

Loran Data Modulation: A Primer

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Loran has provided navigation service since 1958. Though not originally designed with data broadcast capabilities, Loran's versatility has enabled data to be broadcast with great benefits. Research in the last two decades has resulted in a tremendous increase in the data capacity of Loran thereby increasing its utility. Currently, a modernized Loran is being evaluated for its capability to backup GPS and data modulation is an integral part of this Loran design. An overview and analysis of Loran modulation techniques is provided.

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I. INTRODUCTION

Loran or long range navigation is one of the earliest radionavigation systems established, becoming operational in 1958. It has been long used for maritime and timing applications. In addition, it has been used to provide supplementary navigation in aviation. Loran was considered the premier system for position, navigation, and timing (PNT) until the late 1980s when satellite navigation using the Global Positioning System (GPS) came to prominence commercially. Thus, while PNT applications have become increasingly integrated into our economy and society, the significance of Loran's PNT capabilities, until recently, has diminished due to increasing usage of the more modern GPS. In light of recent studies on GPS vulnerability and the prevailing use of GPS in safety and economically critical infrastructure, a backup has been deemed essential [1, 2]. As a result, there is resurging interest in Loran as a back up for GPS. Loran can provide similar capabilities in many of the applications that are critically dependent on GPS. For Loran to meet these needs, some upgrades and modifications are required. Many of these upgrades are either planned equipment upgrades or operational changes and hence are not significantly challenging technically. One essential change is for the Loran signal to carry digital data. The addition of "high" rate data on Loran provides the turnkey solution for Loran to serve maritime harbor entrance approach, aviation landing, and timing and frequency needs.

Data modulation on Loran is not a new concept. Low data rate modulation (less than 6 bit/s) was achieved by the United States Coast Guard (USCG) in the late 1960s [3–5]. In the 1990s, a team from Delft University improved upon some of the earlier modulation ideas and achieved data rates of 30 bit/s and more with forward error correction (FEC) [6]. This system, known as Eurofix, has been operating for some years on several European stations. Even more recently, there has been research in modulating data onto Loran for differential Loran (dLoran) corrections and wide area augmentation system (WAAS) broadcast [7–9].

As of early 2005, no U.S. Loran stations were continuously transmitting any data while only a few European stations were providing low data rate (~ 30 bit/s) transmissions.¹ However, it had been recognized that a data capability on Loran is necessary to meeting future PNT needs. Many modulation designs were created and implemented. A requisite of the modulation design is that it provides a data

¹The FAA and Coast Guard began conducting ninth pulse communications (NPC) on-air tests in July of 2005. Loran Station Seneca is currently transmitting NPC. (see <http://www.navcen.uscg.gov/loran/9th-pulse-modulation-ldc.html>).

rate that can adequately support the desired critical application. It also needs to maintain a signal that can meet the PNT requirements of the application. Another important characteristic is that compatibility with transmitter equipment, legacy receivers, and Loran bandwidth requirements.

This paper details the Loran modulation techniques being used today. It describes the important considerations and constraints as well as provides some tools for designing and assessing Loran modulation. Basic background on Loran is provided first. The basic signal structure is discussed in relation to the effects that modulation may have on it. The body of this paper focuses on the design and analysis of Loran modulation methods that are legacy equipment compatible. A follow on paper continues the discussion provided in this paper [16]. It details some of the more recent modulation proposals and the choices and constraints that led to these designs. One of these proposals will be an integral part of the next generation Loran system.

II. BACKGROUND

A. Fundamentals of using Loran for PNT

This section provides the fundamental background necessary for understanding the design of the Loran signal for PNT. First, it gives an overview of how Loran has historically provided PNT services. Then, the discussion builds up from the fundamental transmitted Loran pulse to signal propagation to the design of the grouping, sequence, and order in which these pulses are broadcast.

Using Loran for PNT applications relies on being able to determine the arrival time of the Loran pulse. This arrival time is used to determine either a relative time difference (TD) for hyperbolic positioning and timing, frequency or, if absolute time is reasonably known, a range measurement. Since absolute time is generally not known with the required precision, a TD of arrival is used. It is generated from measuring the differential arrival times of signals from two transmitters. One TD establishes a hyperbolic line of position (LOP). Measuring the differential arrival time of a signal from a third station yields another hyperbolic LOP. The intersection of those two LOPs then yields horizontal position. Positioning using TD is the traditional method of navigating using Loran.

To facilitate TD positioning, Loran stations operate in geographical groups of 3–5 stations termed chains. Within a chain, there is a master station and several secondary stations. At specified intervals, each Loran-C station broadcasts a group of eight pulses with each pulse transmitted 1 ms apart. The master station identifies itself by transmitting an additional pulse 2 ms after the start of the eighth pulse. The interval between repetitions of the set of eight or

nine pulses is unique to each chain and known as the group repetition interval (GRI). Convention has the GRI denoted in increments of 10 μ s. Loran is a time division multiple access (TDMA) system and careful selection of transmission times ensures that stations operating in a chain do not interfere with each other within the defined coverage area of the chain.

Positioning and navigation using multiple chains is facilitated by time of arrival (TOA) measurements. TOA measurements can be made if the receiver has an accurate knowledge of absolute time. Time information can be derived from an external source or it can be derived from Loran directly. For any given location, the signal pattern from stations with different GRI (or a dual rated station) repeats roughly every 2–3 minutes (for two different GRIs). If time is known within that accuracy, absolute time can be resolved by examining this pattern of the signals of the two different GRI.² Regardless, tracking and determining the timing for TOA measurements is essentially the same as those for TD. Hence factors that effect TD measurements will have the same effect on TOA and timing.

B. The Loran Pulse

The nominal Loran pulse is seen in Fig. 1 and is given by

$$s(t) = (t - \tau)^2 \exp\left(\frac{-2(t - \tau)}{65 \mu\text{s}}\right) \sin\left(2\pi \frac{t}{10 \mu\text{s}}\right) \quad (1)$$

where $(t - \tau)^2 \exp[-2(t - \tau)/65 \mu\text{s}]$ represents the envelope of the signal and $\sin(2\pi(t/10 \mu\text{s}))$ defines the carrier with the signal transmitted on a 100 kHz carrier. Each pulse is approximately 250 μ s in duration. The design meets the requirement to maintain 99% of its energy between 90 and 110 kHz. τ is the difference between the envelope and carrier. It is known as the envelope-to-cycle difference (ECD).

Tracking of the Loran pulse is usually done by examining a given zero crossing, usually the sixth (30 μ s after the start of the pulse). The zero crossing is selected as a compromise between signal power and interference mitigation. Since the maximum signal power is located at 62.5 μ s, tracking later than the sixth zero crossing can increase signal power. However, this increases the likelihood and strength of skywave signals, a form of multipath interference. The correct cycle for the zero crossing is usually identified using the pulse envelope. Measuring the zero crossing is enhanced by averaging the signal which reduces the relative effects of noise and interference. Averaging is limited by dynamics and clock stability. Additionally,

²One motivation for having Loran data transmission is to have a time and station information message that allows for an easier and more direct way of determining absolute time with Loran.

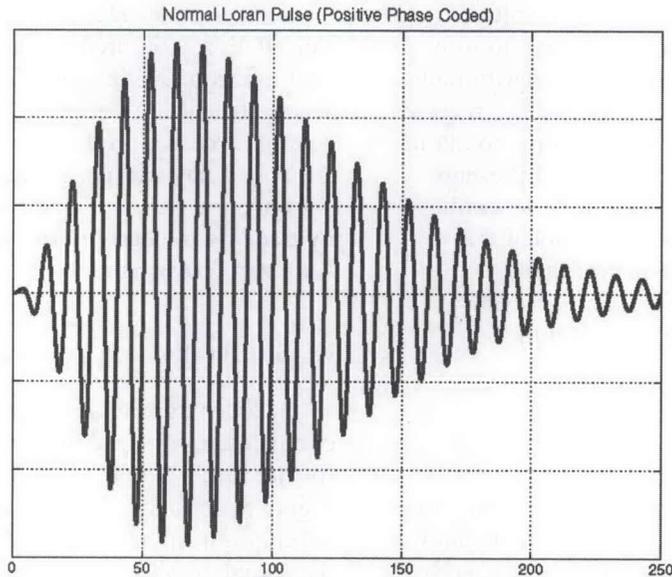


Fig. 1. Normal Loran pulse (positive phase code).

TABLE I
Loran Phase Code Error [10]

| GRI Interval | Master Transmitter | Secondary Transmitter |
|--------------|--------------------|-----------------------|
| A | + + - - + - + - + | + + + + - - - + |
| B | + - - + + + + - - | + - - + + + - - |

there is also a limit on the ability of averaging to reduce interference, particularly from other Loran signals.

C. Pulse Groups

One additional term is necessary to define a transmitted pulse. Relative to the normal pulse defined in (1), a transmitted pulse can have an initial phase shift of 0 or 180 deg. Known as phase coding, this is equivalent to an initial sign of +1 or -1 on the nominal pulse $s(t)$. This is seen in (2) where x_i is the i th phase code. The coding is used to mitigate the effects of interference from other undesired Loran pulses such as skywave interference and signals from other radio systems. Under Loran-C, the phase code repeats every phase code interval (PCI) which equals two GRIs. There is one code for the first pulse group for the specified GRI and another code for the subsequent group. The letters A and B are often used to denote these sets of pulses. So, for a given GRI, the station alternates between transmitting pulses with the A and B phase coding. Table I shows the phase code sets for Loran-C. When the signal is processed over two consecutive GRI intervals, most Loran derived interference average to zero. The autocorrelation of a two GRI sequence is zero except when both sequences have the same phase code and are offset by less than one pulse length (250 μ s). The

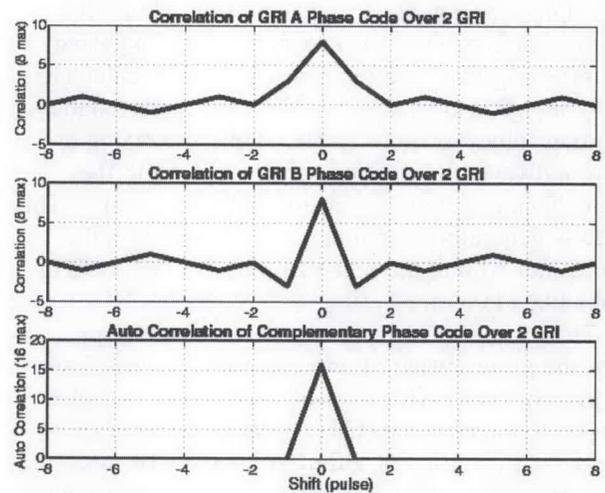


Fig. 2. Correlation of A and B phase code and auto-correlation of complementary phase code (secondary) over two GRI period.

autocorrelation function is shown in Fig. 2. Hence, phase coding is complementary over two GRIs. The master and secondary stations use different phase code sets. Since the phase code repeats over two GRI, (3) offers a representation that can be used as a proxy for all nominal transmissions.

$$p_i(t) = x_i s(t) \quad (2)$$

$$p(t) = \sum_{i=1}^8 x_i s(t - iT) + \sum_{i=9}^{16} x_i s(t - iT). \quad (3)$$

A potential deviation from the nominal pulse $s(t)$ is "blink coding." This can occur on the first two pulses of a pulse group from a secondary station. Under blink, the first two pulses are alternatively transmitted then not transmitted. Blink is a form of integrity check that warns users if a Loran signal is unreliable and

should be excluded from the navigation solution. There is a proposal for modernized Loran to turn off all pulses from a station when an integrity fault is detected at the station. This would aid the receiver in detecting integrity faults within the time to alarm for aviation. Hence (3) still adequately represents these conditions with $x_i = +1, 0,$ or -1 depending on conditions. However, it should be noted that phase coding is part of the nominal Loran transmission while blink and off air occur only when the system is not performing within tolerance.

D. Propagation of the Loran Signal

Thus far, the discussion has focused on the pulse and its format. The performance of the modulation is also affected by Loran propagation as it affects signal strength and the interference environment. There are two ways that Loran signals propagate. The signals propagate as a groundwave along the Earth's surface. They also propagate as a skywave by reflecting from the ionosphere. Time is calculated using the groundwaves since they are more reliable and their phase is more stable. While traversing as a groundwave, the envelope shifts relative to the cycle. This is accounted for in (1) by the term τ . This ECD is generally known and compensated by the transmitter so that the received pulse in the far field looks like (1) with $\tau = 0$.

The groundwave from a station can interfere with the groundwave of another station if it is from a different chain. This type of interference is called cross rate interference (CRI) because it is due to stations transmitting at different intervals or rates.

A skywave Loran signal traverses a longer path than the groundwave. Thus, it is delayed relative to the received groundwave signal with delays typically exceeding $30 \mu s$. The delay fluctuates and is not known a priori. As a result, it is generally not used and as such is considered interference. Skywave signals can be received 1500 km or more from their origin. The power of the skywave signal relative to the groundwave also generally increases with distance.

The cross rate and skywave represent the primary sources of man-made interference on Loran. For Loran navigation, techniques have been developed to mitigate the interference. Since the Loran-C waveform is known, cross rate cancellation techniques can be relatively effective [11]. However this does not work well with Loran communications since the waveforms change depending on the data transmitted. Such waveforms generally have to be "blanked" which eliminates an entire block of time from the processing. This reduces the number of pulses available for averaging over a given time. Cancellation is possible

if a data wipe off algorithm is employed prior to cancellation. The process can prove difficult and lead to errors. Mitigation of skywave interference is accomplished in two ways. The complementary phase coding discussed in the previous section mitigates the long delay (delays greater than one Loran pulse) skywave interference. Most other skywave is mitigated by tracking the early portion of the pulse such as the sixth zero crossing.

E. Specifications on the Loran Signal

In order to provide a common basis for comparison, standard definitions and specifications for the Loran signal need to be used. The Loran-C Signal Specifications [12] provide this. One important definition is signal strength. This is important since the signal amplitude varies throughout the pulse. As defined in the signal specifications, signal strength is based off the standard sampling point which is $25 \mu s$ after the start of the pulse. The sampling point is 5.91 dB less than the maximum signal power. Another important standard is the conversion of noise calculations to a 30 kHz noise equivalent bandwidth (NEBW). The standard is both reasonable and useful since the passband of a typical Loran receiver is roughly 30 kHz. For calculating signal-to-noise ratio (SNR), the post 30 kHz filter noise power is used. The analysis presented uses these standard definitions.

F. Outline of the paper

The rest of the paper is devoted to understanding and assessing Loran modulation techniques. Section III presents an overview of these techniques, outlines the benefits and costs, and presents some examples. This is followed by assessments of the performance and cost of implementing the techniques. It is divided into three sections: data performance, effect on receivers and navigation, and effect on transmitter equipment.

III. LORAN MODULATION TECHNIQUES

A. Modulation Techniques

All data modulations implemented on Loran have been variations on three basic methods. The methods can be used individually or combined to form a more complicated but higher data rate signal design. These methods and their analyses were presented in [13]. This section expands upon that analysis. The first technique is pulse position modulation (PPM). In PPM, the Loran pulse is time advanced/delayed. The second technique is

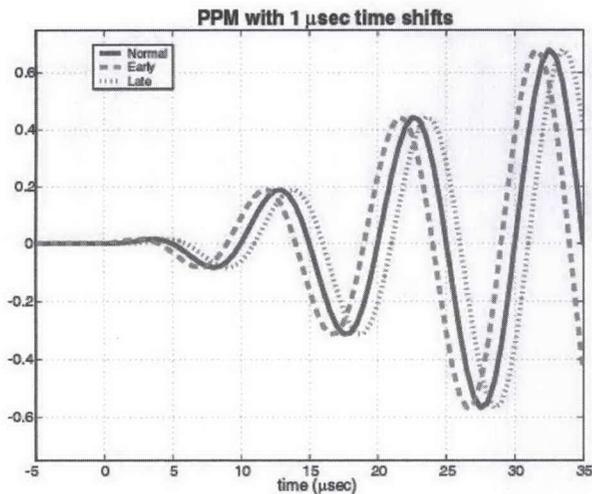


Fig. 3. Example of PPM.

intrapulse frequency modulation (IFM) whereby modulation is encoded within the pulse by slowly frequency shifting the signal. The third method is supernumerary interpulse modulation (SIM) whereby new pulses are generated in between the current pulses.

B. Pulse Position Modulation

PPM is a system where Loran pulses are time shifted to code information. An example of PPM is shown in Fig. 3. Data is coded by displacing the pulse relative to the arrival time of the nominal pulse. The number of different possible displacements dictates the overall data rate while the minimum spacing between displacements dictates the error rate at any given SNR. Eurofix and Two Pulse Communications (TPC) are examples of the technique [6, 14].

One primary advantage of PPM is that it has a high degree of legacy compatibility.³ For example, in Eurofix, the last six pulses are modulated such that the code is balanced over every GRI. Under “balanced coding,” every time shifted pulse is offset by an equal and opposite time shift over a given interval. That ensures that the signal offset is averaged to zero by legacy receivers. Furthermore, PPM can be reasonably implemented on most operating Loran transmitters.

Small time shifts are necessary so as to not greatly disturb the pulse from its nominal position. This helps to preserve navigation capabilities for legacy users. Hence typical implementations limit the maximum time shift to roughly 1–2 μ s. This limits the data capacity per pulse to effectively 2–3 bits if the pulse is necessary for navigation. For pulses not used for navigation such as the proposed Ninth Pulse Communications (NPC) signal, PPM can provide much higher data capacity per pulse.⁴

C. Intrapulse Frequency Modulation

IFM changes the carrier frequency within a Loran pulse to produce a phase offset from the nominal Loran pulse. The resultant phase offset can be constant or it can be varying. It has been found that holding a constant phase offset results in better performance. The change should occur after the sixth zero crossing to reduce the effects of the coding scheme on the navigation performance. Fig. 4 shows

³In this paper, the term legacy or existing receiver is used to denote receivers currently on the market that do not recognize modulated signal. Nonlegacy or modern receiver denotes a receiver designed to utilize the modulation.

⁴NPC is a modulation scheme that adds one additional pulse to each GRI. The pulse is for data only and is modulated using a variation of PPM. See [9].

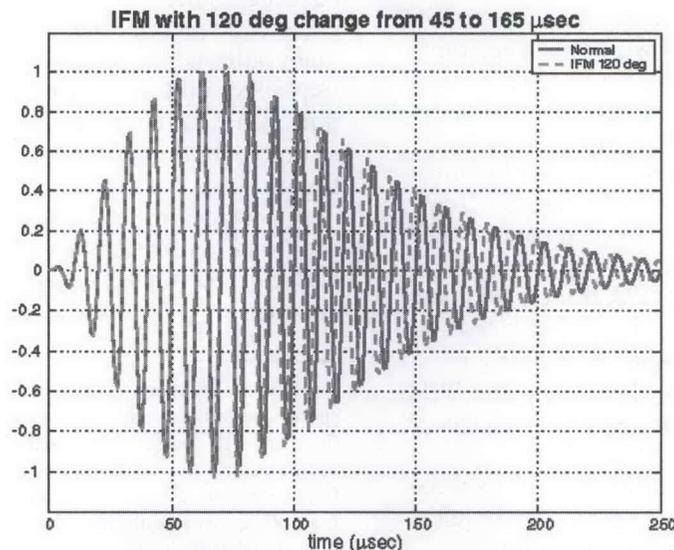


Fig. 4. Loran pulse and IFM Loran pulse.

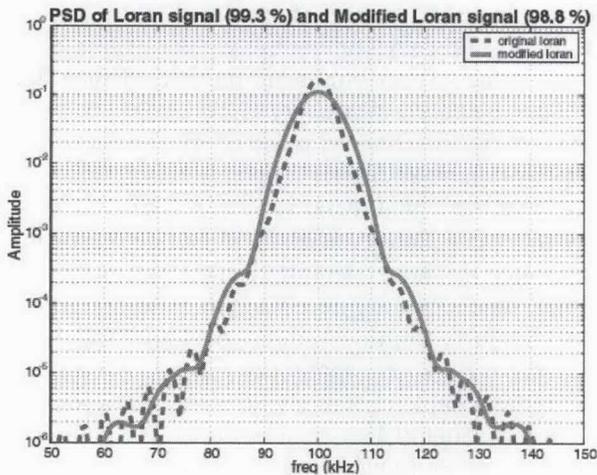


Fig. 5. Four level IFM spectrum.

an example IFM pulse and a standard Loran pulse. Some designs and implementations of IFM help illustrate the idea. One simple implementation uses a gradual change in frequency to achieve a phase shift of up to 90 deg in 100 μ s. The rate of the shift is limited to maintain spectrum requirements.

A more spectrally efficient implementation is to have the frequency change result in a final frequency of 100 kHz [5]. By gradually increasing and then decreasing the frequency (or vice versa), the various symbols are created. The frequency change offsets the phase from a normal Loran pulse by a specified fixed value. The three level design used phase shifts of -120 , 0 , and $+120$ deg and the four level design had phase shifts of -135 , -45 , $+45$, $+135$ deg. Fig. 4 shows the three level design. The greater phase separation between the waveforms relative to the simple 90 deg design results in better data performance. As a result of the greater phase separation, spectrum requirements are slightly exceeded. As seen in Fig. 5, four level IFM has 98.8% of its energy within the required spectrum versus 99.3% for the normal Loran signal. Three level IFM maintains 99.0% of its energy within the spectrum.

Another implementation is a 16 level IFM signal. This signal is capable of providing the full WAAS bandwidth (on most dual rated stations). It was designed and implemented to demonstrate Loran capability to provide the WAAS message [7]. The system is termed the high speed Loran data channel (LDC) and is discussed more in part II.

The primary advantage of IFM is increased data capacity per pulse. As such, a higher data rate can be achieved or fewer IFM modulated pulses are necessary to transmit the same data vice using PPM. This second point is significant because it means fewer pulses have to be blanked for CRI.

However, the costs are significant. IFM requires a more complicated receiver design. It also requires more modifications to implement on a solid state transmitter (SSX) than PPM. Furthermore,

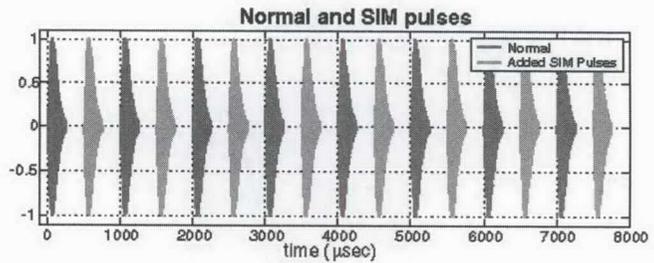


Fig. 6. SIM example.

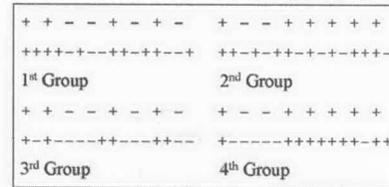


Fig. 7. Example extension of master station phase code for Loran-D (original phase code shown first, interleaved Loran-D pulses phase code shown below).

implementing IFM could also require additional spectrum.

D. Supernumerary Interpulse Modulation

SIM places additional pulses between current Loran-C pulses. One method to accomplish this is to use a concept similar to Loran-D [15] and interleave pulses between current Loran pulses. This is seen in Fig. 6. In the extreme case, each station transmits 16 pulses spaced 500 μ s apart per GRI. The additional pulses, since they are not used for navigation can be modulated in various manners such as PPM or IFM or by changing the phase coding. Frequency, amplitude, and other forms of modulation may be employed. An example of SIM that may become very important to the future use of Loran is the NPC concept. Under NPC, every station has an additional pulse 1000 μ s after the eighth pulse in a GRI. This ninth pulse is the only data modulated pulse and uses PPM. NPC is the turnkey solution to providing differential Loran, aviation warning, and time and station identification data for the next generation Loran.

Under the interleaved or 16 pulse system, the additional pulses can be coded so that a normal Loran-C receiver is oblivious to the existence of the new pulses.⁵ Fig. 7 shows an example of how this is

⁵Making the additional pulses unobservable to legacy users one chooses any phase coding pattern for the interleaved pulses in the first two GRIs (pulse groups) and then uses the opposite phase coding pattern for the interleaved pulses in the next two GRIs. Hence there are 2^{16} phase coding patterns that do not disturb Loran-C receivers with time constants of four GRIs or more [3]. This comes from having two choices for each of 16 pulses (2 GRIs).

accomplished. The pulses can also be phase coded for skywave rejection. Receivers with the ability to utilize the additional pulses could repeat benefits such as greater averaged signal strength since they will receive twice as many pulses per second [15]. While transmitters have limits on the number of pulses that can be transmitted per second, most current Loran-C transmitters should be able to support 16 instead of 8 pulses per pulse group.

The advantage of having and modulating supernumerary pulses is that they can be modified more than the standard pulses since they are not used for navigation. One can modulate by using PPM and/or IFM. If PPM is used, much larger time shifts can be used. Additionally, amplitude modulation (AM) and other forms of modulation may be used if the pulse is not used for navigation. Furthermore, every supernumerary pulse can be modulated.

There are several costs. Supporting legacy receivers would require that the extra pulses be designed so that they are ignored by legacy receivers. Regardless, CRI will increase due to the increased number of transmitted pulses. Again, modulated pulses are not easily cancelled. The broadcast of additional pulses increases power consumption. Furthermore, transmitters are limited in the number of pulses that can be transmitted per to 300 pulses/s. This limit could be exceeded on some dual rated stations (stations operating as part of two different chains) if 16 pulse SIM were implemented.

IV. MODULATION PERFORMANCE AND DATA RATE

As mentioned earlier, there are three significant areas that are considered when designing data modulated Loran signals. This section focuses on data performance with the metric of interest being data rate. However, the data rate needs to be qualified by the SNR necessary to achieve a desired error rate. This section develops models for assessing the performance of each Loran modulation technique in the presence of various noise sources. Some Loran noise sources, such as atmospheric noise, can be assumed to be Gaussian. Other sources, such as CRI and skywave, are not. Also, the receiver bandwidth must also be considered. Hence the model must be able to account for different types of noise. The end result is a matrix containing data rate with the corresponding SNR necessary to achieve a given probability of symbol error. Discussion of FEC is left for the examples paper [16].

A. Union Bound

An estimate of the probability of symbol error can be developed using a union bound. Formulas for the union bound for a M -ary modulation scheme can be found in text such as [17]. From such text we

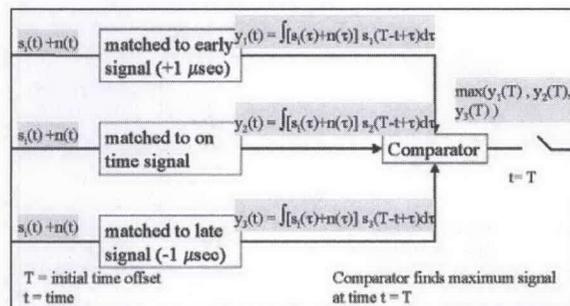


Fig. 8. PPM matched filter.

derive (4) where γ is the normalized received SNR, $1/(N_o/2)$, d_{jk} represents the Euclidean distance, and F_{norm} is the cumulative density function (CDF) for the standard normal variable [18]. The equation for Euclidean distance is given in (5) where the terms s_j and s_k represent different transmitted pulses or symbols. The formula provides a quick means of calculating performance with the important metric being the Euclidean distance.

$$P_e \leq \frac{1}{M} \sum_{i=1}^M \sum_{j=1, j \neq i}^M F_{\text{norm}}(-d_{ij} \cdot \sqrt{\gamma}) \quad (4)$$

$$P_e \leq \frac{1}{M} \sum_{i=1}^M \sum_{j=1, j \neq i}^M F_{\text{norm}}\left(-d_{ij} \cdot \sqrt{\frac{1}{N_o/2}}\right)$$

$$d_{jk} = \sqrt{\int [s_j(t) - s_k(t)]^2 dt} \quad (5)$$

The bound assumes additive white Gaussian noise (AWGN) and does not incorporate filter effects. Hence, a union bound based on a matched filter model was derived to model non-Gaussian interference and filter effects.

B. Deriving a Model from a Matched Filter Implementation

The limitation of using (4) is that it is derived assuming ideal correlation and detection assuming AWGN. We want to develop a model that can be used to assess performance of modulation techniques in the presence of nonwhite noise and interference. A model based on a matched filter implementation was developed as matched filters are practical means of implementing Loran demodulation in the receiver. They are a specific implementation of the correlation and detection idea used to generate the results in (4). It allows us to model the effect of a bandlimited filter. The result is a relationship between SNR and bit error rate. The model can be used to assess the performance of the system in the presence of the primary Loran interference sources—skywave and cross rate. Fig. 8 and Fig. 9 show representative models for PPM and IFM, respectively.

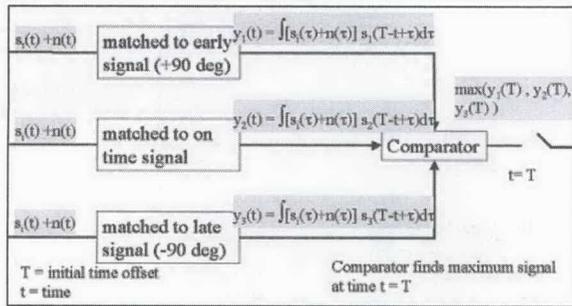


Fig. 9. IFM (three level) matched filter set up.

A matched filter performs convolution of a time-reversed version of a desired signal with the input [17]. The result is the same as multiplying the signal with a time-shifted version of the desired signal and integrating the product. The noise entering the receiver is atmospheric noise. This is generally assumed to be AWGN before passing through any filters in the receiver front end. After passing through the front end, the noise is typically bandlimited to about 25–30 kHz about 100 kHz. Since the noise is stochastic, the derivation results in the statistics of random output. Random variables are designated with a superscript “~.” Additional interference, such as undesired Loran signals such as skywave interference, can be analyzed with the model, as seen in Section IVD.

Assume that the transmitter can transmit M distinct pulse waveforms. The index i is used indicate each transmitted waveform. The receiver has M matched filters, indexed by the letter j or k , matched to each possible waveform. Let \tilde{y}_j represent the output from filter j and $\tilde{y}_{j|i}$ represent this random variable given that waveform i is transmitted. The received signal $x(t)$ is the sum of the transmitted waveform i , $s_i(t)$, and signal noise $n(t)$. For simplicity, it is assumed that the received signal is normalized such that $s_i(t)$ is unity. Assuming that the received signal has passed through matched filter j , denote the expected value and variance of $\tilde{y}_{j|i}$ by $y_{j|i}$ and $\sigma_{j|i}^2$, respectively. The classical result is obtained if AWGN is assumed.

$$y_{j|i} = E \left\{ \int [s_i(t) + n(t)] s_j(t) dt \right\} = \int s_i(t) s_j(t) dt \quad (6)$$

$$\sigma_{j|i}^2 = E \left\{ \int [s_i(t) + n(t)] s_j(t) - s_i(t) s_j(t) dt \cdot \int [s_i(z) + n(z)] s_j(z) - s_i(z) s_j(z) dz \right\} \quad (7)$$

$$\sigma_{j|i}^2 = E \left\{ \int n(t) s_j(t) dt \int n(z) s_j(z) dz \right\}.$$

For AWGN,

$$E\{n(t)n(\tau)\} = \frac{N_0}{2} \delta(t - \tau), E\{n(t)^2\} = \frac{N_0}{2}. \quad (8)$$

Hence, for AWGN,

$$\sigma_{j|i}^2 = \frac{N_0}{2} \int s_j(t) dt \int s_j(z) \delta(t - z) dz = \frac{N_0}{2} E_{s_j} \quad (9)$$

where signal energy, $E_{s_j} = \int s_j^2(t) dt$.

Assuming that the filters contribute negligible noise to the signal, the signal noise seen by each filter should be the same. Since the signal noise is the same for each filter and represents the only random element that passes through the filters, the output from each of the three filters is correlated. The demodulated symbol is chosen from the matched filter with the largest output. The variance of the difference of the outputs from each filter pair now can be derived.

Examine waveform i after it passes through filter j and filter k . Subtract the output from filter k from the output from filter j and denote the result by $\tilde{z}_{jk|i} = \tilde{y}_{j|i} - \tilde{y}_{k|i}$. The variance of $\tilde{z}_{jk|i}$ is given by $E\{(z_{jk|i} - \bar{z}_{jk|i})^2\}$

$$= E \left\{ \int ([s_i(t) + n(t)][s_j(t) - s_k(t)] - s_i(t)[s_j(t) - s_k(t)]) dt \cdot \int ([s_i(\tau) + n(\tau)][s_j(\tau) - s_k(\tau)] - s_i(\tau)[s_j(\tau) - s_k(\tau)]) d\tau \right\} \quad (10)$$

$$= E \left\{ \int (n(t)[s_j(t) - s_k(t)]) dt \int (n(\tau)[s_j(\tau) - s_k(\tau)]) d\tau \right\} \quad (11)$$

$$= \iint [s_j(t) - s_k(t)][s_j(\tau) - s_k(\tau)] * E\{n(t)n(\tau)\} dt d\tau. \quad (12)$$

Hence $\tilde{z}_{jk|i}$ is equal to the variance of the difference of the output from filter j and filter k after the transmitted signal i and noise has been processed by these filters. Since there is no dependence on the transmitted signal, denote $\tilde{z}_{jk} = \tilde{z}_{jk|i}$ and σ_{jk} as the standard deviation of $\tilde{z}_{jk|i}$.

In discrete form, the integral (12) can be approximated as

$$= \sum_{l=0}^M \sum_{m=0}^M \{ [s_j(m\Delta t) - s_k(m\Delta t)][s_j(l\Delta t) - s_k(l\Delta t)] \cdot E\{n(m\Delta t)n(l\Delta t)\} \Delta t^2 \}. \quad (13)$$

The correlation function $E\{n(m\Delta t)n(l\Delta t)\}$ can be made to reproduce the effect of filtering. For example, if a 30 kHz bandpass filter is used, the time domain correlation of the noise is a phase shifted sinc function. This is because a (phase shifted) sinc function is the Fourier transform of an ideal rectangular (bandpass) filter.

The expected values and variances are used to calculate a bound on the probability of error given a signal i . The bound is the sum of the probability of the receiver selecting each incorrect symbol or $P(\text{error} | i)$.

$$\begin{aligned}
P(\text{error} | i) &\leq \sum_{j=1, j \neq i}^M P(\tilde{y}_j > \tilde{y}_i | i) = \sum_{j=1, j \neq i}^M P(\tilde{y}_j - \tilde{y}_i > 0 | i) \\
&\leq \sum_{j=1, j \neq i}^M P(\tilde{z} = (\tilde{y}_{ji} - \tilde{y}_{ii}) - (y_{ji} - y_{ii}) > (y_{ii} - y_{ji})) \\
&\leq \sum_{j=1, j \neq i}^M Q\left(\frac{y_{ii} - y_{ji}}{\sigma_{ji}}\right) = \sum_{j=1, j \neq i}^M F_{\text{norm}}\left(\frac{y_{ji} - y_{ii}}{\sigma_{ji}}\right).
\end{aligned} \tag{14}$$

\tilde{z} is zero mean. Q and F_{norm} are the complementary CDF and cumulative density function (cdf) for the standard normal variable, respectively. The union bound on the average probability of error assuming equally likely signals is then

$$P_e \leq \frac{1}{M} \sum_{i=1}^M \sum_{j=1, j \neq i}^M F_{\text{norm}}\left(\frac{y_{ji} - y_{ii}}{\sigma_{ji}}\right). \tag{15}$$

Now let's see what happens if we assume AWGN at the input. First examine σ_{jk} .

$$\begin{aligned}
E\{(z_{jk} - \bar{z}_{jk})^2\} &= \frac{N_o}{2} \int [s_j(t) - s_k(t)]^2 dt = \frac{N_o}{2} d_{jk}^2 \\
&\cong \frac{N_o}{2} \sum_{m=0}^M [s_j(m\Delta t) - s_k(m\Delta t)]^2 \Delta t.
\end{aligned} \tag{16}$$

Again, d_{jk} is the Euclidean distance. Using (16) in (15), we get (17)

$$P_e \leq \frac{1}{M} \sum_{i=1}^M \sum_{j=1, j \neq i}^M F_{\text{norm}}\left(\frac{\int (s_j(t) - s_i(t))s_i(t)dt}{\sqrt{\frac{N_o d_{ij}^2}{2}}}\right) \tag{17}$$

$$P_e \leq \frac{1}{M} \sum_{i=1}^M \sum_{j=1, j \neq i}^M F_{\text{norm}}\left(\frac{\sqrt{\gamma} \int (s_j(t) - s_i(t))s_i(t)dt}{d_{ij}}\right).$$

Equation (17) is similar to the simple union bound formulation seen in (4). An examination of the quotients for each equation shows they are nearly identical. Let R be the ratio between the quotients in the two equations.

$$R = \frac{\text{quotient eq (17)}}{\text{quotient eq (4)}} = \frac{\int (2s_j s_i - 2s_j^2) dt}{\int (2s_j s_i - s_j^2 - s_i^2) dt}. \tag{18}$$

R is 1 for PPM and nearly 1 for IFM. If each signal has the same power (s^2), the quotients would be exactly the same. However, the IFM signals have the slightly same different power. This results in the slight difference for IFM. In the derivation of the matched filter receiver, it is assumed that the filters are based on orthonormal functions. If this is the case, the result would be the same as the union bound result.

The benefit of using the matched filter derivation is that one can use (12) or (13) to analyze the performance under conditions other than AWGN. By using a noise correlation that is reflective of the noise characteristic or filter effects, the resulting variance of the difference of the output from filter j and filter k can be determined. Since this is the variance of concern in determining probability of error, the effects of filtering and colored noise can be incorporated and assessed.

C. Error Probabilities in the Presence of Noise

Fig. 10 shows the bound for a three level PPM versus SNR along with simulation results. It is assumed that the noise is correlated between each matched filter (and hence correlated error). Additionally, the gain at 100 kHz was matched for the models. The union bound bounds the simulation and both curves converge at lower error rates. The same analysis is conducted for five level PPM with time shifts of 0.5 and 1 μ s. Results are shown in Fig. 11.

Fig. 12 and Fig. 13 show the results of the analysis for the proposed four and three level IFM methods, respectively. The results confirm that IFM can be accomplished with error rates comparable to or better than PPM for a given SNR and data rate. The three level IFM scheme performs slightly better due to the increased phase separation but is not preferred due to the lower data rate. Using the four level instead of the three level technique results in a 26% increase in data rate (12 versus 9.5 bits per GRI). In addition, the four level technique results in an integral number of bits per pulse, which is easier to process. Again, simulation and analytic results are well matched.

D. Error Probabilities in the Presence of Loran Interference and Noise

The previous section analyzes the performance of the modulation methods in the presence of noise. However, we also need to understand the performance of the modulation in the presence of interference from other Loran signals. This section uses the matched filter model to analyze the error rates due to the addition of interference from another Loran station.

The interference due to another Loran pulse can be examined by adding the interfering pulse to the signal received by the matched filter. The interfering pulse modifies the shape of the desired pulse. In addition, the signal is still affected by noise. So the modified pulse along with noise passes through the matched filters and is compared with the unadulterated data pulses. This is represented in Fig. 14 if the pulse with CRI is used as the input.

If the only source of interference is the undesired Loran pulse, then knowledge of the relative arrival time of the pulse and the relative signal strength

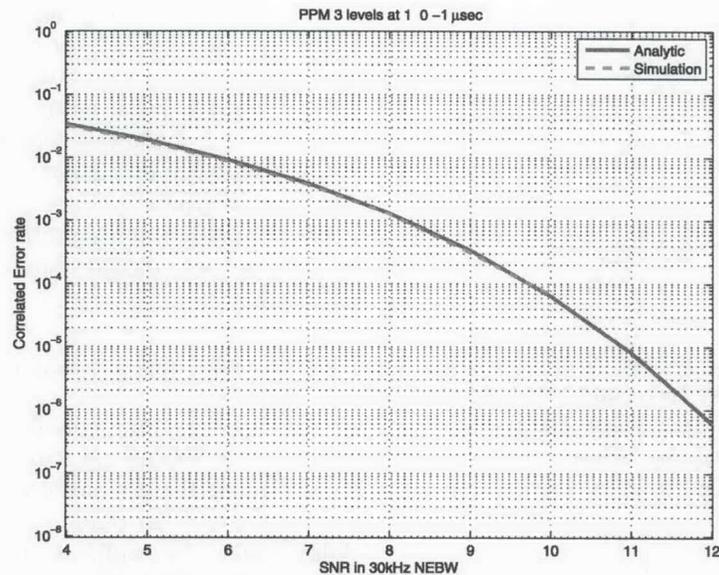


Fig. 10. PPM probability of error versus SNR (3 level).

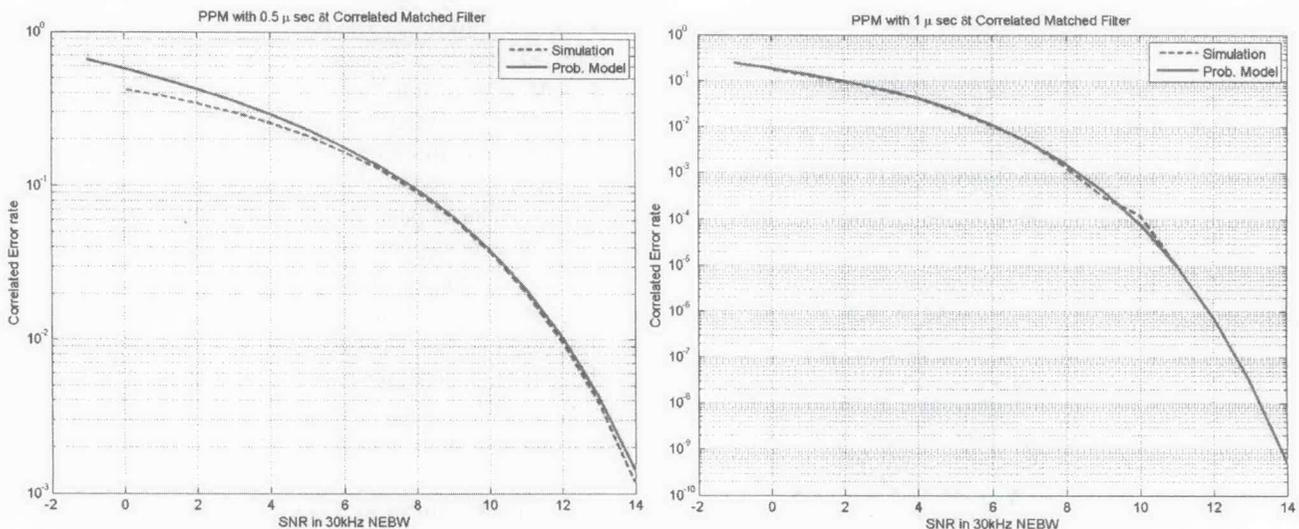


Fig. 11. Five level PPM (0.5 and 1.0 μ s spacing).

of the pulse would make the matched filter output deterministic. However the relative arrival time varies due to changes in the propagation path characteristics. To model the variability, a range of arrival times is analyzed with a given probability for each value in the range. Probability of error is derived with noise and arrival time being the underlying random variables. The arrival time is modeled as a uniform random distribution. The interference power is deterministic for a given location. The code phase of the desired Loran signal relative to the interfering cross rate is modeled as a random variable with equal probability of being the same or out of phase.

Interference from another Loran pulse can affect the primary signal in different ways. For a primary signal with SNR around 10 dB, ambient atmospheric noise has little effect on the data modulation and the major determinant of error will be the relative strength

of the interfering pulse. Fig. 15 shows a plot of data error versus relative SNR difference for interference from another Loran pulse. The difference in SNR is termed the signal to interference ratio (SIR). No noise is added since noise is assumed to have a negligible effect. This curve is different from the previous SNR versus error plots because it averages over all possible pulse interference times. The plot assumes that all interference times are equally likely. Fig. 15 is only useful for illustration purposes since, for a given SIR, the probability of error can vary greatly depending on the time offset between the desired and interfering pulse. Given the location of the user, the offset is calculated. Since the actual delay can vary by a few microseconds, the analysis models the variation using a uniform distribution from -2 to 2μ s. For single hop skywave, the deviation is expected to be greater than the groundwave variation. Fig. 16 shows a curve for a

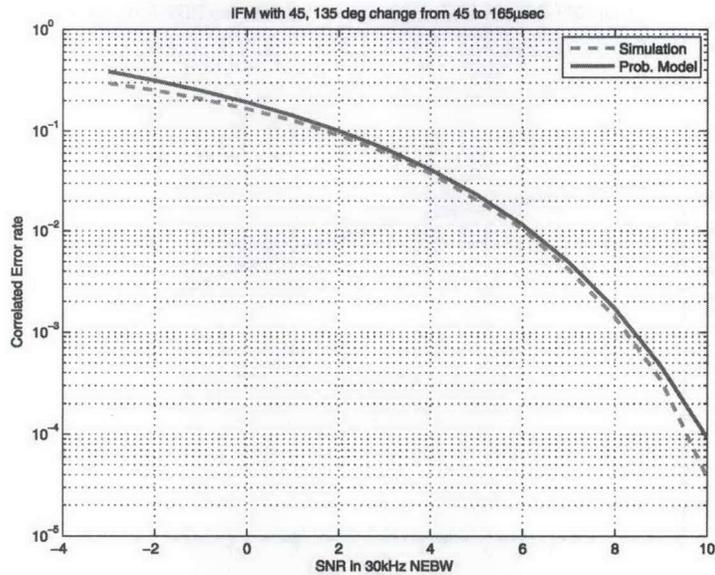


Fig. 12. IFM probability of error versus SNR (4 level).

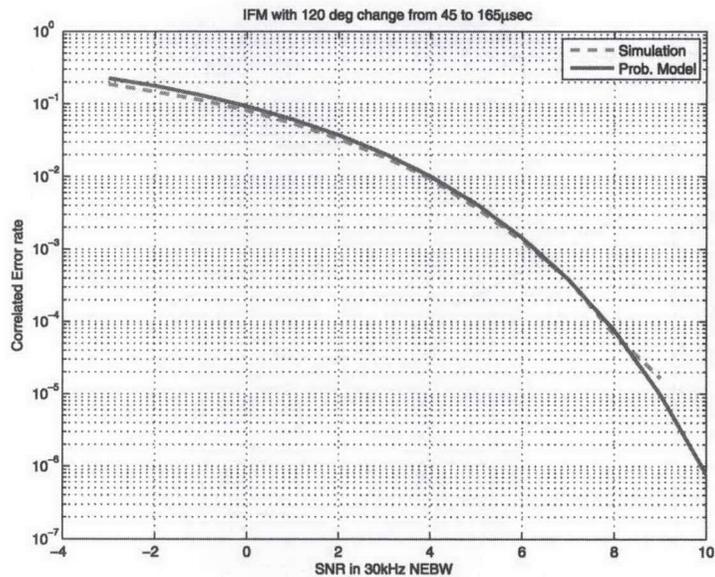


Fig. 13. IFM probability of error versus SNR (3 level).

specific noise level and specific interference level for a pulse that interferes at $70 \mu\text{s}$ after the beginning of the primary pulse.

E. Data Performance

The modulation techniques discussed can be modified and combined in many ways resulting in many different implementations. The matched filter model allows us to qualify the data rate performance of each implementation. This allows us to pare down the variations to a handful of choices that offer data transmission with an acceptable error rate. First, the data rate for each implementation must be calculated.

The calculation of data rate for any implementation is relatively straightforward. Effectively it is dependent

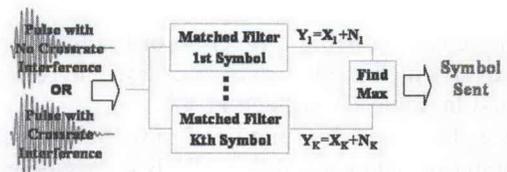


Fig. 14. General match filter for interference and noise on Loran.

on the number of modulation levels per pulse and the number of modulated pulses per GRI. It is assumed that the first two pulses of a GRI are not modulated to preserve blink. Additionally, some consideration must be given for design features that mitigated degradation of navigation performance. For PPM, balanced coding is often employed to reduce navigation error for legacy users. This effectively reduces the data rate

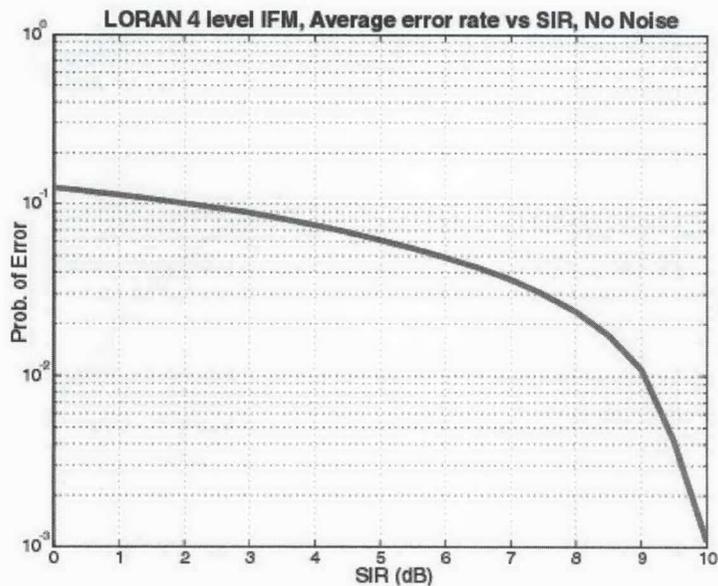


Fig. 15. Average error rate for interference from another Loran signal.

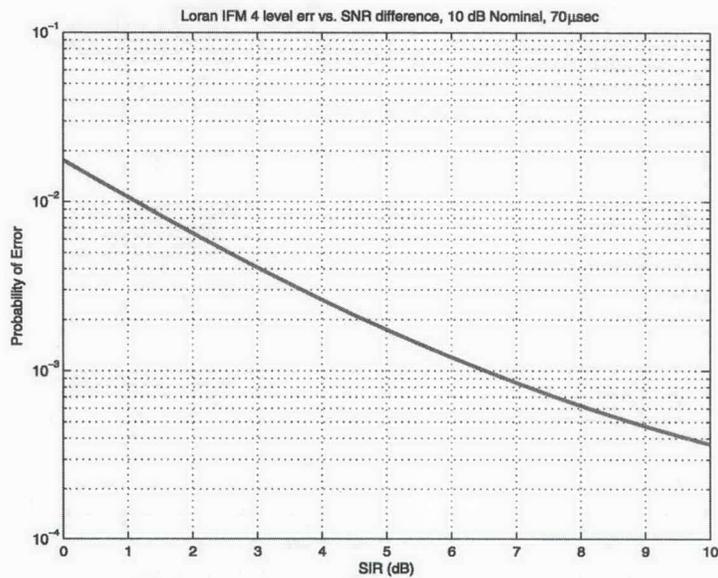


Fig. 16. Error rate versus SIR for a 10 dB Loran signal and an interfering signal delayed 70 µs.

with the reduction being a function of the number of pulses over which the balancing occurs. The result is presented in Table III in Section VA.

Since IFM code does not need to be balanced, the amount of data carried per GRI is $6 * n$ where n represents the number of bits per pulse. A combination scheme can use both IFM and PPM on the same pulse resulting in an increased data rate. The performance when combined will suffer when compared with each modulation used individually and the degradation depends on the type of IFM and PPM used. The effect can be mitigated if the receiver estimates one form of modulation, removes the effect of that modulation and then estimates the other modulation.

SIM data rate depends on several factors. There are several ways of encoding data on the additional pulses. One method would be to include PPM and/or IFM on the additional pulses. For example, one can create a hybrid data scheme with a data rate of over 330 to 380 bit/s if IFM (three or four level) and PPM (three level) are used on a pulse train of 16 pulses. Furthermore, more extreme modulation can be used since the pulse is not used for navigation. However, adding six or eight SIM pulses per GRI effectively doubles the CRI. Since it is modulated, it cannot generally be cancelled. A reasonable compromise is to add and modulated only one pulse per GRI. This is what is done under NPC [9].

Table II presents the data rate achieved with each modulation. To make each data rate comparison equal,

TABLE II
Summary of Modulation Data Rate

| Modulation Method | Data/GRI | Data Rate (bps, 10 GRI/s) |
|---------------------------|----------|------------------------------|
| PPM 3 level | 7.1396 | 71.396 |
| PPM 5 level | 10.7740 | 107.740 |
| IFM 3 level | 9.5098 | 95.098 |
| IFM 4 level | 12.0000 | 120.000 |
| IFM 16 Level | 24 | 240 |
| IFM 3 level & PPM 3 level | 16.6488 | 166.488 |
| IFM 4 level & PPM 3 level | 19.0196 | 190.196 |
| SIM IFM 3 & PPM 3 | 35.1906 | 351.906 |
| SIM IFM 4 & PPM 3 | 40.1710 | 401.710 |
| SIM IFM 4 & PPM 5 | 48.2252 | 482.252 |
| 9th Pulse PPM | 5 | 50 |

it is assumed that there are only 10 GRI/sec and that FEC requires half of the bits. Another assumption is that SIM doubles the number of modulated pulses (from six to twelve). Notice that using PPM and SIM results in greater than double the data rate using the same PPM without SIM. This results from extending the balanced modulation from six to twelve pulses (see Table III).

This data rate should be qualified by a measure of the error rate. One possibility is to use SNR required to achieve a specific error probability such as 0.1% symbol error. This can be done with the models presented. The comparison is shown in the summary. The addition of that factor helps complete the story on data performance. However, data performance is only one of three important considerations. The next two sections discuss the other considerations and develop a full matrix to design decisions.

V. NAVIGATION RECEIVER

While it is desirable for the signal design to yield a high data rate, the design must maintain the position fixing capabilities of Loran. An ill conceived design may result in increased interference, poorer phase tracking and lower received signal power for navigation Loran users. This degrades the utility and performance of Loran for navigation. This consideration is particularly important given that the primary role for Loran in the 21st century is to provide redundancy to GPS applications. Another consideration is the effect of a design on legacy users. If possible, some utility to legacy users should be retained.

The following section discusses and analyzes the impacts that various communication schemes may have on navigation performance. The changes may alter Loran receiver phase and such induced phase errors would lead to greater positioning errors. Many of these factors are discussed in [19] and hence will only be briefly touched upon. Analysis of phase offset, skywave rejection, and received power helps to

TABLE III
Balanced Sequences

| Pulses per Set | Levels | Balanced Sequences | Bits | GRI/set | Bits/GRI |
|-------------------|--------|-----------------------|--------|---------|----------|
| 6 | 3 | 141 | 7.140 | 1 | 7.140 |
| 6 | 5 | 1751 | 10.774 | 1 | 10.774 |
| 7 | 3 | 393 | 8.618 | 1 | 8.618 |
| 7 | 5 | 8135 | 12.990 | 1 | 12.990 |
| 12 (6/GRI) | 3 | 73789 | 16.171 | 2 | 8.086 |
| 12 (6/GRI) | 5 | 19611175 | 24.225 | 2 | 12.113 |

quantify the effect of the modulation on the navigation capabilities of existing receivers.

A. Induced Phase Offsets

One way by which modulation design may degrade navigation performance is by inducing phase offsets. While a phase offset on a single pulse is acceptable, a good design should eliminate phase offsets averaged over the pulses in a few GRIs or PCIs. This is because a Loran navigation receiver integrates over several GRIs forming an average pulse. This average pulse is used to determine timing by examining the phase to find the standard tracking point. For PPM, if the total time advances are not equal to its total time delays over the averaging period, a phase error is induced. As a result, balanced coding is generally used.

Balanced coding is a method where the time/phase shifts are coordinated so that the sum of the time/phase shifts is zero over a specified period. In the case of Eurofix, this period is one GRI with only the last six of the eight pulses per GRI used. Eurofix modulates each pulse advancing the pulse 1, 0, or $-1 \mu\text{s}$ allowing for three different symbols per pulse. Eurofix has 141 (7.1 bits, i.e., $2^{7.1} = 141$) balanced sequences over one GRI. If we balanced an evenly spaced five level modulated code, 1751 (10.8 bits, i.e., $2^{10.8} = 1751$) balanced sequences are possible. The calculation can be done for longer pulse sequences. Twelve modulated pulses may be formed using two GRIs for Loran-C or one GRI for Loran-D or SIM. Loran receivers typically have integration times that equal or exceed two GRIs so being balanced over one or two GRIs is adequate for eliminating the effect of phase offsets. The balanced sequence can be translated to bit sequences through the use of a look-up table. Table III shows the number of balanced sequences for different combinations of levels per pulse and pulses per set.

Using increasingly complex balanced coding schemes necessitates larger look-up tables for balanced sequences. The additional information requires the receiver to have more storage and processing power. So while using a code that is balanced over two GRIs yields better data

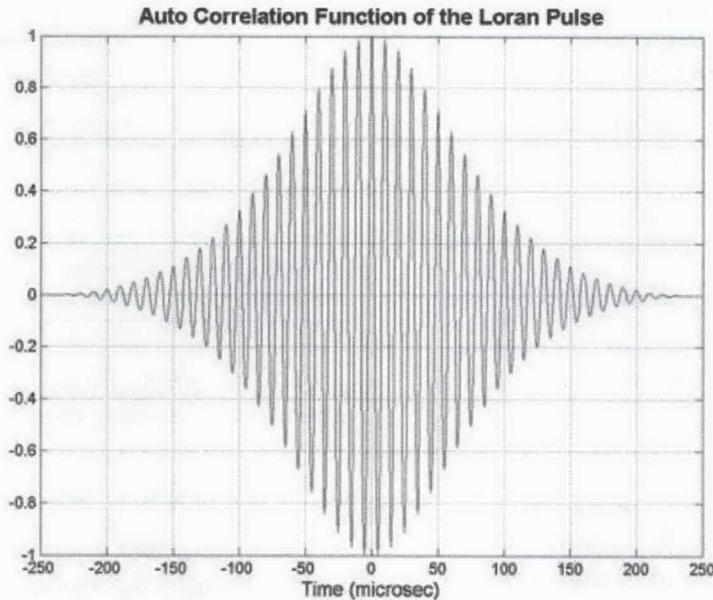


Fig. 17. Autocorrelation function of a Loran pulse, $c(\tau)$.

performance than using two sets of codes balanced over one GRI, it requires more powerful receiver hardware.

Balanced coding is not necessary for IFM since the timing of the Loran pulse is not altered. The phase of the pulse at the 25–30 μs point is unchanged for all IFM schemes. The average three or four level IFM Loran pulse does not greatly differ from a normal Loran pulse until well past the 70 μs point. This can be seen in Fig. 20.

B. Skywave Interference

The Loran pulse structure is designed to minimize the effects of skywave interference. Such interference could lead to a bias error in the receiver, particularly for a legacy receiver.

Recall skywave interference is caused by a delayed copy of the Loran signal (the skywave) arriving coincident with the ground wave. The delay can be decomposed into an integer number of pulse separations or “coarse delay,” d , plus a remainder term or “fine delay.” The fine delay time should not exceed the time between two consecutive pulses (1 ms). The fine delay of the skywave, represented by τ , is the relative phase between the skywave and ground wave. This definition allows us to divide skywave into two categories. Short delay skywave are skywaves that have no coarse delay while long delay skywaves have a coarse delay of one or more pulses intervals.

Interference from short delay skywaves is mitigated by tracking early on the pulse. This mitigation is equally effective for PPM and IFM since the early portion of the pulse is not altered. SIM may cause increased amount of skywave which may adversely affect legacy receivers that are not designed

to handle the additional interference. A subcategory is early skywave which are delayed less than 30 μs . This interference is particularly problematic for all receivers. Fortunately, it is reasonably rare.

Interference from long delay skywaves is mitigated by complementary phase coding. Recall (3) which represented the received groundwave or skywave over one PCI (two GRIs). Assume that the signal passes through a set of correlators or filters that are matched to the expected signals over one PCI. The correlators are aligned with the arrival of nominal signal. These correlators perform like the match filters in the previous section. As derived in [19], the skywave interference results in the output I_s shown by (19), after passing through the receiver correlators. Generally, $c(\tau)$ is the correlation function between the receiver correlator and the Loran pulse $s(t)$. Since the correlators are matched to the standard Loran-C pulse, $c(\tau)$ is the autocorrelation function of the Loran signal. The Loran autocorrelation function is an even periodic function with a period of 10 μs as seen in Fig. 17. I_s , normalized per pulse, represents the average effect of skywave interference over one PCI. Hence this is a quantity of interest for examining the effect of modulation on skywave rejection in legacy receivers.

$$I_s = \sum_{i-d=1}^{i-d=8-d} x_i x_{i-d} c(\tau) + \sum_{i-d=9}^{i-d=16-d} x_i x_{i-d} c(\tau) \quad (19)$$

$$I_s = \sum_{j=1}^{j=8-d} x_{j+d} x_j c(\tau) + \sum_{j=9}^{j=16-d} x_{j+d} x_j c(\tau)$$

$$I_s = c(\tau) \left[\sum_{j=1}^{j=8-d} x_{j+d} x_j + \sum_{j=9}^{j=16-d} x_{j+d} x_j \right], \quad d \in [1, \dots, 7]. \quad (20)$$

As seen by Fig. 2 and (21), the phase code is complementary over two GRI so that the summation is zero for $d > 0$. Hence, I_s is nominally zero.

$$\sum_{j=1}^{j=8-d} x_{j+d}x_j + \sum_{j=9}^{j=16-d} x_{j+d}x_j = 0 \quad \text{for } d \in [1, \dots, 7]. \quad (21)$$

Equation (19) needs to be modified for assessing the skywave rejection performance of legacy receiver for various Loran modulation schemes. While the receiver correlators remain the same, the skywave signal has modulation and the resulting correlation function $c(\tau)$ will depend on the interfering modulated skywave signal. The groundwave of this signal arrived " d " pulses prior. Hence we get (22) where $c_i(\tau)$ is the correlation between the normal Loran pulse and the i th Loran pulse in the PCI. The i th pulse may be data modulated.

$$I_s = \sum_{i-d=1}^{i-d=8-d} x_i x_{i-d} c_{i-d}(\tau) + \sum_{i-d=9}^{i-d=16-d} x_i x_{i-d} c_{i-d}(\tau). \quad (22)$$

The question of concern is how each modulation affects a legacy receiver's ability to reject long delay skywave. The next two subsections examine that question for PPM and IFM, respectively.

1) *PPM Long Delay Skywave Rejection:* For PPM, skywave induced offsets are dependent on the size of the time shifts. Assume the correlation is made to nominal (unmodulated) pulses. Let n_j be the coding on the j th signal and Δt is the size of the time shift. Then $c_i(\tau) = c(\tau + n_i \Delta t)$, the autocorrelation of the Loran pulse, and (22) becomes (23)

$$I_s = \sum_{i-d=1}^{i-d=8-d} x_i x_{i-d} c(\tau + n_{i-d} \Delta t) + \sum_{i-d=9}^{i-d=16-d} x_i x_{i-d} c(\tau + n_{i-d} \Delta t) \quad (23)$$

$$I_s = \sum_{j=1}^{j=8-d} x_{j+d} x_j c(\tau + n_j \Delta t) + \sum_{j=9}^{j=16-d} x_{j+d} x_j c(\tau + n_j \Delta t).$$

Since we are analyzing the effect of modulation on the complementary phase code, only skywaves with coarse delays of at least one pulse length are examined. These are the skywaves that the phase coding is designed to eliminate. As such, for each GRI, we assume that there are at most seven pulses having skywave interference, i.e., $d = 1$.

Now we can calculate the worst case value of I_s . The worst case can be constructed as such. Have

TABLE IV
Selected PPM Skywave Rejection Calculations

| Max $n_j \Delta t$ (μ s) | Max I_{sn} | Mean I_{sn} |
|-------------------------------|--------------|---------------|
| 0.5 | 0.1929 | 0.0416 |
| 1.0 | 0.3669 | 0.0791 |
| 2.0 | 0.5934 | 0.1279 |

Note: At worst case: $\tau = 2.5 \mu$ s and $d = 1$.

all positive phase code products ($x_{j+d} * x_j$) pulses associated with one modulation with correlation c_p . Have all negative phase code products associated with a different modulation with correlation c_n . Denote the cross correlation function of a normal Loran pulse with each of these signals as c_p and c_n , respectively. If this is done for a coarse delay of 1 and assuming six modulated pulses, we get (24). The worst case generally occurs when c_p and c_n are large in magnitude and have opposite signs.

$$I_s = \sum_{j=1, j \neq 9}^{15} x_j x_{j+1} c_j(\tau) = \sum_{j=1}^6 (c_p(\tau) - c_n(\tau)). \quad (24)$$

For ease of comparison, calculate a normalized I_s value such that the nominal value is one for no delay, i.e., $\tau = d = 0$ and $c(0) = 1/16$. Denote this value as I_{sn} . One interpretation of the value is as the ratio of skywave interference to signal power given that the skywave and signal were at the same power. This implies the maximum value of I_{sn} using (24) is 0.75.

Table IV shows some of the effects of PPM on long delay skywave rejection. The result show that the maximum value occurs at a fine delay of 2.5μ s and a coarse delay of 1. The second and third columns are the maximum values determined using the worst case and the average of balanced data combinations, respectively. The average skywave rejection, as seen from Fig. 18 and Table IV, is generally far lower than the worst case. The table illustrates some trends: interference gets worse as we increase spacing (Δt) or code levels (n_j). Modulation spacing governs how easily the receiver can identify the different symbols (hence, probability of error) and code level governs how much data can be included in one pulse. Hence the table shows that higher data rates or modulation spacing (large $n_j \Delta t$) come at the expense of decreased long delay skywave rejection in legacy receivers.

Modern receivers can have better long delay skywave rejection since they can use processing to mitigate the effects of modulation. Since the groundwave data bits arrive first, they are known prior to the arrival of the skywave. This knowledge of data can be used to set the correlators so that the time shift on the signal is eliminated.

2) *IFM Long Delay Skywave Rejection:* The IFM modulation techniques will also result in skywave offset. An analysis method similar to that used for PPM can be employed. Since the modulated pulses

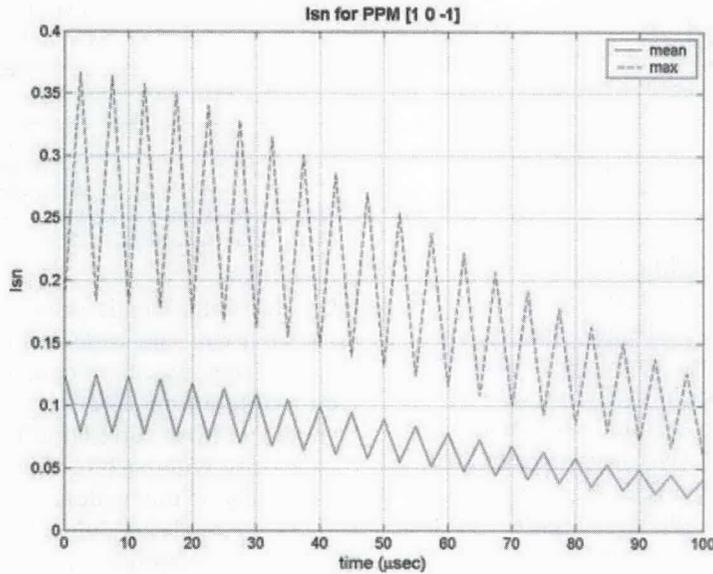


Fig. 18. PPM 3 I_{sn} for various τ and $d = 1$.

will likely differ from each other, we denote the each individual pulse by a subscript. This is necessary since the modulated pulses are not time-shifted versions of the same pulse but rather different waveforms. Denote the i th IFM Loran pulse as $s_i(t - iT)$. The received signal or skywave is

$$p(t) = \sum_{i=1}^8 x_i s_i(t - iT) = \sum_{i=1}^{16} x_i s_i(t - iT). \quad (25)$$

Let $c_j(\tau)$ be correlation function of the j th IFM Loran pulse with a nominal Loran pulse. The correlator output after using the phase coding can be written as

$$I_s = \sum_{i-d=1}^{i-d=8-d} x_i x_{i-d} c_{i-d}(\tau) + \sum_{i-d=9}^{i-d=16-d} x_i x_{i-d} c_{i-d}(\tau) \quad (26)$$

$$I_s = \sum_{j=1}^{j=8-d} x_{j+d} x_j c_j(\tau) + \sum_{j=9}^{j=16-d} x_{j+d} x_j c_j(\tau).$$

If all the IFM signals were coded the same way, the result of the skywave interference would be zero since Loran phase code is complementary and $c_j(\tau) = c_i(\tau)$ for $\forall i, j$. The worst case condition for IFM is similar to that for PPM.

In Table V, a three level format ($-120, 0, 120$ deg) and a four level format ($-135, -45, 45, 135$ deg) are compared. The "average" value is a weighted combination of this I_{sn} for each possible phase shift. Fig. 19 shows the mean and maximum I_{sn} for the four level IFM with different fine delays between skywave. As expected, worst case long delay skywave rejection for the four level scheme is worse than that of the three level. However the average for the four level IFM is not significantly worse than that of three level IFM. The examination shows that for both PPM and IFM skywave performance, there are some residual

TABLE V
Selected IFM Skywave Rejection Calculations

| IFM Phase Shift in $120 \mu s$ | Max I_{sn} | Mean I_{sn} |
|-----------------------------------|--------------|---------------|
| 3 level | 0.1903 | 0.0506 |
| 4 level | 0.2059 | 0.0514 |

Note: Max and Max Mean I_{sn} occur at different skywave offsets.

errors for long delay skywaves. While the average effect is generally small unless extreme forms of modulation are used, there are circumstances where the effect can be large. This may not be acceptable for a Loran system that supports aviation since extreme cases are necessary considerations.

C. Cross Rate Interference

CRI is another consideration. As previously mentioned, CRI can be cancelled out if there is no data modulation. If CRI cannot be cancelled, the interference must be "blanked out" thereby eliminating all signals in that interval. This reduces the number of pulses per second available for navigation. The result is a degraded SNR and hence more error on the navigation signals. With data modulation, cancellation is currently not possible without knowledge of the data. Knowing the bits allows for the modulation to be removed or "wiped off." One method is for getting the data bits is to have all stations to transmit the same data. Decoding one station, usually the strongest, is used to provide the data bits. For example, this can be used if all stations transmitted the WAAS message. A delay is incurred from having to process the entire message before data wipe off is possible.

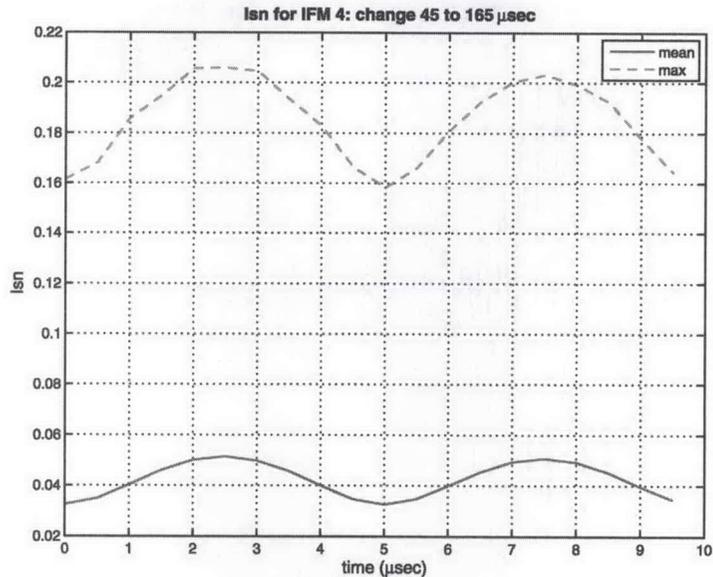


Fig. 19. IFM 4 I_{SN} for various τ and $d = 1$.

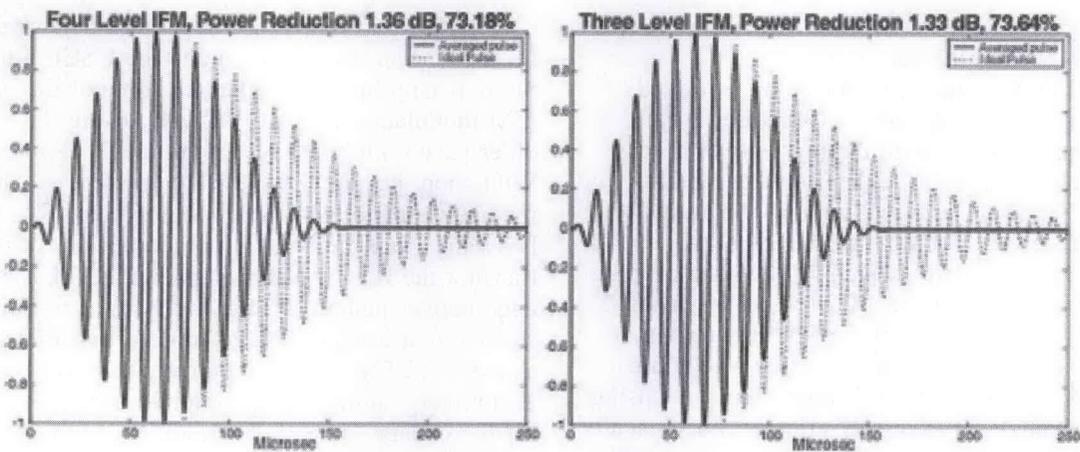


Fig. 20. Four and three Level IFM average pulse.

However, in many applications envisioned, stations will be transmitting different data. For example, differential Loran will either require high data rates (> 250 bit/s) or have each station limit data transmission to only corrections applicable to its coverage region. In order to wipe off data from the navigation signal, the data must be decoded and validated. Cancellation is not possible for low power signals since demodulation is difficult. It also incurs a delay. In this case, the best way to reduce CRI effects on navigation is to minimize the number of pulses modulated.

CRI cancellation is a relatively new technique and is not employed in many legacy receivers. For legacy users, modulating existing pulses does not increase the deleterious effect of CRI. However, the addition of new pulses will increase CRI.

D. Received Power

Another outcome in altering Loran pulses is a potential reduction in received power for legacy

receivers. The modulation generally does not reduce transmitted power nor does it change the pulse shape before the $60\text{--}70\ \mu\text{s}$ —traditional tracking locations. However, there can be a reduction in average received power. A reduction in received power can result in either reduced Loran coverage or reduced Loran availability or accuracy. Experiments and analyses can be conducted to determine the expected loss to receivers that are unaware of the modulation. Since different receivers use different methods for acquiring and tracking the signals, they will respond to the modulation techniques differently. However, the analysis of average signal power will determine a baseline for the loss a legacy user can expect.

Field tests of Eurofix have shown that there is some degradation on normal Loran-C receiver signal power due to PPM time shifts but these changes are small (0.79 dB) [20]. Future systems with knowledge of the PPM can eliminate this loss. While smaller PPM time shifts will reduce undesirable loss in

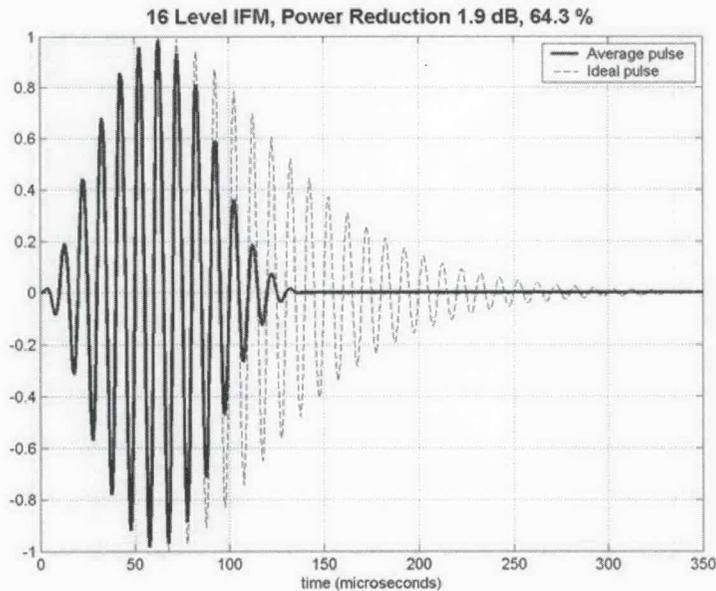


Fig. 21. 16 level IFM average pulse.

tracking signal power, it will also increase the SNR necessary to decode the data bits.

The degradation due to IFM can be examined by looking at the average pulse. Generally Loran navigation receivers average the pulses over at least one GRI. Assuming that the data bits are randomly distributed, the average signal seen by the Loran navigation receiver should resemble the average of all possible pulse waveforms. Fig. 20 shows the average pulse for three and four level IFM, respectively. The amplitude is normalized to the maximum amplitude. A receiver utilizing the sixth zero crossing should not be adversely affected. Nor should receivers using envelope detection be adversely affected since usually only the rising envelope is used. However, matched filters will experience an average power degradation of 1.33 dB and 1.36 dB for the two modulation techniques. This power degradation can be mitigated with knowledge of the transmitted waveform. Hence it only has effect on legacy receivers. Fig. 21 shows the average pulse for the 16 level modulation technique. The loss is more severe. Increasing the envelope of the 16 level pulse does help mitigate the loss. This comes at the expense of increased power consumption.

VI. TRANSMITTER

The feasibility of the modulation techniques and the costs of making the required changes to the transmitter are of great importance to the design. PPM has been demonstrated at many Loran stations. The USCG deployed one form of PPM, TPC, in the late 1970s. Eurofix PPM is currently operating in the Northern Europe Loran System (NELS). Ninth pulse and Eurofix have both been demonstrated using the Wildwood, NJ transmitter as well as numerous other transmitter sites.

IFM poses a more significant technical challenge for the transmitter. Peterson and the USCG Loran Support Unit have implemented some of the proposed IFM modulation (4 level, 16 level) using the older vacuum tube type transmitters (TTX) at the Wildwood, NJ and Tok, AK Loran stations with few modifications to the physical equipment. The difficulty lies with using the SSX. It is theoretically possible to transmit the IFM waveforms using SSX. Discussions and analysis generated after the introduction of IFM suggest that it is possible to modify transmitters such as the AN-FPN/64 to successfully transmit IFM. Preliminary studies of the requirements and costs for the change indicate that additional room and equipment will be necessary. Rough cost estimates for IFM on all U.S. transmitters is about \$40 Million.

SIM should be feasible using current Loran transmitters. While some Loran stations are limited in the number of pulses transmitted for a given period, any single rated station should not exceed this limit even when transmitting supernumerary pulses.

VII. SUMMARY

No single modulation method is clearly better—each has strengths and weaknesses. The modulation method implemented will depend on the relative desirability of each factor and relative strengths of each design.

The assessments presented allow for the comparison of the designs by creating a matrix of the major associated benefits and costs. Each implementation should be compared by data rate and SNR necessary (for a given error rate). Other factors such as receiver SNR degradation, receiver compatibility and transmitter modifications should also be factored in. Some examples are compared in

TABLE VI
Data Modulation Assessment Summary

| Modulation Method | Data (bps at 10 GRI/sec) | SNR for $Pe < 10^{-3}$ | SNR degrade (dB) | Skywave Rejection | Transmitter Modifications |
|--------------------------------------|-----------------------------|---------------------------|------------------|----------------------|------------------------------|
| PPM 3 level | 71.4 | 8.32 | 1.18 | 2 (Good) | 2 (Small) |
| PPM 5 level (1 msec) (1 msec) | 107.7 | 8.49 | 3.7774 | 3 (Fair) | 2 (Small) |
| PPM 5 level (0.5 msec) (0.5 msec) | 107.7 | 14.39 | 0.8768 | 2 (Good) | 2 (Small) |
| IFM 4 level | 120.0 | 8.57 | 1.36 | 2 (Good) | 4 (Significant) |
| IFM 16 level | 240.0 | 7.78 | 1.9173 | 3 (Fair) | 4 (Significant) |
| Ninth Pulse | 50.0 | 7.97 | 0 0.02* | 1 (V. Good) | 2 (Small) |

Table VI. There is not necessarily an apples to apples comparison. For example, the SNR degradation for ninth pulse modulation is calculated based on two models. Ninth pulse, unlike the other modulation methods, results in no change to the navigation signals so nominally there is no degradation. Another way of examining ninth pulse effects on SNR degradation is to consider its effect in increasing CRI. This lowers SNR since modulated pulses need to be blanked to eliminate CRI thus reducing the number of navigation pulses available. This SNR degradation, only examined for NPC, is estimated by assuming interference from 10 stations each transmitting 10 modulated pulses per second. For the receiver and transmitter compatibility, the table uses a quantitative rating with from 1 to 5 where 1 is very good. In terms of transmitter modification, 1 represents minimal modification and 5 represents an extreme amount of modification.

VIII. CONCLUSIONS

For Loran, as with GPS and other navigation systems, the ability to carry data is becoming increasing vital. A Loran data capacity allows it to meet many current and coming needs, particularly for safety of life and commercially critical services. This paper detailed and described the major techniques being used to modulate data on Loran. It derived analytical tools and developed guidance for the modulation designs. The analyses and discussions help determine the benefits and costs of each design. No design is perfect for all situations. In the companion examples paper [16], an IFM design and a modified PPM (on a supernumerary pulse) design are presented. Each of these designs represented the best solution to a given requirement with given constraints. The follow on paper shows some examples of how the modulation concepts developed in this paper may be put into practice.

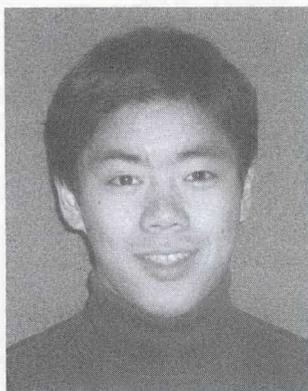
ACKNOWLEDGMENT

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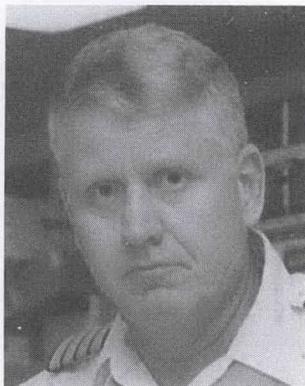
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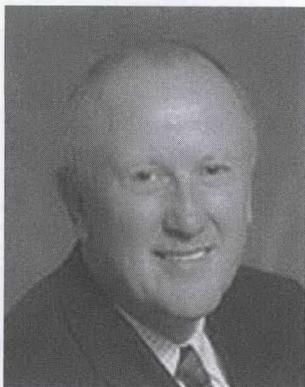
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