MULTI-H CPM RECEIVER
FOR AERONAUTICAL TELEMETRY

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ABSTRACT
Multi-H Continuous Phase Modulation is a highly bandwidth efficient constant amplitude modulation scheme. Because of these qualities it was selected as the Advanced Range Telemetry (ARTM) tier II waveform. In the past, two demodulation techniques have widely been proposed, coherent detection and non-coherent detection. This paper presents a receiver design that implements a hybrid, partially coherent detection scheme that takes advantage of the positive aspects of both coherent and non-coherent detection. Because complete phase recovery is not required, the hybrid receiver performs better in environments with fast fading, strong phase noise, and multi path when compared to the traditional coherent receiver. The hybrid receiver can also acquire and reacquire signals much faster than conventional coherent receivers. The hybrid receiver design implements a partial carrier detection scheme that utilizes phase information that performs much better in AWGN environments than typical non-coherent receivers. Simulation results show that the hybrid receiver has low implementation loss compared to the optimal Maximum Likelihood Sequence Estimation (MLSE) receiver.

KEYWORDS
Multi-H CPM, ARTM Tier II, Demodulation, Hardware Implementation, Coherent, Non-Coherent, Blind Recovery, Synchronization

INTRODUCTION
As the amount of data transmitted, and the number of users both continue to increase for aeronautical telemetry, the amount of available telemetry spectrum space is becoming scarce. For this reason, considerable research in the community has focused on creating bandwidth efficient modulation schemes. Also, because most telemetry transmit amplifiers are run in saturation mode, a modulation scheme that is power efficient (i.e. no amplitude variations) is desired as well. Multi-H CPM has been selected as the (ARTM) Tier II modulation because it satisfies both of these requirements.
Multi-H CPM is three times more spectrally efficient than PCM/FM, but suffers from high complexity due to intricate synchronization and demodulation techniques. The synchronization techniques of other modulation schemes involving the openings of the “eye diagram” do not work, as the eye diagram for Multi-H is almost unrecognizable. Because of this, other unique synchronization algorithms must be used. The MLSE demodulator for Multi-H CPM suffers from large complexity and is not feasible to implement with existing FPGA resources. For this reason, reduced complexity techniques are called for.

In this paper, we first summarize mathematically the Multi-H CPM signal. We then discuss several synchronization techniques that are robust and which provide fast acquisition and reacquisition times in a variety of impaired environments. These synchronization techniques allow partial recovery of the carrier which can be used by the demodulator. Then a reduced complexity demodulator is presented which uses the partial carrier information. Finally, we conclude the paper by discussing our simulation and implementation results.

**SIGNAL MODEL**

The standard CPM signal consists of a digitally modulated carrier where the phase is constrained to be continuous. The ARTM tier II waveform is a specific form of CPM and its complex baseband signal representation is given by the following equations:

$$s(t, \alpha) = \sqrt{\frac{E}{T}} \exp \left( j \psi(t, \alpha) \right)$$  \hspace{1cm} (1)

$$\psi(t, \alpha) = 2\pi \sum_{i=0}^{\infty} h_i \alpha_i q(t - iT)$$  \hspace{1cm} (2)

Where $E$ is the symbol energy, $T$ is the symbol duration, $\{ h_i \} \in \{4/16, 5/16\}$ is the set of modulation indexes, $\alpha = \{\alpha_i\}$ are the modulated symbols taking values in the $M = 4$ element alphabet \{-3, -1, 1, 3\}, and $q(t)$ is a phase shaping pulse defined by the following equations:

$$q(t) = \int_{0}^{t} f(\tau) d(\tau)$$  \hspace{1cm} (3)

$$f(t) = \begin{cases} \frac{1}{2LT} \left(1 - \cos \left(\frac{2\pi t}{LT}\right)\right) & 0 \leq t \leq LT \\ 0 & otherwise \end{cases}$$  \hspace{1cm} (4)

Where $L=3$ and the phase pulse $q(t)$ and frequency pulse $f(t)$ are depicted in the following figure:
The phase of the Multi-H CPM signal can be broken down further into a phase trajectory consisting of the three most current symbols $\alpha_n$ and a phase state $\theta_{n-3}$ consisting of 32 possible values.

\[
2\pi \sum_{i=0}^{n} h_i \alpha_i q(t-iT) = 2\pi \sum_{i=n-2}^{n-3} h_i \alpha_i q(t-iT) + \pi \sum_{i=0}^{n-3} h_i \alpha_i
\]  

(5)

The baseband eye diagram for both the real and imaginary parts of the signal is shown below in Figure 2:
Another useful visual representation of the Multi-H CPM signal comes from the FM eye diagram. This is very useful in partially coherent cases and is depicted below in Figure 3:

![Figure 3: FM Eye Diagram of Multi-H CPM](image)

Once the signal is transmitted through the channel and received by the receiver its new complex baseband signal representation becomes:

\[
r(t) = \sqrt{\frac{E}{T}} \exp\left(j \psi(t + \tau, \alpha)\right) \exp\left(j(2\pi f_c t + \theta)\right) + n(t)
\]

(6)

Where \( \tau \) is the symbol delay, \( f_c \) is the frequency error, and \( \theta \) is the phase shift between the transmitter and receiver. In the above equation \( n(t) \) is the Additive White Gaussian Noise (AWGN) with single sided power spectral density \( N_0 \). All of these impairments are introduced by the channel.
CARRIER SYNCHRONIZATION

Frequency synchronization is critical for any coherent or partially coherent receiver. In our proposed algorithm we attempt to minimize the following cost function:

\[ J_F = \text{avg} \left\{ \left( \frac{d}{dt} \arg(r(t, \alpha, f_c, \theta, \tau)) \right)^2 \right\} \tag{7} \]

Where \( r(t) \) is the received signal which consists of a carrier frequency error \( f_c \), a carrier phase error \( \theta \), a timing error \( \tau \), and the transmitted symbol sequence \( \alpha \). The goal is then to minimize this function with respect to \( f \). Assuming that the phase error \( \theta \) is uniformly distributed between \((-\pi, \pi)\), the timing error is evenly distributed between \((0,T)\), and that each symbol sequence \( \alpha \) has equal probability of occurring, we can average the cost function \( J_F \) with respect to each of these variables. After doing this the cost function will still have a minimum when the received frequency matches the transmitted frequency (\( f_c = 0 \)) and therefore a blind carrier tracking algorithm can be obtained.

Using the above cost function, a gradient decent algorithm that tracks the carrier frequency is achieved which uses the derivative of the cost function \( J_F \) with respect to \( f \). The error for the gradient decent algorithm is then given by the following equation:

\[ e_F(k) = \frac{\partial J_F(k)}{\partial f} \tag{8} \]

The carrier tracking algorithm, like other PLL and timing loops, employs a parameter that controls the loop bandwidth which balances convergence speed with residual RMS noise error. The algorithm in the proposed receiver can have a convergence speed of up to several hundred symbols while still maintaining reasonable residual RMS noise error.

Although this algorithm performs frequency synchronization, the task of coherent demodulation also requires that phase synchronization be achieved. Unfortunately, most phase tracking algorithms for Multi-H CPM require some form of decision-directed phase estimation. For a decision-directed algorithm to work it needs some form of initial convergence information before the algorithm can even be started. If there is no initial convergence, the algorithm will most likely use bad initial decisions and the algorithm will diverge instead of converging. Because of this, the acquisition time in a coherent receiver increases even more because frequency lock must be achieved before any decision-directed phase lock loop can start. In telemetry applications, where the vehicle might only transmit for a tiny segment of time and the signal might quickly fade away and reappear, acquisition time becomes a critical factor and a major deterrent for using coherent detection.
The advantage of the frequency tracking algorithm presented here is that it is completely blind. Therefore, the algorithm can lock even if symbol synchronization has not been obtained yet. Since the algorithm is blind the time for convergence can be very fast as it does not rely on other loops to settle before starting. This is really important in telemetry applications where the signals can fade out many times during a flight and fast acquisitions and reacquisition are a must.

SYMBOL SYNCHRONIZATION

Timing synchronization is very difficult for Multi-H CPM. Due to the temporal length of the 3RC filter, there is a smearing of the signal and along with the large number of phase states; the resulting eye diagram becomes almost unrecognizable as can be seen in Figure 2.

As Figure 2 shows, timing synchronization for Multi-H CPM is a very challenging task. Timing synchronization must sample the signal at the right instant as well as correct for any baud rate offsets. Unlike most receivers, Multi-H CPM provides a unique challenge in that the super baud rate must also be synchronized due to the alternating $h_i$ modulation indices. The cost function that we attempt to maximize is given in [1] and is expressed by the following equation:

$$ J_S = \int_{0}^{L_0 T_s} \int_{0}^{L_0 T_i} r(t_1) r^*(t_2) F(t_2 - t_1, t_2 - \tau) dt_1 dt_2 $$

(9)

Where $T_S$ is the symbol interval, $L_0$ is the observation interval, and $F(\Delta t, t)$ contains the expectation over the hypothesized data sequence $\alpha$ and is defined as

$$ F(\Delta t, t) = E_\alpha \left\{ e^{j[\psi(t, \alpha) - \psi(t - \Delta t, \alpha)]} \right\} $$

(10)

Similar to the Frequency Tracking Loop, this cost function can be used in a gradient decent algorithm to track the baud offset along with the baud frequency. The tracking algorithm uses the derivative of the cost function $J_S$ and the error term is given by the following equation:

$$ e_S(k) = \frac{\partial J_S(k)}{\partial \tau} $$

(11)

The symbol tracking algorithm, like the Frequency Lock Loop (FLL), employs a parameter that controls the loop bandwidth which balances convergence speed with residual RMS noise error. Similar to the FLL, the symbol tracking loop in the proposed receiver can have a convergence speed of up to several hundred symbols while still maintaining reasonable residual RMS noise error.
Because the cost function resembles a correlation with the input stream it naturally produces an algorithm for super baud timing recovery as well. Therefore, resources can be saved while implementing the algorithm in hardware because only one circuit will be needed to produce metrics for both the symbol and super baud recovery.

Throughout the literature there are many tracking algorithms that use decision aided techniques to provide symbol recovery. These techniques are shown in the following reference [2]. A major reason we used the proposed symbol recovery scheme is because it is completely blind and is not dependent on carrier recovery. This means that the algorithm can lock right away and doesn’t have to wait for the carrier tracking loop to lock before it can start tracking. Also, since the algorithm is not decision aided, no knowledge of previous symbols is required. As mentioned in the previous section, the blind nature of the algorithm allows the receiver to perform fast acquisition which is critical in telemetry applications.

DETECTION METHODS

The optimal MLSE detection of Multi-H CPM consists of picking the symbol that maximizes the following (see e.g. [6]):

\[
\hat{\alpha} = \arg \max_{\alpha} \left\{ \text{Re} \left( \int r(t) s^*(t, \alpha) \, dt \right) \right\}
\]  

(12)

This detection method is achieved in a recursive manner using the Viterbi Algorithm. The following recursion is shown below:

\[
\text{Re} \left( \int_{0}^{(n+1)T} r(t) s^*(t, \alpha) \, dt \right) = \text{Re} \left( \int_{0}^{nT} r(t) s^*(t, \alpha) \, dt \right) + \text{Re} \left( \int_{nT}^{(n+1)T} r(t) s^*(t, \alpha) \, dt \right)
\]

(13)

The branch metric term above can be broken down further into the following:

\[
\text{Re} \left( \int_{nT}^{(n+1)T} r(t) s^*(t, \alpha) \, dt \right) = \text{Re} \left( e^{j \theta_{n-3}} \int_{nT}^{(n+1)T} r(t) e^{j \theta(t; \alpha_{n-2}, \alpha_{n-1}, \alpha_n)} \, dt \right)
\]

(14)

Where \( \theta_{n-3} \) is the phase state and \( \theta(t; \alpha_{n-2}, \alpha_{n-1}, \alpha_n) \) is the phase trajectory given in equation (5). Therefore, the optimal MLSE detector is a coherent detector where the phase states \( \theta_{n-3} \) build distance between competing paths in the Viterbi Trellis.
For the case of non-coherent detection we assume that the phase at the receiver is unknown and that this phase will be slowly varying over time. The detection method for non-coherent receivers then tries to maximize the following:

\[
\hat{\alpha} = \arg \max_{\alpha} \left\{ \left| \int r(t) s^* (t, \alpha) dt \right|^2 \right\}
\]

This is identical to the metric for the coherent case except that the non-coherent case uses the magnitude squared of the correlation while the coherent case uses the real part of the correlation.

A typical non-coherent receiver is shown in below in Figure 4:

Different non-coherent detection methods have been proposed in the literature [5] [6]. Each of these techniques tries to achieve performance similar to coherent detection by using correlations across many symbol intervals instead of one. Since the phase is still varying in a non-coherent receiver, there needs to be a limit on the amount of multi symbol correlations because this slowly varying phase will eventually degrade the performance if the correlation interval is too high. Therefore, a trade off emerges between having a long enough interval to increase distance between competing paths and having an interval short enough so that the phase changes over time will not affect performance.
For the case of the partially coherent receiver (where the carrier frequency is tracked), the amount of phase change over time is a lot less compared to that of a typical non-coherent receiver. Consequently, the correlation interval length can be increased and performance close to that of the coherent detector can be achieved. In summary, partially coherent receivers use detection methods similar to non-coherent methods, but performance is better than that of non-coherent receivers because a longer correlation interval length can be maintained.

SIMULATION AND IMPLEMENTATION RESULTS

A couple of simulation and implementation results are presented below in Figure 5. The partially coherent receiver employing techniques similar to those mentioned in the previous section is shown (red line) in the Figure. The resulting performance is within 1.5 dB of the optimal coherent MLSE detector. At the time of this study, a simple FM Detector demodulator using a 16 state trellis was implemented in hardware to achieve a working solution to test the tracking algorithms. That performance result is shown (blue line) in the Figure. With the FM Detector demodulator successfully implemented in hardware, the implementation of the partially coherent receiver in hardware will begin in the near future.

![Figure 5: BER Performance of various Multi-H CPM Receivers](image-url)
CONCLUSION

In this paper, a partially coherent demodulator for Multi-H CPM was presented. The receiver uses two completely blind synchronization techniques to achieve both carrier and symbol synchronization. These completely blind, independent algorithms allow the receiver to achieve fast acquisition and reacquisition times important for many telemetry applications. The receiver utilizes phase information garnered from the carrier tracking loop, thus aiding in the demodulator process. The resulting demodulator achieves performance comparable to that of optimal MLSE coherent detection.

REFERENCES


