Loran Data Modulation: Extensions and Examples

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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. This paper details some recent Loran modulation designs and ideas.

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I. INTRODUCTION

Loran (or long range navigation system) has been an operational navigation system since 1958. Unlike the Global Positioning System (GPS), the preeminent navigation of the last decade, it is a terrestrial, low frequency, high power system. These features make Loran a good complement to providing backup to GPS in position, navigation, and timing (PNT) applications. As GPS plays an increasingly important role in the national safety and economic infrastructures, having a backup has become more vital. This point is highlighted by the findings of the Volpe National Transportation Safety Center (VNTSC) report on GPS vulnerability [1] and by a recent Presidential Decision Directive [2]. In fact, the capabilities of Loran give it the potential to provide backup in many applications from aviation approaches to maritime harbor entrance approach (HEA) to timing and frequency. Loran is one of the few existing systems considered capable of serving such diverse PNT applications. The requisite question to answer is "how can the current Loran system, with reasonable modifications, provide the maximal benefit for backing up the critical PNT infrastructure?"

That question is certainly difficult to answer. A Loran evaluation team commissioned by the Federal Aviation Administration (FAA) has spent the last few years answering the question. As a result, Loran was scrutinized to an unprecedented level due to the strict requirements of many of the desired applications. Many changes were determined to be necessary to create a Loran system that meets the requirements of nonprecision approach (NPA), HEA, and Stratum 1 frequency and timing on the level of tens of nanoseconds [3]. One important addition is that of a data message on Loran capable of supporting these operations. Previous papers discussed Loran modulation [4–6] and these will be relied upon to give the necessary background for this discussion. Background on Loran can be found in [7]. This paper focuses on some of the designs developed by the Loran evaluation team.

Three data modulation design ideas for the next generation Loran in the United States are discussed here. These designs are generically termed Loran data channel (LDC). The first is the high speed Loran data channel (HS-LDC). This design was created to support broadcast of the wide area augmentation system (WAAS) message [8] with some compatibility with legacy Loran receivers. Transmitting the full WAAS message requires a data rate of at least 250 bit/s. WAAS provides increased accuracy for GPS and the integrity necessary for landing aircraft with vertical guidance. The second LDC design is ninth pulse communications (NPC) which is used to support Loran for NPA, HEA, and timing and frequency. This design minimizes interference on the navigation

pulses by placing all the modulation on one additional pulse. The result is a data rate of roughly 50 bit/s or more. The third idea for implementing LDC is termed scalable Loran modulation (SLM). The goal of the design is to generate a system that can scale up in data rate thus giving legacy users a transition period to upgrade to new receivers.

II. REVIEW OF DATA MODULATION

This section provides a quick overview of Loran modulation. While many forms of data modulation exist, there are only a few that are suitable for Loran. The data modulation design must work with the Loran transmitter equipment, retain navigation/timing capabilities, minimize interference, and maintain spectrum. Additionally, maintaining some legacy receiver compatibility is preferable. These modulations are discussed and assessed in great detail in [4] and are only briefly described here. These modulations can be divided into three basic techniques.

Pulse position modulation (PPM) shifts the arrival time of the Loran pulse to transmit data. Small time shifts, when implemented correctly, generally have little adverse effect on navigation receivers. It is also relatively reasonable to implement in nearly all Loran transmitters. This includes both the solid state and tube type transmitters (SSX and TTX, respectively). However, since time shifts must be small, this generally limits the data capacity.

Intrapulse frequency modulation (IFM) places data on the Loran pulse by altering the frequency to induce a phase shift. Having this frequency change occur after the standard tracking point minimizes its effect on navigation and tracking. The frequency shift is selected to maximize phase separation between different symbols while maintaining navigation and spectrum requirements. IFM allows for more data per pulse with similar or better performance than PPM. However, implementing IFM, while reasonable for TTXs, is nontrivial for SSXs. It requires extensive changes to the output network as well as modifications to the current timing frequency equipment (TFE) and transmitter control console (TCC).

Supernumerary interpulse modulation (SIM) is the addition of pulses to the standard set of eight pulses per group repetition interval (GRI). The pulses can be modulated using PPM, IFM, or other means (amplitude modulation, phase code change, etc.) if these pulses are not also used for navigation. Care must be made in modulating these pulses so that legacy compatibility is maintained and interference is minimized.

As a means for assessing and comparing data modulation, Euclidean distance is a reasonable measure. It is certainly true for Loran communications

[4]. The Euclidean distance d_{jk} is defined by (1) where $s_j(t)$ and $s_k(t)$ represent two transmitted symbols

$$d_{jk} = \sqrt{\int [s_j(t) - s_k(t)]^2 dt}. \qquad (1)$$

III. HIGH SPEED LORAN DATA CHANNEL

A. Background

Recent years have seen efforts to both modernize Loran and increase the capabilities of Loran. One goal was for Loran to broadcast satellite based augmentation system (SBAS), such as the U.S. system, WAAS, messages which provide integrity and correction to GPS for aviation. SBAS messages are currently transmitted on GPS L1 (1575.42 MHz) from geostationary satellites. The addition of the message on Loran would allow the SBAS message to reach users at high latitudes and in areas occluded from the geostationary satellites. The occlusion can occur because of foliage, mountainous terrain, aircraft banking, and canyons. An SBAS message channel on Loran can increase the coverage of SBAS in these areas thereby extending its benefits. This is important as the benefits of SBAS are many and can be enjoyed by numerous user groups. There are many nonaviation users of the SBAS message since it increases GPS accuracy by a factor of four.

The SBAS concept is a global one. Many systems are being built such as the European Geostationary Navigation Overlay System (EGNOS), the Japanese Multifunctional Transport Satellite Augmentation System (MSAS) and the Indian GPS and Geostationary Augmented Navigation (GAGAN). SBAS on Loran can thus be a global solution to providing redundancy for GPS SBAS.

The SBAS message is a 250 bit message transmitted at 1 Hz. Half-rate convolutional code is added to mitigate the random errors experienced from the geostationary transmission. The result is an overall transmission rate of 500 bit/s which is decoded to the 250 bit, 1 s message. For Loran to broadcast the WAAS message, it must be able to broadcast 250 bits plus additional bits necessary for forward error correction (FEC) per second. In addition, maintaining the ability to support legacy users was also initially deemed necessary. Different means of data transmission were examined [4, 5] and it was decided that IFM offered the only solution that met these constraints.

B. Design

A 16 level IFM design was created to achieve the desired data bandwidth. Six pulses per GRI are

modulated with each pulse carrying four bits. It is assumed that the station broadcasts a minimum of 20 GRI per second. This results in a throughput of 60 bytes per second. The data rate is necessary to carry the 250 bit WAAS message (parsed into 32 bytes) and 28 bytes of FEC at 1 Hz.

The design involves several changes to the nominal Loran-C operations. First, the frequency must be shifted to achieve the desired phase difference. Fig. 1 shows the phase difference between the pulses representing each of 16 symbols and the nominal Loran pulse for the HS-LDC tested in 2001. Note that some symbols of the modulation need to be initially offset from the nominal Loran pulse to achieve the desired phase change. Another modification necessary is an increased pulse envelope and maximum power band as shown in Fig. 2. This guarantees that the power spectrum is maintained and provides additional separation between each symbol. One additional condition is that there should be 20 or more Loran groups (one group = 8 pulses) or GRIs per second. This results in a data rate of 480 bit/s or more. Currently, only dual-rated stations and stations with a GRI of 5000 (0.05 s) or less can meet this requirement. However, for a dual-rated station, the target is to have 21 or more Loran groups per second since Loran groups are occasionally not transmitted because of cross-rate blanking at the station. Cross rate blanking is the nontransmission (or blanking) of the pulses of one GRI during a given time interval. This is done by a dual-rated station when the station has to transmit the pulses for both rates at roughly the same time. Not all dual-rated stations will meet the 21 Loran groups per second requirement. If a dual-rated station transmits for two chains each with GRI greater than 9500 (0.095 s), then the average number of Loran groups transmitted per second will be less than 21. Only Searchlight, Nevada (GRI 9610 and 9940) falls into this category. It transmits data on an average of 20.4 Loran groups/second. Most dual-rated stations average at least 23 pulse groups/second leaving at least three pulse groups which can be blanked during any given second. As a result of these changes, the 16 level IFM technique may not be fully compatible with all legacy users.

The IFM design modulates using two stages with four different phase change combinations for each stage. Analysis shows that this is better than having one stage with 16 different levels. The one-stage implementation, seen in Fig. 3, implements the frequency modulation from 40 to 140 μs after the start of the pulse. This is compared with a simple two-stage design with no initial phase offset. The phase offset for the design is seen in Fig. 15. It implements frequency change from 40 to 140 and 150 to 250 μs after the start of the pulse. The bit error rate versus signal-to-noise ratio (SNR) for the one-stage and two-stage designs is seen in Fig. 4. However,
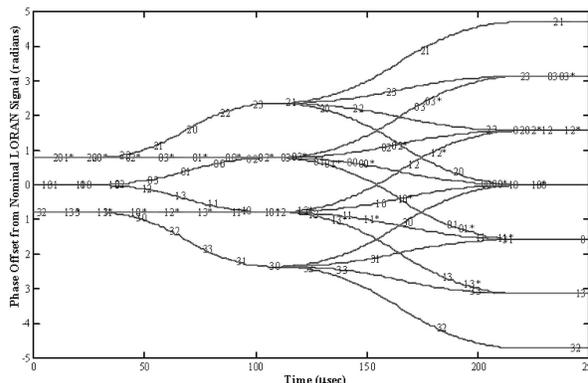
Notes: A. 1st numbers are 2 bits due to 1st phase, 2nd number 2 bits due to 2nd phase.

B. Phase levels are gray coded.

C. Some offset in phase in leading edge to minimize frequency spreading.

D. '00' and '00*' are same data, picked to zero TOA bias to legacy users.

Fig. 1. Two stage 16 level IFM phase offset relative nominal Loran pulse (as tested in Alaska in August 2001).
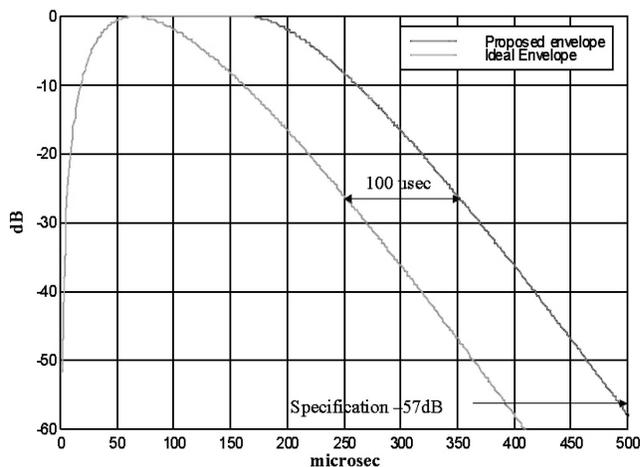

Fig. 2. Proposed 16 level IFM envelope modification.

TABLE I
IFM Spectrum

Design	IFM 4 (40–140 μs)	IFM 16 1 stage (40–140 μs)	IFM 16 2 stage (40–140, 150–250 μs)
Spectrum (90–110 kHz)	97.6%	95.5%	98.2%

as seen in Table I, the one-stage implementation has significantly more energy out of spectrum. For the two-stage implementation, the best performance occurs when the energy in the first stage is balanced with that in the second stage so that the stages have comparable error rates (versus SNR).

The message design is not discussed in this paper. The message is the WAAS message plus six unused bits. The WAAS message is defined in the WAAS minimum operational performance specifications
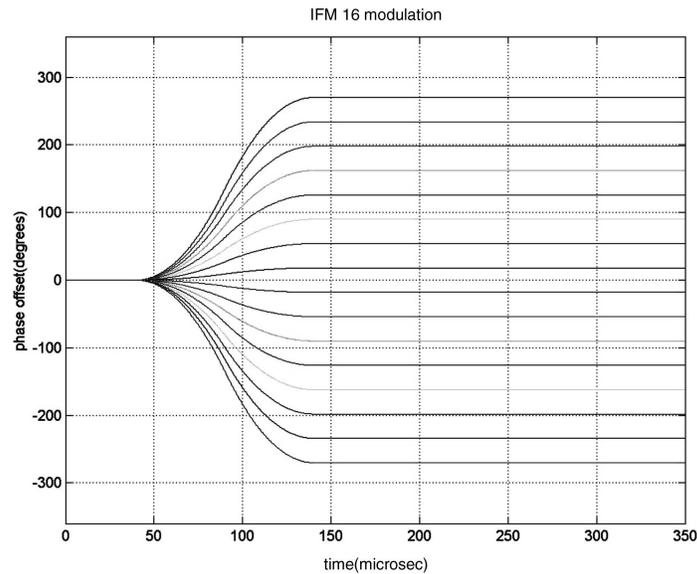
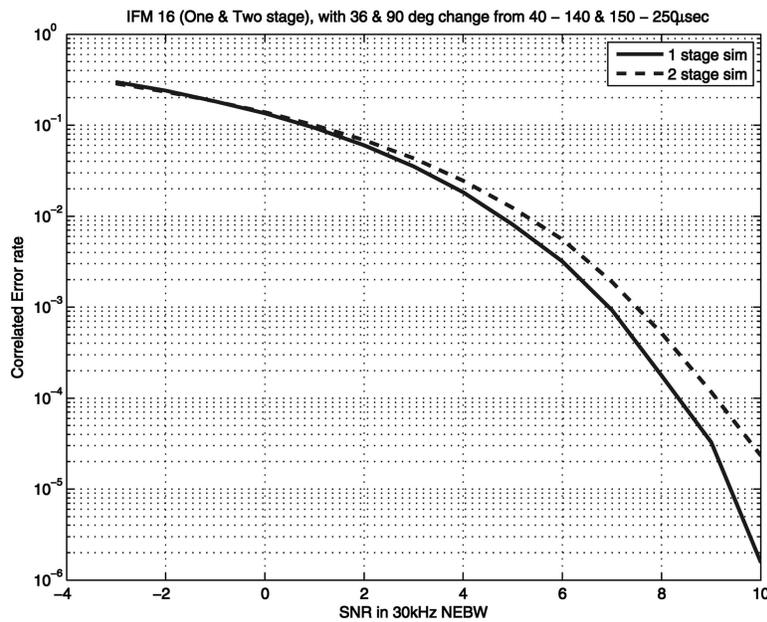Fig. 3.   One-stage 16 level IFM phase offset.

Fig. 4.   Comparison of decode error rate for two IFM 16 designs.

(MOPS) [8]. The message itself provides a six bit preamble which repeats every three messages allowing for synchronization.

C.   Forward Error Correction

The major forms of interference and noise in the Loran band cause burst outages. As a result, Reed Solomon (RS) error-correcting codes (ECCs) are natural choices for FEC. Other ECC such as convolutional coding with interleaving can be used though RS seems to be best suited. The advantages and disadvantages of both are discussed in [9]. Since RS is used by HS-LDC, ninth pulse, and Eurofix, a brief description is presented. Further details

about RS codes are presented in [10] and other similar texts.

RS codes are block codes and can be used for nonbinary formats. They are effective against burst errors and hence are employed in systems such as digital video, compact discs, etc. For encoding, data is parsed into k blocks or symbols of equal lengths. These blocks pass through the RS encoder, which generates an additional $n - k$ blocks for data redundancy. The combination results in one codeword of length n. An RS code is typically characterized by the ordered pair (n, k), which represents the number of overall blocks and the number of data blocks, respectively. The data redundancy allows for error
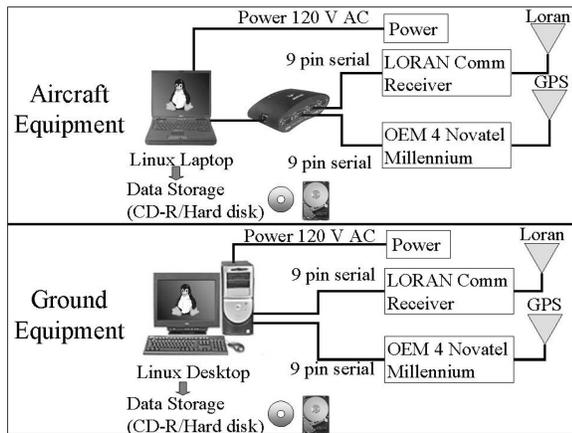
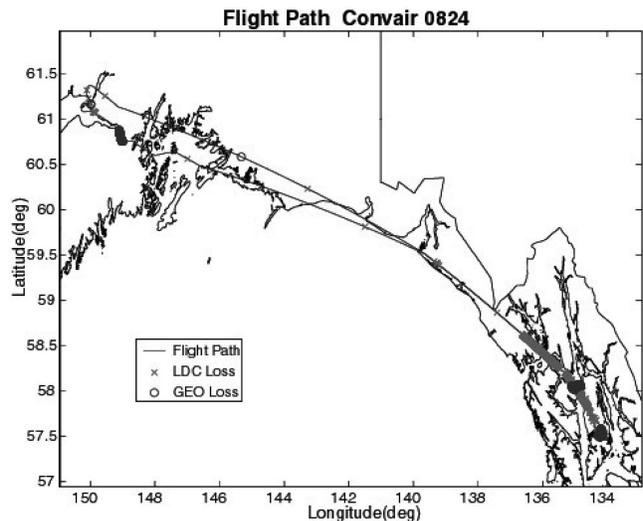Fig. 5. Set up of WAAS user and data collection equipment.

Fig. 6. Flight history of the Convair with message loss history (August 24, 2001).

Fig. 7. Flight history of the King Air with message loss history (August 24, 2001).

detection and correction if some of the blocks are missing or in error.

Hence RS is useful for channels where a burst error may take out a block of data. The size of k relative to n determines the degree to which errors can be corrected. The amount of data within a given block determines the maximum codeword length. Typically, $n = q - 1$ where q is a prime or a power of a prime. Since many systems are binary, q is often a power of 2. Suppose there are m bits of data per block, $n = q - 1 = 2^m - 1$. The code can be shortened by a given amount of data blocks as needed. The "trick" is to assume that these s blocks are exactly zero (and not transmitted then since both parties know this information). Hence, the code has parameter $(n - s, k - s)$ with $k - s$ blocks of data.

The HS-LDC implemented for broadcasting the WAAS message on Loran in Alaska utilized a block length of 8 bits or 1 byte. Thus each symbol is formed from two pulses. The result was to use the RS(255,227) code shortened to RS(60,32). This can correct up to 14 symbol errors.

In addition to RS, the WAAS message incorporates a 24 bit cyclic redundancy check (CRC). Both RS and CRC are cyclic codes.

D. Assessment and Testing

Trials of the HS-LDC were conducted in Alaska on August 23–24, 2001 using the TTX Loran transmitter in Tok, AK. Static testing was conducted in Anchorage on both days as well as a calibration/set up day on August 17th. The static site was previously surveyed allowing for a truth reference. Flight trials were conducted using two aircraft, an Ohio University Beechcraft King Air and an FAA Technical Center Convair 550. The user equipment setup for the test is seen in Fig. 5. Details of the test setup and results are presented in [11], [12].

At static ground site, the HS-LDC signal performed well. The few missing messages on the HS-LDC resulted from the Tok station not broadcasting them. The HS-LDC WAAS message losses tallied 368 losses out of 11477 on August 17th and 1131 losses out of 23280 on the 23rd. The losses are attributable to problems at the transmitter and are not due to the data channel.
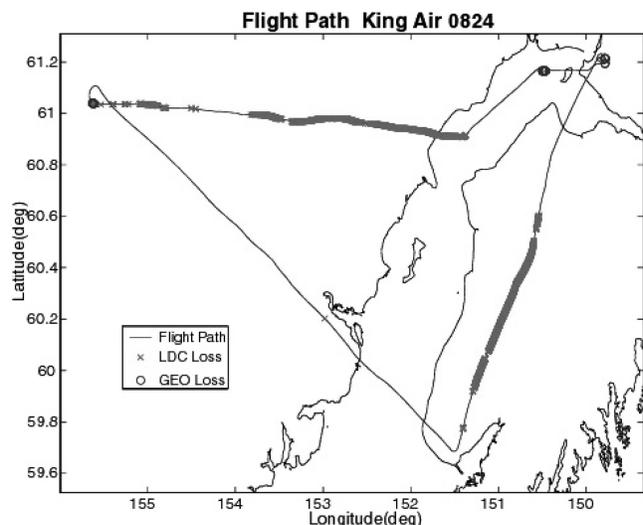
The flight tests illustrated both the potential and pitfalls of data communications on Loran (see Figs. 6 and 7). The data taken from the Convair 550 from the 24th WAAS message had 94.41% availability. This is not bad considering that the majority of LDC lost messages occurred at the full extent of the predicted Tok LDC coverage area. Furthermore, the WAAS message from the geostationary satellite had an availability of 96.95%. Those outages occurred when the aircraft banked away from the geostationary satellite. The King Air had significantly more outages due to installation issues and weather.

TABLE II
Three Implementations for Broadcasting WAAS on Loran

Modulation	Intrapulse Frequency Modulation (IFM)	PPM w/Extra Pulses	Pulse Position Modulation (PPM)
Legacy Support	Support All Legacy rcvrs	Support Timing Legacy rcvrs	Support No Legacy rcvrs
Development	Tested in 2001 on TTX	Least Mature Concept	Similar to earlier, lower data rate schemes
Phased Implementation	Possible	Possible	Impossible
Costs	• Most complex signal format, modulator, and receiver (highest technical risk). • $30–40M to modify SSX. • May need new spectrum to recover performance loss.	Will need TBD additional pulses increasing energy costs and cross rate interference (CRI)	• Cost to legacy users. • Need for international coordination.
Additional Benefits	Trade off between implementation costs and legacy compatibility	• Best performance for new users. • On-air rcvr testing using operational Loran stations is possible.	

E. Alternative Designs

While 16 bit IFM was implemented to demonstrate the capability of Loran to broadcast the WAAS message, other alternatives exist. In addition, there are a few different ways to achieve the required data rate using 16 bit IFM. However, the described method is the only method devised that can achieve the broadcast while maintaining some navigation and timing capabilities for legacy users. There are also other considerations such as performance, maturity of technology, and costs. Table II shows three alternatives for achieving the broadcast and their associate benefits and costs. PPM can be used. However, it provides little or no legacy compatibility for navigation users. It does provide a more cost-effective means of implementing the modulation on Loran. Two forms of PPM have been suggested. The first has less modulation per pulse and uses extra pulses to achieve the desired data rate. Hence, the timing shifts on the pulse can be small and the design should be compatible with legacy timing receivers. The second design places greater modulation on each pulses resulting in greater timing distortions. The benefit is that no additional pulses are necessary.

Another possibility is a phased implementation. The idea is to have the ability to implement the modulation in stages. The initial phases will implement lower data rates that maintain some legacy compatibility. The final configuration will be the full data bandwidth with the accompanying benefits. It will be the least legacy-compatible form. This allows legacy users to have a transition period while providing some data channel benefits. Such an implementation hopefully will induce legacy users to upgrade to attain the full benefits.

The means to achieving this are discussed in Section V.

IV. NINTH PULSE COMMUNICATIONS

A. Background

In the last few years (late 1990s and early 2000s), as GPS vulnerabilities were being examined critically, the need for redundancy for GPS in many safety or economically critical applications was made clear. The convergence of the need to back up GPS in various modes of transportation made this the primary concern in developing Loran. Rather than using the data channel to provide redundancy in WAAS, it was more important that it be designed to support Loran in operating as a stand-alone system capable of acting as an operational backup. This is a fundamental shift in philosophy with data rate being subordinate to protecting the navigational accuracy and integrity of the system.

For the FFA, the key issue for Loran is whether it can support NPA thus providing operational redundancy. The preferred NPA is Required Navigation Performance 0.3 (RNP 0.3). Fulfilling this requirement will probably require the transmission of a low data rate message providing integrity alarms and warnings for hazards such as early skywave [3, 13]. The United States Coast Guard (USCG) needs to provide a backup for GPS for HEA. HEA requires a high level of accuracy—roughly 20 m. This level of accuracy can only be met with differential Loran corrections. Timing acquisition can be greatly aided by station identification. Hence a data capability proves to be integral to meeting these requirements. This modernized Loran system is termed enhanced Loran or *eLoran*.
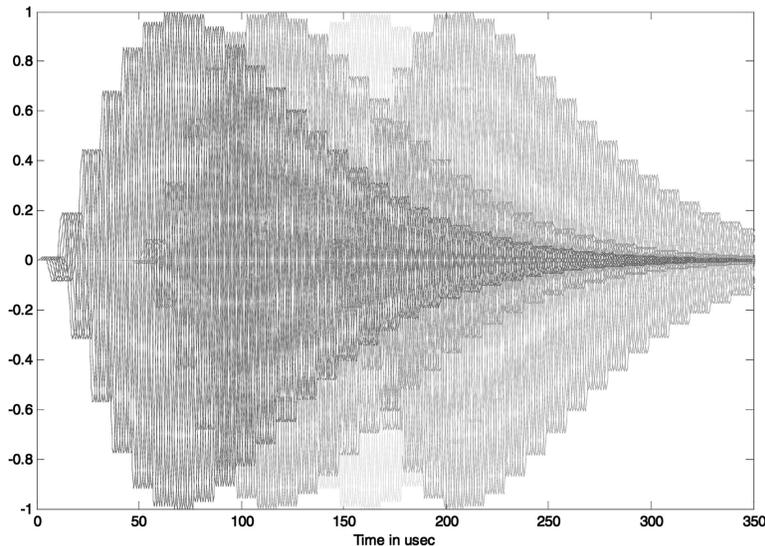
Fig. 8.   Time domain of 32-state PPM.

B.   Design

Initial analysis indicated that a data rate of roughly 30–50 bit/s per second is capable of supporting all three needs. This could be achieved using Eurofix whereby six pulses per GRI are modulated [14]. However, meeting integrity is greatly enhanced by separating the modulation from the navigation pulses. Hence, one pulse was added to the traditional signal solely for data. It would have to achieve a data rate of 50 bit/s without FEC for the slowest GRI. The requirement translates to five bits per pulse or 32 possible states per pulse.

The NPC concept modulates only one pulse every GRI. The ninth pulse concept proposes that all stations broadcast a pulse nominally 1000 μs after the standard group of eight pulses. This pulse is to be used solely for data broadcast. This leaves the eight navigation pulses unmodulated which is vital for integrity. Under Loran-C, secondary stations transmit only eight pulses per GRI while the master station transmits a ninth pulse (2000 μs after the standard eight) for identification purposes. Under ninth pulse, the master and secondary broadcast nine and ten pulses per GRI, respectively.

The idea is appealing since it minimizes the number of pulses that need to be modulated. This is important since each modulated pulse creates cross-rate interference (CRI) that is difficult to process out. Furthermore, since it is not used for navigation, the pulse can be changed in a more significant manner thus allowing a higher amount of data per pulse.

The ninth pulse is nominally an additional pulse after the standard eight pulses. For simplicity of implementation, a positive phase code is always used. NPC has 32 states resulting from a combination of eight phase shift or (PPM) states and four envelope shift states. We can express this mathematically.
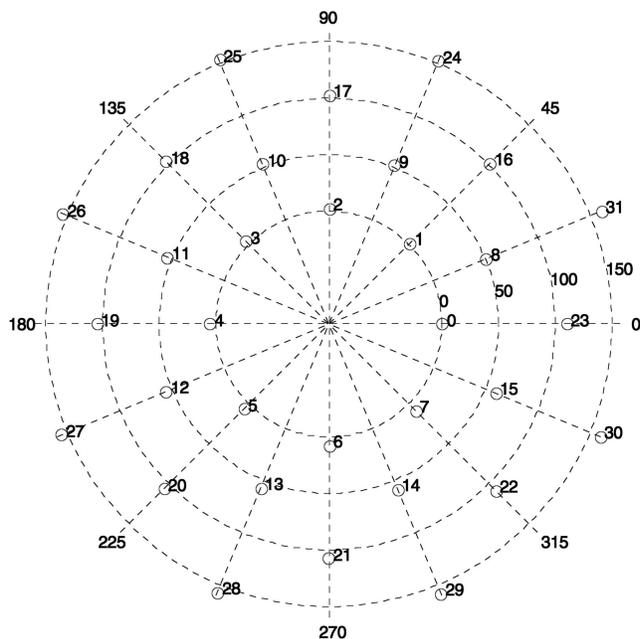

Fig. 9.   Symbol space view of 32-state PPM.

Denote the phase shift state by $n = [0, 1, \ldots, 7]$ and the envelope shift state by $m = [0, 1, 2, 3]$. The ideal delay of each modulated pulse state is given by (2)

$$\text{State}(8^*m + n) : 50.625^*m + 1.25^*n \ \mu\text{s}. \qquad (2)$$

Fig. 8 and Fig. 9 show the 32 states in time domain and symbol space, respectively. For the eight-state PPM, a phase shift of 1.25 μs was used. This divided one cycle into eight evenly spaced divisions. The use of the 5 MHz clock present in the current transmitter system results in some rounding (see Table III). The rounding results in the actual delay being slightly different from the ideal delay. The envelope shift was selected to have the same minimum Euclidean distance as phase shift. A separation of
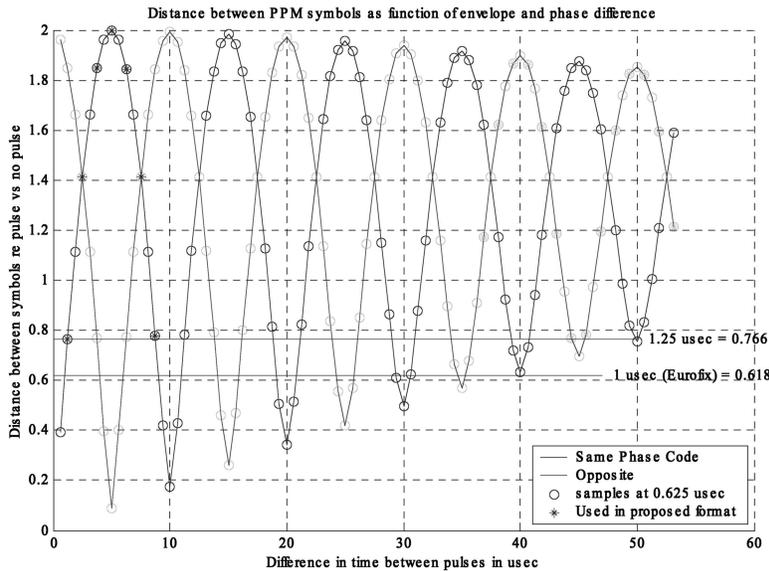
Fig. 10. Determination of minimum envelope delay between groups of 8 symbols.

50.625 μs resulted in the desired Euclidean distance if positive phase code is always used. This is seen in Fig. 10.

The message on the ninth pulse is in the proof-of-concept stage. It is being transmitted in test mode from a few operational Loran stations. For testing and development, 24 GRI was used as the message length. This results in a maximum message length of 2.38 s based on the current United States Loran rates. The limit on the message length allows the receiver to meet time to alarm requirements for aviation and maritime. As work continues on modernizing Loran to serve the critical infrastructure needs of the United States, the ninth pulse message standards and messages will be further defined. As of the writing of this paper, the two messages have currently been defined for testing differential Loran (dLoran) [15].

Synchronization to the start of the message is achieved using the FEC. In fact, the RS FEC on NPC is currently envisioned to play three roles: 1) provide error correction, 2) provide integrity that the message is correct, and 3) provide synchronization [15].

C. Forward Error Correction

RS code is also used on ninth pulse modulation. The current implementation uses an RS(31,16) code shortened to RS(24,9) by setting seven symbols to zero.

Different types of decoders can be used for the message. However, since integrity dictates that it is better not to have a codeword (message) than to incorrectly decode a codeword, a bounded-distance decoder is used. A bounded-distance decoder decodes all signals with t errors or less. Signals with more than t errors are ignored. A common decision basis is to

TABLE III
Symbols Time Delay with Respect to Symbol 0

Symbol	Ideal Delay	Actual Delay (5 MHz Clock)	Symbol	Ideal Delay	Actual Delay (5 MHz Clock)
0	0	0	16	101.25	101.2
1	1.25	1.2	17	102.5	102.6
2	2.5	2.6	18	103.75	103.8
3	3.75	3.8	19	105	105
4	5	5	20	106.25	106.2
5	6.25	6.2	21	107.5	107.6
6	7.5	7.6	22	108.75	108.8
7	8.75	8.8	23	110	110
8	50.625	50.6	24	151.875	151.8
9	51.875	51.8	25	153.125	153.2
10	53.125	53.2	26	154.375	154.4
11	54.375	54.4	27	155.625	155.6
12	55.625	55.6	28	156.875	156.8
13	56.875	56.8	29	158.125	158.2
14	58.125	58.2	30	159.375	159.4
15	59.375	59.4	31	160.625	160.6

use the closest codeword (nearest neighbor) which should be half of the minimum (Hamming) distance. This results in setting $t \leq \lfloor (n-k)/2 \rfloor$ to eliminate any decoding ambiguities for a received sequence. This divides the receiver's observation space into areas (shown by the circle) where the received sequence can be decoded to a valid codeword and other areas where no decoding is possible. A representation of decoder's operation is shown in Fig. 11. The stars ("*") represent exact (error free), valid codewords. The circle around a given valid codeword ("*") is the locus of possible sequences, given by the dots, that can be decoded as that specific codeword. No decision can be made about the area that is not encompassed by a circle.

For the implementation used in 2003, the number of errors corrected was set to a value less than the
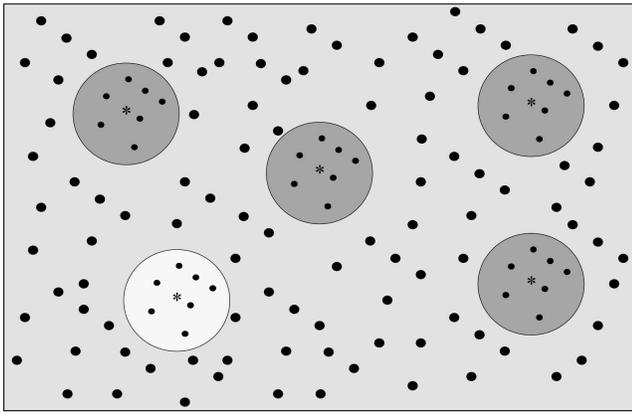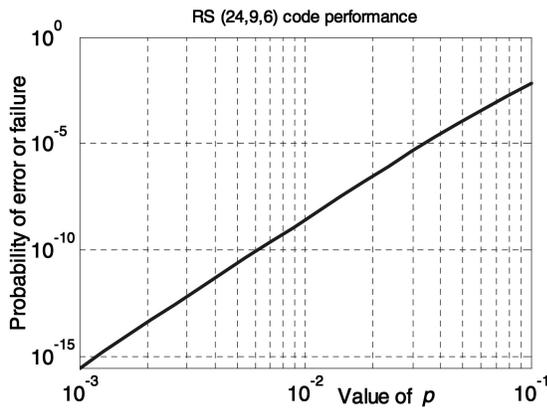
Fig. 11. Representation of decoder's operation.

Fig. 12. RS(24,9) code performance with $t = 6$.

maximum of seven. Typically, t was selected to be six. This aided with synchronization (discussed later) and integrity performance. The sum of the probabilities of incorrect decoding and decoding failure is equal to the probability that the received codeword is outside a circle of t (Hamming distance). This is equivalent to the probability of falling outside the yellow circle. The overall result is seen in Fig. 12 for $t = 6$. The equation for each probability is given below

$$\text{Pr(error or decoder failure)} = \sum_{j=t+1}^{n} \binom{n}{j} p^j (1-p)^{n-j}$$

(3)

where p is the symbol error probability of the channel. The probability of undetected error can be derived

$$\text{Pr(undetected error)}_{RS} = \frac{(q^k - 1) \sum_{j=0}^{t} \binom{n}{j} (q-1)^j}{q^n}.$$

(4)

The numerator is the product of the number of valid, but incorrect, codewords $(q^k - l)$ times the number of correctable patterns per codeword. The denominator is the total number of received sequences.

D. Synchronization

In order to receive and decode the messages, the system must synchronize to the beginning of each message cycle. This allows for the correct parsing of the received data stream into individual blocks of message symbols. Often times this is accomplished using a sequence of synchronization bits. Instead, the RS code itself was used for this purpose thus saving the bandwidth for data. The idea is that if incorrect parsing of the received symbols is modeled as sending random sequences of symbols to the RS decoder, then the analysis of integrity for random data presented in [15] suggests that only with low probability will an incorrectly parsed message be decodable. If the model is correct, for a synchronization error, the RS decoder would typically fail to decode unless the transmitter and receiver frames matched. Further, the probability of multiple symbols in an incorrectly parsed data stream all being decodable is much smaller and the decoder could adjust its framing until multiple messages are decodable.

Unfortunately, the shortened RS code is not well modeled as random data due to its near cyclic nature. Reference [15] also shows an example of why such a synchronization process does not work. The near cyclic nature of the code is preserved (with the error correction capability of the bounded-distance decoder) even when there are small synchronization errors. To solve the synchronization problem, we need to cause synchronization errors to have a disruptive effect on the cyclic nature of the shortened codewords.

The solution is to employ a coset code in which the original code is modified by adding (modulo q) a fixed vector c^* to each codeword. (Note that c^* itself cannot be a codeword as that would not modify the code, just permute the codeword to information symbols assignment.) Coset codes employ the same decoding algorithm as the original code (just first subtract out c^*, again modulo q, before decoding) and have the same error and integrity performance.

The synchronization issue for perfect channels has been addressed [16] and it has been shown that detection of synchronization failures is achievable provided the code's data rate (k/n) is less than one-half. The synchronization of coset codes with error correction has received much less attention in the literature. For example, it is unclear if there exists a coset vector c^* that guarantees synchronization detection for all offsets given that the decoder employs error correction. However, we have had good success with simple coset vectors such as

$$c^* = [0, 1, 2, 3, 4, 5, 6, 7, 8, 9, 10, 11, 12, 13,$$
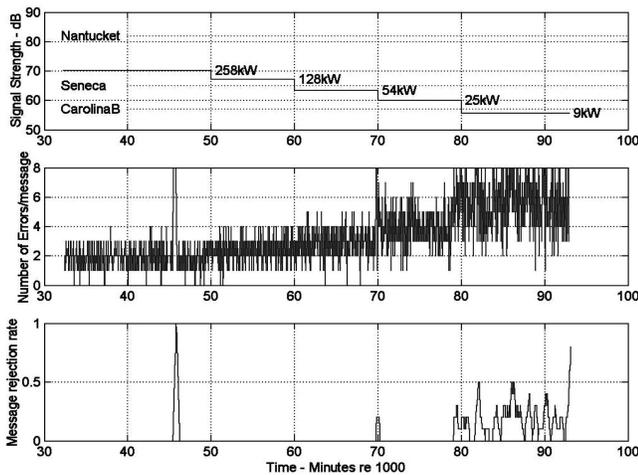$$14, 15, 16, 17, 18, 19, 20, 21, 22, 23].$$

Fig. 13. Transmission testing results.

E. Assessment and Testing

The first significant on-air testing of the ninth pulse signal was conducted in September 2003 using the Loran Support Unit (LSU) tube type transmitter located in Wildwood, NJ. A prototype ninth pulse TTX control software program was developed for this test. Modification to software for the new timing and frequency equipment of the solid state transmitter has been made so that these transmitters can generate the ninth pulse signals.

The test involved a full end-to-end test with the message on the ninth pulse transmitted from Wildwood, NJ to Waterford, CT. The signal was received with the prototype communication receiver that uses an E-field antenna. There was no CRI canceling and the decoder corrected for errors only. Fig. 13 shows the results of the test. It should be noted that the transmitter was temporarily turned off at time=1046 (46 on the plot).

The top graph displays the received signal strength of LSU compared with the three other strongest signals. The second graph shows the number of errors per message (RS code can correct up to six errors). The bottom graph is the percentage of messages lost per block of ten messages.

Only Loran station Nantucket had stronger signal strength than LSU when transmitting full power. Even as the second strongest signal, the message errors were well below the capability of the RS decoder and no messages were lost. As the power of LSU signal was decreased, the errors per message increased, but no messages were lost until LSU signal strength dropped below that of three other stations. Note that the smallest cross-rate pulse that can cause an error is −7 dB in relative strength.

F. Conclusions

NPC is an elegant design. It separates the navigation and communications function of Loran.

This greatly aides the integrity of the navigation signal while providing the bandwidth necessary to support HEA, RNP 0.3, and timing. Furthermore, it should be compatible with most legacy receivers.

V. SCALABLE LORAN MODULATION

A. Background

The utility of data on Loran can be significant. The utility increases with higher data rate. A modulation technique that can be scaled up in data capacity can allow for a graceful transition to higher and higher data rates. It will allow some users to immediately reap the benefits of the modulated data on Loran while still keeping legacy support. This provides time and motivation for users to upgrade to new receivers. New receivers will be developed with the knowledge of the transition path and hence can take advantage of higher data rates as they come online.

This transition strategy gave birth to the concept of SLM. An SLM design is one where the signal can initially be modulated with a low data rate in a legacy-compatible manner. This "basic" modulation can then be scaled up in data rate though it may lose significant legacy compatibility. Additional constraints on the design are that it allow receiver to both easily identify the different "scalings" of modulations and reuse as much of the demodulator hardware as possible.

An initial design was developed to test the feasibility of the concept. A design was done using IFM with the low data rate being a four-state IFM and the high data rate being a 16-state LDC. The reader will recognize the high data rate design since this is very similar to the high data rate LDC design seen in Section IIIB. This design only represents a feasibility test. Scalable PPM was also created and tested.

B. Design of a Scalable IFM

The scalable IFM was implemented using a 4 level IFM (IFM 4) and the 16 level IFM used by the high data rate LDC (IFM 16). Fig. 14 shows the phase relative to a normal Loran pulse of the IFM 4 (2 bit) modulation design. For the HS-LDC [9], the design used IFM 16 on each individual pulse with the phase shift being accomplished in two stages. The pulse was elongated to 350 μs for improved performance. A two-stage IFM 16 phase diagram is shown in Fig. 15. Analysis indicated that two stages performed better than a one-stage implementation. Generally, the modulation occurs after the Loran signal tracking point, hence it does not interfere with legacy receiver tracking.

IFM 4 should be implemented without changing the GRI or the number of pulses per GRI. Since the

TABLE IV
Selected Scenarios for Scaling IFM

|	Initial Low Data Rate	High Data Rate	Higher Data Rate	Potential Higher Data Rate
Scenario 1	IFM 4, same GRI	IFM 16 (2 stages), same GRI	IFM 16 (2 stages), dual rate/add'l pulses	IFM 32 (N stages), dual rate/add'l pulses
Data Rate	60–120 bit/s	120–240 bit/s	250–300 bit/s	500–600 bit/s
Scenario 2	IFM 4, same GRI	IFM 16 (1 stage), same GRI	IFM 16 (1 stage), dual rate/add'l pulses	IFM 32 (N stages), dual rate/add'l pulses
Data Rate	60–120 bit/s	120–240 bit/s	250–300 bit/s	500–600 bit/s
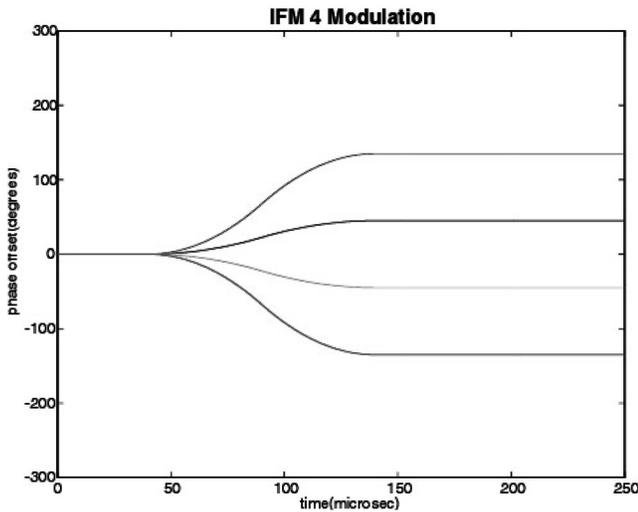
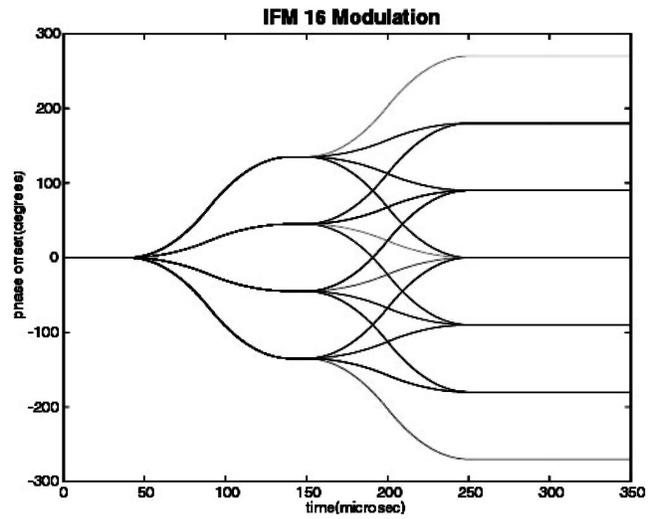Fig. 14.   Phase offset of IFM 4 levels (in comparison to nominal IFM).

Fig. 15.   Phase offset of IFM 16 levels (in comparison to nominal IFM).

existing GRI are used, the signal should be fully legacy compatible. The higher data rate signal(s) are not subject to the constraint. If existing GRIs are used, the IFM 16 data rates will range from 120 to 240 or more bit/s. This implementation is legacy compatible though the increased pulse length will result in greater noise and interference. Hence legacy performance will be degraded. If 250 bit/s is required, more pulses per second will be necessary. This can occur by reducing the GRI of high GRI chains, increasing the pulses per GRI, or dual rating all stations. The first technique is not legacy compatible while the last two may be legacy compatible though the change will result in significantly decreased performance for legacy users.

The scaling can be extended further resulting in IFM 32 or higher. However, there is a trade off since the increased levels decrease the symbol separation resulting in a higher error rate than IFM 16 at a given SNR. The IFM 32 can be implemented over multiple stages. Table IV shows potential scenarios for scaling IFM.

C.   Scalable IFM Analysis

There are three critical factors in implementing SLM: 1) low data rate legacy compatibility,

2) transmitter equipment compatibility, and 3) ability to reasonably detect change to higher data modulation. The first two factors are considerations in any data modulation design and are discussed in [4]. The paramount question in scaling is the third factor—whether the user can identify which modulation is being used.

An analysis of the user's ability to determine the form of modulation transmitted is conducted using match filters. The assumption is that the receiver knows of all modulation designs that will be used. Given that assumption, we examine the receiver's ability to differentiate between IFM 4 and IFM 16 assuming that one of those modulation signals is being transmitted. The analysis uses a match filter matched with every transmitted waveform. The probability that the receiver misidentifies the modulation on the received signal is determined. Fig. 16 shows the probability that the system will mistake an IFM 4 pulse for an IFM 16 pulse. Fig. 17 shows the probability that the system will mistake an IFM 16 pulse for an IFM 4 pulse. No analytic overbound is shown because it is overly conservative. This is due to accumulation of the conservative estimate over an examination of 16 different pulses (for IFM 16). The identification is done using only
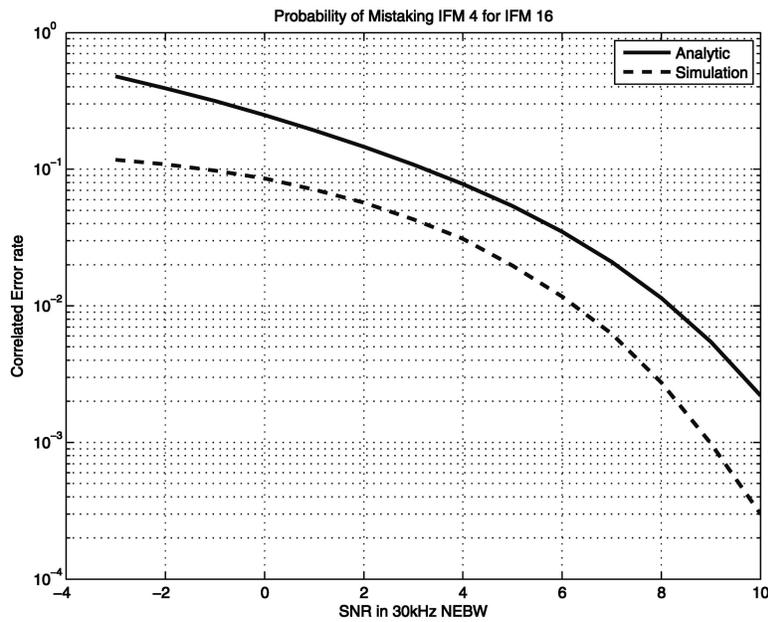
Probability of Mistaking IFM 4 for IFM 16

Fig. 16. Probability of misidentifies IFM 4 signal as IFM 16.

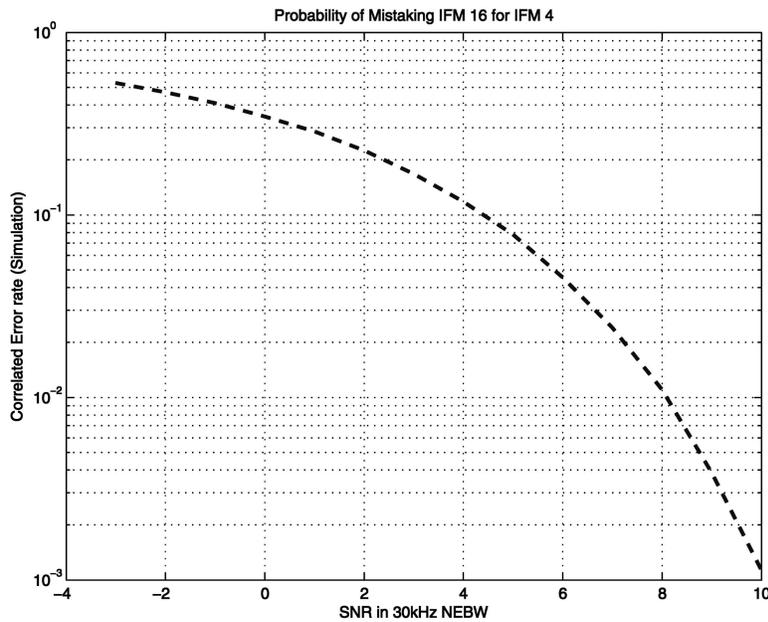**Probability of Mistaking IFM 16 for IFM 4**

Fig. 17. Probability of misidentifies IFM 16 signal as IFM 4.

one pulse. Better performance can be achieved if multiple pulses are used. In addition, the low SNR cases are not typical. The user only needs to determine the modulation design used by one station since all stations will be using the same modulation design. Hence, the receiver will choose the station with the highest signal strength.

The higher data rate signal, since it does not need to be legacy receiver compatible, can be made different enough such that identification is reasonably easy. Match filters were used for the detection analysis since it can be used for decoding either modulation. This allows for "reuse" of hardware. There are other

ways of detection—for example, there can be an algorithm that examines the signal past 250 μs.

One concern with IFM is the use of spectrum and this goes hand in hand with design and optimization of the signal. For IFM 4, there are at least two parameters that can be selected. This is the start and end time of the frequency change. Assuming that the phase offsets are equally spaced in phase and that the frequency change is constant, the frequency rate is then determined by the start and end time. There are even more parameters to select if these assumptions are not followed. Similarly, for IFM 16, there are at least four parameters that can be selected. These
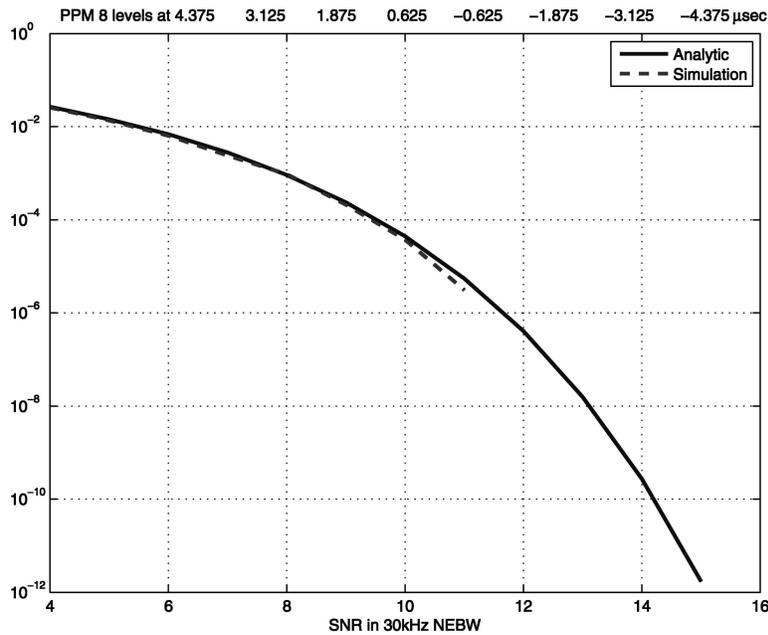
Fig. 18. SNR versus probability of symbol error for PPM 8 (uniform 1.25 μs spacing).

TABLE V
PPM Design

	Time Shift Between Symbol	Time Shift (off nominal Loran)
PPM 3 Eurofix	1 μs	+1, 0, −1 μs
PPM 8	At least 1 μs	Various

changes affect performance (in terms of error rates versus SNR) and spectrum. The search space for ideal IFM designs, particular IFM 16, is large. However, the search space is limited by some design criteria such as restricting the first phase change to occur after the signal tracking point (30 μs).

D. Design of a Scalable PPM

PPM uses different magnitude timing shifts of the Loran signal relative to the nominal Loran pulse to broadcast information. PPM has been demonstrated in a variety of ways, most notably in Eurofix which uses a three level PPM (PPM 3) [14].

The distance between symbols and hence the performance depends on the magnitude of the time shifts. For legacy compatibility, there are limitations

on the shifts. First, the coding of the shifts needs to be balanced. Balanced coding is a scheme where every time/phase shift is coordinated so that the sum of the time/phase shifts is zero over a specified period. This is discussed in [4].

Second, the magnitude of the timing shift should not be too great. Any timing shift will result in some degradation of performance. 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) [17]. Larger timing shifts result in larger variations in the tracking point thus reducing the ability of the receiver to acquire and track of the signal. Hence, legacy compatibility limits our ability to significantly modulate the Loran pulse.

If legacy compatibility is not a concern, larger timing shifts and nonbalanced codes can be used. Such changes would increase the data rate and utility of PPM. One possible design is an eight level PPM which would yield three bits per pulse. This effectively increases the data rate by at least a factor of 2.25. The spacing for PPM 8 needs to be increased to provide greater "distance" between each symbol or level. Table V presents the spacing used from PPM 3 (Eurofix) and PPM 8. Table VI shows a scenario

TABLE VI
Selected Scenarios for Scaling IFM

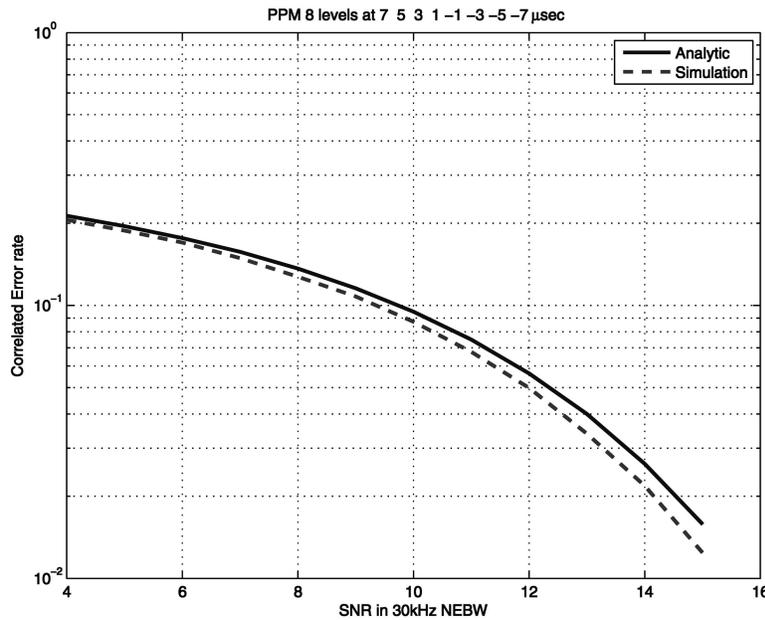|	Initial Low Data Rate	High Data Rate	Higher Data Rate	Potential Higher Data Rate
Scenario	PPM 3, balanced same GRI	PPM 8 not balanced, same GRI	PPM 8 not balanced, dual rate/add'l pulses	PPM 16, dual rate/add'l pulses
Data Rate	40–114 bit/s	90–256.5 bit/s	256.5+bit/s	500–600 bit/s

Fig. 19. SNR versus probability of symbol error for PPM 8 (uniform 2 μs spacing).

for scaling PPM. The initial low data rate is legacy compatible with some performance degradation. Legacy users may be able to use the second case though with significant degradation or timing error. The remaining designs are not legacy compatible.

E. Scalable PPM Analysis

Matched filter analysis performed on the various PPM designs provides an indication of the performance of these designs in the presence of white noise. An example of PPM 8 with 1.25 μs shifts between each level is shown in Fig. 18.

In designing PPM, the time offset or spacing between each level is an important factor. Generally a greater spacing results in better data channel performance. This, of course depends on the method of demodulating the data. If matched filters are used to determine signal, it is possible that larger spacing may result in poorer performance. If there are two levels that are approximately a cycle (10 μs) apart, the matched filter output for the two levels will be strong since they are both in phase. With the addition of noise, there is a significant chance that the matched filter output for the level that is a cycle off may be larger than the matched filter output for the actual signal. An example of this case is seen in Fig. 19. Compare this figure with Fig. 18 which has a smaller time spacing but does not have levels that are a cycle apart.

Performance can be greatly improved if the correct cycle can be identified. If such misidentifications are removed, the result is seen in Fig. 20 which shows significant improvement. The performance seen in the figure for the 2 μs spacing is better than the 1.25 μs spacing. Achieving a 1 percent error rate requires an

SNR of 1 dB versus an SNR of 4 dB for the 1.25 μs design.

Again, we need to determine whether the user can identify a change in modulation. If the PPM 8 design uses some of the same levels (time offsets) as the PPM 3 design, then those signals are identical and the user cannot tell them apart. However, the number of common levels is at most three and if the other levels of PPM 8 are significantly different enough, the user should be able to determine the modulation used with high confidence. Fig. 21 shows the result of this analysis for distinguishing between PPM 3 and PPM 8 given that PPM 8 modulation is used.

F. Conclusions

Preliminary examination of SLM indicates that it is a feasible concept. Given current design and requirements on *eLoran*, it does not offer obvious benefits. However, if additional data requirements arise, the concept provides options to transition to higher data rates. As such, it may prove useful for rapid implementation of the *eLoran* system.

VI. CONCLUSION

Data modulation on Loran is an increasingly important consideration in the modernization of the system. The HS-LDC demonstrated that Loran can support a number of applications including the WAAS message. The NPC design is an integral part of the *eLoran* design. It provides the data bandwidth necessary so that Loran has the integrity and/or accuracy necessary for aviation RNP 0.3 approach and maritime HEA. It also provides information for time synchronization. Retaining the existing user
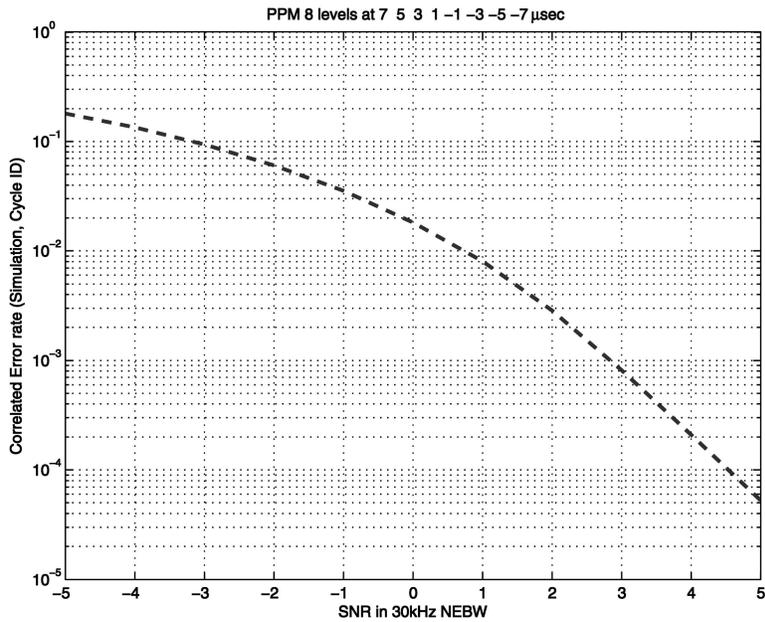
Fig. 20. SNR versus probability of symbol error for PPM 8 (uniform 2 μs spacing), assuming cycle ambiguity can be resolved.
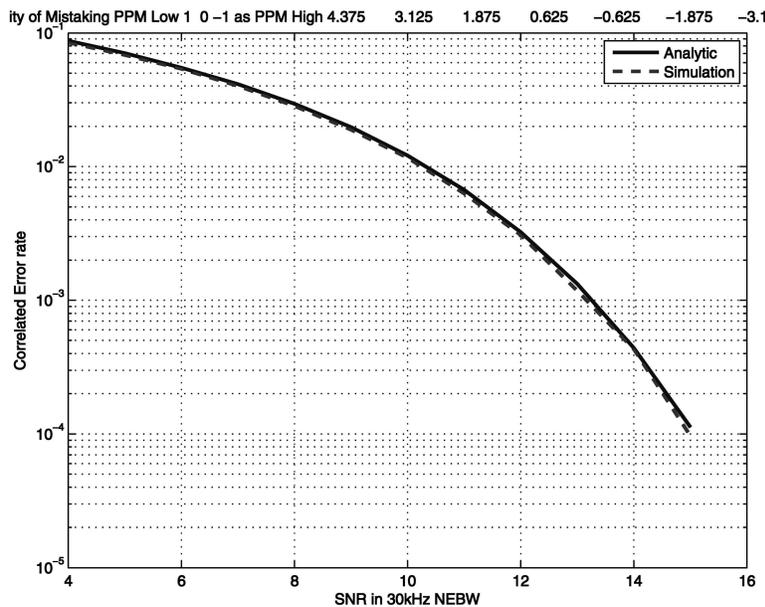

Fig. 21. Probability of decoding a PPM 8 signal and believing that it was a PPM 3 signal.

base is still important consideration. One means of achieving the goal is to provide a reasonable transition period for users to upgrade. However, this should not come at the expense of delaying the introduction of useful data broadcast. The SLM concept allows for a migration path to higher data rates. It provides immediate benefits to users who have upgraded while allowing the legacy community time to migrate to newer technology.

Even in this age of GPS and Galileo, Loran still has a lot of value. It can supplement global navigation satellite systems (GNSS) in areas such as urban canyons, where there is poor availability. It can provide backup to satellite navigation in a variety of manners. It can also distribute high accuracy timing and frequency. Data capacity on Loran only enhances the value by allowing Loran to provide additional features such as increased accuracy and signal authentication.

ACKNOWLEDGMENT

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