PURDUE UNIVERSITY
SCHOOL OF ELECTRICAL ENGINEERING
ELECTRONIC SYSTEMS RESEARCH LABORATORY

SEMI-ANNUAL REPORT OF RESEARCH

PERFORMED UNDER GRANT NsG–553

July 1, 1966 through December 31, 1966

LAFAYETTE, INDIANA

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TABLE OF CONTENTS

Foreword

This report summarizes work carried out at the Electronic Systems Research Laboratory of Purdue University under NASA Grant NsG-553 during the period July 1, 1966 through December 31, 1966.

In keeping with NASA's policy for administration of research grants the report has been kept as concise as possible, and when appropriate reference has been made to interim reports, internal memoranda, and technical papers resulting from research carried out under this grant.

The format of the report consists of a listing of technical papers and internal memoranda which have been submitted to NASA during the period covered by the report, abstracts of interim reports submitted during the reporting period, and appropriate extracts from the current Purdue University, School of Electrical Engineering, Semi-Annual Research Summary.

John C. Lindenlaub

John C. Lindenlaub,
Principal Investigator

Clare D. McGillem

Clare D. McGillem, Director
Electronic Systems Research Laboratory

I. Technical Papers and Internal Memoranda

Two conference paper abstracts were submitted to NASA Headquarters on November 30, 1966

i) "Families of Equal-Length Shift Register Sequences Obtainable from a New Class of Cyclic Codes," by D. R. Anderson. Presented at the International Communications Symposium, Philadelphia, June, 1966.

ii) "Simulation of a Tropospheric Scatter Channel," by C. C. Bailey and J. C. Lindenlaub. Presented at the International Scientific Radio Union meeting, December 7-9, 1966, Palo Alto, California.

A technical journal reprint of "Nonsupervised Sequential Classification and Recognition of Patterns," by E. A. Patrick and J. C. Hancock, IEEE Transactions on Information Theory, Vol. IT-12, No. 3, July, 1966, was submitted November 30.

Electronic Systems Research Laboratory Memorandum Report 66-1, "Some Experimental Results on Communication Systems Subject to Intersymbol Interference," by J. C. Lindenlaub and C. C. Bailey was sent to NASA September 16, 1966.

Electronic Systems Research Laboratory Memorandum Report 66-2, "Tapped Delay Line Simulation of Randomly Time-Variant Channels," by J. C. Lindenlaub and C. C. Bailey was submitted on November 30, 1966.

II. Abstracts of Interim Reports

Two interim technical reports were completed during this reporting period. The first report is entitled "On The Optimization of Mixture Resolving Signal Processing Structures," by J. C. Hancock and W. D. Gregg. Research for this study was partially supported by NsG-553. An abstract of this report is given below:

This report is concerned with the development of the structural form of optimum linear detection operators when the pulse waveform and noise parameters are unknown and are to be obtained by mixture resolving estimation; and with the development of the mixture resolving estimators to learn or extract this parametric information from the noisy signal pattern mixture, in order to obtain the elements for the structure. The observation signal model consists of a discrete, multidimensional, binary (two category) gaussian mixture.

An eigenvalue approach is taken for the development of the structural form of the detection operator; the criterion of optimality being the minimum average conditional probability of error at the Nth stage, conditioned upon the mixture resolving estimating category. The mixture resolving categories developed consist of an optimized, time-varying-weighted decision-directed category and a moment method category. These categories differ from related work in that; (1) In the former, the initial reference is extracted from the signal mixture by correlating the first observation with the next and updating the result with successive-time-varying-weighted combinations of the separated time slots optimized to minimize a measure of distance and dispersion with a subsequent "maximization" of convergence rate and (2) in the latter, no apriori knowledge of either the pulse waveform or noise parameters is required.

Extensive experimental studies, via digital computer simulation, of the performance characteristics of these signal processing algorithms are carried out and compared with the conventional decision directed and Bayes matched filter algorithms under identical input conditions. A complete formulation of this approach, to include verification of the theory by digital computer simulation experimentation, however, is carried out only for the bi-polar case ($\underline{\theta}_2 = \underline{\theta}_1$) in the equiprobable situation ($p_1 = p_2 = \frac{1}{2}$). An analysis for the general case is carried out and the difficulties encountered by the lack of specific apriori information are discussed.

Both algorithms developed in this work converge for negative db. values of SNR at a rate considerably higher than that of the conventional decision-directed algorithm and are bounded from above in performance by the conventional decision-directed algorithm over the entire SNR range investigated. In addition, the weighting in the optimized-weighted-decision directed ("rate maximized") algorithm, over the SNR range investigated, is dominated only by the observation stage, N, and the constraint coefficient, γ; and, in that sense, is non-parametric in the mixture pulse waveform and noise parameters. A signal processing interpretation of the digital computer simulation of the numerical experimentation implies that knowledge of the required signal dimensionality is available.

The second technical report by John C. Lindenlaub and John J. Uhran is entitled "Threshold Study of Phase Lock Loop Systems." The abstract of this report appears below:

This research has been concerned with the study of two aspects of phase lock loop systems. The first is the effect that the phase comparator, a non-linear device operating on the phase difference between reference and signal has on overall performance. The second directly related to the first, is the study and defining of the threshold phenomena.

Several comparators were chosen ad-hoc and analyzed using a Fokker-Planck model. Because the model is only approximate in the threshold region a carefully designed experimental evaluation of a first and second order sine, tanlock and tanlock squared comparator was undertaken.

It is shown that the additional complexity involved in using other comparators rather than the relatively simple sine comparators can be justified particularly when additional tracking range is required. For second order systems, it is shown that critical damping provides optimum performance in terms of tracking and threshold. The inclusion of a section on the limiter is given for comparison purposes.

The threshold is defined in terms of the rate of cycle slipping i.e., as output spikes/sec. It is an easily measurable quantity and the acceptable level once chosen is independent of modulation, noise and loop filter. It also provides the easiest method of comparing the thresholds of various systems since the curves are all parallel to each other. It is also shown that in many cases of interest the output spikes follow a poisson time distribution.

Copies of these interim reports have been sent to NASA Headquarters and to reviewers at NASA centers.

III. Research Summaries

N67 16932

A. Digital Simulation of a Randomly Time-Variant Communication Channel

J. C. Lindenlaub

C. C. Bailey

The program in simulation of communication systems[1] is being extended to the simulation of systems employing randomly time-variant channels. The first phase of this program is the development of digital computer programs to simulate the effects of randomly time-variant channels. A set of programs is now available which can provide simulation of a general class of channels. This simulation system has been applied to the case of tropospheric scatter communication channel.

The basic theory for the simulation system is due to Stein[2]. This requires that the channel model to be simulated obey the following assumptions:

1) The channel is a linear time-variant system. Thus it has a time-varying impulse response $h(\xi,t)$ and an associated time-varying transfer function $H(f,t) = F_{\xi}\left\{\xi h(\xi,t)\right\}$

2) The channel impulse response, $h(\xi,t)$, is stationary complex Gaussian random process.

3) Only bandlimited signals are used as inputs to the channel.

4) The correlation function of the channels' time-varying transfer function, $R_H(\Omega,\delta) = E\left[H(f,t)H^* \,(f+\Omega,t+\delta)\right]$ is factorable into a function of Ω and a function of δ, i.e.

$$R_H(\Omega,\delta) = v(\Omega)q(\delta)$$

The tropospheric scatter channel model used for the testing of the simulation system is the Sunde model[3]. For this model, the channel correlation are

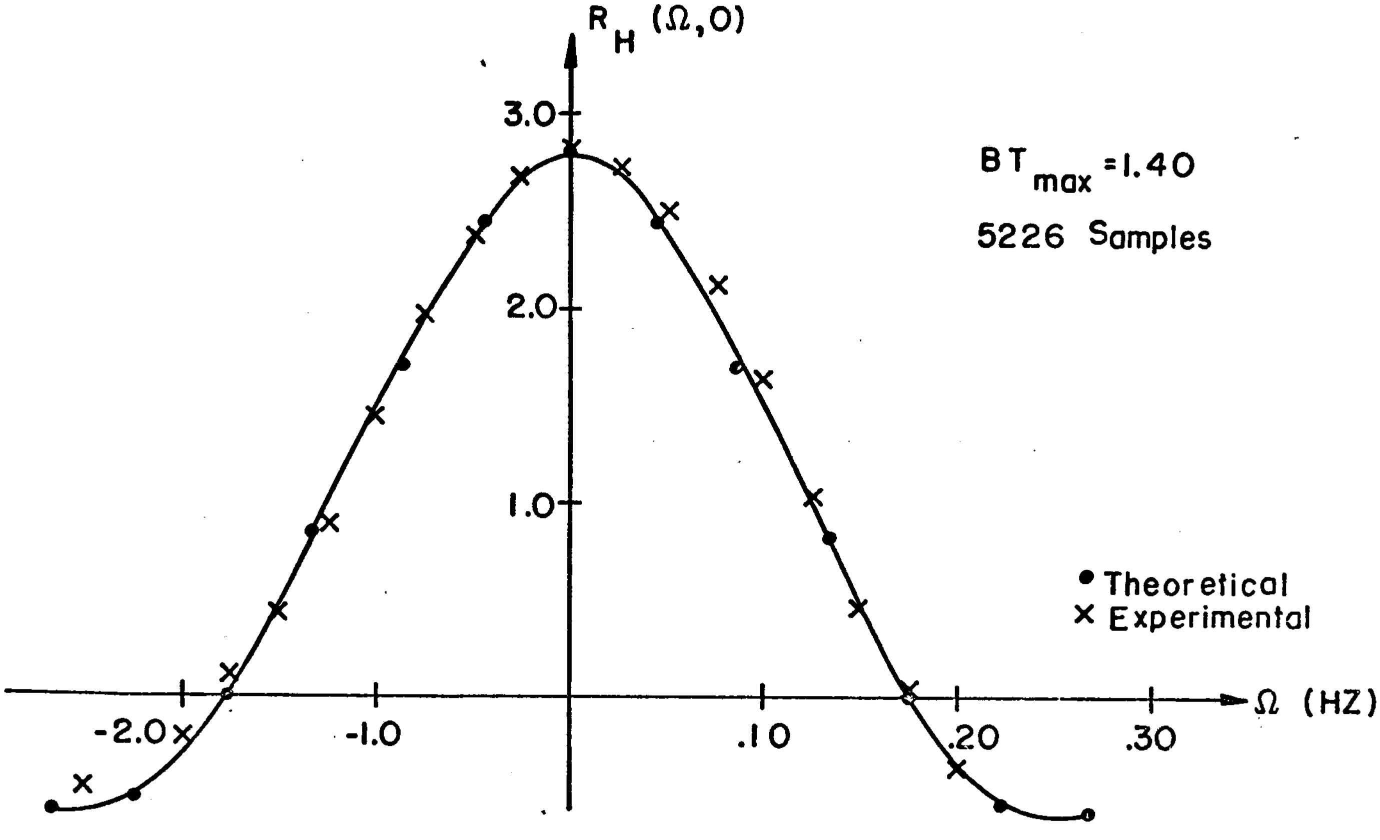
$$R_H(\Omega,0) = v(\Omega) = 2R_o \frac{\sin 2\pi\Omega\, T_{max}}{2\pi\Omega\, T_{max}}$$

$$R_H(0,\delta) = q(\delta) = \exp\left[-\frac{\sigma^2\delta^2}{2}\right]$$

where T_{max} is the maximum departure from the average transmission delay over the channel, R_o is the mean square value of the channel gain, and σ is a measure of the rapidity with which channel fluctuations take place. This channel model has been simulated for the following parameter values

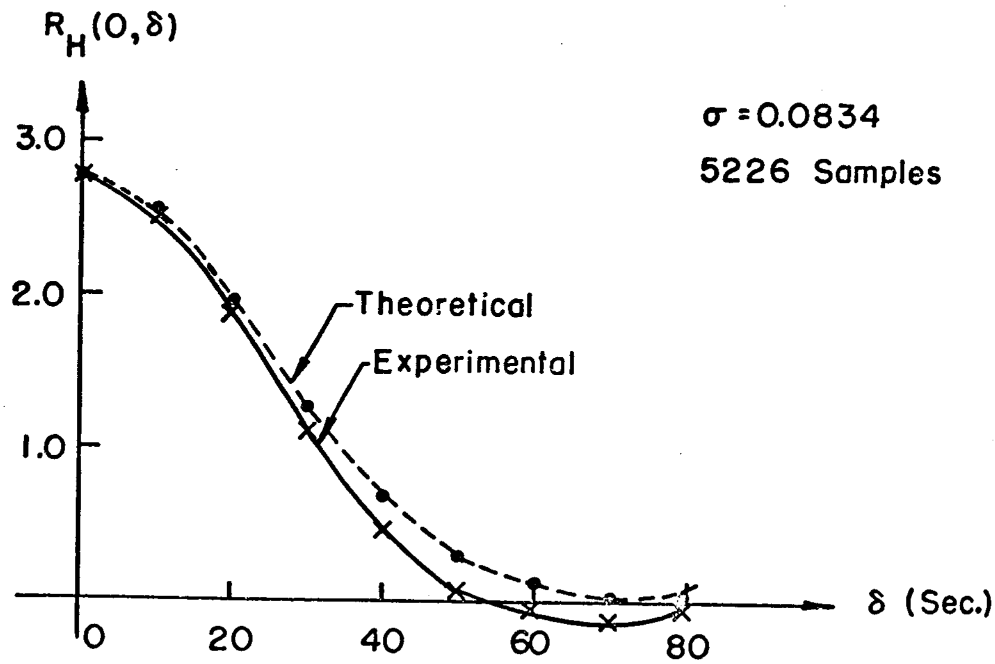
1) input signal bandwidth = $\frac{1}{2}$ Hz.
2) T_{max} = 2.8 sec.
3) R_o = 2
4) σ = .0834 sec^{-1}

Experimental measurements of $R_H(\Omega,0)$ and $R_H(0,\delta)$ have been made with this simulation system. The measurement of $R_H(0,\delta)$ is performed by exciting the channel with a sinusoid, demodulating the channel output with a synchronous detector, and autocorrelating the output of the detector. The measurement of $R_H(\Omega,0)$ is performed by exciting the channel with two sinusoids whose frequencies differ by Ω Hz, demodulating the channels' response to each sinusoid with synchronous detectors, and crosscorrelating the outputs of the two detectors. Figures 1 and 2 show theoretical and experimental graphs of $R_H(\Omega,o)$ and $R_H(0,\delta)$ for this system.



Frequency Correlation Function $R_H(\Omega,0)$ vs. Ω

Figure 1



Time Correlation Function $R_H(0,\delta)$ vs. δ

Figure 2

References

1. Bailey, C. C., and J. C. Lindenlaub, "Experimental Results on Communication Systems Subject to Intersymbol Interference," Fourth Semi-Annual Research Summary, School of Electrical Engineering, Purdue University, Lafayette, Indiana.

2. Stein, Seymour, "Theory of a Tapped Delay Line Fading Simulator," First IEEE Annual Communications Convention, pp 601-607, June, 1965.

3. Sunde, E. D., "Digital Troposcatter and Modulation Theory," BSTJ, Vol. 43, pp 143-214, Jan., 1964, (Part I).

B. Phase Lock Loop Studies

J. C. Lindenlaub

J. J. Uhran

The study to determine the effects of the class of $n^{\underline{th}}$ order tanlock phase detector characteristics upon the design parameters of first and second order phase lock loops has been completed. During the past six months the first order system results, which were reported on in the last semi-annual report, have been augmented with second order system results and a technical report summarizing the work has been prepared (TR-EE 66-19).

The main findings of the study are that the lock range capabilities of the $n^{\underline{th}}$ order tanlock systems exceed those of the sine comparator for high signal to noise ratios but fall off faster than the sine comparator as threshold is approached. Synchronization times of the $n^{\underline{th}}$ order tanlock systems are superior to the ordinary phase lock loop. When normalized to equivalent noise bandwidth the tanlock systems exhibit a higher threshold, but when normalized to lock range the tanlock systems have a slightly lower threshold. We have defined threshold in terms of the rate of cycle slipping. Curves showing these characteristics may be used to aid in the design of tanlock or ordinary phase lock loop systems.

C. A Phase Lock Loop System with a Modulo $2n\pi$ Phase Detector N67 16933

J. C. Lindenlaub

D. P. Olson

As a continuation of the study to determine the effects of phase detector characteristics on phase lock loop design parameters discussed above, a study has been initiated on a phase lock system with a phase detector characteristic that is linear and periodic with period $2n\pi$ radians instead of the usual 2π radians. Of particular interest is the threshold behaviour of the modulo $2n\pi$ system. The mathematical or experimental anlaysis of the threshold performance of such a system has not been noted in the literature.

It is felt intuitively that such a phase lock loop has a lower threshold and intermodulation distortion than the conventional phase lock loop. It is believed that the larger phase error range will decrease the probability of cycle slipping. This will decrease the noise due to 2π impulses in the output and hence decrease the threshold. In addition, the linearity inherent in this phase detector will decrease the intermodulation distortion at the output.

A phase lock loop using this type phase detector ($n = 2^i$, $i=1$ to 9) is currently under construction. Figure 1 is a block diagram of the system. In area 1 of this figure the input signal, consisting of a frequency modulated carrier plus band limited noise, is sampled by a locked reference oscillator. This is done by two sample and hold circuits operating in quadrature. The yield is two signals proportional to the sine and cosine of the phase error between the input signal and the reference oscillator. In area 2 the sign of the sine and cosine voltages is detected and used to control a gating network that feeds two adders in area 3. The output of the upper adder is

$$e_n = A\left[\mathrm{SIGN}(\sin\theta_e)\cos\theta_e - \mathrm{SIGN}(\cos\theta_e)\sin\theta_e\right]$$

and for the lower adder is

$$e_d = A\left[\mathrm{SIGN}(\cos\theta_e)\cos\theta_e + \mathrm{SIGN}(\sin\theta_e)\sin\theta_e\right]$$

$$= A\left[\,|\cos\theta_e| + |\sin\theta_e|\,\right]$$

where A > 0 is the input signal amplitude. The output of the adders enter area 4 where an analog divider calculates

$$\frac{e_n}{e_d} = -\tan\left[\frac{\pi}{4} + \theta_e - (K-1)\frac{\pi}{2}\right]$$

where K is the quadrant of θ_e modulo 2π. It can be shown that

$$A/\sqrt{2} < e_d < A\sqrt{2}.$$

Thus the analog divider denominator is never zero and always positive for A > 0. Therefore the divider never saturates, also the divider output in independent of A.

In area 5 an inverse tangent operation upon the output of the analog divider produces the phase, θ_e, modulor $\pi/2$. The non linear inverse tangent operation is approximated to within $\pm$ 1 % by 7 line segments over the $\pm\,\pi/4$ range. A signal determined by the quadrant of the modulo 2π phase is added to the modulo $\pi/2$ phase to produce the modulo 2π phase. The quadrant is determined by a logic circuit connected to the sign (A sin θ_e) and sign (A cos θ_e) detectors in area 2. Also a signal proportional to the integral number of cycles slipped is added to the modulo 2π phase to obtain the modulo $2n\pi$ phase.

In area 2 the A sin θ_e and A cos θ_e signals are individually operated upon by single pole low pass filters to improve the signal to noise ratio. Then the sign of each filtered signal is determined.

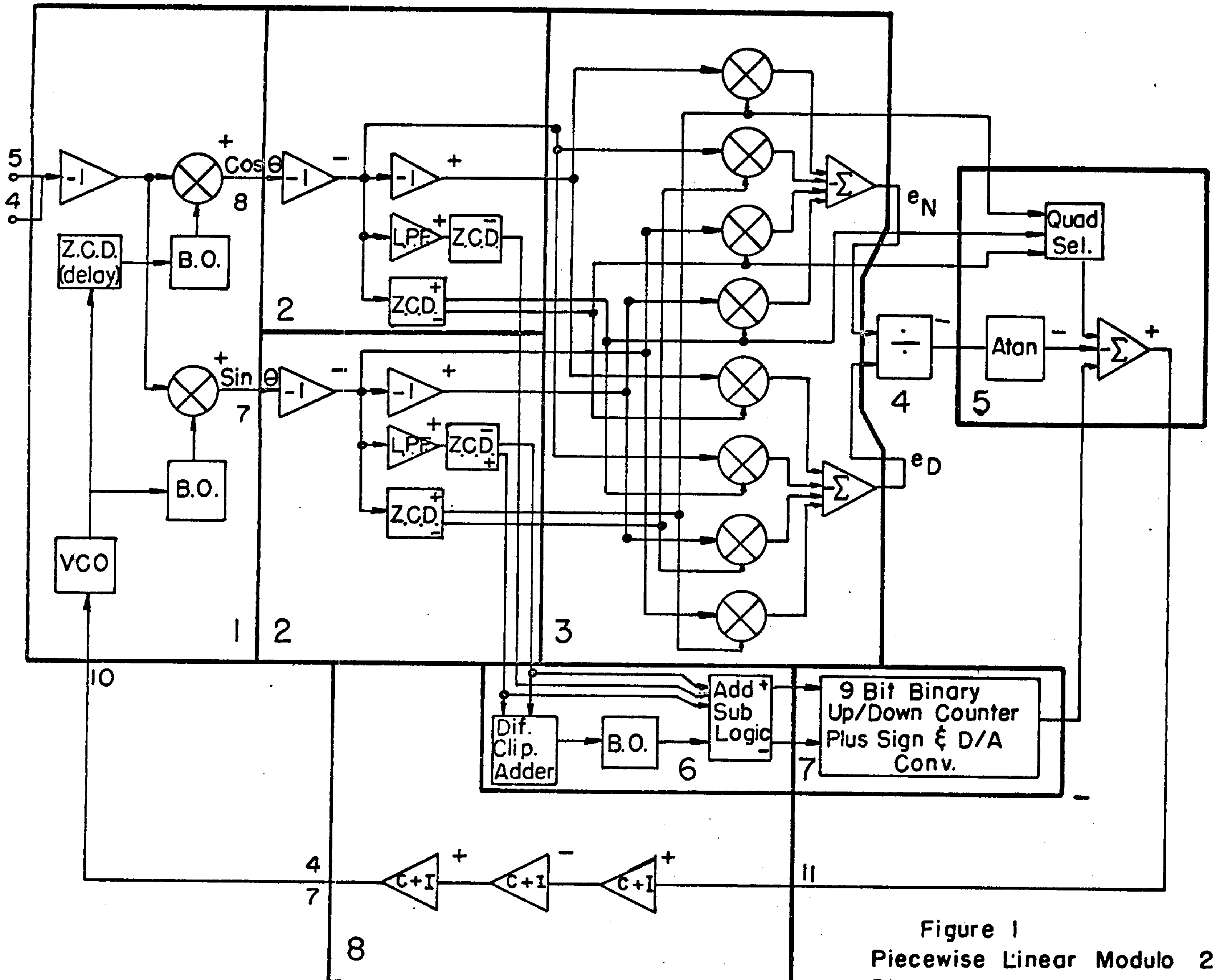


Figure 1
Piecewise Linear Modulo 2nΠ
Phase Lock Loop

In area 5 a logic circuit operating on the sign of the filtered sine and cosine is used to determine the occurance and direction of cycle slipping. That is if $F(\sin \theta_e)$ becomes positive while $F(\cos \theta_e)$ remained negative a negative cycle slip has occurred. (Here $F(\cdot)$ denotes the filtered signal.) Similarly if $F(\sin \theta_e)$ becomes negative while $F(\cos \theta_e)$ remains negative a positive cycle slip has occurred.

Area 7 is a 9 big up-down counter plus sign bit and digital to analog converter. If a positive cycle slip occurs, one is added to the count stored in the counter. Similarly, if a negative cycle slip occurs, one is subtracted. The analog output signal is proportional to the count including its sign. This signal is added to the modulo 2π phase in area 5 as mentioned above.

Area 8 is the loop filter in which any combination of poles and zeros up to 3 each can be realized thru use of 3 operational amplifiers connected as ideal integrators. The output of the filters is used in area 1 to control the frequency of the reference oscillators.

The prototype models of areas 2,3,5 and 8 have been constructed in the lab and work individually. Area 4 will be a Philbrick analog divider. The sample and hold circuits in area 1 are presently under development.

D. Correlation Detector

P. A. Wintz
R. A. Markley

It is well known that the optimum detector of known signals in white gaussian noise is a correlation detector. The correlation detector must correlate the received data with replicas of the known transmitted signals, and decide in favor of the signal having the largest correlation.

The design and testing of a binary detector using Hall effect multipliers has been started. Circuitry for generating and synchronized sample and dump pulses, an analog integrator and dump circuitry, and a sampler have been designed and testing is nearly complete. Work now in progress includes testing and adapting the Hall multipliers to the system and design at the decision logic circuitry.

E. M-ary Signal Sets for Scatter-Multipath and Other Asynchronous Communication

D. R. Anderson

As indicated by Anderson[1] the optimum equal-energy, common bandwidth occupancy binary signal set for the Turin and Price scatter-multipath channel models is characterized by the following property: the peak of the envelope of the cross-correlation is essentially a constant multiple of $(2TB)^{-\frac{1}{2}}$, TB being the common time-bandwidth product. This condition also characterizes the optimal binary signal set with the same constraints for non-synchronous, phase-incoherent reception in the presence of additive white noise. In the case of M-ary signal sets the condition for optimality when there is common frequency occupancy is not known. However, both Lerner[2] and Yates and Cooper[3] have found constant-envelope M-ary signal sets with common frequency occupancy and common TB-product equal to M for which the peak of the envelope of the cross-correlation of every pair is at most $\ln M/M^{\frac{1}{2}}$. In view of the condition for optimality in the binary signal-set case, these results appear to be close to optimal. A current paper exhibits a method of using shift register sequences[4] to obtain constant envelope p-ary signal sets for every prime p and every $n \geq 1$. These signal sets have the property that:

1) every signal set is generatable by a p-ary shift register.
2) all the members of a given set have a TB product equal to p.
3) for a given p^n-ary signal set, all envelopes of crosscorrelation functions are at most $\ln(p^n)/p^{\frac{1}{2}}$.

References

1) Anderson, D. R., "On Optimum Binary Signal Sets for Certain Scatter-Multipath Channel Models," IEEE International Symposium on Information Theory, Los Angeles, California, February, 1966.

2) Lerner, R. M., "Signals Having Good Correlation Functions," 1961 WESCON Convention.

3) Yates, R. D., and G. R. Cooper, "Design of Large Signal Sets with Good Cross-correlation, IEEE International Symposium on Information Theory, Los Angeles, California, February, 1966.

4) Anderson, D. R., "Distinguishable Synchronization Sequences Arising From Cyclic Codes," International Conference on Communications, Philadelphia, Pennsylvania, June, 1966.

F. Use of Channel Simulator

D. R. Anderson

S. E. Nykanen

A description of the channel simulator has been previously given by Wintz and Markley in previous semi-annual reports. During this period, the amenability of the physical channel simulator to the use of broad-band signals has been investigated. The purpose of this work is to investigate experimentally some of the effects of multipath, scatter, and doppler shifts on a transmitted chirp signal and to attempt to design signals which minimize (in some sense) their effects.

A major part of the effort has gone into the design, construction and testing of the electronic circuitry comprising the interface between the chirp signal generator and the channel simulator.

G. Signal Deblurring

C. D. McGillem

S. S. C. Yao

When a signal is passed through a system the signal is frequently altered as a result of interaction with the system. This process may be thought of as "blurring" of the signal. In many instances the exact nature of the blurring is known and the question is raised as to the possibility of recovering the original signal by suitable processing of the blurred signal. Some examples will illustrate the nature of the problem. Consider the recording of an electrical signal on a magnetic tape. Due to the finite width of the airgap in the recording head the signal is simultaneously applied to a finite length interval on the tape. Accordingly then the intensity of recording at any point on the tape is the integrated value of the signal that occurred while the tape passed under the airgap and is thus a "running average" of the actual instantaneous signal. Mathematically the recorded signal can be expressed in terms of the actual signal as

$$y(t) = \int_{t-T/2}^{t + T/2} x(\xi)\, d\xi \tag{1}$$

where $x(t)$ is the original signal, $y(t)$ is the recorded signal and T is the time required for a point to move through the airgap. It can be readily shown that the integral above is equivalent to convolving the signal with a rectangular pulse of unit height, thus

$$y(t) = x(t) * h(t) \tag{2}$$

where $h(t)$ is the blurring function--in this case a rectangular pulse of width T centered at the origin. Ideally one would solve a problem of this sort by using the Fourier or Laplace transform. Formally this leads to

$$Y(\omega) = X(\omega)\, H(\omega)$$

$$X(\omega) = Y(\omega) \cdot \frac{1}{H(\omega)}$$

$$x(t) = y(t) * F^{-1}\left\{\frac{1}{H(\omega)}\right\} \tag{3}$$

Unfortunately $F^{-1}\left\{\frac{1}{H(\omega)}\right\}$ does not in general exist for functions, $h(t)$, of interest. In the present case

$$H(\omega) = F\{h(t)\} = T \frac{\sin \omega T/_2}{\omega T/_2} \tag{4}$$

$$\frac{1}{H(\omega)} = \frac{1}{T} \frac{\omega T/_2}{\sin \omega T/_2} \tag{5}$$

This is not a well behaved function since it is unbounded as $\omega \to \infty$ and furthermore has infinite discontinuities at $\omega = \frac{2n\pi}{T}$, $n = 1,2,\ldots$ There is, accordingly, no inverse transform. There are never-the-less ways to handle problems of this sort and it is the purpose of this research to investigate a number of these methods. Before discussing the methods some additional examples where deblurring would be important should be mentioned. Photographs that are blurred due to image motion or defocussing are an example of process where the blurring operation is known precisely and recovery of the original would be highly important. In such cases the deblurring is two dimensional. Transmission of waveforms through channels of limited bandwidth causes distortion that can be removed by deblurring techniques. Beam sharpening in radar and crispening of T.V. pictures are other examples of deblurring.

Ultimately the limitation on deblurring will be determined by noise present in the system or in the processing itself. However much improvement is possible even with noise present.

Several methods of deblurring are being investigated. In one method the reciprocal of the transform of the blurring function is approximated by various infinite series. The inverse transformation is then carried out on a term by term basis and deblurring is obtained by convolution of the blurred signals with partial sums of these terms. It has been demonstrated that such procedures are valid in many cases of interest but generally appear to be limited to signals with only modest amounts of blurring. In the presence of heavy blurring the partial sums are divergent.

Another method that is being studied is to solve the integral equation (2) for particular blurring functions (kernels) by working in the time domain. For example, solutions for rectangular pulses, single loop cosine and raised cosine pulses, and Gaussian pulses have been obtained. Such solutions involve derivatives of the blurred signal. In the case of the rectangular pulse, only the first derivative is required to obtain an exact recovery of the signal whereas for the Gaussian pulse derivatives of all orders are required. This method of approach appears very promising and further studies are being made.