EFTS RECEIVER WITH IMPROVED PERFORMANCE

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ABSTRACT

The PAM representation was used to formulate a reduced-complexity detector for the Enhanced Flight Termination System (EFTS) whose performance is 5.6 dB better than limiter-discriminator detection when no phase noise is present and 3.4 dB better in the presence of expected phase noise in EFTS.

KEY WORDS

CPM, PAM Decomposition, Flight Termination, Range Safety

INTRODUCTION

Most range safety programs use a Flight Termination System (FTS) to bring stray airborne test vehicles to crash at a preselected location, accomplished by bringing the vehicle into a state of zero lift and zero thrust. Methods used to meet this goal include parachute deployment and detonation of explosive charges that destroy the test vehicle[1]. The test range initiates flight termination by sending a radio signal on a dedicated channel to the stray test vehicle. The first FTS, brought into use in the 1950s, modulates an FM carrier with different frequency audio tones [2, 3, 4]. The aircraft completes the "arm" and "terminate" commands after the arrival of a predefined sequence of these tones from the ground-based receiver [2]. The modified high-alphabet system [5] adds a security feature by using a predefined sequence of tone *pairs* to encode the commands.

With the increase in flight altitudes came an incident where the terminate signal sent by one test range inadvertently terminated the flight of a vehicle at a nearby test range [1, 6]. In response to this incident, the Range Safety Group of the Range Commanders Council created a committee in April 2000 to enumerate the requirements of the next generation FTS that would search for techniques to deal with the aforementioned situation. In January 2002, this group chose bi-phase pulse-coded modulation/frequency modulation (PCM/FM) as the modulation for the next generation FTS, named Enhanced Flight Termination System (EFTS) [7]. The justification behind this choice of modulation included the important reason that the ground-based FTS transmitter (i.e., the current hardware infrastructure) required an AC-coupled input to the FM modulator [8]. The EFTS standard chose a more “digital” route for the EFTS waveform which opened the door for better security (3DES encryption) and a higher level of reliability (Reed-Solomon error control coding).

The selection of bi-phase PCM/FM included the assumption that the airborne vehicle would use a limiter-discriminator for a demodulator because the limiter-discriminator has a history of reliability in aeronautical applications. Unfortunately, this limiter-discriminator detector has a much poorer BER than an optimal receiver would have. The tradeoff for using an optimal receiver is increased complexity. There are two main sources of compleity. First, an optimal receiver requires matched filters and a maximum-likelihood sequence estimator. Second, the optimal receiver requires full phase coherency which is difficult to attain in the presence of excessive amounts of phase noise resulting from the high levels of shock and vibration typical in airborne vehicles that use a FTS.

In this paper, a reduced complexity receiver architecture is described for use with EFTS, which receiver uses the PAM decomposition of continuous phase modulation (CPM), of which PCM/FM is a special case. The simplified receiver limits the inevitable increase in BER relative to that of optimal while allowing a receiver to be used that is less complex than an optimal receiver. Indeed, the simplfied receiver performs 5.6 dB better than a limiter-discriminator detector when no phase noise is present. However, in the presence of expected phase noise present in EFTS applications, the simplified receiver still performs 3.4 dB better than the limiter-discriminator.

The PAM decomposition for EFTS is presented in the first section and is used to build a simplified PAM-based receiver in the second section.

PAM REPRESENTATION OF WEAK CPM

The complex baseband representation of a binary CPM signal may be expressed as [9]

$$s(t) = \exp\left\{ j\phi\left(t; \boldsymbol{\alpha}\right)\right\} \tag{1}$$

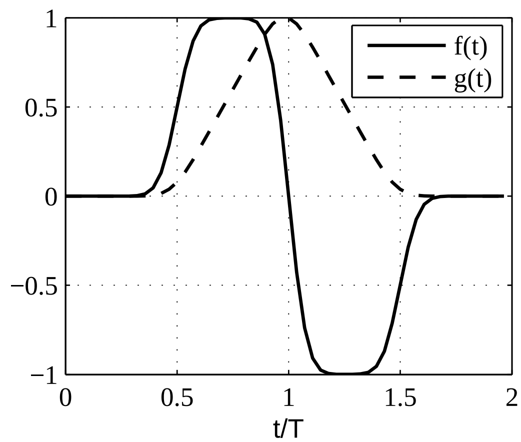

Figure 1: The frequency and phase pulses for the EFTS weak CPM. (Note: the pulses have been normalized to unit peak amplitude for display.)

where

$$\boldsymbol{\alpha} = \ldots, \alpha(0), \alpha(1), \ldots \tag{2}$$

$$\phi\left(t; \boldsymbol{\alpha}\right) = 4\pi f_d T \sum_{n=-\infty}^{\infty} \alpha(n) g(t - nT) \tag{3}$$

and where T is the bit time, $\alpha(n) \in \{-1, +1\}$ is the n-th binary symbol, and $g(t)$ is the phase pulse. Also, f_d is the peak frequency deviation, which is specified in the EFTS standard to be 60 kHz; the frequency pulse is normalized to ensure this target peak frequency deviation is not exceeded.

The EFTS phase pulse is the time-integral of a frequency pulse $f(t)$ which is zero outside the interval $0 \leq t \leq LT$:

$$g(t) = \int_0^t f(x)dx. \tag{4}$$

Since EFTS uses a bi-phase frequency pulse, the phase pulse is time-limited as follows:

$$g(t) = \begin{cases} 0 & t \leq 0 \\ 0 & t > LT \end{cases}, \tag{5}$$

which means that bi-phase PCM/FM is one form of weak CPM. The EFTS standard dicates that the bi-phase frequency pulse shall be filtered by a 4^{th}-order low-pass Bessel filter with a cutoff frequency of 15 kHz. This filter was simulated using a Gaussian low-pass filter with the same cutoff frequency; this filter choice made the mathematics reasonably tractable. The frequency-pulse/phase-pulse pair is shown in Figure 1.

The PAM representation of binary CPM was introduced by Laurent [10]. (His results were later extended to M-ary CPM by Mengali [11], to multi-h CPM by Perrins and Rice [12], and to weak CPM by the authors [13].) PAM representations of CPM transform the non-linear CPM into a linear PAM modulation by pushing the non-linearity out of the modulation into the data symbols. The most important reason for making this representation is that it allows the design of a demodulator to perform sub-optimal detection as explained later.

In its standard form, CPM uses a frequency pulse which is normalized to to have an area of 1/2 [9]. Since EFTS uses a bi-phase frequency pulse, which has an area of zero, the phase trellis possesses properties that do not allow a direct application of Laurent's technique to produce the equivalent PAM representation. Taking a different approach yields a PAM representation of EFTS CPM, expressed in the following equation

$$s(t) = \sum_{N=-\infty}^{\infty} \left(e^{j\pi h \alpha(n)} q_0(t-NT) + e^{j\pi h[\alpha(n)+\alpha(n-1)]} q_1(t-NT) + q_2(t-NT) \right), \qquad (6)$$

where the pulses $q_K(t)$ are modulated by functions of the data symbols which functions multiply each pulse. (Equation (6) shows that $q_2(t)$ is independent of the data-symbols.) Note that the lengths of the pulses are $2T$, T, and T for $q_0(t)$, $q_1(t)$, and $q_2(t)$, respectively, as can be seen in a plot of these pulses shown in Figure 2; the spectra of the three pulses is illustrated in Figure 3.

REDUCED COMPLEXITY DETECTOR FOR EFTS CPM BASED ON THE PAM REPRESENTATION

An optimum PAM-based detector uses all of the symbol-dependent PAM pulses in a bank of matched filters, whereas a reduced complexity PAM-based detector for CPM neglects one or more of the symbol-dependent PAM pulses in its bank of matched filters. The PAM representation of EFTS, found in (6) shows that $q_2(t)$ can be neglected with no reduction in performance since this pulse is symbol-independent. Thus, in simplifying the receiver, the only pulse to negelct is $q_1(t)$, keeping only the most energetic PAM pulse, $q_0(t)$. An approach to building a simplified receiver using the PAM representation of weak CPM (CPM using the phase pulse defined in (5)) has also been studied by the authors [13]; the same approach is applied here to build a simplified PAM-based receiver for EFTS.

The simplified EFTS detector and the limiter-discriminator are shown in Figure 4. In this case, the signal model used by the detector consists of only the first term in (6). As can be seen, only a single matched filter, corresponding to the main PAM pulse, $q_0(t)$, is used in the simplified detector. The

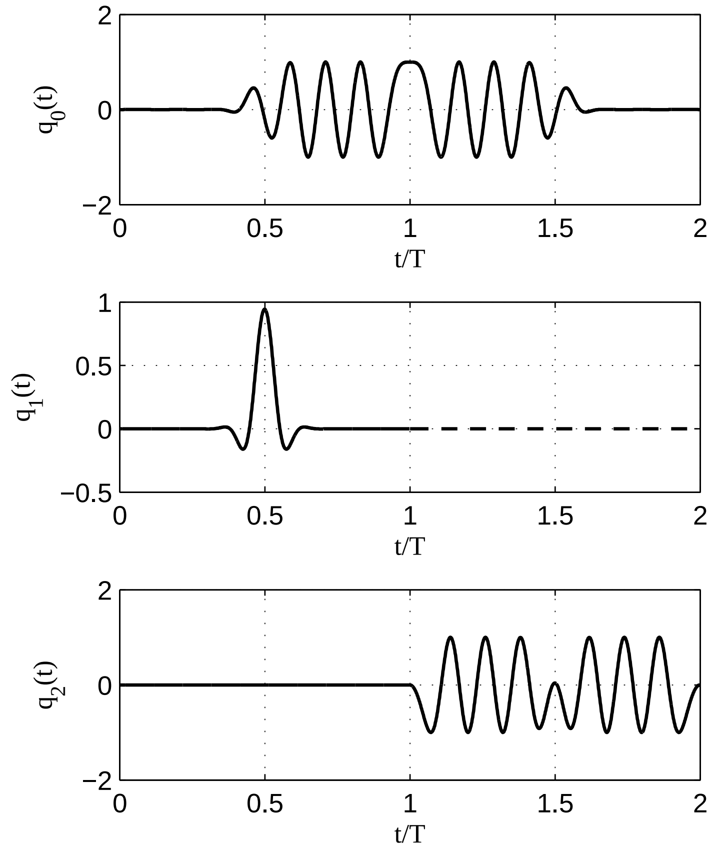

Figure 2: The three PAM pulses $q_0(t)$, $q_1(t)$, and $q_2(t)$ in the PAM representation of the EFTS waveform.

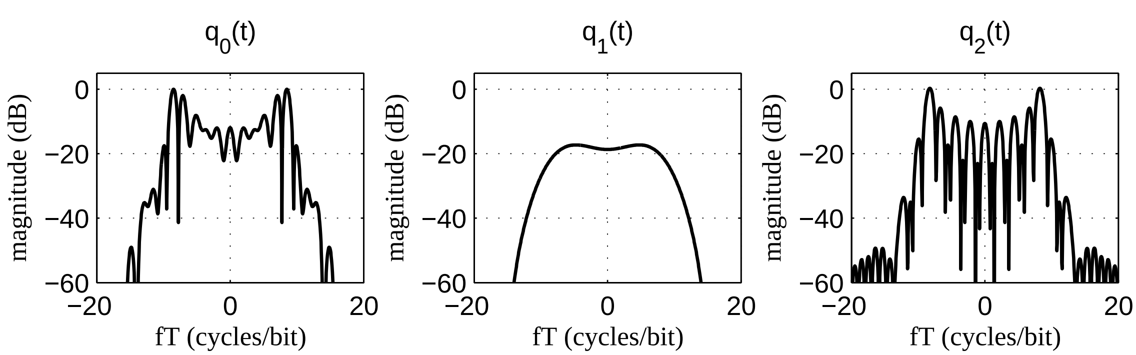

Figure 3: The spectra of the PAM pulses shown in Figure 2.

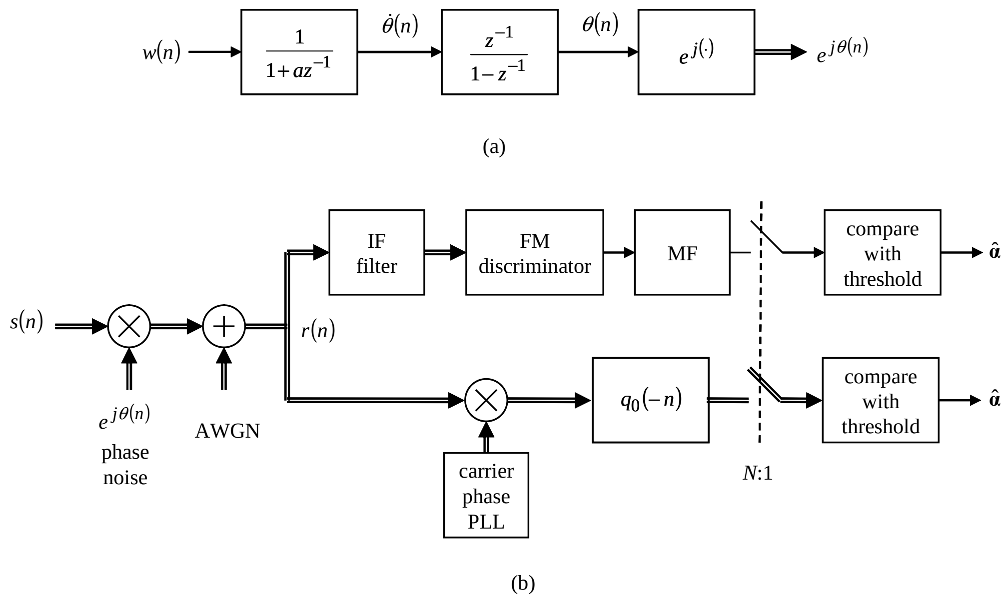

Figure 4: This figure shows the simulation setup used to simulate the receivers (in discrete time): (a) Phase-noise generator, injected with white Gaussian noise ($w(n)$); (b) PAM-based simplified receiver and FM limiter-discriminator. Note that both receivers were simulated with and without the phase noise included.

decision can be based solely on the current output of this matched filter since the dependence on $\alpha(n-1)$ suggested by (6) was eliminated when $q_1(t)$ was removed. Thus, the decision rule is

$$\hat{\alpha}(n) = \begin{cases} +1 & \text{Re}\left\{x_0(nT)e^{-j\pi h}\right\} \geq \text{Re}\left\{x_0(nT)e^{j\pi h}\right\} \\ -1 & \text{otherwise} \end{cases} \tag{7}$$

Using $h = 50/3$, the latter decision rule can easily be reduced to

$$\hat{\alpha}(n) = \begin{cases} +1 & \text{Im}\left\{x_0(nT)\right\} \geq 0 \\ -1 & \text{otherwise} \end{cases}. \tag{8}$$

Thus, in the detector, the matched filtering is followed by a comparator to make a final decision on each bit.

The performance of this detector, compared to a limiter-discriminator, in an additive white Gaussian noise (AWGN) channel and in an AWGN channel with phase noise is illustrated in Figure 5 where it can be seen that the simplified PAM-based detector performs, with no phase noise present, 5.6 dB better than the limiter-discriminator at a P_b of 10^{-5}. When phase noise is injected, the simplified receiver still performs 3.4 dB better than the limiter-discriminator at a P_b of 10^{-5}; the Appendix discusses the setup for the phase noise simulation. There is only one BER curve for the limiter-discriminator, since it was not affected by the phase noise.

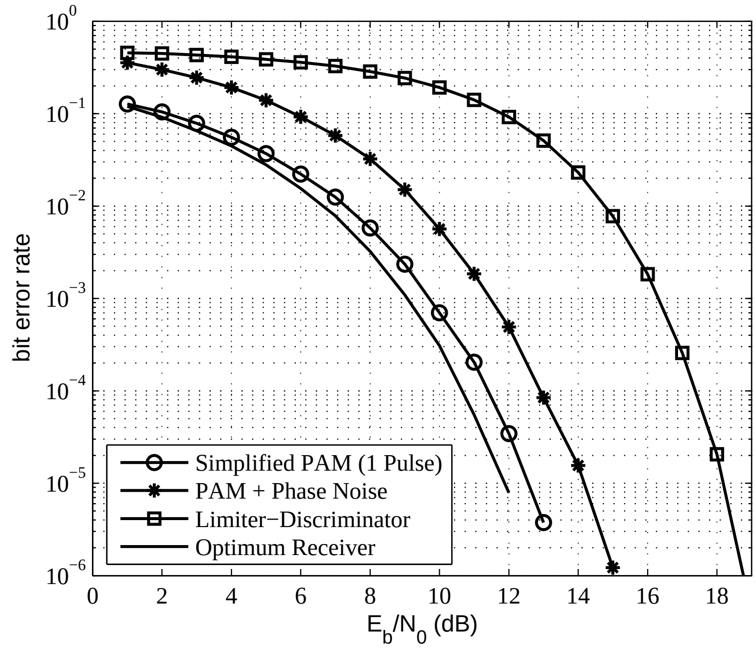

Figure 5: Bit-error rate simulations for the two detection methods for the EFTS modulation, with and without phase noise present. The limiter-discriminator simulation, which used an 8-th Chebychev IF filter with cut-off frequency $0.75/T$ and maximum ripple-width of 0.5 dB, was not affected by phase noise. The BER curve for an optimal EFTS receiver in the AWGN channel is included for comparison.

CONCLUSION

A method for computing the PAM representation for EFTS CPM was used to formulate a reduced-complexity detector for the EFTS modulation that was shown to perform 3.4 dB better than a limiter-discriminator in the presence of phase noise levels expected in EFTS applications. This performance comes at the cost of a carrier phase PLL.

APPENDIX

The phase noise that is being simulated is due to the large amounts of vibrations found experimentally to be present in flight vehicles using the EFTS. This phase noise has an RMS frequency deviation of 4 kHz and an equivalent noise bandwidth of 2.5 kHz [14].

The following simple recursive filter, where a was chosen so that the phase noise had the desired RMS frequency deviation and equivalent noise bandwidth,

$$H_{\mathrm{PN}}(z) = \frac{1}{1 + az^{-1}}, \tag{9}$$

was used to generate the phase noise as

$$\dot{\theta}(n) = h_{\text{PN}}(n) \star z(n) \tag{10}$$

where $\star$ denotes convolution. Also, $h_{\text{PN}}(n)$ is the time-domain version of the filter in (9), $\dot{\theta}(n)$ is the simulated frequency noise, and $z(n)$ is zero-mean white Gaussian noise. The filter output was then sent through an integrator to produce the phase noise, $\theta(n)$, which noise was then added to the phase of the signal as follows:

$$r(n) = s(n)e^{j\theta(n)} + w(n) \tag{11}$$

with

$$\theta(n) \approx T_s \sum \dot{\theta}(n)\, dt, \tag{12}$$

where $w(n)$ is AWGN and T_s is the sample time.

The simulation in a simple AWGN channel assumed ideal phase coherency, so a phase-lock loop (PLL) was not actually built into the simulation. A PLL must be included to simulate the effect of PLL tracking errors in the presence of phase noise. This PLL simulates the effect of a PLL due to transients while assuming phase-lock is achieved.

The phase noise was added to the phase of the received signal as previously noted. The phase noise was also filtered by a filter that was equivalent to the frequency response of a PLL, in order to simulate the effect a PLL would have on the phase noise. The PLL was chosen to have a damping factor, $\zeta = 1$, and a noise-bandwidth, $B_n = 18$ Hz. The following equation was used for the equivalent transfer function of the PLL:

$$H_{\text{PLL}}(z) = \frac{(K_1 + K_2)z^{-1} - K_1 z^{-2}}{1 + (K_1 + K_2 - 2)z^{-1} + (1 - K_1)z^{-2}} \tag{13}$$

with

$$K_1 = \frac{\frac{4\zeta}{N}\left(\frac{B_n T_b}{\zeta + \frac{1}{4\zeta}}\right)}{1 + \frac{2\zeta}{N}\left(\frac{B_n T_b}{\zeta + \frac{1}{4\zeta}}\right) + \left(\frac{B_n T_b}{N\left(\zeta + \frac{1}{4\zeta}\right)}\right)^2} \tag{14}$$

$$K_2 = \frac{\frac{4}{N^2}\left(\frac{B_n T_b}{\zeta + \frac{1}{4\zeta}}\right)^2}{1 + \frac{2\zeta}{N}\left(\frac{B_n T_b}{\zeta + \frac{1}{4\zeta}}\right) + \left(\frac{B_n T_b}{N\left(\zeta + \frac{1}{4\zeta}\right)}\right)^2} \tag{15}$$

where N is the number of samples per bit used in the simulation and $T_b = \frac{1}{R_b}$ is the bit time.

The phase noise could be added to the phase of the received signal, as well as filtered by the PLL-equivalent filter, the results of which being subtracted off the phase of the signal, all before the signal is match-filtered. However, in this configuration, the noise bandwidth of the PLL could be cranked up to eliminate the phase noise altogether. Therefore, after the phase noise was added to the received signal but before it was filtered by the PLL, it was modified by adding the quadrature component of the AWGN ($w(n)$ in (11)) in order to get a more realistic result as shown in the following equation:

$$\tilde{p}(n) = (\theta(n) + w_Q(n)) \star h_{\mathrm{PLL}}(n) \tag{16}$$

where $h_{\mathrm{PLL}}(n)$ is the time-domain version of the filter in (13) and $w_Q(n) = \mathrm{Im}\{w(n)\}$. Thus the filtered phase noise will cancel some of the phase noise, simulating the imperfect yet still effective operation of a PLL through the following equation:

$$r(n) = s(n)e^{j\theta(n)}e^{-j\tilde{p}}(n) + w(n). \tag{17}$$

It should be noted that all of this assumes a priori that a PLL could be designed that functions in the presence of such severe phase noise as that found in EFTS applications.

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