1. Introduction

Implementation

The paper fundamentally describes a novel method for enhanced communication (IMC) with applications in various fields. The IMC approach utilizes advanced techniques to improve data transmission and reception. This section provides an overview of the implementation details and highlights the key components of the IMC method.

Abstract

The novel method for enhanced communication (IMC) introduces significant advancements in data transmission and reception. This paper presents a detailed explanation of the IMC approach, including its theoretical foundations and practical implications. The IMC method offers enhanced performance and reliability, making it a promising solution for future communication systems.
suppressed carriers are used and since the 
load loop S/F is a non-linear function of the
and a relatively high bit rate, significant data loss must be avoided. The S/N ratio is 
required to be greater than 10 dB to 0.6 , the bit error rate is less than 10^-6. 
These conditions are met for the low EMI mission with 2 symbol per second (64k/s) and for the high EMI mission with 128 symbols per second (512k/s). 

The applications of the MIL extend to control missions to provide command, data 
link and navigation information. 

In order to operate the communication link below the threshold of control receivers, for example, 

consider NASA's 5th Generation Computer. The key to its weight saving is to use a

distributed computing architecture that enables a multi-process implementation. The 
computational bandwidth is equal to the bandwidth of a single wave substituting in 
reduced individuality and (d) interaction between symbols lost during reception and reproduction, (e) the minimized receiving

The ability to use some information about the signal and control navigation in the 
airplane to process the signal to reduce acquisition time and enable reproduction.
symbol SNR, significantly reducing loss results in the loop. Additionally, since the downlink symbol rate can be programmed to take advantage of the antenna apertures on the ground, several data rate changes can occur in a pass or a single day, each resulting in data loss due to resynchronization in the ground receiver. At present, data lost during acquisition or cycle slips is not recoverable because existing receivers in NASA’s Deep Space Network (DSN) process III( signal in real-time and do not buffer or store the received signal or reprocessing.

The Buffered Telemetry Demodulator (BTD) is a coherent receiving system which overcomes these problems by recording the signal and then processing and reprocessing the samples. Once acquisition occurs, the BTD operates on the past and current estimates of the signal state (such as phase, frequency, and drift rate) to basically go back in time and recover those symbols lost during acquisition. This is referred to as smoothing where future data is used to provide an estimate in the past. The BTD is designed mainly for 10% symbol rate applications and can be implemented on a general purpose workstation. In order to reduce the computational requirement, the signal is reconstructed on a subcarrier harmonic basis, thus rendering the BTD independent of the subcarrier frequency. This results in a computational throughput requirement based on the symbol rate rather than on the bandwidth of the transmitted signal. Consequently, when antenna arraying is employed, only a fraction of the total signal bandwidth need to be transmitted between antennas, resulting in a significant reduction in transmission requirements.

Section 2 of this paper briefly reviews the existing coherent receiver which demodulates the downlink signal in real time. Sections 3 and 4 describe the recording and demodulation functions of the BTD. It is shown that the BTD performance exceeds that of existing receivers since no data is lost due to reprocessing. The BTD implementation on various workstations is evaluated in section 5, and smoothing strategies are discussed in section 6.
2 A Review of Current Receivers

In deep space communications, baseband symbols are typically first modulated onto a square-wave subcarrier and modulated onto a sine-wave carrier. This allows transmission of a residual carrier component whose frequency may not coincide with the data spectrum. Consequently, at the receiving end the deep space signal is demodulated using a carrier, subcarrier, and symbol synchronizer loop. This section briefly reviews the Advanced Receiver (ARX) [1] carrier and subcarrier loops so that they can be compared to the carrier and subcarrier loops of the BTD. The ARX is a receiving system that has been developed to digitally demodulate and process signals from open spacecraft. It is a broad band system for the Block V receiver, which will eventually replace various receivers currently used in the DSN [2]. The main difference between the ARX and BTD is that the ARX carrier and subcarrier loops process the received signal after it has been open loop down converted to an intermediate frequency (IF) function, the BTD loops process the received signal after it has been open loop down converted to baseband and subcarrier harmonic basis. All the other receiving functions, e.g., symbol synchronization, symbol detection, lock detection, etc., are implemented the same way in both the ARX and BTD. Consequently, they are not discussed in this article. Instead, the interested reader is referred to [1] for a complete review of these functions.

A block diagram of the ARX carrier and subcarrier loops is shown in Fig. 1. The received signal, which is assumed to be a typical (1 + 1) space signal, after down conversion to an appropriate IF can be represented as [2]

\[ r(t) = \sqrt{2P} \cos(\omega) \sin[\Psi_c(t)] + \sqrt{2P} \sin(\omega) d(t) \sin[\sin[\Psi_{sb}(t)]] \cos[\Psi_c(t)] : n(t) \]  

where the first term represents the residual carrier component and the second term represents the data modulation. The total power \( P \) in Watts (W) is (ii \( \omega \)) between the residual carrier and data by controlling the modulation index, \( \Lambda \). Specifically, the carrier power \( P_c = P \cos^2(\Lambda) \) and the data power \( P_d = P \sin^2(\Lambda) \), respectively. The quantity \( n(t) \) is the
total carrier phase and \( \Psi_{su}(t) \) is the total subcarrier phase, both in radians. The data, \( d(t) \), is given by

\[
d(t) = \sum_{k=-\infty}^{\infty} a_k p(t - kT) \tag{2}
\]

where \( p(t) \) is the unit-power square pulse limited to 1 second and \( \{a_k\} \) represents the independent and equally likely binary (±1) symbols. The noise process, \( n(t) \), is modeled as bandpass with a flat one-sided power spectral density \( (1 \, \text{s} \, \text{Hz}) \) level equal to \( N_0 \, \text{WHz} \). By expressing the square wave subcarrier as an infinite sum, i.e.,

\[
\sin[\Psi_{su}(t)] = \sum_{n=0}^{\infty} \frac{1}{n!} \sin[j \Psi_{su}(t)]
\]

the digitized signal which is sampled at rate \( f_s = 1/T_s \) can be written as

\[
v(kT_s) = \sqrt{2} P_s \sin[\Psi_c(kT_s)] + \sqrt{2} P_d(kT_s) \left( \sum_{n=1}^{\infty} \frac{1}{n!} \sin[j \Psi_{su}(kT_s)] \right) \cos[\Psi_c(kT_s)] + n(kT_s) \tag{4}
\]

where only harmonics up to the \( J^{th} \) subcarrier harmonic \((J \text{ is an odd integer})\) are assumed to be present at the A/D converter output in Fig. 2. Samples of the bandpass noise \( n(kT_s) \) can be expressed as

\[
n(kT_s) = \sqrt{2} n_1(kT_s) \cos[\Psi_c(kT_s)] - \sqrt{2} n_2(kT_s) \sin[\Psi_c(kT_s)] \tag{5}
\]

where the random variables \( n_1(kT_s) \) and \( n_2(kT_s) \) are samples of the baseband noise processes \( n_1(t) \) and \( n_2(t) \) which are assumed to be statistically independent. The Gaussian noise processes with a flat one-sided \( (1 \, \text{s} \, \text{Hz}) \) level equal to \( N_0 \, \text{WHz} \) within a one-sided bandwidth equal to \( 1/(2T_s) \, \text{Hz} \). As a result, \( n_1(kT_s) \) and \( n_2(kT_s) \) are zero mean with variance \( \sigma^2_{n_1} = \sigma^2_{n_2} = N_0/(2T_s) \). When appropriate, the time index \( kT_s \) will be omitted from now on to allow a simpler notation.

The residual carrier locking loop control signal, \( U_q \) in Fig. 1, is derived by mixing the digitized \( \Pi \) signal with its carrier quadrature component, low-pass filtering the mixer output,
and then accumulating the filtered samples over the loop update interval. Mathematically,

\[ U_q = \sqrt{P_c \sin(\phi_c)} \cdot n U_q \tag{6} \]

where \( \phi_c \) is the carrier phase error, \( n U_q \) is a zero mean Gaussian random variable with variance \( N_0/(2T_u) \), and \( T_u = M_1 T_s \) is the loop update interval in seconds. Note that the data has been neglected in writing (6), since it is assumed that the accumulator in the carrier quadrature arm averages over several subcarrier cycles (i.e., \( f_s T_u \gg 1 \)) so that the data sum is approximately zero. It will be shown in section 4 that the BTJ carrier loop control signal has the same signal power and noise statistics as \( \omega \), and therefore, both loops will have the same tracking performance.

Next, let’s consider the subcarrier tracking loop. As shown in Fig. 1, the input to the subcarrier loop is the quadrature component of the residual carrier. The subcarrier control signal is obtained by forming the product \( Z_i Z_{q,su} \) where \( Z_i \) is the output of the subcarrier in-phase arm and \( Z_{q,su} \) is the output of the subcarrier quadrature arm. Note that the subscript ‘su’ in \( Z_{q,su} \) is used to distinguish it from \( Z_{q,su} \), the quadrature arm of the suppressed carrier loop. The signals \( Z_i \) and \( Z_{q,su} \) obtained after respectively mixing the residual carrier quadrature component with the subcarrier in-phase and quadrature references and then accumulating these samples over a symbol duration (perfect symbol synchronization assumed), are given as [1]

\[ Z_i = \sqrt{P_d a_1 F_I(\phi_{su}) \cos(\phi_c)} \cdot n Z_i \tag{7} \]

and

\[ Z_{q,su} = \sqrt{P_d a_1 F_Q(\phi_{su}) \cos(\phi_c)} \cdot n Z_{q,su} \tag{8} \]

where \( \phi_{su} \) is the subcarrier phase error and the functions, \( F_I(\phi_{su}) \) and \( F_Q(\phi_{su}) \), are periodic functions of \( \phi_{su} \) (period \( 2\pi \)) given in the principal phase interval \( [-\pi, \pi] \) as follows [1]

\[ F_I(\phi_{su}) = 1 - 2 \left| \frac{\phi_{su}}{\pi} \right|, \left| \phi_{su} \right| \leq \pi \tag{9} \]
and

\[
P_Q(\phi_{su}) = \begin{cases} 
(2/\pi)[\%3, (\phi_{su}) & (\phi_{su}) < (n/2)W_{su} \\
\text{sgn}(\phi_{su})W_{su} & (n/2)W_{su} \leq |\phi_{su}| < n(1 - W_{su}/2) \\
2\text{sgn}(\phi_{su}) \cdot (2/n)\phi_{su} & (1 - W_{su}/2) \leq |\phi_{su}| \leq n 
\end{cases}
\]  

(10)

The parameter \(W_{su}\), ill fractions of cycles, is the width of the subcarrier window (\(W_{su} < 1\) and \(W_{su} > 1\) corresponds to no window). It is used in the quadrature arm to improve the square-wave tracking performance [3]. The noise \(n_{Z_i}\) and \(n_{Z_{q,scn}}\) in (7) and (8) are independent with respective variances \(N_0/(2T)\) and \(N_0 W_{su} / (2T)\) where \(T\) denotes symbol time. It will be shown in section 4 that the in-phase and quadrature accumulator outputs have the same signal and noise statistics as (7) and (8). Consequently, both subcarrier loops will have the same control signal and therefore, the same tracking performance.

Finally, let's consider the suppressed carrier with subcarrier case (\(A = 90^\circ\) in (1)). In this scenario, the suppressed carrier loop control signal is obtained by forming the product \(Z_{s}Z_{q,scn}\) as shown in Fig. 1. The signal \(Z_{s}\), which was shown to be the subcarrier in-phase component, now becomes the suppressed carrier in-phase component because it is already proportional to \(\cos(\phi_c)\) and \(F_1(\phi_{su})\) as indicated by (7). The suppressed carrier quadrature component \(Z_{q,scn}\) is generated by mixing the digitized IF signal with the quadrature carrier reference and the in-phase subcarrier reference, and then accumulating the resulting samples \(0/(T)\) a symbol duration. Hence, the suppressed carrier in-phase and quadrature signals are given as [1]

\[
Z_i = \sqrt{P_d}a_1F_i(\phi_{su})\cos(\phi_c) \cdot n_{Z_i} 
\]  

(11)

and

\[
Z_{q,scn} = \sqrt{P_d}a_1F_i(\phi_{su})\sin(\phi_c) \cdot n_{Z_{q,scn}} 
\]  

(12)

where \(n_{Z_i}\) and \(n_{Z_{q,scn}}\) are independent with variance \(N_0/(2T)\). It will be shown in section 4 that the in-phase and quadrature signals of the BTDCD suppressed carrier loop have the same statistics as in (11) and (12). Hence, both suppressed carrier loops will have the same error
signal and tracking performance.

3 Baseband Recording

As indicated earlier, the BT operation processes the received signal after it has been open-loop (tou'llc oil'teJt(1° to baseband, and subcarrier harmonics basis and recorded on tape. This section describes two methods for recording the received signal at baseband. The first recording scheme, depicted in Fig. 2a, distributes the sampled IF signal into 2 carrier channels and 4n subcarrier channels where n denotes the number of subcarrier harmonics present at the A/D output and is related to L (the highest recorded subcarrier harmonic) through \( L = 2n - 1 \). Recording of the carrier requires two channels because baseband demodulation requires both in-phase and quadrature samples. Similarly, recording of each subcarrier harmonic requires 4 channels, 2 channels (1 in-phase subcarrier channel and 1 quadrature subcarrier channel) per carrier channel (in-phase and quadrature). Table 1a depicts the loss in data power as a function of the highest recorded subcarrier harmonic L.

Each band of the sampled IF signal is recorded at baseband after mixing it with tones tuned to its predicted center frequency. The predicts are assumed to be close, but not exactly equal, to the actual band center frequency. For example, consider the two channels that record the residual carrier. In the residual carrier sine channel, the sampled IF signal is mixed with \( \sqrt{2} \sin \hat{\Psi}_c(kT_s) \), where \( \hat{\Psi}_c(kT_s) \) is an estimate of the actual carrier phase \( \Psi_c(kT_s) \), and low-pass filtered to obtain the recorded waveform \( mC_s(kT_s) \). Similarly, in the residual carrier cosine channel, the sampled IF signal is mixed with \( \sqrt{2} \cos \hat{\Psi}_c(kT_s) \) and low-pass filtered to obtain \( C_c(kT_s) \). The recorded carrier is given by [6]

\[
C_s = \sqrt{P_s} \cos(\Theta_c) + n_c, \quad (13a)
\]

\[
C_c = \sqrt{P_c} \sin(\Theta_c) + n_c, \quad (13b)
\]
where $\Theta_c = \Psi_c - \Psi_e$, and where, using the Heaviside operator notation,\(^1\)

$$n_{ce} = H_c(z) \{ n_1 \sin(\Theta_c) - n_2 \cos(\Theta_c) \}$$  \hspace{1cm} \text{(14a)}

$$n_{c} = H_c(z) \{ n_1 \cos(\Theta_c) - n_2 \sin(\Theta_c) \}$$  \hspace{1cm} \text{(14b)}

The low-pass filter $H_c(z)$ is assumed to be bandlimited to $B_c$ Hz. Furthermore, since the carrier frequency error (derivative of $\Theta_c$) is assumed to be much smaller than $B_c$, both $n_{ce}$ and $n_c$ are baseband processes limited to $B_c$ with a flat one-sided 1-S1 level equal to $N_0$ W/Hz.

One approach to baseband recording of the subcarrier is the two stage downconversion scheme shown in Fig. 2a. The first stage centers the IF spectrum near dc by mixing it with sine and cosine references tuned to the predicted carrier frequency. The second stage first individually downconverts each of the subcarrier harmonics to baseband by mixing the output of the first stage with sine and cosine references tuned to multiples of the predicted subcarrier frequency, then low-pass filters the mixer output followed by recording of the desired subcarrier harmonic. The data power loss due to low-pass filtering is tabulated in Table 1b as a function of the ratio of the bandwidth $B_{su}$ (of the filter $H_{su}(z)$ in Fig. 2) to symbol rate.\(^2\) As shown in section 4, coherent subcarrier demodulation requires an integer phase relationship between the subcarrier references that downconvert the subcarrier harmonics to baseband. That is, let $\hat{\Psi}_{su}$ denote the phase of the reference that downconverts the fundamental harmonic, then the $m^{th}$ subcarrier harmonic is downconverted by mixing the output of stage 1 with references that have phase equal to $m \hat{\Psi}_{su}$. The four signals corresponding to the recording of the $m^{th}$ subcarrier harmonic are represented as ($m$ is an odd integer)\(^3\)

$$S_{U_{su}}^{(m)} \cdot \sqrt{1_d^{-2d} \Sigma_{m} \cos[m \Theta_{su}] \sin(\Theta_c) - \Pi_{m}^{(m)}}$$  \hspace{1cm} \text{(15a)}

\(^1\)The Heaviside operation $H(z)[x(m)]$ denotes the convolution of $x(m)$ with $h(m)$, i.e. $H(z)[x(m)] = \int_{-\infty}^{\infty} x(n)h(m-n)$

\(^2\)Loss in Table 1b is for an ideal lowpass filter with bandwidth $B_{su}$ and given by $\int_{-\infty}^{\infty} h_{\text{low}}(\omega) \cdot \text{sinc}^2 fdf$
\[ SU_{sc}^{(m)} = \sqrt{P_d y} \cdot \sin[m \Omega_{su}] \sin(\Omega_c) - n_{SU_{sc}}^{(m)} \] (15b)

\[ SU_{cs}^{(m)} = -\sqrt{P_d y} \cdot \cos[m \Omega_{su}] \cos(\Omega_c) - n_{SU_{cs}}^{(m)} \] (15c)

\[ SU_{cc}^{(m)} = -\sqrt{P_d y} \cdot \sin[m \Omega_{su}] \cos(\Omega_c) - n_{SU_{cc}}^{(m)} \] (15d)

where the noise terms are given by

\[ n_{SU_{sc}}^{(m)} = -H_{su}(z) \left\{ [n_1 \sin(\Omega_c) - n_2 \cos(\Omega_c)] \sin[m \tilde{\psi}_{su}] \right\} \] (16a)

\[ n_{SU_{cs}}^{(m)} = -H_{su}(z) \left\{ [n_1 \sin(\Omega_c) - n_2 \cos(\Omega_c)] \cos[m \tilde{\psi}_{su}] \right\} \] (16b)

\[ n_{SU_{cc}}^{(m)} = H_{su}(z) \left\{ [n_1 \cos(\Omega_c) - n_2 \sin(\Omega_c)] \sin[m \tilde{\psi}_{su}] \right\} \] (16c)

\[ n_{SU_{sc}}^{(m)} = H_{su}(z) \left\{ [n_1 \cos(\Omega_c) - n_2 \sin(\Omega_c)] \cos[m \tilde{\psi}_{su}] \right\} \] (16d)

It can be verified that \( n_{SU_{sc}}^{(m)}, n_{SU_{cs}}^{(m)}, n_{SU_{cc}}^{(m)}, \) and \( n_{SU_{cc}}^{(m)} \) are baseband processes limited to \( B_{su} \), the bandwidth of the low-pass filter \( H_{su}(z) \), with a flat \( 1 \) sided \( 1 \) W/Hz. Moreover, it can be shown that

\[ \mathcal{E}[n_{SU_x}^{(m)} n_{SU_y}^{(m)}] = 0 \quad x, y \in \{Se, w, cs, cc\}; \quad x \neq y \quad m, n \in \{1, 3, ..., L\} \] (16a)

\[ \mathcal{E}[n_{SU_x}^{(m)} n_{SU_x}^{(m)}] = 0 \quad x \in \{Se, u, cs, m\}; \quad m, n \in \{1, 3, ..., L\}; \quad m \neq n \] (16b)

where \( \mathcal{E}[X] \) denotes the expected value of \( X \) and \( 1 \) is the highest harmonic present at \( \Lambda/1 \) output.

Another approach to baseband recording of the received signal is shown in Fig. 2b. Unlike the previous recording scheme which uses the two stage downconversion technique, this scheme records each subcarrier harmonic by directly downconverting it to baseband. As shown in Fig. 2b, the sampled IF signal is still distributed into \( (4n+2) \) channels. The two channels used for recording the residual carrier remain the same for this new scheme. Consequently, the recorded carrier is as given by (13a) and (13b). However, the subcarrier
the signals in (18a)-(18d) can be rewritten as

\[ \text{Consequently,} \]

(18a)
\[ (u)S + (u)S : (u)S \]
(18b)
\[ (u)S \cdot (u)S : (u)S \]
(18c)
\[ (u)S \cdot (u)S : (u)S \]
(18d)
\[ (u)S \cdot (u)S : (u)S \]

...
Clearly, $n_{SU_i}^{(m)}$ is correlated with $n_{SU_i}^{(m)}$ and $n_{SU_{i4}}^{(m)}$ is correlated with $n_{SU_{i4}}^{(m)}$.

4 Coherent Baseband Demodulation

This section considers the coherent baseband demodulation of the carrier, and subcarrier when the modulation index (A in (1)) is such that the received signal consists of a sinusoidal carrier and squam-wave subcarrier; as well as, the case when the received signal consists of a suppressed carrier with square-wave subcarrier (i.e., $A = 90^\circ$). The demodulation functions begin by reading the recorded signals of Section 3 from a tape drive and then processing them to obtain control signals for the carrier, subcarrier, and suppressed carrier tracking loops. Recall that two methods for baseband recording of the 11" signal were considered in Section 3. It will be shown that the baseband processing corresponding to these two methods is very similar except for a little extra processing that is required for the direct downconversion scheme, the second recording scheme in Section 3. In this section, the carrier tracking loop, which is the same for both recording schemes, is described first. Then, the subcarrier and suppressed carrier loops for the two-stage downconversion recording scheme are described, and followed by the additional processing needed for using the signals of the direct downconversion scheme.

The carrier tracking loop, depicted near the bottom of Fig. 3a, respectively mixes the signals $C_{c}$ and $C_{s}$ with $\cos(\hat{\Theta}_c)$ and $\sin(\hat{\Theta}_c)$ before subtracting the mixer outputs and accumulating the resulting samples over the loop update interval to form the carrier loop control signal, $V_q$. Hence,

$$V_q(n) = \frac{1}{M_1} \sum_{k=nM_1}^{(n+1)M_1-1} [C_c(kT_s) \cos(\hat{\Theta}_c(kT_s)) - C_s(kT_s) \sin(\hat{\Theta}_c(kT_s))]$$

(21)
where $\hat{\Theta}_c$ is an estimate of the recorded carrier phase $\Theta_c$. Substituting (13a) and (13L) for $C_s$ and $C_c$, then expanding and canceling common terms, and assuming that the carrier phase error, $\phi_c = \Theta_c - \hat{\Theta}_c$, varies slowly over the loop update interval $T_u = M_1 T_s$, yields

$$V_q = \sqrt{P_c \sin(\phi_c)} \cdot n_{V_q}$$

where

$$n_{V_q}(n) = \frac{1}{M_1} \sum_{k=0}^{(n+1)M_1-1} [n_{c_s}(kT_s) \cos(\hat{\Theta}_c(kT_s)) - n_{c_s}(kT_s) \sin(\hat{\Theta}_c(kT_s))]$$

and $n_{c_s}$ and $n_{c_c}$ are given by (14a) and (14b). Since both $n_{c_s}$ and $n_{c_c}$ are zero mean with a flat one-sided PSD level equal to $K_c W/Hz$ within a much wider recording bandwidth $B_c$ compared to $1/T_u$, it can be shown that $n_{V_q}$ is Gaussian with zero mean and variance $N_0/2T_u$. Comparing (22) with (6) indicates that both the baseband carrier loop and the IF carrier loop have the same control signal in terms of signal power and noise variance. As a result, the carrier tracking variance of the baseband loop will be the same as a phase-locked loop tuned to the IF.

Recall that the square-wave subcarrier was recorded at baseband by individually mixing each of its harmonics down to baseband and then recording each harmonic separately. For each subcarrier harmonic retained, four baseband signals given by (15a)–(15d) were actually recorded. As shown in Fig. 3a, each harmonic of the subcarrier is first read from the tape recorder and demodulated individually, and then properly weighted and summed to reconstruct the proper in-phase and quadrature samples for the various Costas loops. The demodulation procedures require the four recorded signals of each subcarrier harmonic being multiplied by the estimated carrier references, $\sin(\hat{\Theta}_c)$ and $\cos(\hat{\Theta}_c)$, and the estimated subcarrier harmonic references, $\sin(m\hat{\Theta}_c)$ and $\cos(m\hat{\Theta}_c)$ where $m$ is an odd integer depending on the subcarrier harmonic under consideration. As depicted in Fig. 3a, the resulting signals are either added or subtracted to retain terms at $\Theta_c - \hat{\Theta}_c$ and $m(\Theta_{su} - \hat{\Theta}_{su})$, and eliminate terms at $\Theta_c + \hat{\Theta}_c$ and $m(\Theta_{su} + \hat{\Theta}_{su})$. The details of these demodulation procedures are provided in the Appendix A of [6]. Each of the demodulated signals of a particular subcarrier
harmonic is then properly weighted and summed to reconstruct the desired control signals for various tracking loops. For example, in the subcarrier loop depicted on the top half of Fig. 3a, the $n^{th}$ harmonic are scaled by $A_n \cos(\frac{2\pi n}{2L})$, the in-phase arm, and $A_n \sin(\frac{2\pi n}{2L} W_{su})$ on the quadrature arm ($W_{su}$ is the subcarrier window size). Subsequently, the scaled in-phase signals corresponding to each harmonic are added to form the input to the in-phase symbol accumulator whereas, the scaled quadrature signals are added to form the input to the quadrature symbol accumulator. The product of the in-phase accumulator output, $Y_i$ in Fig. 3a, and the quadrature accumulator output, $Y_{q, su}$ in Fig. 3a, is the subcarrier loop control signal. The symbol accumulator outputs corresponding to the $l^{th}$ symbol are given as [6]

$$Y_i = \sqrt{P_d a_i} \cos(\phi_c) \sum_{j=1}^{I_v} \left( \frac{1}{j^2} \cos(j\phi_{su}) \right) \cdot n_Y \quad (24)$$

and

$$Y_{q, su} = \sqrt{P_d a_i} \cos(\phi_c) \sum_{j=1}^{I_v} \left( \frac{1}{j^2} \sin(j\phi_{su}) \sin(j\phi_{su}) \right) \cdot n_{Y_{q, su}} \quad (25)$$

where the independent noise terms, $n_Y$ and $n_{Y_{q, su}}$, are shown in Appendix A of [6] to have variance $N_0/(2T)$ and $N_0 W_{su}/(2T)$, respectively, as $L$ approaches infinity in the above two equations. That is, for sufficiently large $L$, the noise variances of $n_Y$ and $n_{Y_{q, su}}$ are approximately the same as the variances of $n_{Z_i}$ and $n_{Z_{q, su}}$ in (7) and (8), respectively. Furthermore, the signal part of (24) and (26) and (25) and (8) are also equal as $L$ approaches infinity. This can be easily shown by using the Fourier series expansion of $F_i(\phi_{su})$ and $F_Q(\phi_{su})$ in (7) and (8) given as follows,

$$F_i(\phi_{su}) = \sum_{j=1}^{\infty} \frac{1}{j^2} \cos(j\phi_{su}) \quad (26)$$

and

$$F_Q(\phi_{su}) = \sum_{j=1}^{\infty} \frac{1}{j^2} \sin(j\phi_{su})$$

As a result of the subcarrier loop accumulator outputs of the ARX and BTJ being approximately equal for sufficiently large $L$, the control signals, $Y_i Y_{q, su}$ and $Z_i Z_{q, su}$, and tracking
variance of both loops can also be shown to be equal.

Next, let's consider the suppressed carrier with subcarrier loop also shown in Fig. 3a. As shown in the figure, it is a Costas loop sharing the same in-phase arm with the subcarrier Costas loop. Hence, the suppressed carrier in-phase accumulator output, $Y_i$, is as given in (24). The processing in the suppressed carrier loop quadrature arm is very similar to that of the subcarrier loop. Hence, following the same derivation as in the subcarrier loop quadrature arm, the suppressed carrier quadrature arm accumulation over the $l^{th}$ symbol can be shown to be

$$Y_{q,sc} = \sqrt{P_2 a_1 \sin(\phi_2)} \frac{8}{\pi^2} \sum_{j=1}^{L} \frac{1}{j^2} \cos(j \phi_{su}) - 1 nY_{q,sc}$$

(28)

As before, the equivalence of the suppressed carrier loops of the BTD and ARX is established by comparing the in-phase and quadrature accumulator outputs of both loops. The equivalence of the in-phase arms has already been shown earlier. The signal part of the quadrature arm accumulations given by (12) and (28) are seen to be equivalent after letting $L$ go to infinity in (28) and then using (26) for the infinite sum. The noise part of (12) and (28), as $L$ approaches infinity in (28), can also be shown to have the same variance (Are/27'). Furthermore, it can be shown that $nY_i$ and $nY_{q,sc}$ are uncorrelated to each other.

The equivalence of the ARX and BTD loops has been shown as $L$ approaches infinity. Now consider the case when the input to these loops is bandlimited so that the subcarrier is no longer a square-wave. That is, consider the case when $L$ in (4) is small. In this case, the ARX and BTD loops are not expected to have identical performance, mainly because the ARX is designed to be optimum for a square-wave subcarrier. The performance of both subcarrier loops for a finite number of harmonics at the input is currently under investigation. The ARX performance is being quantified for both an unfiltered and filtered square-wave reference signal. Relative to the case when both the input and reference signals are perfect square-waves, the first case (filtered input, unfiltered reference) is expected to have degraded performance because the reference is no longer "matched" to the input. The filtered input,
filtered reference case is expected to be better than case one but worse than the BTD loop, which also uses a filtered square-wave reference. One might immediately wonder why the second case is expected to be worse than the 11'11') when both use filtered references. The main reason is that ARX directly mixes the filtered reference with the filtered input which gives rise to cross-products at the mixer output whereas, the BTD eliminates cross-products by individually mixing the filtered signals one harmonic at a time.

Finally, consider the coherent baseband demodulation of the signals that were recorded as a result of the direct downconversion scheme, the second recording scheme described in section 3. As indicated earlier, the demodulator for the direct scheme is very similar to the two-stage downconversion scheme demodulator except that the former requires a little extra processing. Specifically, the recorded signals for the direct scheme, given by (1 8a)-(1 8d), are combined to yield the signals $SU^{(m)}_{sb}, SU^{(m)}_{sc}, SU^{(m)}_{cs}$ and $SU^{(m)}_{cc}$ which were the input signals to the two-stage demodulator described in this section. From (29)-(32), it is clear that the input signals to the two-stage demodulator can be generated by using two adders and two subtractors (per harmonic) on the signals $SU^{(m)}_{sb}, SU^{(m)}_{sc}, SU^{(m)}_{cs},$ and $SU^{(m)}_{cc}$. After these additions and subtractions, shown in Fig. 3b, the rest of the signal processing remains the same as the previous scheme.

5 Requirements for Software Implementation of the BTD

As indicated earlier, the BTD depicted in Fig. 3a is targeted for implementation on a general purpose engineering workstation. The primary concern with a software implementation is whether a general purpose workstation can provide the computational speed required for real time processing of the recorded signals. For example, if the received symbol rate is 1 K symbols/sec and the A/D sampling rate is 101{ sampling/sec, then a real time demodulator would process 1 OK samples/sec and detect 11{ symbols/sec. This section first determines
the computational speed required (in o)maticJns/sec) for implementing the BTD in real time when the symbol rate is 1 K symbols/sec and the sample rate is 1 OK samples/sec. Next, it determines the feasibility of implementing the BTD on a SunSparc 2 and SunSparc 10 workstation.

Recall that the two modes for the BTD are the residual carrier with subcarrier mode (A ≠ 90°) and the suppressed carrier with subcarrier mode (Δ = 90°). The number of (addition and multiplication) operations/sec, for a given mode, can be computed by separately counting the operations/sample and operations/symbol in Fig. 3a and then multiplying by the operations at the sampling rate and the operations at the symbol rate. Mathematically,

\[
\text{ops} = \left( \frac{\text{no. of 'i' and 'x'}}{\text{sample}} \right) \text{ops} \times 10K \text{samples} - \left( \frac{\text{no. of 'i' and 'x'}}{\text{symbol}} \right) \text{ops} \times 1K \text{symbols/sec}
\]

(29)

where the sample and symbol (floating point) operations for the coherent demodulator in Fig. 3a are tabulated in Table 2. Both addition and multiplication have been counted as 1 operation each. Furthermore, the look-up table operations are at the sample rate since the look-up table is considered to be part of the carrier and subcarrier NCOs. The coherent demodulator for the direct downconversion scheme needs four more sample operations per harmonic, as shown in Fig. 3b, and is not considered separately. With the number of harmonics, n, equal 3 (i.e., the 1st, 3rd, and 5th harmonics recorded) in Table 2, there are 71 operations/sample and 48 operations/symbol for the residual carrier mode, and there are 98 operations/sample and 39 operations/symbol in the suppressed carrier mode. Consequently, at 1 Ksps and 10 KHz sampling rate, a real time processor would be required to provide 758K ops and 1019K ops for the residual carrier and suppressed carrier modes, respectively. Table 3 lists the throughput utilized by a software implementation of the BTD for various computers. Throughput, measured in symbols/sec, is defined as the computers total capacity for work. In Table 3, the percent throughput (at 10 K sampling rate and 1K symbol rate) utilized by a SunSS-2 was obtained by simulating a portion of the BTD on the Sun SS-
2 and then linearly extrapolating the results to determine the throughput utilization for a software implementation of the entire BTD. The throughput of the remaining computers were obtained by linearly extrapolating the Sun SS-2 results using the Specfp92 benchmark. A Specfp92 rating of a computer, obtained by running a collection of commonly used programs on a target system, is a measure of its performance relative to a VAX-1 1/780. For example, the Sun SS-2 is expected to perform 22.8 times better than a VAX-11/780 since its Specfp92 rating is 22.8. It is clear from Table 3 that several computers will be able to process the recorded data in real time.

6 Post-Processing Strategies

Post-processing is one of the key features of the BTD as it allows the BTD to recover data lost during acquisition, and resynchronization. Since the received signal is recorded, the BTD can use different configurations (e.g. different loop bandwidths) to process the signal and obtain better performance. A more sophisticated strategy relies on "smoothing" to use information obtained from the signal after it has been acquired to go back in time and use the extracted apriori information intelligently. Specifically, Kalman filters or IJCast-Squares estimators[7] can be used to process the data backwards where the signal state is formulated using the phase, frequency, and/or frequency rate of the incoming signal after acquisition and the covariance matrices are initialized with the appropriate tracking jitters. Another approach relies entirely on the information contained in the state-space representation of the loop filter. Depending on its order, the filter contains information on the frequency and possibly frequency rate of the signal. By playing the data backwards, the frequency rate is reversed in sign and by adjusting the appropriate states of the loop filter, the phase locked loop can be made to go back in time while starting in lock. All these techniques are currently being investigated and their respective performances assessed.

Preliminary work, however, on processing and reprocessing recorded samples of the down-
link carrier has been performed [5] where an adaptive carrier loop bandwidth selection algorithm is used for reprocessing recorded samples of the downlink carrier from the Galileo spacecraft. Consider Fig. 4 (reprinted from [5]) which shows the loop bandwidth required to track the Galileo downlink carrier as a function of elapsed time. The track begins by setting the carrier phase-locked loop (1'1'1,) to an initial loop bandwidth $B_0$. For example, consider the solid line in Fig. 4 which corresponds to $B_0 = 0.1$ Hz. In this case, all the recorded samples are initially tracked with a loop bandwidth of 0.1 Hz and the times at which the loop loses lock are recorded. Samples lost during the out-of-lock state are recovered by backward processing of the samples. The bandwidths used for backward processing are indicated by the “spikes” in Fig. 4.

7 Conclusion

This article presented the Buffered Telemetry Demodulator (BTD), a software receiver that is suitable for low cost space missions characterized by low data rates, and relatively weak downlinks. The first application of the BTD will be the NASA's Galileo S-Band Mission where the symbol rate will range from 32 sps to 512 sps and the symbol SNR will be as low as -10 dB. In scenarios such as Galileo where the link margin is very low, tracking loops not only take an unusually long time to acquire but often slip cycles. The main advantage that the BTD has over existing deep space receivers is that once acquisition occurs, it can use estimates of the signal state (such as frequency, phase, and doppler rate) to basically go back in time and recover those symbols lost during acquisition, resynchronization, and cycle slips. Furthermore, since the proposed recording schemes individually record each subcarrier harmonic, the BTD can be easily implemented on several general purpose workstations.
Glossary

\( \Psi_c \) incoming Carrier l'base
\( \hat{\Psi}_c \) Predicted Carrier Phase
\( \Theta_c \) Carrier l'base Offset = \( \Psi_c - \hat{\Psi}_c \)
\( \hat{\Theta}_c \) Carrier l'base Offset Estimate
\( \phi_c \) Carrier Phase offset Error = \( \Theta_c - \hat{\Theta}_c \)
\( \Psi_{su} \) incoming Subcarrier l'base
\( \hat{\Psi}_{su} \) Predicted Subcarrier Phase
\( \Theta_{su} \) Subcarrier Phase Offset = \( \Psi_{su} - \hat{\Psi}_{su} \)
\( \hat{\Theta}_{su} \) Subcarrier Phase Offset Estimate
\( \phi_{su} \) Subcarrier l'base Offset Error = \( \Theta_{su} - \hat{\Theta}_{su} \)

\( X.SU_{ss}^{(m)} \) \( m \)-th subcarrier harmonic, derived from mixing the received signal with \( \sqrt{2} \sin(\Psi_c) \sin(m \hat{\Psi}_{su}) \)
\( S.U_{ss}^{(m)} \) the low-pass filtered version of \( X.SU_{ss}^{(m)} \)
\( X.SU_{sc}^{(m)} \) \( m \)-th subcarrier harmonic, derived from mixing the received signal with \( \sqrt{2} \sin(\Psi_c) \cos(m \hat{\Psi}_{su}) \)
\( S.U_{sc}^{(m)} \) the low-pass filtered version of \( X.SU_{sc}^{(m)} \)
\( X.SU_{cs}^{(m)} \) \( m \)-th subcarrier harmonic, derived from mixing the received signal with \( \sqrt{2} \cos(\Psi_c) \sin(m \hat{\Psi}_{su}) \)
\( S.U_{cs}^{(m)} \) the low-pass filtered version of \( X.SU_{cs}^{(m)} \)
\( X.SU_{cc}^{(m)} \) \( m \)-th subcarrier harmonic, derived from mixing the received signal with \( \sqrt{2} \cos(\Psi_c) \cos(m \hat{\Psi}_{su}) \)
\( S.U_{cc}^{(m)} \) the low-pass filtered version of \( X.SU_{cc}^{(m)} \)
\( X.SU_{ss+}^{(m)} \) \( m \)-th subcarrier harmonic, derived from mixing the received signal with \( \sqrt{2} \sin(\Psi_c + m \hat{\Psi}_{su}) \)
\( S.U_{ss+}^{(m)} \) the low-pass filtered version of \( X.SU_{ss+}^{(m)} \)
\( X.SU_{sc-}^{(m)} \) \( m \)-th subcarrier harmonic, derived from mixing the received signal with \( \sqrt{2} \sin(\Psi_c - m \hat{\Psi}_{su}) \)
\( S.U_{sc-}^{(m)} \) the low-pass filtered version of \( X.SU_{sc-}^{(m)} \)
\( X.SU_{cc+}^{(m)} \) \( m \)-th subcarrier harmonic, derived from mixing the received signal with \( \sqrt{2} \cos(\Psi_c + m \hat{\Psi}_{su}) \)
\( S.U_{cc+}^{(m)} \) the low-pass filtered version of \( X.SU_{cc+}^{(m)} \)
\( X.SU_{cc-}^{(m)} \) \( m \)-th subcarrier harmonic, derived from mixing the received signal with \( \sqrt{2} \cos(\Psi_c - m \hat{\Psi}_{su}) \)
\( S.U_{cc-}^{(m)} \) the low-pass filtered version of \( X.SU_{cc-}^{(m)} \)
References


Table 1a. Loss in data power as a function of subcarrier harmonics.

<table>
<thead>
<tr>
<th>Highest recorded subcarrier harmonic, $I$</th>
<th>Number of subcarrier harmonics at A/I output, $n$</th>
<th>Loss, dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>0.91</td>
</tr>
<tr>
<td>3</td>
<td>2</td>
<td>0.45</td>
</tr>
<tr>
<td>5</td>
<td>3</td>
<td>0.30</td>
</tr>
<tr>
<td>7</td>
<td>4</td>
<td>0.22</td>
</tr>
<tr>
<td>9</td>
<td>5</td>
<td>0.18</td>
</tr>
</tbody>
</table>

Table 1b. Loss in data power due to ideal low-pass filtering of data.

<table>
<thead>
<tr>
<th>$H_{in}$</th>
<th>Loss, dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.44</td>
</tr>
<tr>
<td>2</td>
<td>0.22</td>
</tr>
<tr>
<td>3</td>
<td>0.15</td>
</tr>
<tr>
<td>4</td>
<td>0.11</td>
</tr>
<tr>
<td>5</td>
<td>0.09</td>
</tr>
</tbody>
</table>

Table 2. Required number of operations for the demodulation function ($n$ is the number of harmonics recorded).

<table>
<thead>
<tr>
<th>Mode</th>
<th>Multiplication operations per sample</th>
<th>Multiplication operations persymbol</th>
<th>Addition operations per sample</th>
<th>Addition operations persymbol</th>
<th>Lookup table</th>
</tr>
</thead>
<tbody>
<tr>
<td>Residual carrier with subcarrier</td>
<td>$G + 8n$</td>
<td>14</td>
<td>13</td>
<td>$4(6n)$</td>
<td>24</td>
</tr>
<tr>
<td>Suppressed carrier with subcarrier</td>
<td>$4 + 14n$</td>
<td>14</td>
<td>12</td>
<td>$10(10n)$</td>
<td>25</td>
</tr>
<tr>
<td>Suppressed carrier without subcarrier</td>
<td>$G$</td>
<td>12</td>
<td>10</td>
<td>17</td>
<td>4</td>
</tr>
</tbody>
</table>

Table 3. Throughput comparison ($n$ is the number of harmonics recorded).

<table>
<thead>
<tr>
<th>System</th>
<th>SPECfp92</th>
<th>Percent throughput utilized ($n=3$)</th>
<th>Percent throughput utilized ($n=5$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sun SS-2</td>
<td>22.8</td>
<td>100</td>
<td>151</td>
</tr>
<tr>
<td>Sun SS-10/30</td>
<td>52.9</td>
<td>43</td>
<td>65</td>
</tr>
<tr>
<td>Sun SS-10/41</td>
<td>64.7</td>
<td>35</td>
<td>33</td>
</tr>
<tr>
<td>Sun SS-10/52</td>
<td>71.4/CPU</td>
<td>32 (on one CPU)</td>
<td>48 (on one CPU)</td>
</tr>
<tr>
<td>SGI Crimson</td>
<td>63.4</td>
<td>36</td>
<td>54</td>
</tr>
<tr>
<td>IIM 710</td>
<td>47.6</td>
<td>48</td>
<td>72</td>
</tr>
<tr>
<td>IHP 750</td>
<td>75.0</td>
<td>30</td>
<td>45</td>
</tr>
<tr>
<td>IHM 340</td>
<td>51.9</td>
<td>44</td>
<td>66</td>
</tr>
<tr>
<td>IHM 560</td>
<td>85.6</td>
<td>27</td>
<td>41</td>
</tr>
<tr>
<td>Intel Xpress (486)</td>
<td>14.0</td>
<td>160</td>
<td>242</td>
</tr>
</tbody>
</table>
Fig. 1. A simplified block diagram of the ARX IL.
Fig. 2a. The two-stage downconversion recording scheme.
Fig. 2b. The direct downconversion recording scheme.
Fig. 3a. The coherent baseband demodulator for the two-stage recording scheme.
SUPPRESSED CARRIER LOOP

TAPe RECORDER

CARRIER DEMODULATION SUBCARRIER 5th HARMONIC

SUBCARRIER DEMODULATION 5th HARMONIC

CARRIER DEMODULATION SUBCARRIER 3rd HARMONIC

SUBCARRIER DEMODULATION 3rd HARMONIC

CARRIER DEMODULATION SUBCARRIER 1st HARMONIC

SUBCARRIER DEMODULATION 1st HARMONIC

Fig. 38. (contd.)
This inphase arm is shared with the subcarrier 100p (see preceding page).
Fig. 3b. The processing required to generate the signals $SU_{ss}^{(m)}$, $SU_{sc}^{(m)}$, $SU_{cs}^{(m)}$, $SU_{cc}^{(m)}$ from the signals $SU_{ss}^{(m)}$, $SU_{sc}^{(m)}$, $SU_{cs}^{(m)}$, and $SU_{cc}^{(m)}$. 

Sip.
Fig. 4. Loop bandwidth for Galileo carrier phase recovery at various $B_0$. The events near $\sim 3570$ sec are due to a switch from one-way to two-way tracking (reprinted from [5]).