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Modernized *e*Loran:

The Case for Completely Changing Chains, Rates, and Phase Codes

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BIOGRAPHIES

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ABSTRACT

First deployed in the U.S. in 1957, Loran-C dominated radio-based navigation for many years. In 2000 the FAA began a significant recapitalization of Loran in the U.S.; the 2001 Volpe report on the vulnerability of the GPS reinforced the need for a revamped Loran. What emerged was an enhanced or evolved version, so called "*e*Loran," aiming to achieve, for example, 10-20 meter absolute positioning accuracy, RNP 0.3 mile required navigation performance, and stratum 1 time. After 10 years of development, in 2010, this U.S. effort was halted and the U.S. transmitters were silenced; since that time, *e*Loran is still being developed in Europe and deployed in Asia.

Earlier this year U.S. Government interest in *e*Loran has again stirred (evidenced by a U.S. Army request for information and a U.S. Dept. of Transportation request for public comment); the first of these initiated much conversation at the 2015 ION ITM.

The prior U.S. (and continuing European) development of *e*Loran kept many of the 1950's system design choices so as to be compatible with legacy Loran receivers. These include the pulse shape, groups, chains, rates, phase codes, emission delays, etc. Chosen to suit 1950's technology, many of these restrictions are no longer necessary given the advances in transmitter and receiver technology (e.g. software defined radio) over the last half century. It is the opinion of these authors that as Loran, per se, no longer exists in the U.S., any re-emergence of a low frequency radio navigation system need *not* be held to these performance limiting constraints.

In prior work these authors have promoted more significant changes to *e*Loran to improve system performance; specifically, single-rating all stations, reconfiguring the chain/rate structure within the continental U.S., and changing the phase codes. The current paper expands on these prior efforts. Specifically, we propose putting all of the *e*Loran transmitters on the same repetition period and employing unique phase codes for each transmitter. To effectively choose new phase codes for *e*Loran, and assess their performance, we rely on the auto- and cross-correlation metrics. These metrics describe how well a receiver can both acquire and track a specific signal when contaminated by multipath interference, the existence of other signals, and noise. While a "perfect" auto-correlation function, large at zero lag corresponding to the actual arrival of the signal and zero elsewhere, and a "perfect" cross-correlation function, zero for all lags, are preferred, it is impossible to find such codes. However, limiting the size of the window for which we require perfect auto- and cross-correlations, such codes can be found. To create such codes for *e*Loran we adapt results from the CDMA literature on complementary sequences and Large Area Synchronized (LAS) codes.

This paper begins with a brief review of the relevant characteristics of Loran-C, including a discussion of the effects of sky wave and cross rate interference. This is followed by a survey of previously published ideas/concepts on how elements of the system could be changed so as to improve performance. Finally, details on the proposed rate/chain/phase code structure are presented.

The reader should recognize that these ideas and results are not intended to define what the best *e*Loran system would be; rather, if *e*Loran soars again in the U.S., we hope to initiate a dialogue that looks beyond the decisions made in the 1950's.

INTRODUCTION

Loran-C is a radio-navigation signal first deployed in the U.S. in 1957. Initially provided as a service for the U.S. Navy, Loran dominated coastal navigation for many years; in the mid-1970s, additional Loran transmitters were erected to provide positioning service over the continental U.S. Loran has also been deployed in Europe, Asia, and the Middle East; the Russians developed a compatible system, Chayka. The year 2000 saw a significant recapitalization effort in the U.S., with the resulting *e*Loran being seen as a potential backup to the GPS. In 2010 this effort was halted and the U.S. Loran transmitters were silenced; since that time, *e*Loran is still being developed in Europe and deployed in Asia.

Recently it appears that interest in a low frequency, pulsed, radio-navigation system might be re-emerging within the U.S. In January of 2015 the U.S. Army issued a request for information for *e*Loran receivers; this announcement spiked much conversation on *e*Loran at the 2015 ION ITM. In March of 2015 the U.S. Dept. of Transportation requested public comment on *e*Loran.

The prior U.S. (and continuing European) development of *e*Loran kept many of the design choices originally made in the 1950's so as to provide backwards compatibility to legacy Loran receivers. These include (and are further described below) *chains* of nearby Loran transmitters, some of them *dual-rated*, employing *coding/emission delays* to implement time division multiplexing of *groups* of Loran *pulses* repeatedly transmitted at specific *group repetition intervals* with their polarities set by *master* and *secondary phase codes* (on the *phase code interval*) to mitigate *skywave* and *cross rate interference*. Many of these system characteristics were chosen based upon the level of technology available in the 1950's and are no longer necessary. Some system characteristics of Loran-C were changed for *e*Loran; specifically, ones that would improve performance for modern receivers, but would not negatively impact the legacy user.

In prior work these authors (and others) have promoted more significant changes to *e*Loran to improve system performance, arguing that the legacy receiver community was sparse. Notably, these addressed modifying the pulse shape, single-rating all transmitters, re-defining the chains and rates, and modifying the phase codes. These prior efforts are reviewed in the third section of this paper.

The current paper expands on these prior efforts on changing rates, chains, and phase codes. Specifically, while we assume that the signal still consists of discrete Loran pulses spaced a minimum of 1 millisecond apart, we propose putting all of the *e*Loran transmitters on the same repetition period and employing unique pulse patterns (which we will still call phase codes) for each transmitter. To effectively choose new phase codes for *e*Loran, and assess their performance, we rely on the auto- and cross-correlation metrics. These metrics describe how well a receiver can both acquire and track a specific signal when contaminated by multipath interference, the existence of other *e*Loran signals, and noise. The goal is to have a "perfect" auto-correlation function, large at zero lag corresponding to the actual arrival of the signal and zero elsewhere, and a "perfect" cross-correlation function, zero for all lags. Not surprisingly these are opposing characteristics; it is impossible to find such codes. However, by limiting the size of the window about zero lag for which we require perfect auto- and cross-correlation, such codes can be found. For example, the current Loran master and secondary phase codes satisfy this for lags less than 10,000 μsec. Similarly, the phase codes that we presented in [1] met these same limited characteristics; unfortunately with only four of them (and no knowledge on how to find more at that time) we did not further develop the concept. To increase the number of codes to the 15-30 necessary for CONUS-wide coverage, we exploit previously published results for CDMA systems on complementary sequences [2] and Large Area Synchronized (LAS) codes [3].

This paper is organized as follows:

- The next section of this paper contains a brief review of the relevant characteristics of Loran-C with a brief discussion of the changes in *e*Loran. This section includes a discussion on how sky wave and cross rate interference are determined by the rates, chains, and phase codes.
- The third section of this paper briefly reviews previously published ideas and concepts on how other elements of the Loran system could be changed so as to improve performance.
- The paper next includes details on the proposed rate/chain/phase code structure, a review of the methods for designing the codes, and examples for a 8 station system (i.e. 8 codewords).
- The paper closes with comments with further directions for the presented ideas.

The reader should recognize that the ideas and results herein are not intended to define what the best *e*Loran system would be; rather, if *e*Loran soars again in the U.S., we hope that these remarks initiate a dialogue that looks beyond the decisions made in the 1950's.

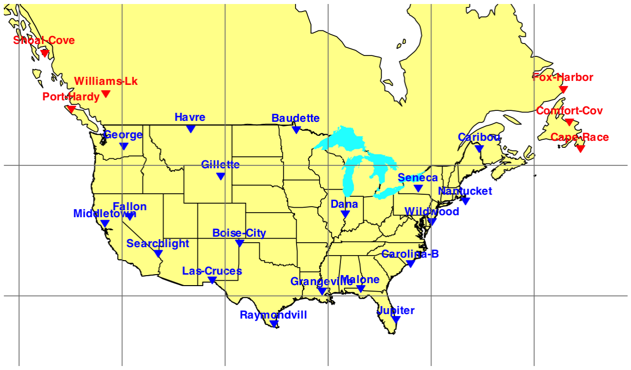

Figure 1: Loran stations in North America, without Alaska, circa 2010.

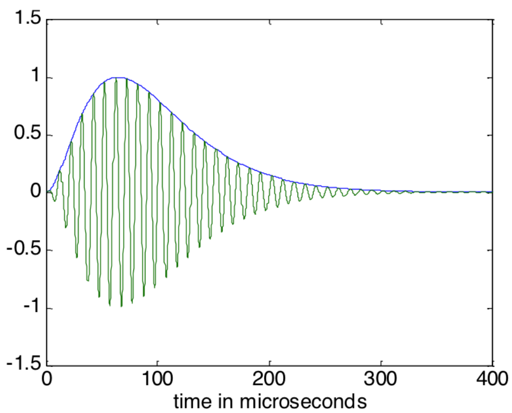

Figure 2: The Loran pulse, $p(t)$.

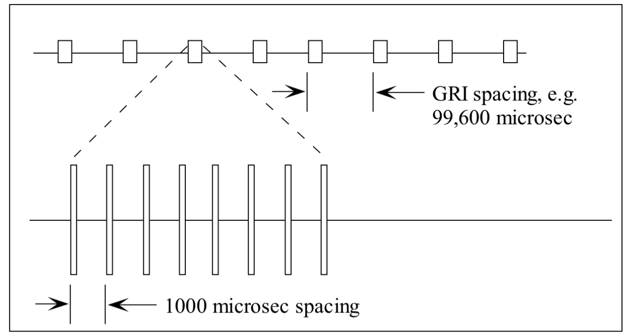

Figure 3: Loran group timing.

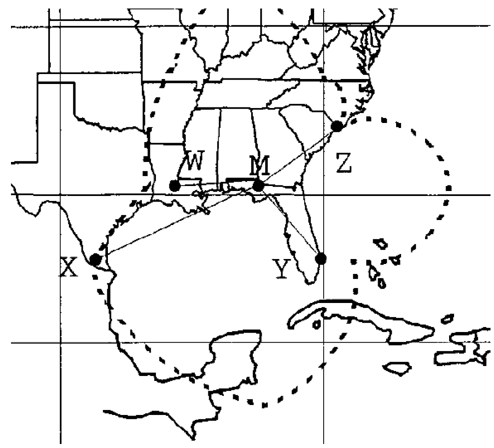

Figure 4: Southeast U.S. chain, from [4].

THE BASICS OF LORAN

There have been a number of papers over the years that have described the history of the Loran system in great detail and the interested reader is referred to the archives of the Wild Goose Association, later the Int'l Loran Association (www.loran.org), and the Institute of Navigation. For the technical details of the Loran transmission, consult the Loran Signal Specification [4] and other reports on the U.S. Coast Guard's NavCen website (www.navcen.uscg.gov). Some basics of the system are described here in order to provide a framework and some notation for understanding the rest of this work.

Loran is a high power, low frequency, long range radionavigation system. When discontinued in 2010 the North American Loran-C system consisted of 18 U.S. stations in the lower 48 states, 6 Canadian stations, and 5 Alaskan stations (Figure 1 shows the geographic layout, without Alaska).

Each station repeatedly transmits the Loran pulse, $p(t)$, shown in Figure 2 (normalized to unit maximum amplitude), a teardrop shaped pulse envelope (shown in blue) modulated by a 100 kHz carrier. The pulse envelope rises from zero to its maximum amplitude at 65 μsec and then decays back to zero with a total duration under 300 μsec. The pulse shape was designed so that 99% of the radiated power is within the allocated frequency band of 90 to 110 kHz.

Each Loran station transmits 8 (or for the master station, 9) of these pulses in a group with the pulses spaced 1000 μsec (1 msec) apart; the group is repeated at the group repetition interval or GRI; Figure 3 suggests this timing for a station on the 9960 rate, a GRI of 99,600 μsec.

Rather then working independently, the transmitters are organized into collections or chains of geographically close stations which share a common GRI. For example, Figure 4 shows the locations of the stations in the Southeast U.S. chain. Within each chain one station is designated as its master (denoted M, it broadcasts first) and the other stations (between 2 and 5) are called secondaries (designated W, X, Y, and Z in this example) and broadcast in that particular order. Figure 5 shows the timing for a chain with master and three secondaries. The secondaries' broadcasts are delayed (TDX, TDY, and TDZ in this diagram) to ensure

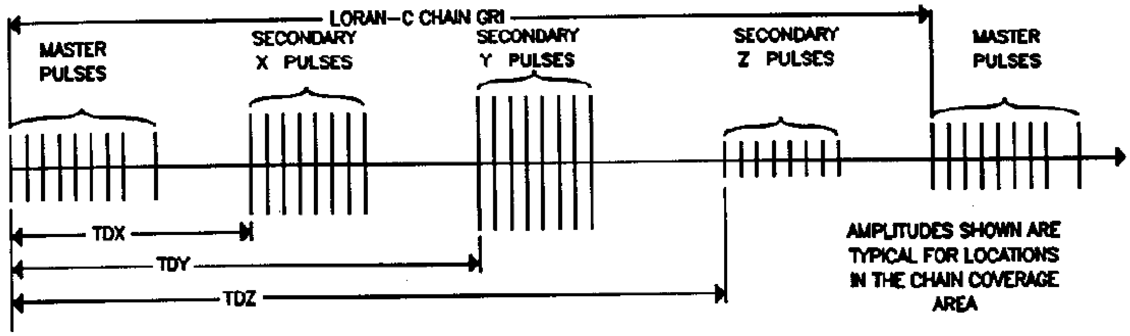

Figure 5: Entire chain timing, from [4].

that a receiver, no matter where it is, observes the signals from the chain in this same order. This allows a receiver to uniquely identify each secondary in the chain and is accomplished by having each delay greater than the time required for the prior stations' signals to have been received at the next transmitter. The master is uniquely identified by sending an additional pulse in each group.

To further distinguish master from secondary, the pulses are polarity encoded (a multiplier of ± 1 on the envelope) by either the master or secondary phase code. The phase codes repeat every two groups, called the phase code interval or PCI. The phase codes are

$$[+1, +1, -1, -1, +1, -1, +1, -1, +1]$$
$$[+1, -1, -1, +1, +1, +1, +1, +1, -1]$$

for master and

$$[+1, +1, +1, +1, +1, -1, -1, +1]$$
$$[+1, -1, +1, -1, +1, +1, -1, -1]$$

for the secondaries. Besides uniquely identifying the master station in a chain, the Loran phase codes aid in the rejection of multipath interference.

Finally, an individual station is assigned to either one (single-rated) or two chains (dual-rated). For example, the station in Malone, Florida (M in Figure 4) was dual-rated, being the master of the shown chain and a secondary in the Great Lakes chain.

With these system characteristics the transmitted signal for a particular rate at a particular (single-rated) station can be written as

$$\begin{aligned} s(t) &= A \sum_{i=-\infty}^{\infty} \sum_{m=0}^{1} \sum_{k=0}^{7} \\ &\quad b_{7m+k}\, p\left(t - i \cdot PCI - m \cdot GRI - 1000k - ED\right) \end{aligned}$$

for some amplitude A, emission delay ED, phase code coefficients b_k (b_0 through b_{15}), and the time related terms in units of μsec. (As written this expression is valid for a Loran secondary since the summation over k indexes the 8 pulses; for a master we would increase the upper limit on k to accommodate the additional master pulse; a dual-rated station would have a duplicate set of terms with a different emission delay.)

Pulse Averaging at a Receiver

The Loran signal propagates as a ground wave at nearly the speed of light, following the curvature of the Earth. A Loran receiver observes this signal delayed in time by τ (due to the propagation distance from the transmitter), and embedded in noise

$$r(t) = s(t - \tau) + n(t)$$

for some finite observation window $t \in [0, T]$. Taking advantage of the known periodicity of the transmission, the receiver can reduce the effect of the additive noise by shifting and averaging segments of the received signal, correlating away the phase code, for an estimate of the delayed pulse from one transmitter

$$\begin{aligned} \widetilde{x(t)} = &\quad \frac{PCI}{16\,T} \sum_{i=1}^{T/PCI} \sum_{m=0}^{1} \sum_{k=0}^{7} \\ &\quad b_{7m+k}\, r\left(t + i \cdot PCI + m \cdot GRI + 1000k\right) \\ = &\quad A\, p(t - ED - \tau) + \widetilde{n(t)} \end{aligned}$$

In this expression the reciprocal of $\frac{PCI}{16\,T}$ is the number of pulses observed in $[0, T]$. The resulting waveform is proportional to a single Loran pulse, delayed by τ and the emission delay of that station, in noise with a smaller variance than the original noise

This operation is identical for each station within one chain; hence, the resulting average $\widetilde{x(t)}$ for $0 < t < GRI$ would consist of one pulse for each station in the chain (indexed by j) with the start time of the pulse equal to its emission delay plus the delay due to the distance from the receiver to each transmitter

$$\widetilde{x(t)} \quad = \quad \sum_j A_j \, p(t - ED_j - \tau_j) + \widetilde{n(t)}$$

Since the emission delays are chosen so that the pulse arrivals never overlap in time wherever the receiver might be, this appears as a sequence of pulses, one for each station in the chain, in the transmitted order (master, W, X, Y, ...). Assuming that signals from at least three stations are received, the Loran receiver estimates the arrival times of these averaged pulses and computes the receiver position using trilateration.

Sky Wave Interference

Besides noise, the Loran receiver typically observes so-called "sky wave" interference, multipath interference due to ionospheric reflections of the desired transmission. Including amplitude scaling and additional delay (another τ, but indexed by k) the reception from one transmitter would be

$$r(t) = s(t - \tau) + \sum_k \alpha_k s(t - \tau - \tau_k) + n(t)$$

Typically the additional sky wave delay, τ_k, ranges from a minimum of about 30-50 μsec (for long range, single hop reflections) to perhaps 2000-4000 μsec for short range, multi-hop reflections [5]. The value of the amplitude scaling, α_k, can be smaller or larger than one and depends upon the lengths and attenuation characteristics of the propagation paths.

The effect of sky wave on a Loran receiver depends upon the amount of delay:

- If $\tau_k < 250$ μsec then the reflections overlap the desired pulses (all 8 of them in a group), distorting each pulse constructively and/or destructively. To eliminate this interference, most Loran receivers estimate the desired time delay τ using the third up-going zero crossing of the Loran pulse at 30 μsec. This works well except for *very early* sky wave (i.e. due to very low ionosphere), for which other techniques are applied (and are beyond the scope of this discussion).
- If 250 μsec $< \tau_k <$ 700 μsec then the reflections do not overlap any of the desired pulses. With pulse averaging, the receiver sees the sky wave pulse *after* the desired pulse within the one GRI averaging window; it's time of occurrence is well before the pulses corresponding to the other stations in the chain due to the choices of the emission delays. In other words, this level of sky wave interference is not a problem; the reflection response is easily ignored.
- If 700 μsec $< \tau_k <$ 7300 then the reflections might overlap later pulses in the same group (e.g. with $\tau_k = 1000$ μsec the multipath of pulse 1 hits pulse 2, the multipath of pulse 2 hits pulse 3, etc.) and add to the averaging. The phase codes guard against this as explained next.

To explain the suppression of longer delayed sky waves consider the case of one reflection so that the observation is

$$r(t) = s(t - \tau) + \alpha_1 s(t - \tau - \tau_1)$$

(we drop the noise for clarity). Since 700 μsec $< \tau_1 <$ 7300 an overlap of 1 to 7 pulses is possible. Define the integer d so that

$$\tau_1 = 1000\, d + \tau_0$$

with $\tau_0 \in [-300, 700]$. Then $d \in \{1, 2, \ldots 7\}$ and $8 - d$ of the pulses are contaminated by sky wave.

With the pulse averaging above, the estimate is

$$\begin{aligned}\widetilde{x(t)} \; &= \frac{PCI}{16\,T} \sum_{i=1}^{T/PCI} \sum_{m=0}^{1} \sum_{k=0}^{7} \\ &\quad b_{7m+k} s\left(t + i \cdot PCI + m \cdot GRI + 1000k - \tau\right) \\ &+ \alpha_1 \frac{PCI}{16\,T} \sum_{i=1}^{T/PCI} \sum_{m=0}^{1} \sum_{k=0}^{7} \\ &\quad b_{7m+k} s\left(t + i \cdot PCI + m \cdot GRI + 1000k - \tau - \tau_1\right) \\ &= A\, p(t - \tau) \\ &+ \alpha_1 \, A \, \left(\frac{1}{16} \sum_{m=0}^{1} \sum_{k=0}^{7-d} b_{7m+k} b_{7m+k+d} \right) p\left(t - \tau - \tau_0\right)\end{aligned}$$

In words, the result is the desired pulse plus a second one delayed by an additional τ_0 seconds. While this second pulse could cause a problem because of the range of τ_0, the scalar in parentheses solves the problem. Specifically, the value

$$\sum_{m=0}^{1} \sum_{k=0}^{7-d} b_{7m+k} b_{7m+k+d} = \sum_{k=0}^{7-d} b_k b_{k+d} + \sum_{k=8}^{15-d} b_k b_{k+d}$$

sometimes called the aperiodic autocorrelation function, can be manipulated by the choice of the phase codes. For the Loran phase codes above, both secondary and master (if you modify the sums to work on the 9 values), this term is identically zero for $d = 1, 2, \ldots 7$.

While this argument dealt with only one reflection it extends directly to the algebraic sum of multiple reflections so that sky wave with delay up to about 10,000 μsec is completely eliminated by the current Loran phase codes (the window can be increased from 7300 μsec because of a 2900 μsec guard band after each group). Further, the effect of sky wave under the use of some other phase code is easily computed by examining the autocorrelation function of the b_k.

Self-Chain Interference

The Loran receiver also observes the transmissions (and, potentially, sky wave) from other stations in the same chain; specifically

$$r(t) = s_{desired}(t-\tau) + \sum_m \alpha_m s_{others}(t-\nu_m)$$

in which the ν_m are the delays of these other signals and are, by definition, greater than 10,000 μsec. The only concern is that the emission delays plus the distance based delays plus any additional sky wave delays not be large enough to push the offending signal into the range of the desired signal. The Loran Signal Specification's 2900 μsec guard window between secondary transmissions (1900 for master to the first secondary), is long enough to eliminate these overlaps for the sky wave delays seen in practice.

Cross-Rate Interference

If all Loran stations broadcast at the same GRI we would be done with the list of system induced interference; however, since Loran has chains of stations at different rates then periodically pulses from other chains are nearly time coincident with the signals of interest, creating so-called "cross rate interference." An analysis of cross rate and selecting rates to minimize the effect is a difficult problem. A frequency domain analysis from [6] can provide some insight.

In brief, recall that the Loran signal is periodic on a PCI; hence, it can be fully described by a Fourier series representation. To develop this representation, start with a single Loran pulse starting at time zero in the PCI. Ignoring the modulation by 100 kHz (we can offset the Fourier results up by 100 kHz later, if so desired) its Fourier series representation is

$$s(t) = \sum_{n=-\infty}^{\infty} d_n e^{j2\pi nt/PCI}$$

with $PCI = 2\,GRI$ representing the repetition time, in μsec, the integer n divided by the PCI is the shift, in Hertz, from 100 kHz for the spectral line of interest (in other words, spectral lines occur with a spacing of 1/PCI Hz), and the series coefficients are

$$d_n = \frac{65^3}{4PCI^2\,(PCI + j65\pi n)^3}$$

As the actual Loran signal consists of 16 replicas of this pulse, shifted in time and multiplied by the phase code $[b_0, b_1, b_2, b_3, b_4, b_5, b_6, b_7]$ and $[b_8, b_9, b_{10}, b_{11}, b_{12}, b_{13}, b_{14}, b_{15}]$, the series expansion for the full Loran transmission is of the same form with modified coefficients

$$\begin{aligned} d_n = \quad & \frac{65^3}{4PCI^2\,(PCI + j65\pi n)^3} \\ & \times \begin{bmatrix} (b_0 \pm b_8) + (b_1 \pm b_9)\,e^{-j2\pi n1000/PCI} \\ + (b_2 \pm b_{10})\,e^{-j2\pi n2000/PCI} \\ + (b_3 \pm b_{11})\,e^{-j2\pi n3000/PCI} \\ + (b_4 \pm b_{12})\,e^{-j2\pi n4000/PCI} \\ + (b_5 \pm b_{13})\,e^{-j2\pi n5000/PCI} \\ + (b_6 \pm b_{14})\,e^{-j2\pi n6000/PCI} \\ + (b_7 \pm b_{15})\,e^{-j2\pi n7000/PCI} \end{bmatrix} \end{aligned}$$

where the plus signs are used for even n and the minus signs for odd n. Note that the $n = 0$ term (i.e. at 100 kHz) is

$$d_0 = \frac{65^3}{4PCI^5} \sum_{k=0}^{15} b_k$$

For the existing Loran phase codes, both master and secondary, we have $\sum_{k=0}^{15} b_k = 4$ so

$$d_0 = \frac{65^3}{PCI^5}$$

distinctly non-zero for all rates. Further, note that different GRIs will share common spectral lines. In the U.S. all GRIs were a multiple of 100 μsec, so the PCIs were all a multiple of 200 μsec and minimally had 200 μsec as their greatest common divisor; the implication is that all U.S. Loran signals had common spectral lines at multiples of 5 kHz. The amplitude at these lines, with $n = 5m\frac{PCI}{1000}$ for integer m, was

$$d_n = \frac{65^3}{4PCI^2\,(PCI + j65\pi n)^3} \left(\sum_{k=0}^{7} b_k \pm \sum_{k=8}^{15} b_k \right)$$

(again, plus for even n and minus for odd n). This is also non-zero for the existing Loran phase codes.

Why do we care about spectral lines? The pulse averaging described above can be viewed as a filter matched to the spectral characteristics of the chain of interest. As the observation length, T, gets large then the averaging becomes a narrow comb filter about the spectral lines of the signal. If the interfering rate has one or more spectral lines very close to ones of the signal of interest, then energy will leak through, create a bias in the averaged pulse, and degrade the eventual positioning performance. To minimize the interference, one could choose rates so that the lines were not too close, and this has been previously discussed in [7]. However, the common lines at $100 \pm 5m$ kHz mean that some energy will always leak through the averaging approach. This interference at 95, 100, and 105 kHz was also noted in [8].

If two nearby chains share common spectral lines, then simple linear averaging methods will not completely cancel both noise and cross rate interference. A modern receiver can avoid this cross rate interference via smart time domain blanking of data when an overlap occurs. Specifically, a receiver could estimate occurrences of cross rate interference and ignore the received data during those periods. However, the reduction in total signal energy due to this data reduction itself results in a lessening of positioning performance. And if the amount of cross rate interference is high, as occurs in the center of the continental U.S., blanking could mean a significant reduction. Further, blanking destroys the ability of the phase code to completely eliminate sky wave interference. In conclusion, we cannot eliminate both sky wave and cross rate interference with the chain based approach.

What's different in *e*LORAN

As mentioned in the Introduction changes were made for *e*Loran under the proviso that legacy Loran receivers would still function normally. Hence, chains, rates, pulses, and phase codes were left unchanged. The system changes include:

- Moving from SAM to TOT control – previously chain timing was under the control of a system area monitor with the result that the emission times for the various stations were not directly tied to UTC and that the chains were not jointly synchronized. The implication was that if a receiver wanted to use signals from two or more chains simultaneously it had to receive the signal from a station in both chains so as to compute the inter-chain offset. With time of transmission control the chains are all synchronized and a modern receiver can be all-in-view meaning that it can effectively use all Loran signals that it can observe from as many chains as possible.

  Of course this change did not impact legacy receivers as they could still solve for the chain offset (zero) and use the additional signals if master was present.

- The Loran Data Channel (LDC) – there has been considerable interest in adding some amount of data carrying capacity to the Loran signal (typical payloads include a time stamp and differential correction data). In the mid-1990's Eurofix modulation was added to Loran transmissions in Europe to provide differential GPS corrections [9]. The method employed is pulse position modulation of 6 of the 8 pulses per Loran group. The amount of modulation was limited so as to have minimal impact on a legacy Loran receiver (appearing like transmitter jitter), yet provide data capacity for a modern receiver. In the 2000's the U.S. experimented with 9^{th}-Pulse LDC, adding an additional pulse at the end of the group (in the inter-station guard interval) [10]. Again the modulation method was pulse position. And since a legacy receiver would not be looking for an extra pulse, the only impact it caused to legacy users was a small increase in cross rate interference.

REVIEW OF PRIOR IDEAS

Our focus in this paper are the system characteristics that have remained unchanged in *e*Loran: rates, chains, and phase codes. There has been prior discussion on these ideas; we review these below.

Change the Rates

This has seen considerable discussion with the aim of decreasing cross rate [7]. No set of choices completely removes it.

Change the Phase Codes

This is not a new idea either. In 1960 Frank replied to early criticism of the Loran codes not being balanced, claiming that the rejection to longer sky wave, while not important for much of the world, still had value in some locations and that the current phase codes were better at CW rejection [11]. (A balanced phase code requires that

$$\sum_{k=0}^{15} b_k = 0$$

and would zero out the spectral line at 100 kHz; this is not satisfied by the current master and secondary phase codes.) In 1974 Roland also noted the value of balanced phase codes for reducing cross rate interference, but at the cost of reduced immunity to longer delay skywave [12]. In 1977 Gressgang et al reconsidered the situation; besides noting the trade-off of reduced sky wave protection versus less cross rate interference, they also mention the need to modify receivers (ah, the beauty of a firmware update!) if the codes were to be changed [13]. At that point in Loran's history, balance lost and their work aimed at choosing the rates to minimize cross rate interference. They do report on experiments of a balanced phase code on a Loran-D mini-chain, showing that it does work.

In 1979 Van Etten suggested that cross rate interference could be eliminated by only tracking specific pulses ("strobing"), essentially converting the current phase code into a balanced subset [14]. For strobing combined with different rates, cross rate interference could be much reduced. The question, of course, is how much SNR loss and sky wave protection loss there is by the reduction in the number of pulses.

At the 2008 IEEE/ION PLANS meeting we focused on the Loran phase codes, arguing that while the master and secondary phase codes completely mitigate long delay skywave, this choice was design overkill; that multipath delays beyond 2 or 3 milliseconds could be safely ignored [1]. Further, we also argued in that paper that the lack of balance in the phase code yielded periodic cross-rate interference that presented itself as a bias in the pseudorange measurements (we presented a detailed treatment of this effect at the 2008 International Loran Association symposium [6]). To further enhance overall system performance we proposed the use of unique balanced phase codes for each chain, mutually orthogonal to the others (basically a form of code division multiplexing). Maintaining the 8 pulse group structure for the transmission, we demonstrated a set of 4 such codes.

Reconfigure the Chains

While retaining the geographic locations of the stations the number of chains and the assignment of specific stations to chains would be quite easy to modify and could result in a significant reduction in cross rate interference.

In 2006 these authors proposed "growing" the chains, increasing the number of stations in each chain, increasing the geographic area covered by each chain, and, hence, reducing the number of chains [15]. Fewer chains at larger distances equates to reduced cross rate interference. Since the Master station might not be visible to a receiver in the enlarged chain, we recognized that some form of positive station ID would be necessary and suggested at that time that it might be achieved via a signature on the 9^{th} Loran data channel.

Spread Spectrum Loran

In 1982 Raab proposed bringing circa 1950's Loran-C into the 1980's, noting that modern receivers can automatically acquire the signal, that spread spectrum modulation is widely used (e.g. in the GPS), and that microprocessors are commonplace in receivers [16]. He proposed staying at 100 kHz, using a 10 kbps pseudonoise biphase modulation (a unique code for each station), and time sequencing the transmissions at each station (basically pulsed modulation to eliminate the near-far problem; he mentions 5 time slots, but provides a figure with 4 labeled M, X, Y, and Z, just like Loran-C). So this is basically a pulsed form of GPS for Loran transmitters, not too different than what is done with GPS pseudolites.

ONE CHAIN, ONE RATE, UNIQUE PHASE CODES

As mentioned in the Introduction, we propose putting all of the *e*Loran transmitters on the same repetition period and employing unique phase codes for each transmitter. To begin the discussion it is convenient to think of Loran as the signal that results from modulation of a ternary code. Specifically, imagine a Loran rate of 50,000 μsec; over 1 PCI the signal consists of pulses starting at the following multiples of 1 msec (offset, of course, by the start time of the PCI):

$$0, 1, 2, 3, 4, 5, 6, 7, 50, 51, 52, 53, 54, 55, 56, 57$$

To view this as a codeword, break the 100 msec PCI into 100 equal width time slots and use the ternary symbols $\{+1, -1, 0\}$ to represent a positive polarity phase coded pulse, a negative polarity phase coded pulse, or no pulse, respectively, in each time slot. Figure 6 shows a sample codeword, with symbols $A[1]$ though $A[100]$; the gaps consisting of 42 zeroes are the quiet periods between groups in the Loran transmission. The codeword for individual secondaries would be right cyclic rotations of this result (shifted by the emission delay in msec); the master station's codeword would have a different set of polarities and two extra non-zero symbols for the master pulses.

To effectively choose new phase codes for *e*Loran we identify their desirable characteristics:

- A receiver must be able to acquire the signal of interest without knowledge of the basic timing of the signal. This will be evaluated by examining the auto- and cross-correlations of the phase codes across all possible lags. We want these low compared to the auto-correlation peak at zero lag to facilitate acquisition in noise.
- A receiver must be able to track each signal of interest to provide an accurate, unbiased pseudorange (unaffected by multipath or interference by the signals from other transmitters). This capability can be evaluated by requiring perfect auto- and cross-correlations for small lags (when the signal is nearly aligned).

For length M codewords A and B with symbols $A[k]$ and $B[k]$, $k = 1, 2, \ldots M$, these metrics are the auto-correlation

$$c_A[k] \equiv \sum_{m=1}^{M} A[m]A[m+k]_{\text{mod } M}$$

and the cross-correlation

$$d_{AB}[k] \equiv \sum_{m=1}^{M} A[m]B[m+k]_{\text{mod } M}$$

$$
\begin{aligned}
\mathbf{c} &= [A[1], A[2], \ldots, A[100]] \\
&= \left[+1,+1,+1,+1,+1,-1,-1,+1,\underbrace{0,0,\ldots 0}_{\text{42 zeroes}},+1,-1,+1,-1,+1,+1,-1,-1,\underbrace{0,0,\ldots 0}_{\text{42 zeroes}} \right]
\end{aligned}
$$

Figure 6: Loran code.

$$
\begin{aligned}
\mathbf{c}_1 &= \left[\underbrace{0,0,\ldots 0}_{\text{10 zeroes}},+1,+1,+1,+1,+1,-1,-1,+1,\underbrace{0,0,\ldots 0}_{\text{42 zeroes}},+1,-1,+1,-1,+1,+1,-1,-1,\underbrace{0,0,\ldots 0}_{\text{32 zeroes}} \right] \\
\mathbf{c}_2 &= \left[\underbrace{0,0,\ldots 0}_{\text{23 zeroes}},+1,+1,+1,+1,+1,-1,-1,+1,\underbrace{0,0,\ldots 0}_{\text{42 zeroes}},+1,-1,+1,-1,+1,+1,-1,-1,\underbrace{0,0,\ldots 0}_{\text{19 zeroes}} \right]
\end{aligned}
$$

Figure 7: Two Loran codewords.

for k ranging from $-M/2$ to $M/2$. As an example, consider two Loran secondaries in a chain with emission delays of 11,000 and 24,000 μsec, respectively (keep the GRI at 50 msec). The corresponding codewords on the 100 slot PCI are shown in Figure 7. The auto- and cross-correlations of these two codewords are shown in Figure 8. This figure shows excellent auto-

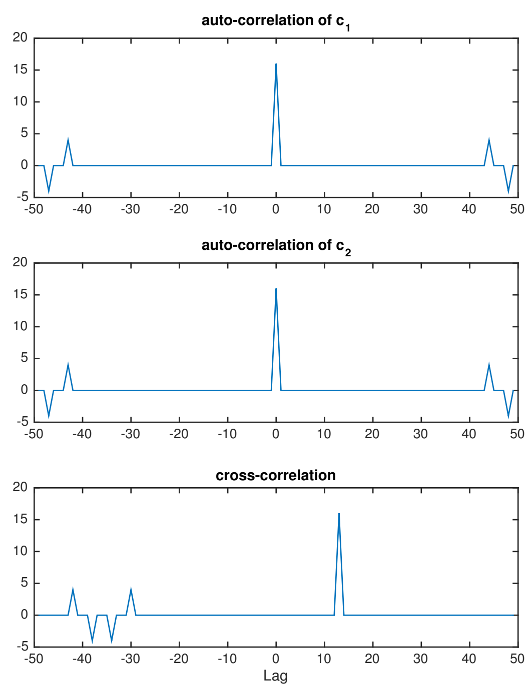

Figure 8: Correlations of two Loran codewords.

correlation characteristics, zero for lags between -42 and $+42$ (excluding the peak at 0), and zero cross-correlation for lags from -29 to $+13$. The zero zone about the origin of the auto-correlation function implies excellent resilience to sky wave; the zero zone of the cross-correlation is the result of the time division multiple access of the emission delays in which (by design) signals within a chain do not interfere with each other for small lags. Unfortunately this codeword representation is not help in characterizing cross rate interference.

Complementary Code Design

Unfortunately there is little known on the general construction of ternary codes. While it is tempting to immediately consider PRN sequences (e.g. Gold codes) they do not appear to be the best solutions due to their non-zero auto- and cross-correlations. Our goal is to completely remove both sky wave and interference from other transmitters even if their signals are much larger (the near-far issue).

To make some headway, we previously proposed extending the group concept; specifically, each codeword would consist of multiple blocks of pulses separated by zero periods [1]. Instead of the two blocks, say A_1 and A_2, of the existing secondary code, we constructed codewords having four blocks (of 8 pulses each) before repeating. Our set of four codewords, found by a search, displayed both excellent auto- and cross-correlations for small lags. Unknown to us at that time, a general design process for such codes is possible.

In 1972 Tseng and Liu developed procedures for constructing sets of mutually orthogonal complementary codes [2]. To employ some of their methods, we review the relevant definitions of sequences and sequence operations:

- Let A be a length n, binary sequence of symbols $+1$ and -1; as an example with $n = 8$, $A = [1, 1, 1, 1, 1, -1, -1, 1]$.
- Let $\widetilde{A}$ represent the reversal of A; for the given A, $\widetilde{A} = [1, -1, -1, 1, 1, 1, 1, 1]$.
- Let $-A$ be the negation of A; again, for the given A, $-A = [-1, -1, -1, -1, -1, 1, 1, -1]$.

Let $(A_1, A_2, \ldots A_p)$ represent a collection of p such sequences, each of length n. The collection of sequences is called a complementary code if the aperiodic autocorrelation sequence is perfect

$$\begin{aligned} c[k] &\equiv \sum_{i=1}^{p} \sum_{m=1}^{n-k} A_i[m] A_i[m+k] \\ &= \begin{cases} np & ; \quad k = 0 \\ 0 & ; \quad k = 1, \ldots, n-1 \end{cases} \end{aligned}$$

We note that formulation is equivalent to the full autocorrelation when the lag is limited to the range 0 to $n-1$ and the groups are separated by zero sequences of length n or greater. Two complementary codes of the same size, say $(A_1, A_2, \ldots A_p)$ and $(B_1, B_2, \ldots B_p)$, are called "mates" if the cross-correlation satisfies

$$d_{AB}[k] \equiv \sum_{i=1}^{p} \sum_{m=1}^{n-k} A_i[m] B_i[m+k] = 0$$

for $k = -(n-1), -(n-2), \ldots, n-1$. As above, this is a portion of the full cross-correlation assuming sufficient length zero sequences between groups.

Tseng and Liu provided methods of constructing new complementary sequences from old ones. They also discussed methods for constructing mutually orthogonal complementary sets. We state without proof several of their results:

- (Application of Theorem 11) Given a complementary pair of sequences (A_1, A_2) then $(\widetilde{A_2}, -\widetilde{A_1})$ is one of its mates. This is, in fact, exactly the case with Loran's master and secondary phase codes with

$$\begin{aligned} A_1 &= [+1, +1, +1, +1, +1, -1, -1, +1] \\ A_2 &= [+1, -1, +1, -1, +1, +1, -1, -1] \end{aligned}$$

- (Application of Theorem 14) If Δ is a matrix whose columns are mates then the columns of

$$\Delta' = \begin{bmatrix} \Delta & \Delta \\ \widetilde{\Delta} & -\widetilde{\Delta} \end{bmatrix}$$

  form a larger collection (twice as many) of mates each with twice as many groups (the notation $\widetilde{\Delta}$ is defined to be the column reversal, but not a sequence reversal, of Δ).

Table 1: 4 codeword example.

#	group 1	group 2	group 3	group 4
1	CA	9F	F9	AC
2	F9	AC	CA	9F
3	06	53	CA	9F
4	35	60	F9	AC

As an example, let's extend the code from the first bullet. Since Δ is defined by

$$\Delta = \begin{bmatrix} A_1 & \widetilde{A_2} \\ A_2 & -\widetilde{A_1} \end{bmatrix}$$

then

$$\Delta' = \begin{bmatrix} A_1 & \widetilde{A_2} & A_1 & \widetilde{A_2} \\ A_2 & -\widetilde{A_1} & A_2 & -\widetilde{A_1} \\ \widetilde{A_2} & A_1 & -\widetilde{A_2} & -A_1 \\ -\widetilde{A_1} & A_2 & \widetilde{A_1} & -A_2 \end{bmatrix}$$

This construction takes two codes, each with two groups of 8 values, and constructs a set of four codes, each with four groups of 8 values and is equivalent to the example presented in [1]. Mapping each $+1$ to 1 and each -1 to 0, the codewords can be written in hexadecimal symbols (two per group) as shown in Table 1.

Extending, we easily can construct a set of 8 codes, each with 8 groups of 8 values, followed by 16 codes, each with 16 groups of 8 values, and then 32 codes, each with 32 groups of 8 values (a large enough code set for the Loran stations in North America).

Tseng and Liu also provide other recursive recipes for building larger sets of mutually orthogonal codes. While also doubling the size of the code by each cycle, the number of values per group also doubles (e.g. 4 codes, each with four groups of 16 values!). Our preference here is to keep the growth smaller.

We envision the new eLoran transmission as being such blocks of pulses, with polarities set by the values of the complementary code. A PCI, then, would consist of p sub-blocks, each of 8 pulses, separated by quiet periods. The codewords, then, consist of p blocks of ± 1 separated by strings of zeros. Figure 9 shows the first codeword of Table 1.

$$\Big[\underbrace{+1,+1,-1,-1,+1,-1,+1,-1}_{\text{group "CA"}},\underbrace{0,0,\ldots 0}_{\text{42 zeroes}},\underbrace{+1,-1,-1,+1,+1,+1,+1,+1}_{\text{group "9F"}},\underbrace{0,0,\ldots 0}_{\text{42 zeroes}},$$
$$\underbrace{+1,+1,+1,+1,+1,-1,-1,+1}_{\text{group "F9"}},\underbrace{0,0,\ldots 0}_{\text{42 zeroes}},\underbrace{+1,-1,+1,-1,+1,+1,-1,-1}_{\text{group "AC"}},\underbrace{0,0,\ldots 0}_{\text{42 zeroes}}\Big]$$

Figure 9: First codeword of the 4-ary code.

Performance of Complementary Codes

By design, complementary codes have perfect auto- and cross-correlation characteristics for small lags. Of interest is what happens for larger lags and how important the results are.

Recall that *e*Loran pulses travel at nearly the speed of light. Assuming a maximum Loran station separation of 5600 km (using existing locations, ignoring Alaska, this is Cape Race in Northeast Canada to Middletown in Southern California, and you'll never hear Cape Race in California) the maximum propagation time difference is approximately 19,000 μsec; in other words, the first pulse in a codeword from one transmitter is received within ± 19 msec (± 19 slots) of the first pulse in the codeword of another transmitter (assuming that the transmitters are synchronous). Hence, a window of at least 28 slots (8 pulse periods for the group transmission itself plus 20 more pulses periods for the differential transmission delay) with zero cross-correlation would suffice to eliminate multiple access interference. In deciding the size of this zero zone we should also account for sky wave from distant stations. So perhaps zero cross-correlation out to 35 slots would be preferred. As to auto-correlation, a zero window of size 8 would suffice to eliminate sky wave interference.

The discussion above assumes that the receiver knows which group is which; i.e. is tracking the signal and the zero auto- and cross-correlations imply no bias in a TOA estimate from either sky wave or other transmitters. For acquisition we care about the entire auto- and cross-correlation sequences. At first glance it might appear that any non-zero cross-correlation might be a problem when acquiring weaker (more distant) signals – the near-far problem. Fortunately, given the Loran time scale, all a receiver need do is acquire the strongest station. Since the desired peak for the weak station is known to be within 28 slots of this point, zero cross-correlation out to lag 28 will allow for correct acquisition of the weaker signal.

For example, let's consider the performance of the 4 codeword code in Table 1. Above we envisioned a GRI of 50 msec, adding 42 zeroes between each group. For this example, let's do the same, creating a total PCI of 200 msec (200 slots) for the four groups of each codeword. The relevant correlation lags, then, will be -100 to $+100$. Figure 10 shows the auto-correlation functions for the four codewords. As expected, each has a peak of 32 at zero lag and otherwise displays zero auto-correlation for lags between -42 and $+42$ (the complementary design only guaranteed -7 to $+7$). These zeroes imply the elimination of sky wave with a delay of up to 34 ($42-8$) slots. The remaining values are bounded within -8 and $+8$, easing the task of acquisition (versus the true peak at 32, 6 dB). Similarly, Figure 11 shows the cross-correlations. The zero correlations for lags from -42 to $+42$ implies that once we acquire the strongest station, all of the others will be found. Of concern are the perfect (value of ± 32) cross correlations between the first and second codewords and the third and fourth. These occur because those codewords are cyclic shifts of one another (and is a function of the code construction procedure); the concern is that a receiver might confuse the codewords. One solution to this problem is to vary the spacing between blocks of pulses.

Large Area (LA) Codes

In 1999 Li presented a technique for creating zero correlation zone codes for application to CDMA [17]. The codes are ternary, consisting of symbols from $\{+1,-1,0\}$. The non-zero symbols of each codeword come from an orthogonal Hadamard matrix (i.e. an K-by-K Hadamard matrix yielding K codewords). The individual codewords are constructed by inserting strings of zeros between the ± 1 values. Notationally, a LA(N,M,K) code consists of K codewords, each of length N, consisting of $N-K$ zeros and K ± 1s. The parameter M represents the shortest length of zeros in the codeword construction (all codewords follow the same pattern). The rules for the zero insertions are:

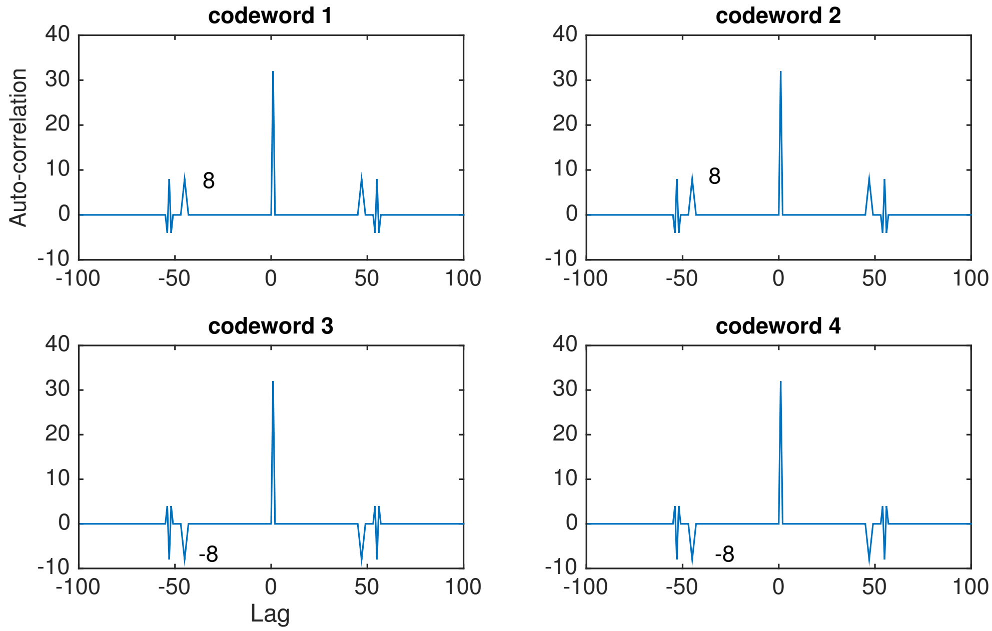

Figure 10: Auto-correlations of the 4 group code.

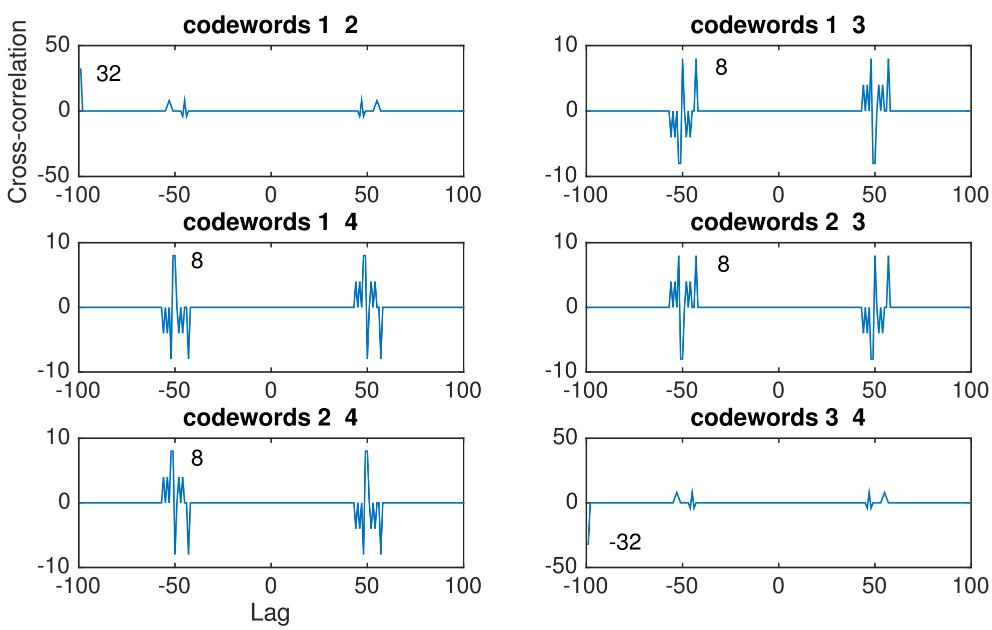

Figure 11: Cross-correlations of the 4 group code.

$$\left[\underbrace{+1,+1,-1,-1,+1,-1,+1,-1}_{\text{group ``CA''}},\underbrace{0,0,\ldots 0}_{\text{27 zeroes}},\underbrace{+1,-1,-1,+1,+1,+1,+1,+1}_{\text{group ``9F''}},\underbrace{0,0,\ldots 0}_{\text{39 zeroes}},\right.$$
$$\left.\underbrace{+1,+1,+1,+1,+1,-1,-1,+1}_{\text{group ``F9''}},\underbrace{0,0,\ldots 0}_{\text{55 zeroes}},\underbrace{+1,-1,+1,-1,+1,+1,-1,-1}_{\text{group ``AC''}},\underbrace{0,0,\ldots 0}_{\text{47 zeroes}}\right]$$

Figure 12: First codeword of the 4-ary code, but with variable spacing.

- The minimum length defines the width of the zero correlation zone; i.e. the auto- and cross-correlations are zero from $-M$ to M.
- All but one of the lengths should be even; only one should be odd.
- Each length should appear only once.
- No length should be the sum of other lengths.
- While the ordering of the zero strings can be permuted, they should be the same for all codewords.

This construction guarantees that for any non-zero shift (i.e. auto- or cross-correlation lag) only one pair of pulses will align; hence, these auto- and cross-correlations are limited to ± 1 and zero. At zero lag the auto-correlation is K and the cross-correlation is zero by the use of a Hadamard matrix.

As an example consider the lengths 6, 8, 10, and 11. A typical codeword would be of length 39 with format

$$\pm 1,\underbrace{0,0,\ldots 0}_{6},\pm 1,\underbrace{0,0,\ldots 0}_{8},\pm 1,\underbrace{0,0,\ldots 0}_{10},\pm 1,\underbrace{0,0,\ldots 0}_{11}$$

This code's auto-correlation at zero lag is only 4 versus ± 1 at a few other points; however, longer codes will provide more gain.

LA codes can be combined with complementary sequences to form LAS codes [3]. For our Loran application each group in the complementary sequence is separated by a different length zero string. For Loran we can think of this as having a fixed PCI with varying GRIs within the PCI.

Let's continue the example with the four codewords in Table 1 using zero strings of lengths 27, 39, 55, and 47, respectively, still resulting in a 200 slot code. Figure 12 shows the construction of the first of the codewords for comparison to Figure 9. Assuming that all 4 codewords have this same staggering, the auto and cross-correlations are shown in Figures 13 and 14. The autocorrelations are slightly worse; while the off peak maximum is still limited to 8, the width of the zero zone is narrower (now only ± 27) and more lags have non-zero auto-correlation. The cross-correlations are much better, however, dropping from a maximum of 32 to a maximum of 8. In LA codes, the worst case auto- (excluding zero lag) and cross-correlations are constructed to be ± 1; using these same ideas for multiple groups, the worst case values should equal n, 8 in this case.

Going to longer (more groups) codes, with proper selection of the zero lengths the maximum auto-correlation increases while the auto- and cross-correlations outside of the zero zone should still be bounded within ± 8. For example, 32 unique codewords with 32 groups each would have a zero lag auto-correlation of 256 (8 times 32) while the non-zero lag auto-correlation and cross-correlations would remain within ± 8. This factor of 32 provides near-far protection for the acquisition process. And, in any case as mentioned above, for acquisition we need only use the strongest station for group timing recovery. Once we know the group alignment, the zeroes in the cross-correlation mitigate any near-far problem.

A Code with 8 Codewords

This same construction can be continued. Tseng and Liu's method easily yields a set of 8 unique phase codes, each composed of 8 groups of 8 pulse amplitudes. Mapped to binary (1/0) symbols, hexadecimal representations for the codewords appear in Table 2. Since they are mutually orthogonal complementary sequences, we have perfect auto- and cross-correlations for small lags. To yield good acquisition we must find a set of lengths for the intervening zero sequences to eliminate overlap. Even spacing does not work well in this case; for example, inserting blocks of 42 (or 52 or 62 ...) zeros results in a worst case auto-correlation of 8 and a worst case cross-correlation of 64, not effective when compared to the zero lag auto-correlation of 64. It is, however, possible to find a set of varying gaps to improve matters. For example the set $\{60, 69, 78, 87, 96, 105, 114, 145\}$ achieves maximum

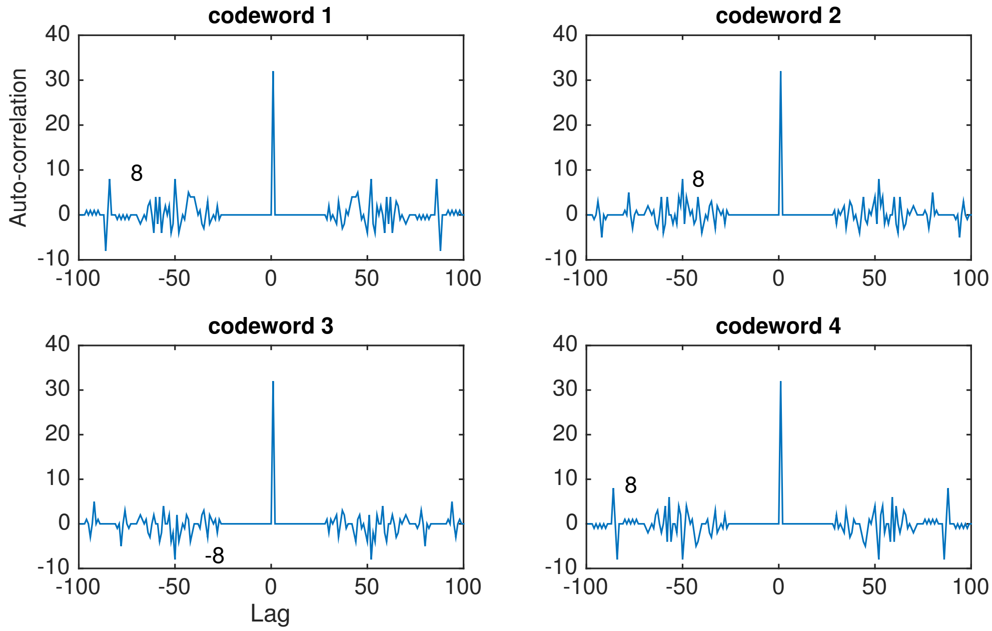

Figure 13: Auto-correlations of the 4 group code using unequal group spacing.

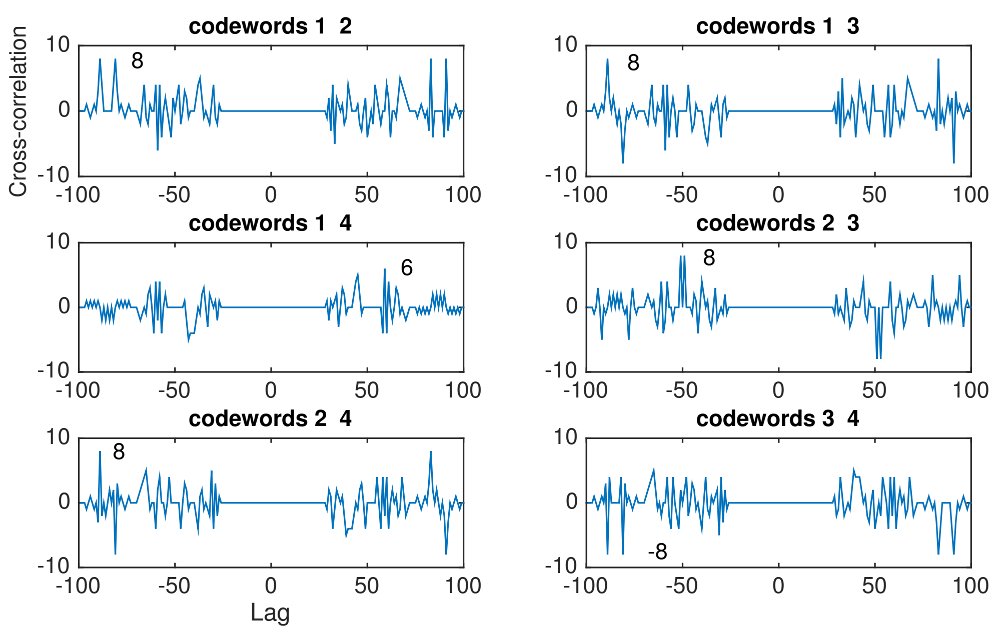

Figure 14: Cross-correlations of the 4 group code using unequal group spacing.

Table 2: Code of size 8.

codeword #	group 1	group 2	group 3	group 4	group 5	group 6	group 7	group 8
1	CA	9F	F9	AC	35	60	F9	AC
2	F9	AC	CA	9F	06	53	CA	9F
3	06	53	CA	9F	F9	AC	CA	9F
4	35	60	F9	AC	CA	9F	F9	AC
5	CA	9F	06	53	CA	9F	F9	AC
6	F9	AC	35	60	F9	AC	CA	9F
7	06	53	35	60	06	53	CA	9F
8	35	60	06	53	35	60	F9	AC

$$\left[\underbrace{+1,0,0,+1,0,0,-1,0,0,-1,0,0,+1,0,0,-1,0,0,+1,0,0,-1,0,0,}_{\text{group "CA"}},\underbrace{0,0,\ldots 0}_{\text{88 zeroes}},+1,0,0,-1,0,0,\ldots\right]$$

Figure 15: The start of the first codeword of from Table 2 assuming 1/3 msec pulse spacing.

(non-zero lag) auto- and cross-correlations of 8. Summing the number of slots yields a PCI of 818 (0.818 seconds assuming 1msec pulse spacing) or an average GRI of 102.25 μsec. This is a little large, we would prefer 15-20 groups per second.

An alternative, so as to reduce the average group duration, is to allow for fractional spacing of the pulses. Recall that the existing *e*Loran pulse has a duration of approximately 300 μsec. If we imagine three time slots per msec, then we can use this fractional spacing to stagger the groups, dramatically reducing out of group overlap. Our current construction maintains the 1 msec minimum spacing of the pulses; equivalently, every ± 1 in the codeword is followed by two 0's so that each group fills 24 slots. To exploit this fractional spacing, consider having 8 zero sequences of lengths

$$88, 175, 80, 185, 121, 224, 173, 262$$

With these lengths the total PCI is duration $88+175+80+185+121+224+173+262+8*24 = 1500$ slots; at 3 slots/msec this is 500 msec for a total of 16 groups per second. Figure 15 shows the first portion of the first codeword to demonstrate the construction.

COMMENTS

To reiterate our view of this work – the reader should recognize that these ideas and results are <u>not</u> intended to define what the best *e*Loran system is; rather, if *e*Loran soars again in the U.S., we hope that these remarks initiate a dialogue that looks beyond the decisions made in the 1950's.

To close we have several comments:

- We believe that through the use of code division multiplexing with all stations broadcasting simultaneously, and with the same timing, can eliminate both sky wave interference as well as interference from the other transmitters, resulting in unbiased time of arrival estimates. This is true whether the signal is being used for position or dissemination of precise time.
- To consider examples we kept the basic pulse shape, pulse separation of 1 msec, binary polarity modulation (± 1), and 8 groups per group:
  - A shorter pulse (< 250 μsec) would be better in that we could stagger pulse blocks more easily as in LA codes.
  - Closer or wider spacing of the pulses could be considered. Closer spacing would allow for more energy transmission per second; wider spacing could help eliminate sky wave and reduce near-far issues.
  - An obvious extension is to allow complex polarity (sign and phase) to ± 1, $\pm j$.
  - Having 8 pulses per group should be reexamined; we kept this for convenience only.
- The last design example resulted in 8 unique codewords although the extension to 16 or 32 is obvious. With a smart selection of power level and reuse, it might be possible to cover North America with 16 or maybe even 8 codewords.

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