DFSE EQUALIZATION OF DUAL-*H* CPM OVER UHF MILSATCOM CHANNELS

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ABSTRACT

To meet the demand for higher data rate MIL-STD-188-181B has added a set of dual-h Continuous Phase Modulation (CPM) waveforms that nearly double the available data rates without requiring additional transmit power. Unfortunately, rates greater than 56 kbps are constrained by severe inter-symbol interference (ISI) induced by narrow satellite crystal filters. In this paper, we study the application of Decision Feedback Sequence Estimation (DFSE) in order to equalize a 64 kbit/s dual-h quaternary full response CPM waveform transmitted over a representative 25 kHz UHF MILSATCOM channel. The performance is evaluated by Monte-Carlo simulation. Results show DFSE equalization offers an effective and computationally manageable approach to higher data rates using legacy waveforms.

I. INTRODUCTION

Ultra High Frequency (UHF) military satellite communications (MILSATCOM) links provide narrowband tactical secure voice and data communications to the mobile warfighter. The current satellite constellation, UHF Follow-on (UFO), uses four pairs of collocated satellites spaced in quadrature about the equatorial plane parked in low-incline geostationary orbit. Each UFO satellite provides one Earth coverage beam and contains a limiting repeater UHF transponder payload that provides only 21-5 kHz and 17-25 kHz "bent-pipe" channels. To support the increasing demand for high data rate service, the Joint Interoperability Engineering Organization (JIEO) and Defense Information Systems Agency (DISA) introduced a high-speed dual-*h* CPM waveform that increased data rates to 56 kbps doubling current capacity without requiring increased transmitter power or bandwidth [1]. Unfortunately, further increases in data rate are not possible due to a massive drop in detection efficiency caused by severe ISI induced by UFO's narrow satellite crystal filters.

With the launch of UFO F-10 in 1999 and the procurement of F-11 to be launched in 2003 the UFO constellation will be in service well into the next decade. There is still a growing need to improve availability and provide expanded service types such as imagery, video telecommunication, and common operation picture. This will require even higher data rates. Since the next generation UHF constellation will not be deployed until 2008 [2] and will not offer full operational capability until 2013, there is a need to further increase data rates using legacy waveforms.

In this paper, we study the application of DFSE [3] in order to equalize a 64 kbit/s dual-*h* quaternary full response CPM waveform transmitted over a representative 25 kHz UFO channel. Computer simulation results indicate DFSE can effectively mitigate ISI offering an attractive approach to higher data rates with only a manageable increase in processing complexity. Section II describes the UFO channel and illustrates the source of ISI. Dual-*h* CPM is introduced in Section III. Section IV details the DFSE algorithm. Performance results, given in terms of bit error probability, are presented in Section V.

II. UFO SATELLITE CHANNEL

Consider Fig. 1, which illustrates the major elements of UFO's transponder channel processing.

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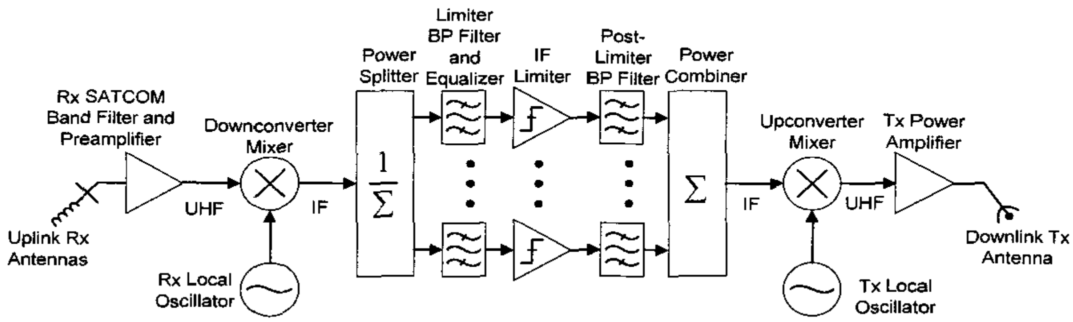

Fig. 1 - UFO Transponder Channel Processing

The uplink signals are group down converted. User carriers are individually channelized and processed by an analog bandpass crystal filter/phase-equalizer assembly, hard-limited, post-filtered, and then combined before being retransmitted. The overall selectivity requirements of the channel are specified in Table I.

TABLE I
Overall Channel Selectivity Specifications

Selectivity	Channel BW	
	25 kHz	5 kHz
≤ 1 dB	± 12 kHz	± 2.4 kHz
≥ 60 dB	± 37.5 kHz	± 7.5 kHz

The pre-limiter and post-limiter crystal filter assemblies set the bandwidth of the channel. Table II summarizes the filter designs.

TABLE II
Limiter Pre-Filter and Post-Filter Design

Filter	Type	Channel BW	
		25 kHz	5 kHz
Pre-limiter	Chebyshev	4 pole, 0.05 dB with 2 pole phase equalizer	4 pole, 0.05 dB with 2 pole phase equalizer
Post-limiter	Chebyshev	2 pole, 0.05 dB	4 pole, 0.05 dB

Although the actual channel uses a two-pole phase equalizer to linearize the overall phase response, we have found a sufficient model, used in this work, to be an unequalized 6-pole Chebyshev filter with cutoff frequencies of ± 12.5 kHz and having 0.05 dB of amplitude ripple. Performance comparisons in Section V show good agreement between this model and actual laboratory measurement.

We can illustrate the source of ISI by superimposing the power spectral density (PSD) of 64 kbps {4/15,5/16} CPM onto the 25 kHz channel filter mask shown in Fig. 2. We see the main lobe will be sharply attenuated by the filter inducing ISI.

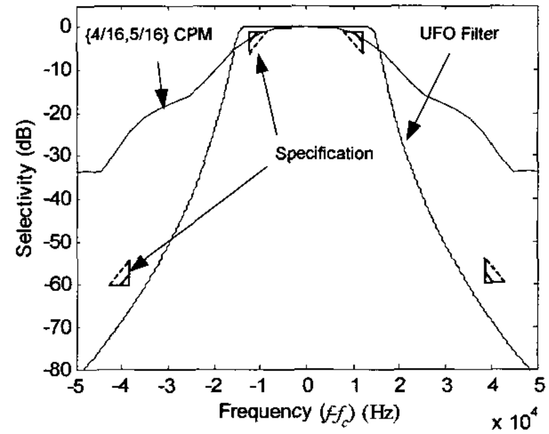

Fig. 2 - UFO 25 kHz Crystal Filter Selectivity and Quaternary {4/16,5/16} CPM PSD

From a time-domain point of view, Fig. 3 shows that the channel impulse response persists even after nine symbol periods adding a significant amount of memory to the system and making equalization a challenge.

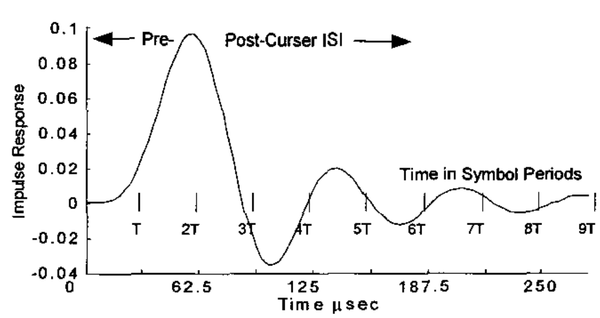

Fig. 3 - UFO 25 kHz Crystal Filter Impulse Response

II. DUAL-*H* CPM

Dual-*h* CPM was selected by JIEO/DISA because of three main qualities; constant envelope, compact power spectrum, and inherent coding gain. Being constant envelope makes it insensitive to the satellite's hard-limited channels that can suppress amplitude information and cause spectral regrowth in post-modulation filtered waveforms. Good coding gain is attributed to CPM's inherent trellis code structure. Analogous to the convolutional code in trellis-coded modulation (TCM) [4], CPM constrains its high order constellation by cycling between a pair of small, rational, modulation indices. This increases the minimum distance of the code by delaying the so-called first merge events that dominate bit error rates at medium to high signal-to-noise ratios (SNR) [5]. As rates increase, the coding gain is traded off by gear shifting through a decreasing set of modulation indices in order to maintain spectral occupancy. Table III lists the pairs of *rate-matched* indices that were selected for MIL-STD-188-181B [14] and the corresponding loss in coding gain. Loss was computed with respect to H = {12/16,13/16}, which is the strongest code achieving a probability of bit error $P_b(e) = 10^{-5}$ at 5.4 dB E_b/N_o.

TABLE III
Rate-Matched Indices

Data Rate (kbps)		Indices	Coding Gain
5 kHz Chan	25 kHz Chan	$H = (h_0, h_1)$	Loss (dB)
4.8	9.6, 19.2	{12/16,13/16}	--
6	--	{7/16,10/16}	-0.7
7.2	28.8, 32	{6/16,7/16}	-1.1
8, 9.6	38.4, 48	{5/16,6/16}	-1.8
--	56	{4/16,5/16}	-2.9

The dual-h CPM waveform has a full-response, $L=1$, rectangular frequency pulse shape (1REC) given by

$$g(t) \triangleq \begin{cases} 1/2LT, & 0 \le t < LT \\ 0, & \text{otherwise} \end{cases}. \qquad (1)$$

The complex low-pass 1REC CPM signal is [5]

$$\tilde{s}(t;\boldsymbol{\alpha}) = \sqrt{\frac{2E_s}{T}} \exp\left\{ j\frac{\pi}{T}\sum_{i=0}^{n} h_i \alpha_i t \right\}, \quad nT \le t < (n+1)T \qquad (2)$$

where E_s is the symbol energy, T is the symbol period, and $\boldsymbol{\alpha} = (\alpha_0, \alpha_1, \cdots, \alpha_N)$ is a length N vector of transmitted symbols chosen from a quaternary alphabet with bit to symbol mapping {00, 01, 10, 11} $\leftrightarrow$ {+3, +1, -1, -3}. It is implied that the starting phase is *zero* and the indices are cycled $n \bmod 2$.

After propagating through a hard-limited satellite channel the downlink-limited received signal can be modeled as

$$r(t) = \tilde{s}(t,\boldsymbol{\alpha}) * h(t) + n(t) \qquad (3)$$

where $(*)$ denotes convolution and $n(t)$ is a zero-mean complex Gaussian random process with noise spectral density N_o W/Hz. The uplink noise is assumed to be negligibly small and its performance impact is left for future analysis.

III. DFSE EQUALIZATION

First introduced in the early 90's by Duel-Hallen and Heegard [6], Eyuboglu and Qureshi [7], and specifically for CPM by Guren and Holte [3], the DFSE algorithm determines the minimum distance path through a reduced state trellis. It cancels *residual* ISI not incorporated by the partial trellis description using Per-Survivor [8] Decision Feedback (PSDF).

It has an MLSE-like structure that is appealing considering that the receiver for dual-h CPM is likely to already be MLSE-based and extending the concepts/signal processing is straightforward. However, unlike MLSE [9] where the trellis grows exponentially with respect to the channel memory, L_c, the complexity can be managed by varying the number of ISI symbols, L_e, incorporated into the trellis and the number of symbols, $L_{df} = L_c - L_e$, cancelled by PSDF. A notable case where $L_e = 0$ and $L_{df} = L_c$, known as Parallel Decision Feedback Equalization (PDFE), leaves the responsibility of removing ISI completely to decision feedback. Results in Section V indicate surprisingly good performance with only a linear increase to complexity. Of course different ratios can be considered making DFSE easily scalable, gradually trading performance for reduced complexity [3].

DFSE uses the Viterbi Algorithm (VA) [10] to efficiently determine the set of minimum distance paths through the reduced state trellis. For the n^{th} trellis stage, the i^{th} state is denoted by the m-tuple $\boldsymbol{\sigma}_n^{(i)} = (\theta_i, \gamma_i)$ in which θ_i is the *starting phase,* taking one of $2p$ discrete values from the set $\{0, \pi/p, 2\pi/p, \ldots, (2i-1)\pi/p\},\ i = 1\ldots p$, and $\gamma_i = \{\alpha_{n-1}, \alpha_{n-2}, \cdots, \alpha_{n-L_e}\}$ is the *correlated state vector* (c.s.v.) representing one of M^{L_e} distinct possible sequences of past channel inputs. The algorithm proceeds as follows. At each trellis stage, modified branch metrics for all valid transitions emanating from the i^{th} state and terminating in the j^{th} state, given by the set $A_n(i,j)$, are computed according to

$$\lambda_n(i,j)=\int_{nT}^{(n+1)T}\left|r(t)-b(t-nT;\theta_i,\gamma_i,\alpha_0)\cdots\right.$$
$$\left.-c(t-nT;\theta'_i,\gamma'_i)\right|^2 dt \tag{4}$$

in which

$$b(t;\theta_i,\gamma_i,\alpha_0)=\int_{-L_cT}^{T}\exp j\left\{\frac{\pi}{T}\sum_{k=-L_c}^{0}h_k\alpha_k\tau+\theta_i\right\}\cdots$$
$$\times\ \hat{g}(t-\tau)d\tau,\ 0\le t<T \tag{5}$$

is a locally generated partial reference signal at the channel output and

$$c(t;\theta'_i,\gamma'_i)=\int_{-L_c-L_{df}T}^{T}\exp j\left\{\frac{\pi}{T}\sum_{k=-L_c-L_{df}}^{-1}h_k\hat{\alpha}'_k\tau+\theta'_i\right\}\cdots$$
$$\times\ \hat{g}(t-\tau)d\tau,\ 0\le t<T \tag{6}$$

is an estimate of the residual channel ISI not described by (5) in which $\gamma'_i=\left\{0,\cdots,\hat{\alpha}'_{n-L_c-1},\hat{\alpha}'_{n-L_c-2},\cdots,\hat{\alpha}'_{n-L_c-L_{df}}\right\}$ is the most probable sequence of past channel inputs determined by *tracing back* through the surviving path that terminated into the j^{th} state and θ'_i is the appropriately time shifted starting phase given by

$$\theta'_i=\left[\theta_i-\pi\sum_{k=-L_c-L_{df}}^{-L_c-1}h_k\hat{\alpha}'_k\right]\bmod 2\pi\cdot \tag{7}$$

The symbol α_0 is determined by the one-to-one relationship with the i,j^{th} transitions, i.e. $\alpha_0\leftrightarrow A_n(i,j)\cdot$

Both (5) and (6) rely on precise knowledge of an unknown and slowly time-varying channel impulse response $\hat{g}(t),\ 0\le t<L_c$ that must be estimated by the receiver.

Fortunately, adaptive channel estimation has been extensively studied and suitable techniques, such as Recursive Least Squares (RLS) or Least Mean Square (LMS), exist; however, in this work it was assumed $\hat{g}(t)$ is known a priori. The performance impacts of these practical channel estimation techniques are currently under investigation, but, for now, the results presented here demonstrate the potential of DFSE under somewhat ideal conditions.

For all states $1\le j\le S=2pM^{L_c}$, the surviving path that minimizes the accumulated path metric

$$\Gamma^j_{n+1}=\min_{i\in A_n(i,j)}\left\{\Gamma^i_n+\lambda_n(i,j)\right\} \tag{8}$$

is stored keeping the associated state and symbol histories. After a sufficiently long delay, d, usually 15 to 20 symbols, bit decisions $\hat{\alpha}_{n-d}$ are determined by tracing back through the state history starting from the state with the minimum path metric.

Fig. 4 illustrates one realization of the DFSE equalizer that would work well with newer MIL-STD-188-181B [14] terminals. As suggested by Guren, et al. [3], discrete time versions of (5) and (6) can be tabulated, updating them only at the beginning of a new transmission or when refined channel estimates are available.

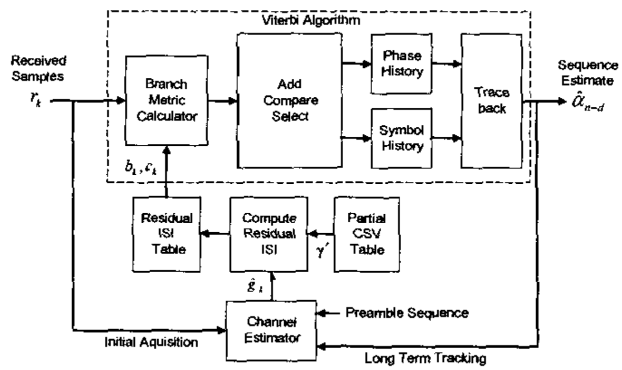

Fig. 4 - DFSE Receiver Block Diagram

IV. SIMULATION RESULTS

Computer simulations have been carried out based on the downlink-limited communication system model shown in Fig. 5.

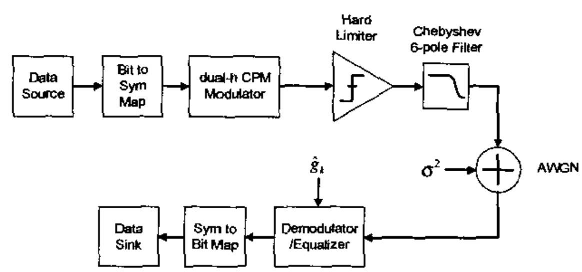

Fig. 5 - Simulation Model of Dual-*h* CPM over UFO

It is assumed that the impulse response and carrier phase are known. Symbol timing is derived using the mean-square-error estimation technique presented in [11] and the channel is represented in complex baseband equivalent form [12]. The noise variance was calculated using [13]

$$\sigma^2=\frac{A^2N_s}{2(E_b/N_0)\log_2(M)} \tag{9}$$

where N_s is the number of samples per symbol and A is the signal amplitude. Waveform level simulations used six samples per symbol and 100,000 symbols per transmission. Runs were allowed to loop until 100 symbol errors were collected at each data point.

Fig. 6 shows the simulation results for the clear-air case, bypassing the channel, and the 25 kHz band-limited case. To validate our model, we compared simulated results with laboratory measurements. The laboratory BER results were collected from a Multiple Output SATCOM Terminal (MOST) running 64 kbps dual-h CPM over a 25 kHz UFO RF channel emulator that contained the actual analog filter circuitry. Notice that the simulated channel introduces the same error flooring effect as the emulated channel. It is also clear that, without equalization, error performance will never reach the specified $P_b(e) = 10^{-5}$ at ≤ 13.44 dB [22]. For reference, the clear air and simulated satellite channel performance results will be repeated in subsequent BER plots.

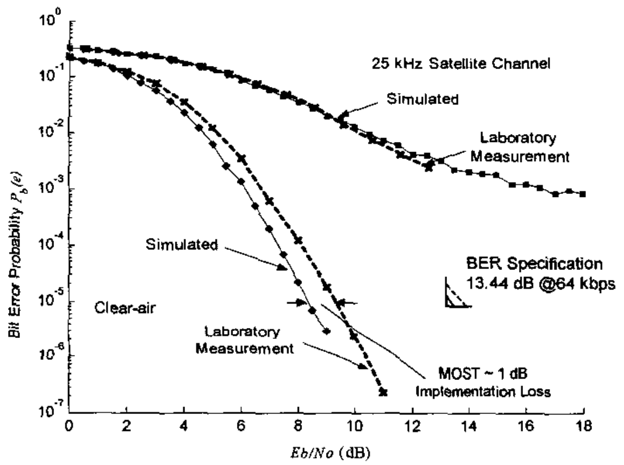

Fig. 6 - BER in Clear-air vs. a 25 kHz UFO Channel

Fig. 7 shows the results using an MLSE equalizer [9] with channel equalizer memory, $L_c = L_e = 1$, 2, and 3. The complexity ranges from 32 states for no equalization ($L_e = 0$) to 128, 512, and then 2048 states, respectively. We observe that, after two symbols of channel equalization, performance is quite good and meets the specifications given in MIL-STD-188-181B [22]. However, computing a 512-state trellis at 64 kbps would require a significant signal processing upgrade even to the latest terminals.

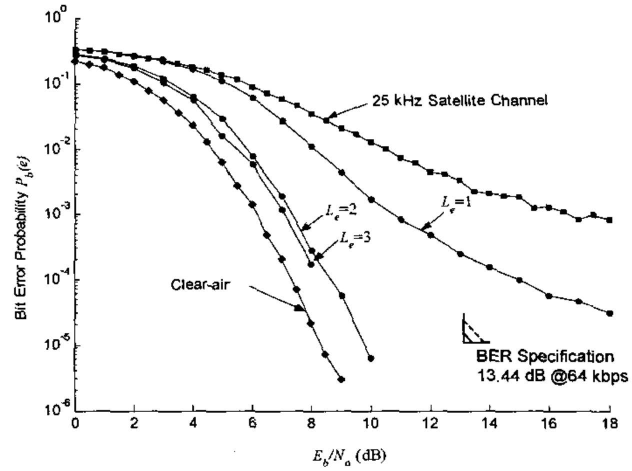

Fig. 7 - BER of MLSE Equalization at $L_e = 1$, 2, and 3 Symbols of Channel Memory

Fig. 8 shows the results of PDFE for varying amounts of decision feedback. Notice that the error floor introduced by ISI is mitigated without expanding the trellis. The overall increase in complexity can be kept small if decision feedback terms are computed and stored only once per reception.

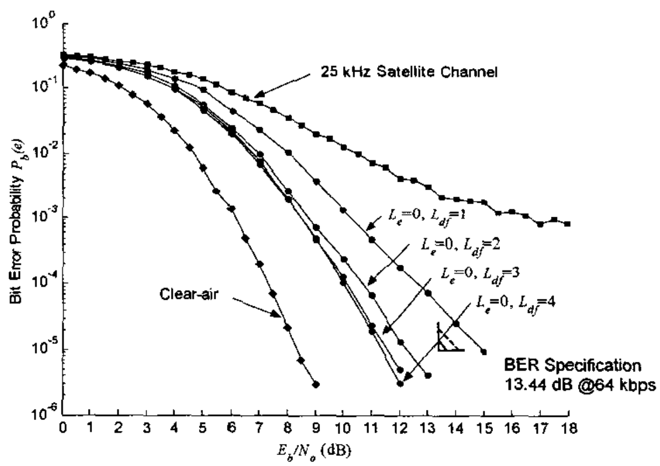

Fig. 8 - BER of DFSE Equalization using PDFE

Of course, by incorporating more memory into the trellis better performance can be achieved. Fig. 9 shows that if $L_e = 1$ and $L_{df} = 2$ or 3, performance is only slightly degraded from the $L_e = 3$ MLSE equalizer. With only a 4× increase to trellis complexity, this is a computationally tractable approach with very good performance.

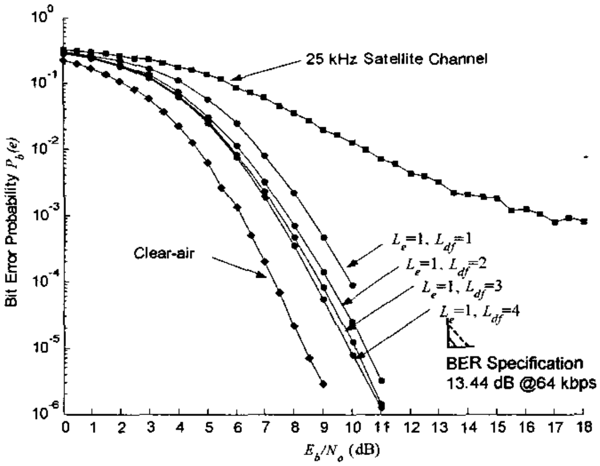

Fig. 9 - BER of DFSE Equalization with One Symbol of Trellis Memory and 1 to 4 Symbols of DF

VI. CONCLUSION

In this paper we applied Decision Feedback Sequence Estimation (DFSE) [3] in order to equalize a 64 kbit/s dual-*h* quaternary full response CPM waveform transmitted over a representative 25 kHz UFO channel. Computer simulation results indicate DFSE can effectively mitigate ISI with a manageable level of complexity offering an attractive approach to higher data rates.

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