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FEASIBILITY STUDIES OF SURVEILLANCE, COMMUNICATION, AND DATA PROCESSING SUBSYSTEMS FOR ADVANCED AIR TRAFFIC MANAGEMENT

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16. Abstract

Analyses are made of waveforms, parameters, codes, error rates, and multi-access noise for proposed communications and surveillance subsystems to be useful for air traffic control in the 1990-2000 time period. The systems represented in these analyses are all based on the use of satellites. Included is a computer sizing comparison for 3 proposed systems.

This report presents a collection of technical study papers as a partial contribution toward the feasibility assessment of the proposed systems.

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1.0 INTRODUCTION

A study team at the Transportation Systems Center has been investigating the technical feasibility of proposed concepts for advanced air traffic management systems, (AATMS). The objective of this effort is to gain a clearer understanding of the problems involved, to determine some of the system constraints and limitations, and to better appreciate the technical risks. To meet this objective, and to review and evaluate the proposed systems properly from a technical standpoint, independent studies are being conducted at TSC.

This report presents results of a part of the technology study effort that deals with several key problem areas relating to asynchronous transmissions for a satellite-based surveillance and communication system. This surveillance technique, which uses difference in time-of-arrival measurements to determine aircraft position, has become commonly identified as the "LIT system" - Location, Identification, Transmission system. Work on the LIT system was accomplished by TRW and reported in "Communication, Navigation, and Surveillance", TRW No. 14671-6007-RO-00, Dec. 1970. LIT-like surveillance systems have also been recommended by North American-Rockwell Autonetics and Boeing Co. in work accomplished under contract to DOT-TSC.[1,2] Study papers prepared to date by TSC on this subject are included in this report. These papers discuss the subject with emphasis on signal structures and data processing and are summarized in the next section.

2.0 SUMMARY DISCUSSION

A typical LIT type surveillance system is shown in Figure 1. The aircraft transmits asynchronously phase coded signals to a central ground station via a constellation of at least four satellites. The position of the aircraft is determined by measuring differences in the time-of-arrival (TOA) of the transmitted signals. The transmitted waveforms also contain the aircraft's identity. Several problem areas and contemplated requirements related to this type of system were studied and are summarized below. Modifications of the LIT system were used as models for these study papers on Surveillance and Communication Signal Structures and on Data Processing.

2.1 SURVEILLANCE AND COMMUNICATION SIGNAL STRUCTURES

2.1.1 Aircraft Surveillance and Back-Link Communication Requirements

The design requirements relating the aircraft peak power required for the surveillance function, and for pulse width and code length bi-phase modulating the transmitted signal is determined in study paper A. With conventional transmitters, the required aircraft peak power for the aircraft to satellite link would be too high. To satisfy the signal power requirements without sacrificing the surveillance accuracy, waveforms with large time-bandwidth products are utilized. In the particular systems under consideration, this is obtained by bi-phase modulating the transmitted signal with a pseudo random code consisting of L code elements. The power requirements also depend on other parameters such as the aircraft and satellite antennas and the required signal-to-noise ratio. With the aid of Figure A-1 the aircraft surveillance requirements are readily obtained. For example, if we let

SNR = 18dB (signal-to-noise ratio)
G_T = 0dB (aircraft antenna gain)
G_R = 21dB (satellite antenna gain; providing 11° coverage)
τ_{ch} = 0.1μsec. (compressed pulse width)

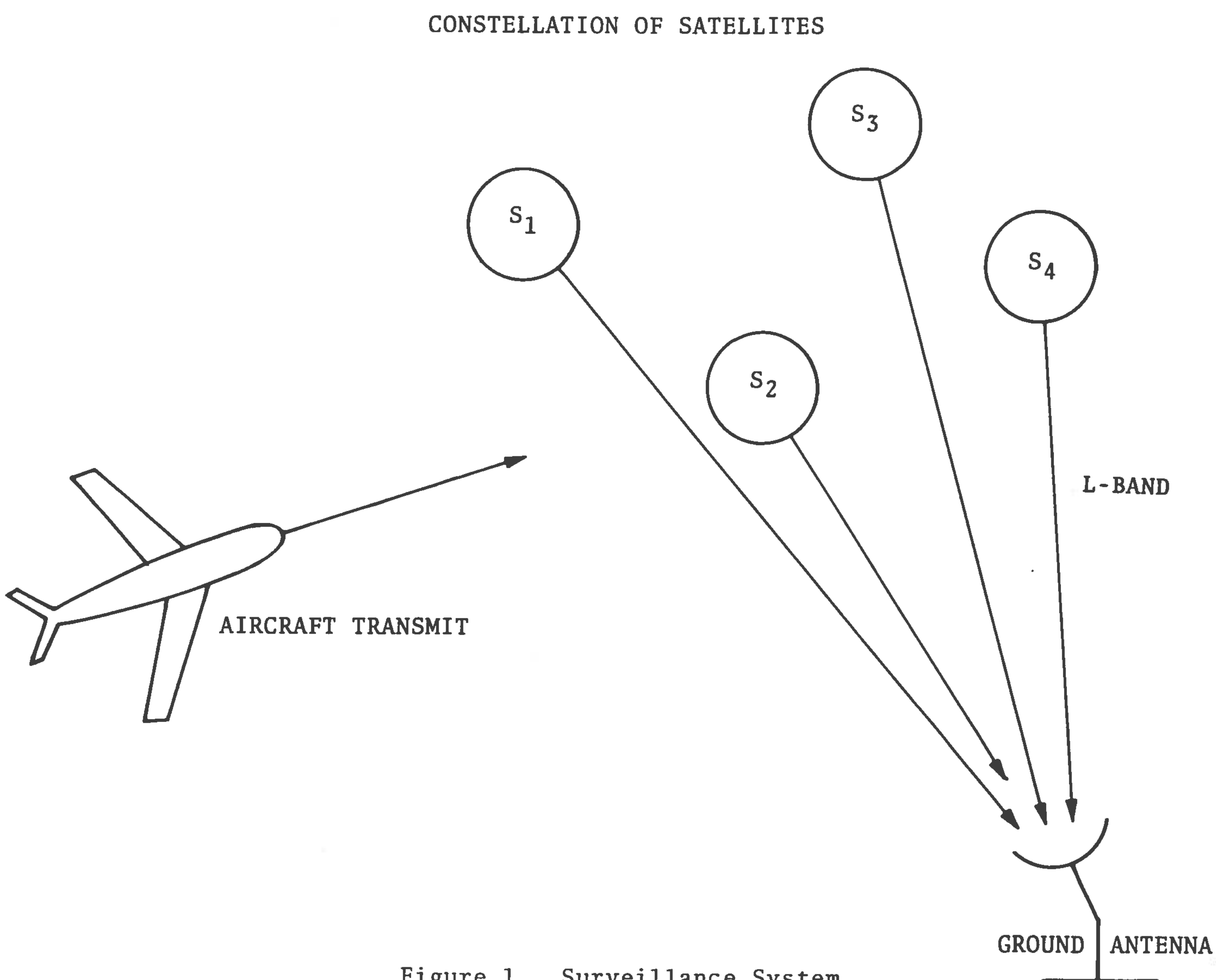

Figure 1. Surveillance System

the following aircraft peak power P_T is required as a function of the number of code elements L.

L	1000	500	250	125
P_T	1KW	2KW	4KW	8KW

The average power requirements are low.

In an advanced air traffic system, command and control information must be provided through a back-link to the aircraft. There is a great burden on the communication satellite to provide the required flow of information to a large number of operating aircrafts. The information rate will depend on the message length and the up-date rate. In study paper B, the requirements relating message length, up-date rate, and satellite coverage power for CONUS were determined. It has been shown that in order to hold the average satellite power to less than 2KW the following bounds must be generally satisfied

$$\frac{\alpha}{T} \leq 1$$

where α is the message length per aircraft and T is the up-date rate. The satellite power is directly proportional to the message length and inversely proportional to the up-date rate T. Both of these quantities must be determined from operational considerations. However, as happens in many cases, the above bounds may influence to a considerable extent the operational characteristic of the system.

2.1.2 Multiple Access Noise

A major limitation in LIT-type systems is multiple access noise. Multiple access noise or cross-talk is created by the signals transmitted by all aircrafts. In the receiver all waveforms not matched to its filter seriously impair the identification of aircrafts whose waveforms are matched to other filter channels. The multiple access noise is a function of several surveillance parameters. To hold the multiple access noise within a tolerable limit, these surveillance parameters must be bounded. In study paper C the impact of multiple access noise on the choice of these parameters is discussed and criteria established. In the presence of multiple access noise the matched filter normalized output signal-to-noise ratio $(SNR)_o$ is given by

$$\frac{(SNR)_o}{L(SNR)_i} = \frac{1}{1+M(SNR)_i}$$

where $(SNR)_i$ is the input signal-to-noise ratio and M is the multiple access noise parameter. The multiple access noise parameter M is defined as

$$M = \left(\frac{MNa}{T}\right) L\tau_{ch}$$

where Na is the number of operating aircraft, m is the number of pulses per transmission, T is the update rate, L is the code length and τ_{ch} the pulse width. As M increases, the signal-to-noise ratio decreases and system reliability deteriorates. In the study paper two parametric curves are drawn, Figures C-1 and C-2, from which suitable boundaries of the surveillance parameters can be established and the impact of proposed waveforms on the multiple access noise and system performance evaluated.

2.1.3 Waveforms and Codes

A number of waveforms with different modulation parameters are being proposed for the surveillance function. In study paper D the signal structures and parameters proposed by TRW, RCA, Autonetics and Boeing are summarized. The identity of an aircraft is contained within the transmitted waveforms. Unique identification of each operating aircraft presents a key and formidable problem. The proposed waveforms utilize pseudo-random codes to aid in that identification. Up to 1,000,000 separate identities may be required in the 1990's for the operating system. The total number of distinct identities (ID) is obtained from

$$ID = nFPRI$$

where n is the number of available unique pseudo-random codes, F is the number of independent frequency channels and PRI is the number of assigned pulse repetition intervals. To provide distinguishability and reliability a large number of pseudo random codes is desired. To be useful, the codes must have both "good" auto-correlation and cross-correlation properties. In study paper B, the availability and characteristics of such codes is also discussed. The availability of suitable pseudo-random codes that are being proposed is limited.

The LIT-type surveillance system presents a coding problem involving both aircraft detection and identification. It requires good radar range codes which at the same time have good distinguishability, one relative to another. This capability, as noted above, is expressed in terms of the "aperiodic" auto- and cross-correlation functions. A fairly rigorous statement of this coding problem, including basic definitions, is contained in study paper E.

As noted, PN codes are being proposed for the application we have in mind. However, the aperiodic auto-correlation and certainly the cross-correlation properties, important for these applications, are not well known. The aperiodic functions cannot be determined analytically and hence brute force computation must be used. Computer results of aperiodic correlations for some of these PN sequences are given in study paper F. The side peak auto-correlation and peak cross-correlation values are tabulated. The results show rather high peak cross-correlation values, typically about 10 to 16dB below the main peak, for PN sequences around 200 to 300 bits long. This is true, for example, for twn 200-bit sequences that might be used in the proposed Autonetics surveillance system. Unless better codes can be found than these, system performance might be in jeopardy from a too high false alarm rate and too low probability of detection.

The codes evaluated in study paper F are binary codes, which employ two phase angles of the carrier signal. It has been proposed that by increasing the number of phase angles to m, $m>2$, we might be able to improve code performance considerably. These m-phase, or polyphase, codes have been studied somewhat to date, for example by Frank and Golomb and Scholtz, but again in the radar range context where one sequence at a time is used. Although the performance obtained by Frank for a single sequence ($\sqrt{n}$ phase angles are needed for a sequence n digits long) is impressive, the subject has not been explored enough yet to say how much better a set of polyphase codes are in terms of cross-correlation. The mathematics involved in this approach and the correlation results of some sets of 6-phase codes are presented in study paper G. The rationale for pursuing this technique is also discussed in that study paper.

Doppler Effects:

The length of a code is constrained by doppler effects. In study paper H these constraint criteria are given. In Figure H-1 a parametric curve is drawn from which the performance criteria and code length can be established. As an example, it is shown that if

$\Delta f = 4 \times 10^4 Hz$ (frequency off-set)

$\tau_{ch} = 0.1 \mu sec.$ (compressed pulse width)

$G = 0.9$ (10% degradation in filter processing gain)

then the code length L must be restricted to $L \leq 400$ code elements.

2.1.4 System Reliability Criteria

The reliability performance criteria of a surveillance system is defined in terms of the probability of false alarm P(FA) and the probability of detection P(D). The threshold of the receiver is set for a low P(FA) and the desired P(D) determines the signal energy requirements. In study paper I these criteria are defined for a single pulse and one receiver channel. Furthermore, the effect of sidelobes generated, for instance, by cross-correlation is evaluated. These sidelobes can appear as true targets, thus generating false alarms. It is shown that with the codes so far available and the expected number of operating aircrafts, the sidelobes would generate a high alarm rate. For example, a sidelobe level of 13dB below the main peak signal value would generate about 70 false alarms per channel.

In study paper J, system reliability performance of multi-pulse waveforms is expressed in terms of pertinent surveillance parameters. With these expressions, the system reliability in terms of the expected false alarm rate can be evaluated. Conversely, by requiring a given reliability performance, the pertinent surveillance parameters can be bounded and optimized. A simple analytical approach is used in deriving these expressions. In all probability, better models and simulations will be required to obtain a high degree of confidence in system reliability performance.

2.2 DATA PROCESSING

Study papers K, L, and M are analyses on the effects of false alarms due to multi-access noise and jamming on data processing, Individual Autonetics and RCA proposed systems are used as models.

Paper study N is a comparison of the data processing requirements of 3 baseline systems: I - Boeing Satellite System, II - Autonetics Hybrid System, and III an Upgraded Third Generation ATC System.

A brief description of each of these study papers is given below.

K. The Effect of False Alarms on Data Processing in The Autonetics 3-Pulse System calculates the false alarm rate and the target dismissal rate as the data enters the position determining the computer, associative memories and digital detectors act as preprocessors to this computer. Not taken into account is the possibility that false data may produce non-realistic positions, and that tracking history could be used to validate targets and cover for missing targets.

The conclusion reached is that the computer is not overburdened and at the same time, not too many actual targets are lost.

It is assumed in this analysis that the receiver signal-to-thermal noise ratio is 16dB, and that the interference caused by multi-access noise is Gaussian distributed. In this manner thermal noise and multi-access noise are added to produce a signal to total noise ratio, 14.35dB. Then from Rice's curves, for a given probability of noise, a probability of signal detection is selected. With these probabilities the false alarm and target dismissal rates are determined. This approach was originated by RCA in their analysis of the SATANS system; it was also used by Autonetics in their 4GATCS Concept Study.

L. The Effect of False Alarms on Data Processing Using a Modified Autonetics 2-Pulse System produces similar results using essentially the same approach as in K.

M. Jamming The SATANS System and Other Satellite Systems and The Effect on Data Processing analyzes what happens to the false alarm rate and the target dismissal rate when a low budget jammer is in action.

N. A comparison of Data Processing Requirements in Future ATC Systems relates MIPS rates for the 3 baseline systems: I - Boeing Satellite, II - Autonetics Hybrid System, III an Upgraded Third Generation ATC System.

REFERENCES

1. Autonetics, "Fourth Generation Air Traffic Control Study", Nov., 1971, prepared under Contract DOT-TSC-304.

2. Boeing, "Study and Concept Formulation of a Fourth Generation Air Traffic Control Study", Nov., 1971, prepared under Contract DOT-TSC-145 and -306.

TECHNICAL STUDY PAPERS

STUDY PAPER A

SYSTEM REQUIREMENTS AND DESIGN RELATING TO AIRCRAFT POWER, PULSE WIDTH, AND CODE LENGTH FOR AN AATMS SURVEILLANCE

B.S. GOLDSTEIN

In the proposed satellite aided 4th GATS surveillance, the required aircraft peak power is of critical importance. With conventional transmitters, the required peak power would be much too high. It is therefore desirable to increase transmitter duty cycle without sacrificing the resolution and accuracy of the surveillance system. This can be accomplished by increasing the number of code elements in the transmitted pulse. In this study paper, the system design parameters relating power, pulse width, and code length are determined.

The aircraft to satellite surveillance link requirements are determined from

$$\left(\frac{P_i}{N}\right) = \left(\frac{S}{N}\right) = \frac{P_T \tau G_T G_R \lambda^2}{(4\pi R)^2 \left(RT_e\right) \ell_s} = \frac{P_T L \tau_{chip} G_T G_R \lambda^2}{(4\pi)^2 R^2 \left(RT_e\right) \ell_s} \qquad (1)$$

where

P_T = aircraft transmitted peak power

G_T = aircraft antenna gain

G_R = satellite antenna gain

R = aircraft to satellite range

K = Boltzman's constant

T_e = equivalent noise temperature

P_i = received signal power

N = receiver noise power

ℓ_s = system losses

λ = operational wavelength

τ = transmitted pulse width

τ_{chip} = compressed pulse width

L = number of code elements per transmitted pulse

It is assumed that the satellite to ground link has ample signal margin because of the availability of large ground antenna.

Some typical aircraft to satellite link parameters are postulated as follows.

$\lambda^2/(4\pi)^2R^2$ (space loss) = -188dBw

ℓ_s (system losses) = 6dB

kT_e = -201.0dBw/Hz (3dB noise figure)

Using these basic parameters, Equation 1 can be expressed as

$$(SNR)_{dB} = P_T L \tau_{chip} + (G_T G_R)_{dB} + 7dB \tag{2}$$

In Figure A-1, the SNR is plotted versus $P_T L\tau chip$ with $G_T G_R$ as a parameter. With the aid of Figure A-1, the surveillance design parameters can be determined. Now suppose that

1. The combined gain $G_T G_R$ = 21dB. This provides an 11° coverage for CONUS and 0dB aircraft antenna.
2. To provide sufficient system reliability, the required signal to noise ratio is taken to be 18dB.

From Figure A-1, it can be seen that to meet these requirements the parameter $P_T L \tau_{chip}$ must be bounded so that

$$P_T L \tau_{chip} \leq 10^{-1} \tag{3}$$

The value of the compressed pulse width $\tau chip$ is determined from the surveillance accuracy requirements. However, if we let τ_{chip} = 0.1μsec., the following table can be constructed.

L = no. of code elements	1000	500	250	125
P_T = aircraft peak power	1KW	2KW	4KW	8KW

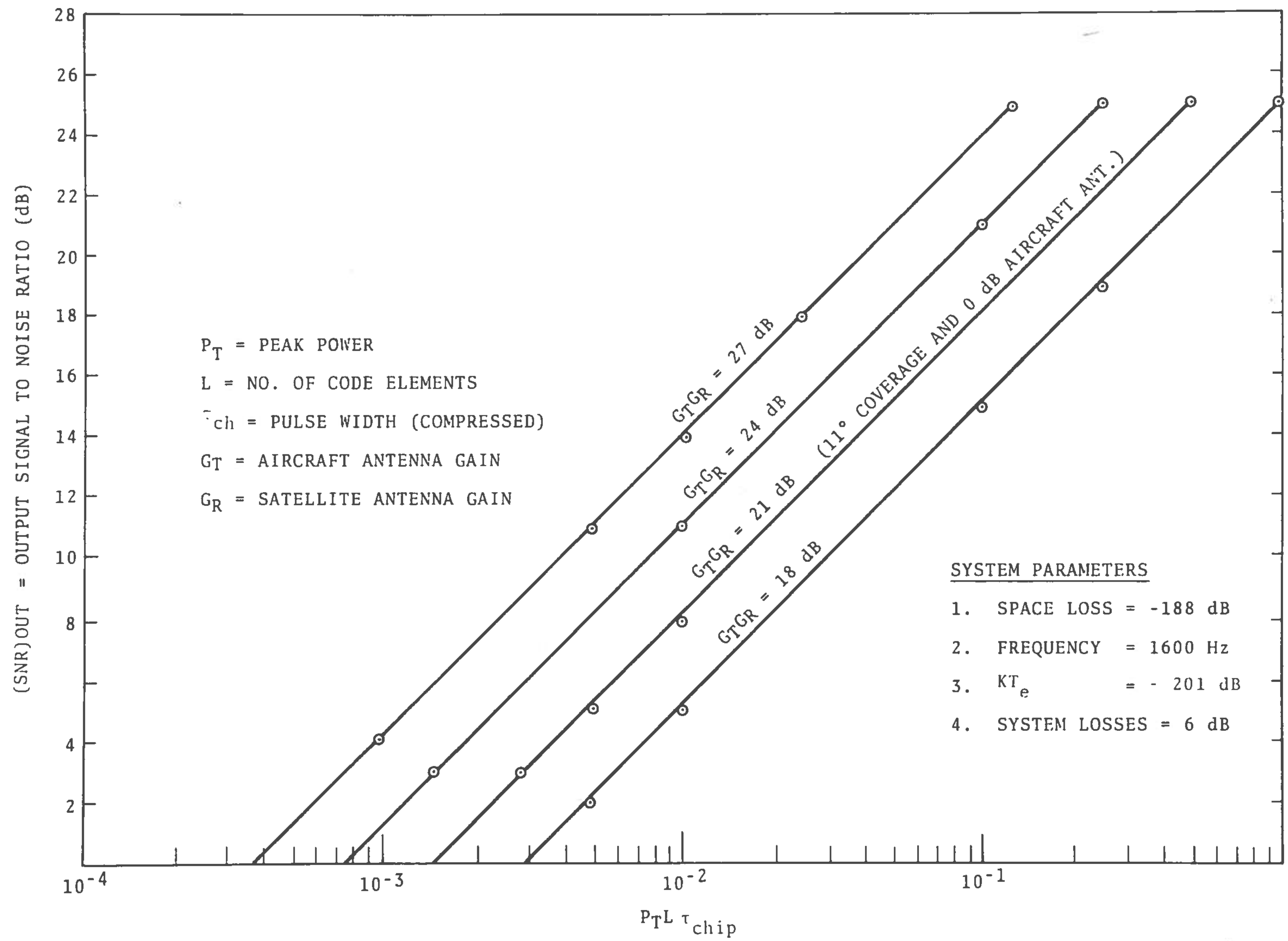

Figure A-1. Output Signal To Noise Ratio versus $P_T L \tau_{chip}$

To reduce the peak power, a large number of code elements is desired. The number of code elements, however, is limited by other constraints such, for instance, as discussed in a separate study paper C "Impact of Multiple Access Noise on the Choice of Surveillance Parameters in 4th AATMS". A code length between 200 and 500 elements is presently anticipated.

STUDY PAPER B

BACK LINK COMMUNICATION REQUIREMENTS AS A FUNCTION OF MESSAGE LENGTH AND UP-DATE RATE FOR A SATELLITE AIDED AATMS

B.S. GOLDSTEIN

In a satellite type aided surveillance system, it will also be necessary to transmit positional and command and control information back to the aircraft. Because this information must be provided to perhaps 100,000 aircraft, the question arises about the feasibility of implementing a satellite back-link to provide the necessary communications to the aircraft. In this study paper, the requirements relating to satellite power, message length, and up-date rates are determined.

SYSTEM REQUIREMENTS

The satellite power requirements can be determined with the aid of the range equation given by

$$(SNR) = \frac{Pev G_R G_T \lambda^2}{(4\pi)^2 R^2 K T_e B \ell_s} \qquad (1)$$

where Pev is average required satellite power, G_R is the aircraft antenna gain, G_T is the satellite antenna gain, λ is the operating wavelength, R is the range, K is Boltzman's constant, Te is the effective noise temperature, B is the bandwidth, ℓ_s are system losses and SNR the signal to noise ratio.

Assuming,

f	=	1600 MHz(operating frequency)
G_R	=	0dB (aircraft antenna)
G_T	=	21dB (Satellite antenna 11° degree coverage)
KT_e	=	-201dB/Hz (3dB noise figure)

$\lambda^2/(4\pi R)^2$	=	-188dB (space loss)
SNR	=	10dB (output signal to noise ratio)
ℓ_s	=	-6dB (system losses)
B	=	Bandwidth (to be computed)

With the above parameters, Equation 1 can be expressed as:

$$\text{Pev (dB)} = \text{B (dB)} - 18 \text{ dB} \tag{2}$$

BANDWIDTH EVALUATION

To evaluate B, we compute the digit rate R. The digit rate R is given by

$$R = \frac{(Na)\alpha}{T} \tag{3}$$

where

Na	=	number of aircrafts
α	=	no. of bits/message/aircraft
T	=	up-date rate

The bandwidth will depend on the choice of modulation. However, if we assume a phase-shift keyed system, we can postulate that at best

$$B \approx R \approx \frac{(Na)\alpha}{T} \tag{4}$$

Using Equation 4, we can rewrite Equation 2 as

$$\text{Pev} \approx \frac{(Na)\alpha}{63T} \tag{5}$$

We now relate the power requirements to the number of aircrafts, message length and up-date rate.

MESSAGE LENGTH

The message type and length still need to be determined. However, it is clear that as a minimum, the back link information must contain the aircraft's address and positional information. To address an aircraft and provide positional information the following message may be required:

Address	30 bits
Synch	11 bits
Latitude (100 ft.)	21 bits
Longitude	21 bits
Heading	10 bits
Velocity	6 bits
	99 bits

In addition, command and control as well as advisory information may be required. Tentatively, it appears that such information could be provided with 10 alpha-numeric characters using the American Standard Code for Information Exchange. In this code, each character is represented by a 7 bit code plus parity. Hence, a total of 80 bits would be required for this function. Thus, a 100 to 200 bit message per aircraft is not unreasonable.

SYSTEM PERFORMANCE PARAMETERS

Equation 5 is plotted in Figure B-1, as Pev vs. the parameter $\frac{(Na)\alpha}{T}$ from which the back-link requirement parameters can be readily assessed. Thus, if $\frac{(Na)\alpha}{T} = 10^6$, about 16 KW of coverage satellite power is required if a single communication satellite is utilized. On the other hand, if we project

$$Na = \text{no. of aircraft} = 10^5$$

$$Pev = \text{average power} = 1.6\text{KW}$$

then we must place a restriction on system parameters such that

$$\frac{\alpha}{T} \leq 1 \qquad (6)$$

where again α is the number of bits/message and T is the up-date rate. The up-date rate must be determined from operational consideration. However, if, for example, Pev is restricted to 2KW and α = 100 bits/message than T must be $\geq$ 100 sec. In the table below, the approximate power requirements as a function of α and T are tabulated.

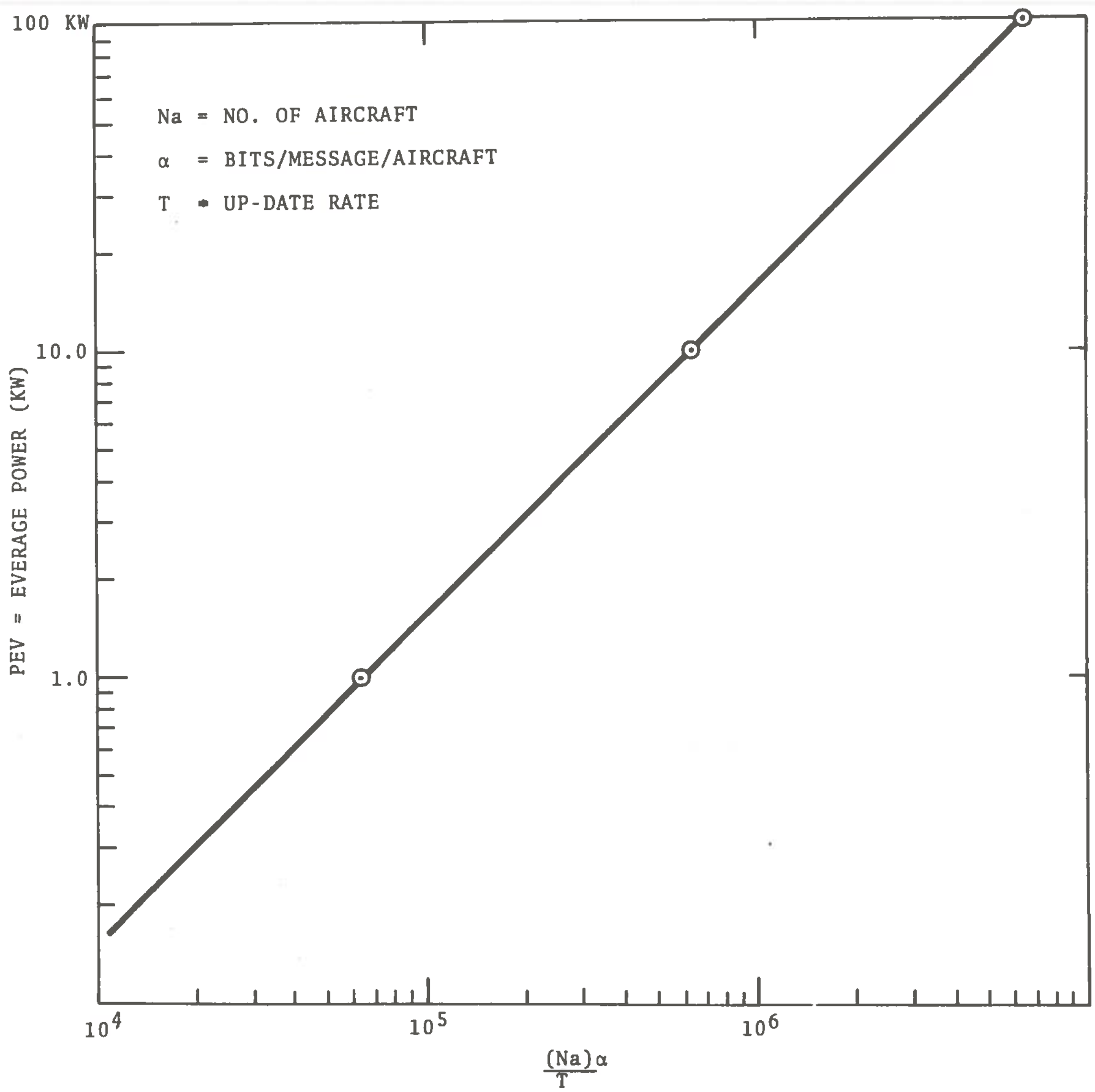

Figure B-1. Average Power Requirement for a Back-Link Communication Satellite as a Function of Parameter $\frac{(Na)\alpha}{T}$

T \ α	200	100	50
200 sec.	1.6 KW	0.8 KW	0.4 W
100 sec.	3.2 KW	1.6 KW	0.8 KW
50 sec.	6.4 KW	3.2 KW	1.6 KW
10 sec.	32 KW	16 KW	8 KW

The power requirements could perhaps be reduced somewhat with sophisticated coding; the aircraft antenna could be designed with a 2 or 3 dB gain. However, if reliable operation is to be provided, potential improvement of more than 3 dB should not be anticipated.

STUDY PAPER C

THE IMPACT OF MULTIPLE ACCESS NOISE ON THE CHOICE OF SURVEILLANCE PARAMETERS IN THE AATMS

B.S. GOLDSTEIN

In the AATMS proposed by Boeing and Autonetics, aircraft transmit phase coded signals via a constellation of satellites to a central ground station from which the position of the aircraft is computed. Each aircraft transmits its own identity, which the receiver attempts to identify uniquely. This identification is seriously impaired by multiple access noise or cross-talk due to the signals transmitted by all the other aircrafts. This cross-talk may reduce the signal to noise ratio to a point where the false alarm rate is too large and the probability of detection or proper aircraft identification too small for reliable system operation. In the attached appendix, the resulting signal to noise ratio as a function of the multiple access noise parameter is given. The multiple access noise parameter M is a function of several key surveillance parameters and is defined as

$$M = \left(\frac{mNa}{T}\right)(L\tau_{ch}) \tag{1}$$

where

Na = no. of aircraft

m = no. of pulses per transmission

L = no. of code elements per pulse

T = up-date rate

τ_{ch} = width of a single pulse chip

In Figure C-1 is shown how the normalized output signal to noise ratio deteriorates as a function of the multiple access noise parameter M. The signal to noise ratio satisfies the relationship

$$(SNR)_o = \frac{L(SNR)_i}{1 + M(SNR)_i} \tag{2}$$

where $(SNR)_o$ is the desired signal to noise ratio at the output of the matched filter, L is the number of elements in a code and $(SNR)_i$ is the receiver input signal to noise ratio. As M increases the

$(SNR)_o$ or system reliability deteriorates. In Figure C-2 is drawn the multiple access noise parameter M versus $L\tau_{ch}$ with (mNa)/T as a parameter. With the aid of these two graphs, it is possible to:

1. Evaluate performance as a function of surveillance parameter such as, for instance, proposed by TRW, Boeing, and Autonetics.
2. Establish bound on surveillance parameters for an acceptable signal to noise ratio and hence reliable operation.

EXAMPLE

CONSIDER THE WAVEFORM PARAMETERS PROPOSED BY BOEING

1. L = no. of code elements = 511
2. τ_{ch} = width of a pulse chip = 0.5μsec.
3. T = update rate = 1 sec.
4. m = no. of pulses per transmission = 1
5. Na = postulated no. of aircraft = 10^5

From these parameters, the multiple access noise parameter M ≈ 25. To use Figure C-1, consider that the output signal to noise ratio from the matched filter $(SNR)_o = L(SNR)_i$, where $(SNR)_i$ is the input signal to noise ratio. If, for an acceptable performance, the $(SNR)_o$ must be 17 dB ≈ 50, then for L = 500, the $(SNR)_i \approx 0.1$. From the graph, it can be seen that for $(SNR)_i = 0.1$, and M = 25, the

$$\frac{(SNR)_o}{L\ (SNR)_i} \approx 0.25$$

implying a deterioration by 6 dB of the output signal to noise ratio and in all likelihood unacceptable system performance. Thus, redesign of surveillance parameters is in order.

REDESIGN OF SURVEILLANCE

Now, suppose that in the above example, we restrict the degradation in the output signal to noise ratio, due to multiple access noise, to (0.9) $(S\ N\ R)_o$ or 0.9 of its full value. Also let the design values be

$Na = 10^5$

$L = 500$

$(SNR)_i = 0.1$

Then, from Figure C-1 and Equation 1, the multiple access noise parameter M must be bounded by $M \leq 0.8$ or

$$M = \left(\frac{mNa}{T}\right) (L\tau_{ch}) = \frac{5 \times 10^7 \tau_{ch}}{T} \leq 0.8$$

This implies, that if T = 1, then the pulse width τ_{ch} must equal 16×10^{-8} sec or 16 nanoseconds to satisfy the above requirements. This is unrealistic in a practical system. On the other hand, if we increase τ_{ch} to 100 nsec., the up-date rate T must be reduced to about 6 seconds. This update rate may not satisfy the system operational requirements.

For the waveform proposed by Boeing with τ_{ch} = 0.5 usec., the update rate would have to be reduced to 30 sec. to satify the postulated requirements. Any other waveform can be similarly evaluated.

In this study paper we discussed multiple access noise as it affects system performance. The multiple access noise is a function of the number of aircraft, number of pulses per transmission, pulse length, code length and up-date rate. To hold the multiple access noise within a tolerable limit, the above surveillance parameters must also be bounded as illustrated in the above example. This is readily accomplished with the aid of the drawn graphs of Figure C-1 and C-2.

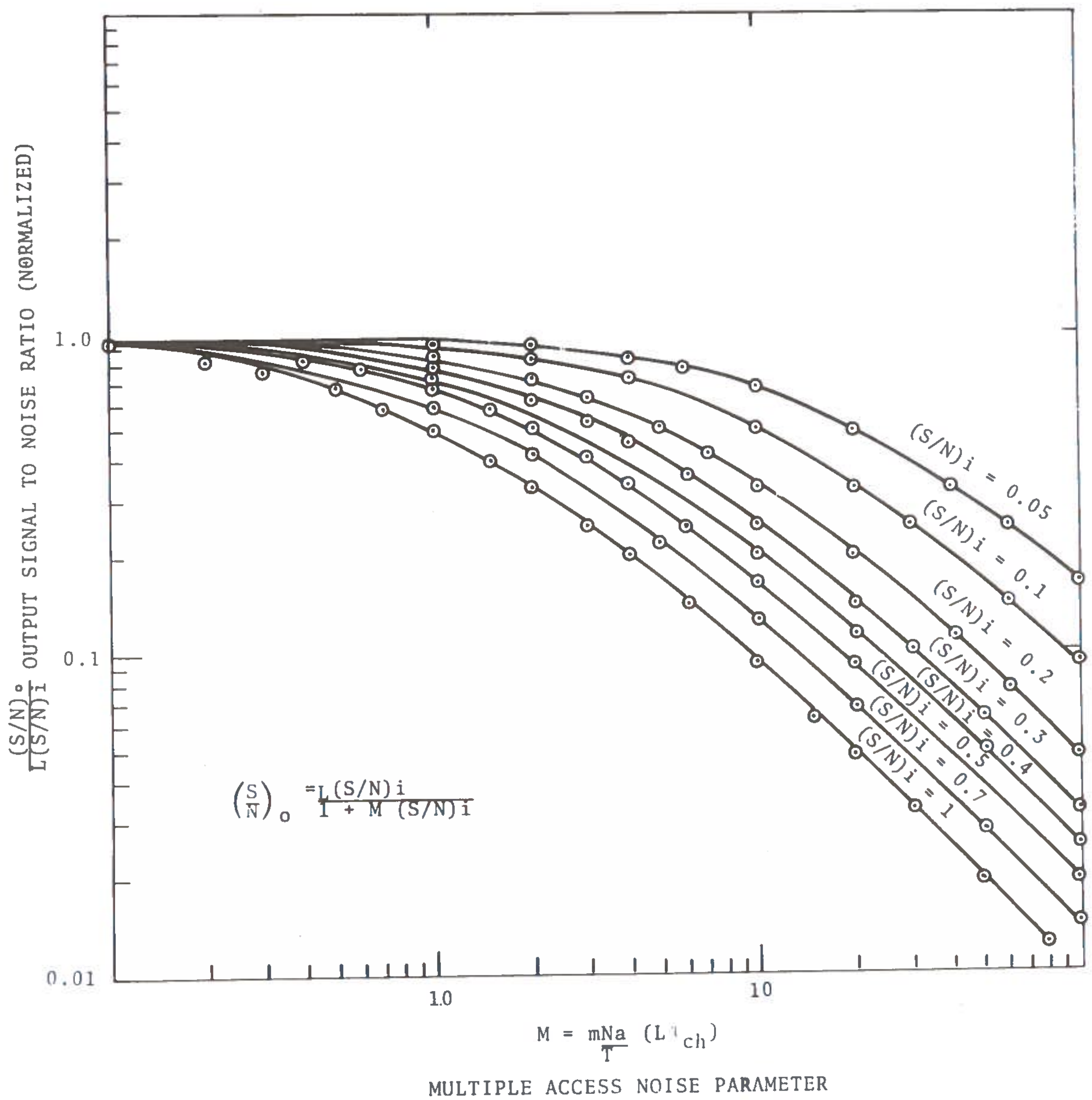

Figure C-1 Output Signal To Noise Ratio vs. The Multiple Access Noise Parameter

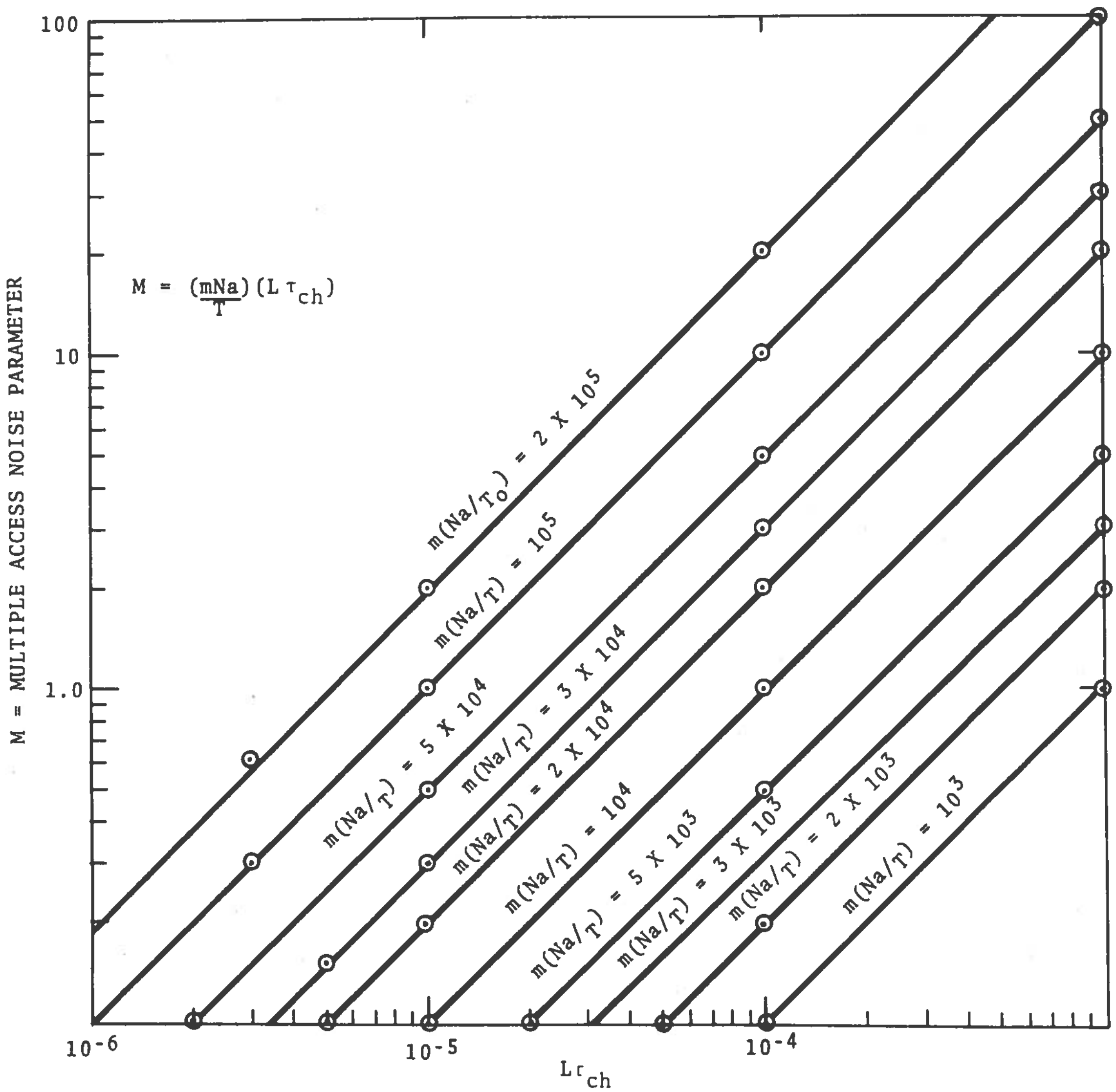

Figure C-2 Multiple Access Noise Parameter M vs. $L\tau_{ch}$

APPENDIX A

MULTIPLE ACCESS NOISE IN 4GATS SYSTEM

In the proposed 4GATS Multiple access noise may limit the available signal to noise ratio and hence system performance. The multiple access noise is a function of several system parameters such as a number of transmitting aircraft, up-date rate, pulse width, code length and input signal to noise ratio. Since these parameters are key ingredients of the system, it is important to examine how they affect the multiple access noise and hence system operation. By adjusting some of the parameters, the multiple access noise can be minimized.

In the surveillance system under consideration, the signal to noise ratio at the output of the matched filter is given by

$$(S/N)_{out} = \frac{S}{N+N_v} = \frac{\text{signal}}{\text{thermal noise + cross-talk noise}} \quad (1)$$

The output signal is given by LS where S is the signal power at the input to the matched filter. L is the number of chips in the phase coded word where a chip, τ_{ch}, corresponds to the basic compressed pulse width. The transmitted pulse width τ is related to L and τ_{chip} by

$$\tau = L\tau_{ch} \quad (2)$$

The cross-talk or multiple access noise due to coded signals which are not matched to a particular filter appears as random noise and is given by

$$N_{\mu} = mNa \frac{(L\tau_{ch})}{T} S \quad (3)$$

where

Na = no. of aircraft transmitting different codes

S = signal

T = up-date rate

m = no. of pulses per transmission

Equation 1 can therefore be written as

$$(S/N)_{out} = \frac{L\ (S/N)_{in}}{1 + \left(\frac{mNa}{T}\right) (L\tau_{ch})\ (S/N)_{in}} \qquad (4)$$

The two cases of small and large effects of multiple access noise deserve special consideration.

CASE A: Small Nu

The effects of multiple access noise will be small if

$$\left(\frac{mNa}{T}\right) (L\tau_{ch})\ (S/N)_i << 1 \qquad (5)$$

In this case, the signal to noise ratio at the output of the matched filter is $(S/N)_{out} = L\ (S/N)_{in}$, as it should. If a 10% loss due to multiple access noise is to be tolerated, then the above parameters must be adjusted so that

$$\left(\frac{mNa}{T}\right) (L\tau_{ch})\ (S/N)_{in} \leq 0.1 \qquad (6)$$

<u>CASE B: Large Nu</u>

The multiple access noise will predominate

$$\left(\frac{mNa}{T}\right) (L\tau_{ch})\ (S/N)_i >> 1 \qquad (7)$$

so that as the input signal to noise ratio becomes large, the output signal to noise ratio in the limit becomes

$$(S/N)_{out} = \frac{L}{\left(\frac{mNa}{T}\right) (L\tau_{ch})} = \frac{T}{mNa\ \tau_{ch}} \qquad (8)$$

which is independent of L. This particular case may be applicable in a ground based system where ample signal power may be available. Suppose now that

$Na = 10^5$

$T = 1$ sec.

$\tau_{ch} = 0.1$ usec.

$m = 1$

Under these conditions, even in the strong signal case the maximum available signal to noise ratio is only on the order of 20dB.

In Figure C-1, a normalized output signal to noise ratio is drawn as a function of the multiple access noise with the input signal to noise ratio as a parameter. In Figure C-2, the multiple access noise is drawn as a function of $L\tau_{ch}$ with mNa/T as a parameter. With the aid of these two graphs, the parameters can be readily adjusted to meet the signal to noise ratio requirements.

Unequal Signals

In the above considerations, all signals were assumed to be of equal strength. If an aircraft signal is weaker than the other signals transmitted from all other aircrafts, the multiple access noise will further degrade signal detectability. If we denote the weaker signal as S', then the output signal to noise ratio is from Equation 4

$$(S/N)_{out} = \frac{L(S'/N)_{in}}{1 + \frac{mNa}{T} (L\tau_{ch}) (S/N)_{in}} \qquad (9)$$

In the strong multiple access noise case this becomes

$$(S/N)_{out} = \left(\frac{T}{mNa\tau_{ch}}\right) \frac{(S'/N)_{in}}{(S/N)_{in}} \qquad (10)$$

where S is the signal strength averaged over all aircrafts. The system will obviously be designed for the weakest expected case. For example, if the average signal strength is 3 dB higher that the weakest signal, then the output signal to noise ratio degrades by 3dB.

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STUDY PAPER D

WAVEFORMS, MODULATION PARAMETERS, AND PSEUDO-RANDOM CODES FOR SURVEILLANCE APPLICATIONS IN THE AATMS

B.S. GOLDSTEIN, E.H. FARR

WAVEFORMS AND MODULATION PARAMETERS

In the proposed surveillance system for an AATMS, coded waveforms are transmitted asynchronously by aircraft to a central ground station, via satellites. By measuring the difference in the time of arrival of the signals from at least 4 satellites, the aircraft position is determined. Figure D-1 illustrates such a satellite aided surveillance system.

The surveillance system under consideration was initially suggested by R. L. Garwin of IBM and subsequently extended and examined more closely by TRW in their Location-Identification-Transmission (LIT) system. Table D-1 and Figure D-2 show typical LIT modulation parameters and the respective waveform. Table D-1 also gives the modified parameters proposed by Boeing. Table D-2 gives the modulation parameters proposed by RCA and Autonetics and the respective waveforms are shown in Figures D-3 and D-4.

A key and formidable problem in this type of surveillance system is the identification of perhaps up to 100,000 transmitting aircrafts. The number of total required identities ID, corresponding to the expected aircraft population, is given by

$$ID = C \times PRI \times F$$

where

C = no. of available codes

PRI = no. of available pulse repetition intervals

F = no. of available channel frequencies

From Tables 1 and 2, it can be seen that

1. TRW proposes 16 codes and 31255 PRI for a total of 500,000 available identities

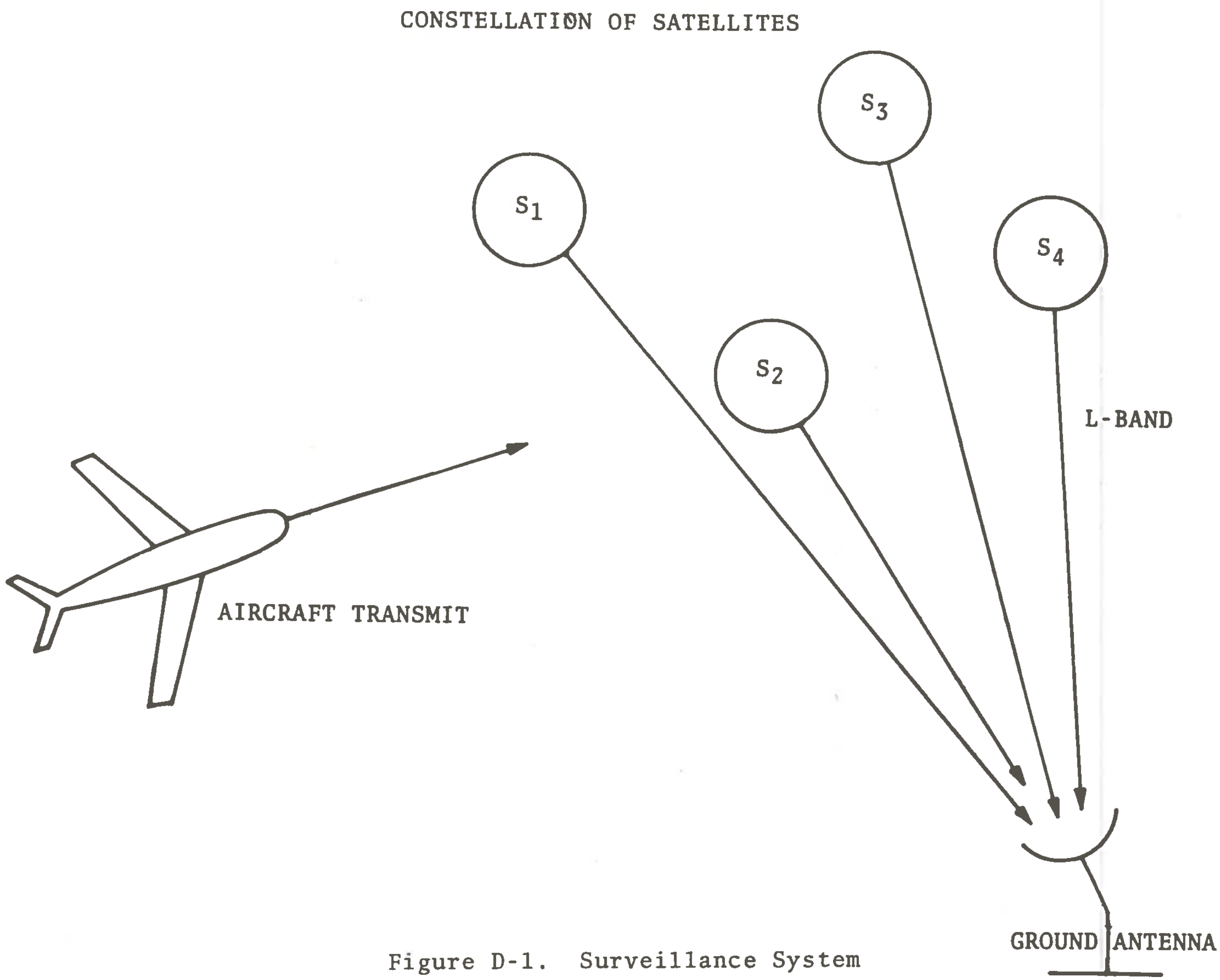


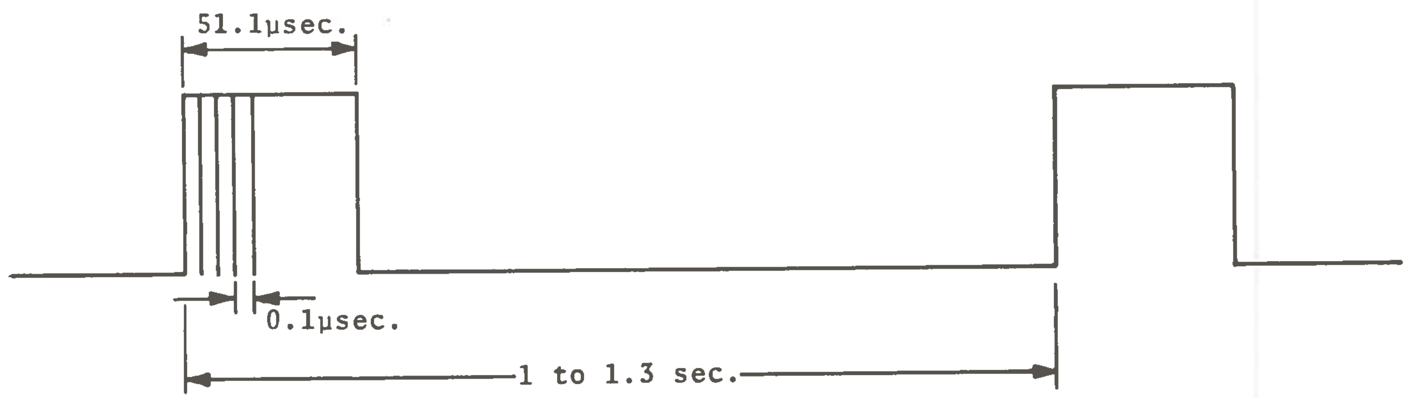
Figure D-1. Surveillance System

TABLE D-1. LIT MODULATION PARAMETERS

PARAMETER	TRW(LIT)	BOEING(LIT)
Waveform	PN code	PN code
Pulse length	51.1 usec.	255 usec.
No. of chips	511	511
Chip width	0.1 usec.	0.5 usec.
No. of codes	16	16
Pulse repetition period	1.0 to 1.3 sec.	1.0 to 1.3 sec.
PRP spacing	10 usec.	50 usec.
No. frequency channels	1	6
Channel bandwidth	20 MHz	4 MHz
No. of users	50,000	60,000
No. of users/channel	50,000	10,000
Sorting positions	3125 out of 31255	625 out of 6251

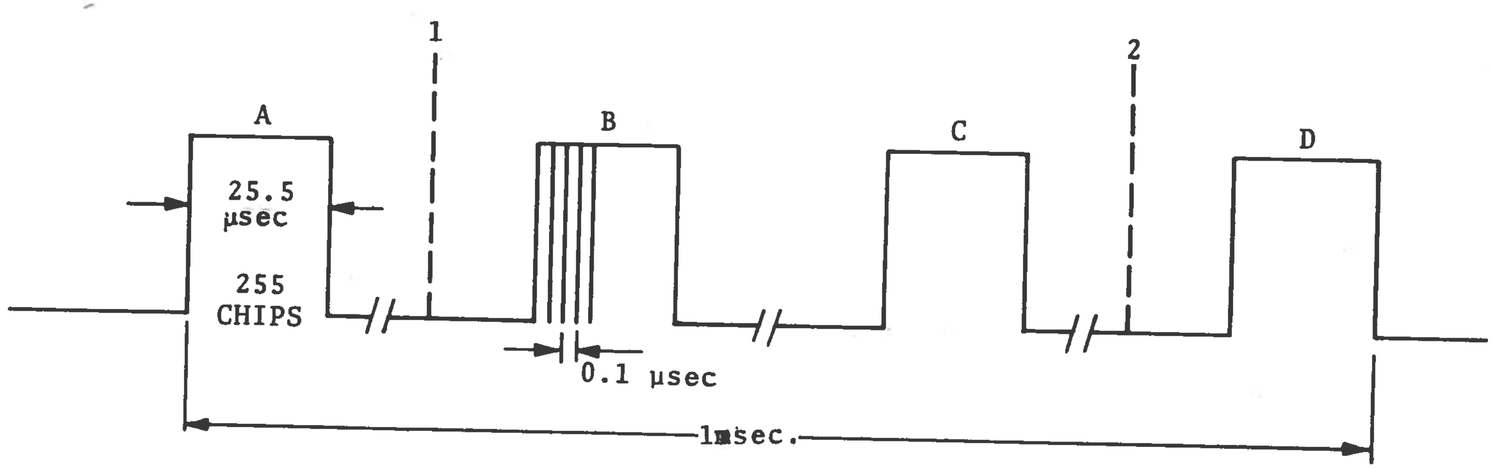
TABLE D-2. RCA AND AUTONETICS MODULATION PARAMETERS

PARAMETER	SATANC (RCA)	INTEGRATED-SATELLITES (AUTON.)
Waveform	PN code	PN code
Pulse length	25.5 usec.	20 usec.
No. of chips	255	200
Chip width	0.1 usec.	0.1 usec.
No. of pulses	4	3
No. of PN codes	8	10
Length of transmission	1 msec. (4 pulses)	100 usec. (3 pulses)
No. of frequency channels	1	10
Pulse repetition period	2.5	2.5 sec.
Pulse spacings	A to B; 512 intervals 1.0 usec. apart C to D; 256 intervals, 1.0 usec. apart	A to B; 100 intervals, 0.1 usec. apart A to C; 100 intervals, 0.1 usec. apart.



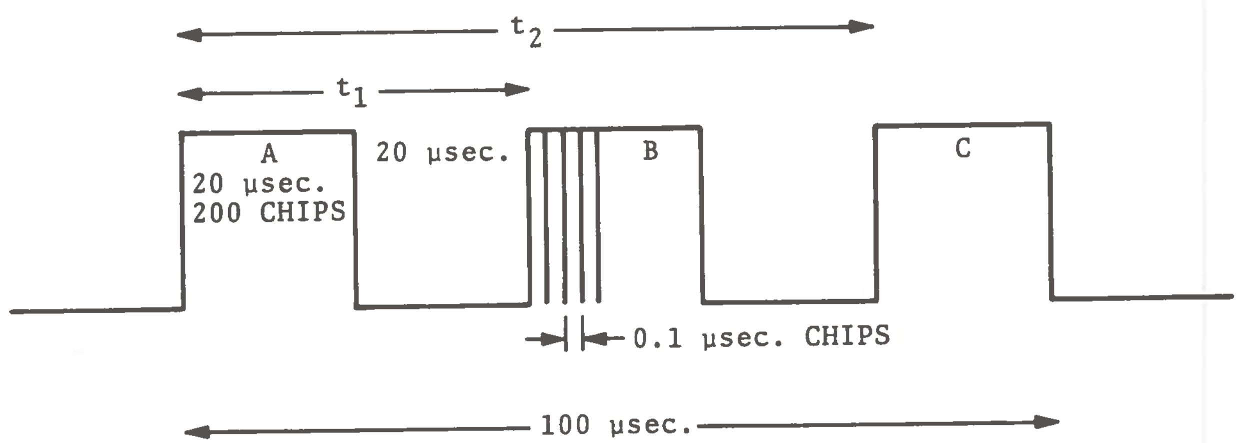
1. NUMBER OF CHIPS L = 511, EACH CHIP 0.1μsec.
2. PRI = 30,000, EACH 10μsec. APART
3. TOTAL NO. OF ADDRESSES = (30,000) X NO. OF PSEUDO-RANDOM CODES

Figure D-2. LIT Type Waveform



1. 4 PULSES PER TRANSMISSION (A,B,C,D)
2. 255 CHIPS PER PULSE - EACH CHIP 0.1μsec WIDE
3. POSITION BETWEEN A & B VARIED ABOUT CENTERLINE 1
   POSITION BETWEEN C & D VARIED ABOUT CENTERLINE 2

Figure D-3. RCA Four Pulse Waveform (Satan)



1. 3 PN PULSES PER TRANSMISSION (A,B,C)
2. EACH CODE 20 μsec. LONG
3. 200 CHIPS PER PULSE - EACH CHIP 0.1 μsec.
4. B HAS 100 POSITIONS WITH RESPECT TO A (0.1 μsec. INTERVALS) C ALSO HAS POSITIONS WITH RESPECT TO A (0.1 μsec. INTERVALS)
5. TOTAL ADDRESSES = 10^6 = (10 PN CODES) X (10^4 POSITIONS) X (10 FREQUENCIES)

Figure D-4. Autonetics 3 Pulse Waveform (Satellite System)

2. Boeing proposes 16 codes, 625 PRI and 6 channel frequencies to obtain about the same number of addresses.

3. RCA proposed 8 codes and 512 x 256 PRI to obtain about 10^6 identities

4. Autonetics proposed 10 codes, 10,000 PRI and 10 channel frequencies to obtain about 1,000,000 identities.

It may be worth mentioning that using F frequency channels, does not mean expanding the bandwidth by a factor F. In a tapped delay line a matched filter is used to compress the phase coded signals, the samples are independent at frequency spacings greater then $1/\Gamma$ where Γ is the length of the delay line. Thus, for 51.1 usec. long delay line, independent channels will be established at frequencies greater than 20KC. Channel spacings of 100KC might be appropriate. However, the overall performance of such operations needs to be investigated.

The codes provide unique distinguishability. The larger the number of such codes, the easier it is to identify an aircraft's identity. On the other hand, as the number of PRI becomes larger, it is more difficult to reliably sort out the identification of these aircrafts. For the proposed surveillance application, it is important that the set of pseudo-random codes have a good capability in the radar acquisition sense as well as being highly distinguishable, one relative to the other. Mathematically, this is equivalent to requiring that the codes have good aperiodic autocorrelation and cross correlation properties respectively. The availability as well as the characteristics of these codes are therefore of considerable importance.

PSEUDO-RANDOM CODES

Pseudo-random codes can be readily generated with shift registers and proper feedback connections. Maximum shift register sequences have the longest available period of length

$$L = 2^n - 1 \qquad (1)$$

where L is the length of the sequence, and n is the number of stages in the shift register.

Pseudo-random sequences have a number of interesting properties.[1,2] However, one of the more important properties of maximal shift register sequences is their two-level autocorrelation c(k) (periodic case) given by

$$c(k) = \sum_{n=1}^{L} a_n a_{n+k} = \begin{cases} L \text{ for } k=0, \pm nL \\ -1 \text{ otherwise} \end{cases} \quad (2)$$

where a_n = +1 or -1. This yields the ideal high peak value and low sidelobes, but for our application we deal with aperiodic or finite length sequences (meaning the sequence is preceded and followed by zeros) for which this property deteriorates, as discussed in the next section.

The number of available maximal length shift register sequences can be calculated with the aid of Euler function ϕ $(2^n - 1)$ in the following manner: If 2^{n-1}, where n is the number of stages in a shift register, is factorable into prime numbers denoted P_i, then the number of sequences is

$$\frac{\phi(2^n-1)}{n} = \frac{(2^n-1)\ (P_i-1)\ (P_2-1)\ (P_3-1)}{n\ (P_1)\ (P_2)\ (P_3)} \quad (3)$$

where each prime is used only once, even if it appears several times. For example, if L = 2^{n-1} = $2^6 - 1 = 1 \times 3 \times 3 \times 7$, then from the above equation, the number of available maximum length sequences is 6. In Table D-3 are tabulated the number of available maximum length shift register sequences of length L from a register of n stages.

Specific pseudo-random codes are generated with the aid of appropriate primitive polynomials which have been tabulated by Peterson[7] up to the 19th degree. Thus, in the table is listed a sixth degree primitive ploynomial 103 in octal notation, corresponding to a maximum shift register sequence of length L = 2^n -1 = 63 and 6 stage register. In binary notation, this polynomial can be expressed as 1,000,011 and represents the polynomial $x^6 + x + 1$. Figure D-5 illustrates the shift register sequence according to the 103 polynomial. The feedback connections are made from the one coefficients of the

TABLE D-3. NUMBER OF AVAILABLE MAX. LENGTH PR SEQUENCES, $\phi(2^n-1)/n$, FROM n STAGE REGISTER

$L = 2^n-1$	n	$\phi(2^n-1)$	$\phi(2^n-1)/n$
7	3	6	2
15	4	8	2
31	5	30	6
63	6	36	6
127	7	126	18
255	8	128	16
511	9	432	48
1023	10	600	60
2047	11	1936	176
4195	12	1728	144
8191	13	8190	630

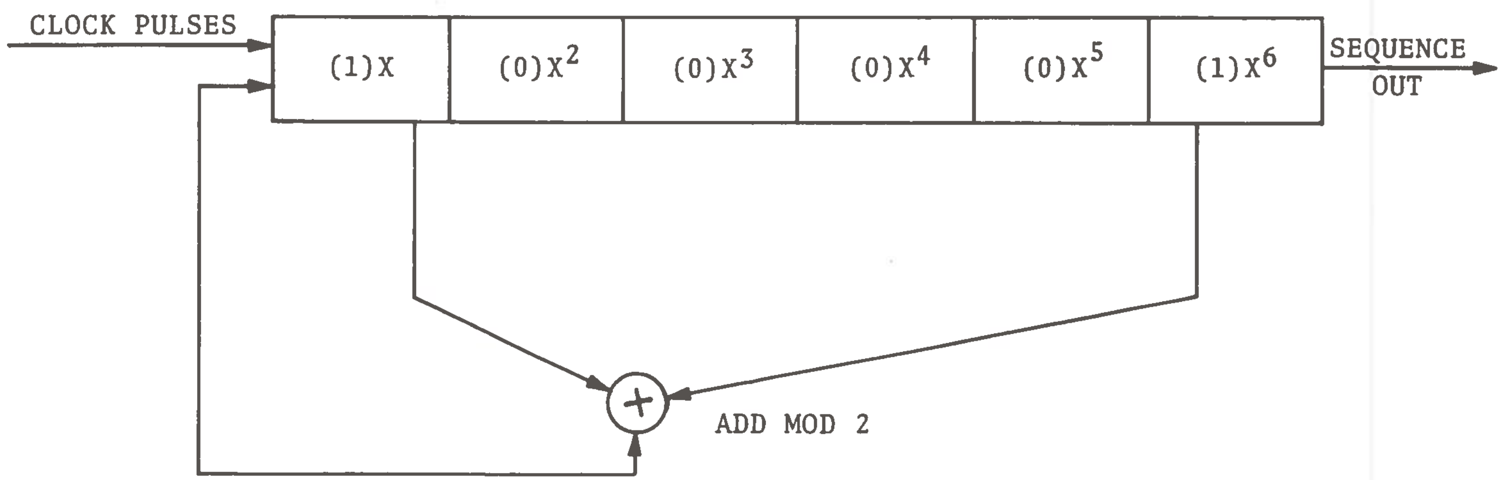


Figure D-5. Shift Register Encoder For Generating A PR Sequence Of Maximum Length $2^n-1 = 63$ From Primitive Polynomial $X^6 + X + 1$.

polynomial and added mod 2. The "one", which appears in every polynomial, represents the feedback connection to the input for closing the loop as shown in the diagram.

PSEUDO-RANDOM CODES FOR SURVEILLANCE APPLICATION

For the proposed 4th GATS surveillance function, a number of codes are required, but aperiodic, that is, of finite length as used in typical pulse radar applications. For finite length codes, however, the two-level autocorrelation functions of Equation 2 deteriorate and high peak sidelobes appear. In Figure D-6 are plotted the highest and lowest peak values for all known codes and starting points. The main peak value corresponds to the code length L, and the first sidelobe is generally in the order of $\sqrt{L}$. In Table D-4 are given the polynomials and starting conditions which investigations have shown to yield lowest peak side lobe amplitudes. The side lobes, if not sufficiently low with respect to the main peak, may appear as targets at the receiver output.

The requirements for low sidelobes also demand that the cross correlation between the transmitted surveillance waveforms be low. At the receiver, all coded signals appear in individual matched filter channels whose number corresponds to the number of different available codes. A high peak output is generated whenever a code is matched to a particular channel; the other codes in that channel should ideally produce only noise-like output. This will be true provided codes with "good" cross-correlation properties can be generated. Gold[3] and TRW[4] investigated to a limited extent such characteristics of pseudo-random codes. Gold has computed the partial correlation of these codes, which is conjectured to have characteristics similar to the cross-correlation. The maximum shift register sequences up to length 4095 were generated from longer polynomials (shorter codes generated from a higher order polynomial seem to have lower cross-correlation sidelobes). The lowest partial correlation values extracted from Gold's computations are tabulated in Table D-5. In the table, ℓ is the length of the sequence from which the shorter sequence of length L was derived. The number of such available "pseudo-random" sequences of length L is $\approx \ell/L$. The tabulated values

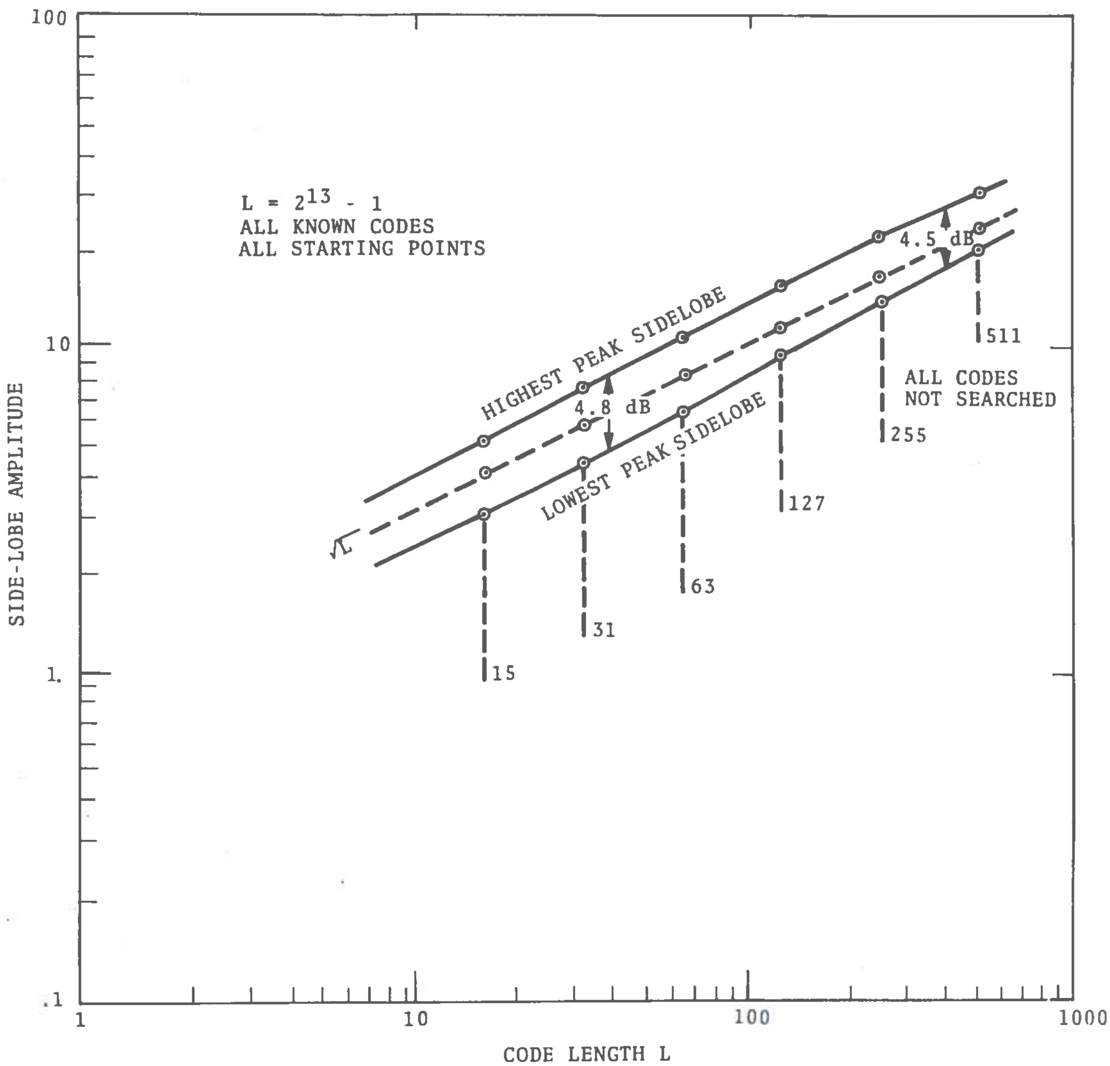


Figure D-6. Sidelobe Amplitude For Maximum Length Shift Register PR Codes Versus Code Length[5]

TABLE D-4. SIDELOBE CHARACTERISTIC OF PR CODES FOR THE SPECIFIED POLYNOMIAL AND INITIAL CONDITIONS.[6]

Degree (number of stages) and length	Polynomial, octal	Lowest Peak sidelobe amplitude	Initial** conditions, decimal	Lowest rms sidelobe amplitude	Initial conditions, decimal
1 (1)	003*	0	1	0	1
2 (3)	007*	-1	1,2	0.707	1,2
3 (7)	013*	-1	6	0.707	6
4 (15)	023*	-3	1,2,6,8	1.39	2,8
			10,11,12		
5 (31)	045*	-4	5,6,26,29	1.89	6,25
			(9 conditions)	1.74	31
			2,16,20,26	1.96	6
6 (63)	103*	-6	1,3,7,10	2.62	32
			26,32,45,54		
			(9 conditions)	2.81	35
			(9 conditions)	2.38	7
7 (127)	203*	-9	1,54	4.03	109
	211*	-9	9	3.90	38
	235	-9	49	4.09	12
	247	-9	104	4.23	24,104
	253	-10	54	4.17	36
	277	-10	14,20,73	4.15	50
	313	-9	99	4.04	113
	357	-9	15,50,78,90	4.18	122
8 (255)	435	-13	67	5.97	135
	453	-14	(20 conditions)	5.98	254
	455	-14	124,190,236	6.10	246
	515	-14	54	6.08	218
	537	-13	90	5.91	90
	543	-14	(10 conditions)	6.02	197
	607	-14	(6 conditions)	6.02	15
	717	-14	124,249	5.92	156
9 (511)	(24 codes)	-19	(1743 polynom.)	8.0	
10(1,023)	(30 codes)	-29	(3023 polynom.)		

* Single Mod 2 Adder Required

** Mirror Images Not Shown

TABLE D-5. PARTIAL CORRELATION VALUES OF PR CODES
(EXTRACTED FROM REF. 3)

Code Length ℓ / Window Size L	31	63	127	255	511	1023	2047	4095	8191
7	5	5							
15	5	7	7	9	9	9	11	11	13
31		7	11	11	13	13	13	15	17
63			9	15	15	17	19	21	23
127				15	21	21	23	25	29
255					19	29	31	35	39
511						27	39	43	47
1023							37	57	61
2047								53	75
4095									75

show the highest partial correlation for the sub-sequences of length L. Thus, if ℓ is 8191, and L = 511, there are 16 pseudo-random sequences whose sidelobe level will be equal or smaller than 47. From the table, it can be seen that the availability of codes with low cross-correlation sidelobes is limited.

If we compare Table D-5 with Table D-4, it can be seen that the auto-correlation sidelobes generated from the given polynomials are much lower than the cross-correlation sidelobes. However, it appears that the sequences generated from the higher order polynomial have also high autocorrelation sidelobes. In fact, computations by TRW indicate that for the sequence of length 511, generated from the higher order polynomial, the autocorrelation sidelobes are equal to or less than 44. This may be compared with Table 4, where for the particular polynomial the highest autocorrelation sidelobe is only 19.

From the above discussion, it can be seen that it is difficult to obtain codes with both low auto-correlation and cross-correlation sidelobes. Consequently, the number of codes available for surveillance applications may be limited. If the number of available codes suitable for surveillance applications is limited, as it appears to be, the number of pulse repetition intervals for aircraft identification may have to be increased, resulting in all likelihood in further degradation of system reliability and increasing of the processor complexity. A research effort for finding suitable codes for the surveillance application is strongly indicated.

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STUDY PAPER E

CORRELATION PROPERTIES OF CODES

EDWIN H. FARR

An important coding problem exists in the surveillance (data acquisition) function of both the Autonetics and Boeing proposed satellite systems. These systems use essentially RCA's SATANS and TRW's LIT systems respectively. In both systems the aircraft's position and identification are given by a combination of code sequence and pulse position. In the Boeing system, 16 PN (pseudo-noise) sequences, 511 bits long are proposed, and in the Autonetics system, 10 PN sequences, 255 bits long are proposed.

The performance of these codes boils down basically to their correlation properties. There are several kinds of correlation functions that can be formulated, and it is important to clearly understand each kind. The purpose of this memorandum is merely to identify and define, rigorously enough for our AATMS purpose, these different correlations. We shall not discuss here the rich mathematical structure that exists for these codes nor the results that have been obtained to date.

The following correlation properties are pertinent to the AATMS work:

1. Aperiodic autocorrelation
2. Aperiodic crosscorrelation
3. Periodic autocorrelation
4. Periodic crosscorrelation
5. Partial correlation

The first two properties, dealing with aperiodic correlation, are the most important for our purpose. And, as fate would have it, these are the properties about which the least is known. The most information is known about property 3, the periodic autocorrelation of PN sequences. These are discussed by Peterson[1] and Golomb[2], for example. Tables of partial correlation results have been prepared by Gold[3]. The last three properties above are more mathematically tractable and there is reason to believe that knowledge of these three will shed some light on the first two. This is the thrust of the paper by Boehmer[4] at least, in which periodic correlation properties are used to infer aperiodic correlation properties.

We shall be concerned with sequences of bits, i.e., sequences of binary digits {0,1}. To perform the correlation calculations, we convert to a sequence of +1's and -1's by the transformation $0 \to 1$, $1 \to -1$. Let $x_0, x_1, \ldots, x_{n-1}$ and $y_0, y_1, \ldots, y_{n-1}$ be two n bit sequences. Then, the aperiodic crosscorrelation function R_i for these two sequences is given by

$$R_i = \begin{cases} \sum_{j=0}^{n-i-1} x_{j+i} y_j & i \geq 0 \\ \sum_{j=0}^{n+i-1} y_{j-i} x_j & i \leq 0 \end{cases}$$

for $i=0, \pm 1, \ldots, \pm(n-1)$. The aperiodic autocorrelation function results when the two sequences are the same.

The periodic crosscorrelation function is given by

$$R_i = \sum_{j=0}^{n-1} x_j y_{j+i} \qquad i=0,\pm 1,\ldots,\pm(n-1),$$

where the subscript addition is performed mod n. The periodic autocorrelation function results when the two sequences are the same.

The partial correlation function deals with calculating correlation values obtained by sliding a sub-sequence along a given PN sequence. Let $z_0, z_1, \ldots, z_{k-1}$ be the given subsequence of k consecutive bits taken from the given PN sequence $x_0, x_1, \ldots, x_{n-1}$, $k<n$. Then the partial correlation function is

$$R_i = \sum_{j=0}^{k-1} z_j x_{j+i}, \qquad i=0,\pm 1,\ldots,\pm(n-1),$$

where again the subscript addition is performed mod n. As the subsequence slides past one end of the PN sequence, it is allowed to cycle back to the other end of the PN sequence. The parameter k is called the "window size".

In these definitions of correlation we should, strictly speaking, normalize the quantities by dividing by the sequence length. This would make the correlation values fall in the range -1 to +1, as they should. However, without this normalizing factor, the coding structure is brought out more clearly and this is why we omit it.

One question that has to be considered is to determine the optimum tradeoff between number of different codes used versus the number of unique pulse positions required. The extent to which this becomes a coding problem depends on the answer to this question. In fact, if there were enough pulse positions available, only one PN sequence would be needed and this would hardly be a coding problem at all. At the other extreme, one can create a horrendous coding problem by requiring a unique code for each aircraft and doing away with the pulse position scheme. It is interesting to observe that here we basically have a communication problem (broadly speaking) that can be formulated as a coding problem, but it need not be. In general, what the job coding does can often be done also by all sorts of non-coding techniques and vice-versa. Thus coding for the AATMS kind of situation, must be justified in relation to and in competition with these other techniques. As a first cut at the problem, there seems to be no reason to believe that the Autonetics and Boeing suggested technique is anything but a reasonable one.

In more conventional coding terms, the problem under consideration involves the use of codes for both synchronization and for error correction. The synchronization codes provide the timing which gives us range information. Good error correcting capability is equivalent to high distinguishability among the different codes with the same pulse position. Most of the coding work done in the past dealt with one or the other kind of codes. By requiring codes that have both capabilities, we tend to move into a relatively new area.

The coding work that needs to be undertaken now is stated rather well in the Autonetics Report on Pages 7-146. This suggests that code lengths and code selections be evaluated in terms of detailed cross- and autocorrelation effects for varied codes and overlap conditions. The results could then be used in conjunction with air traffic density estimates, transmission rates, and total message modulation concepts to derive probabilities of false alarm and false dismissal of pulses. These probabilities can then be related to system performance by evaluating their impact on surveillance capabilities, communication errors, etc.

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STUDY PAPER F

APERIODIC CORRELATION COMPUTATIONS OF AATMS CODES

EDWIN H. FARR

Some computer calculations were made over the past few months to determine aperiodic auto and cross-correlation functions of PN sequences. These codes (i.e., sequences) are being considered for use in performing the surveillance functions of proposed fourth generation ATC systems. The intent of this memorandum is to record and explain the computer results in summary form so that the greatest use can be made of them. Some elementary notions about AATMS codes were discussed in a previous memorandum dated 3 September 1971.

Let $X_0, X_1, \ldots, X_{n-1}$ and $Y_0, Y_1, \ldots, Y_{n-1}$ be two n-bit binary (two-phase) sequences of +1's and -1's. Then the discrete aperiodic cross-correlation function is given by the equation

$$R_i = \begin{cases} \sum_{j=0}^{n-i-1} X_{j+i}Y_j, & i = 0,1,\ldots,n-1 \\ \sum_{j=0}^{n+i-1} X_jY_{j-i}, & i = 0,-1,\ldots,-(n-1) \end{cases}$$

If the two sequences are identical, we get the auto-correlation function. The reason it was necessary at this time to go on the computer is that very little is known about the aperiodic correlation properties of PN sequences, whereas the mathematical structure of the codes guarantees ideal <u>periodic</u> correlation properties. The mathematics produces very little knowledge about aperiodic correlation and, therefore, the computer is necessary. However little is known about the aperiodic auto-correlation properties, and even less is known about aperiodic cross-correlation properties.

The results reported here should not in any way be considered a complete treatment of the problem. It is just a first attempt to get a better understanding of the performance of PN codes. The following table lists the number and type of aperiodic calculations performed:

Code Length (n)	Number of Correlations
7	2 auto and 1 cross-correlation
15	2 auto and 1 cross-correlation
31	6 auto and 3 cross-correlation
63	6 auto-correlations
127	18 auto-correlations
200	10 auto and all cross-correlations
255	16 auto and all cross-correlations
511	48 auto-correlations

The following tables give the maximum and minimum (except the peak value for auto-correlation) of all the correlation data listed above. The length of the code is n and the codes are listed at the left and top of each table. Where no cross-correlations were obtained, the table becomes a one-dimensional listing. The code sequences are identified by their shift register polynomial in the manner of Peterson.

n = 7	13	15
13	1 -2	3 -4
15		1 -2

n = 15	23	31
23	3 -4	6 -4
31		2 -3

n = 31	45	75	67	51	57	73
45	4 -5			11 -11		
75		5 -6				7 -10
67			5 -6		7 -10	
51				4 -5		
57					5 -6	
73						5 -6

n = 63

$103\ {}^{5}_{-6}$

$147\ {}^{7}_{-8}$

$155\ {}^{6}_{-7}$

$141\ {}^{7}_{-8}$

$163\ {}^{8}_{-9}$

$133\ {}^{6}_{-7}$

$211\ {}^{10}_{-11}$

$217\ {}^{10}_{-11}$

$235\ {}^{10}_{-11}$

$367\ {}^{11}_{-12}$

$277\ {}^{11}_{-12}$

$325\ {}^{12}_{-13}$

$203\ {}^{10}_{-11}$

$313\ {}^{9}_{-10}$

$345\ {}^{10}_{-11}$

n = 127

$221\ {}^{10}_{-11}$

$361\ {}^{11}_{-12}$

$271\ {}^{10}_{-11}$

$357\ {}^{11}_{-12}$

$375\ {}^{11}_{-12}$

$253\ {}^{12}_{-13}$

$301\ {}^{9}_{-10}$

$323\ {}^{9}_{-10}$

$247\ {}^{12}_{-13}$

n = 200	453	551	561	607	615	651	703	717	747	765
453	14 -16	37 -31	31 -30	46 -34	32 -30	32 -31	49 -28	40 -33	41 -26	29 -37
551		18 -19	35 -27	42 -38	43 -28	39 -42	55 -33	33 -36	35 -35	43 -34
561			14 -16	46 -30	47 -30	52 -36	51 -28	37 -24	39 -30	31 -30
607				15 -16	33 -41	38 -29	31 -31	33 -35	29 -33	37 -37
615					13 -17	30 -30	40 -29	41 -37	48 -25	36 -31
651						14 -16	45 -30	45 -37	41 -44	29 -34
703							17 -15	39 -31	33 -35	36 -34
717								17 -17	33 -31	50 -30
747									17 -18	52 -24
765										15 -18

n = 255	435	453	455	515	537	543	545	551	561	607	615	651	703	717	747	765
435	16 -17	54 -35	38 -27	43 -36	44 -35	63 -31	48 -30	58 -34	40 -35	62 -27	46 -52	45 -29	91 -35	55 -27	45 -62	33 -31
453		16 -17	39 -54	54 -33	33 -30	42 -38	85 -40	60 -29	50 -28	52 -39	37 -33	31 -35	50 -31	46 -50	56 -35	42 -38
455			19 -20	56 -38	60 -36	61 -27	36 -32	36 -44	56 -35	44 -35	89 -38	63 -29	43 -64	47 -27	56 -35	48 -29
515				14 -15	60 -28	44 -37	35 -40	36 -32	48 -30	52 -35	64 -26	92 -39	49 -31	44 -42	38 -27	47 -38
537					15 -16	40 -37	45 -44	48 -33	32 -29	47 -64	53 -25	47 -35	52 -32	87 -32	57 -31	40 -34
543						15 -16	63 -27	90 -37	48 -51	39 -27	40 -38	38 -28	40 -37	60 -26	47 -39	44 -34
545							14 -15	56 -38	43 -38	45 -31	43 -36	55 -29	48 -39	39 -28	60 -39	58 -28
551								19 -20	40 -27	46 -44	62 -28	42 -59	66 -32	41 -35	45 -26	60 -35
561									14 -15	75 -34	62 -28	59 -39	60 -31	42 -46	58 -29	44 -37
607										14 -15	40 -36	48 -38	33 -27	37 -29	32 -44	60 -29
615											13 -14	42 -40	49 -29	56 -37	49 -29	45 -37
651												18 -19	47 -37	60 -26	46 -61	37 -29
703													15 -16	50 -35	34 -31	43 -40
717														17 -18	31 -32	59 -30
747															20 -21	65 -33
765																16 -17

n = 511

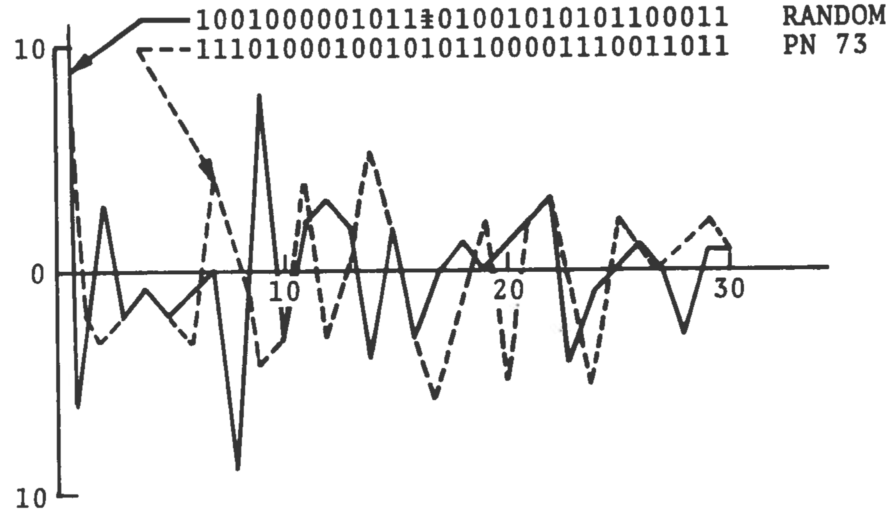
1021	24,-25	1425	26,-27
1033	21,-22	1437	21,-22
1041	21,-22	1443	21,-22
1055	20,-21	1461	26,-27
1063	24,-25	1473	28,-29
1131	22,-23	1517	25,-26
1137	23,-24	1533	24,-25
1151	22,-23	1541	23,-24
1157	26,-27	1553	24,-25
1167	21,-22	1555	22,-23
1175	21,-22	1563	26,-27
1207	27,-28	1577	24,-25
1225	23,-24	1605	27,-28
1243	26,-27	1617	25,-26
1245	20,-21	1665	27,-28
1257	23,-24	1671	23,-24
1267	26,-27	1707	28,-29
1275	26,-27	1713	25,-26
1317	23,-24	1715	24,-25
1321	25,-26	1725	24,-25
1333	24,-25	1731	23,-24
1365	23,-24	1743	23,-24
1371	24,-25	1751	25,-26
1423	23,-24	1773	24,-25

The sequences were started with the same configuration as the corresponding polynomial with the left most bit deleted. For example, for polynomial 1725, the starting sequence is 111010101. The 200 bit sequences are taken from the 255 bit PN sequences. These sequences were chosen because ten 200 bit sequences were recommended for the Autonetics system.

For the PN sequences listed, the sum of the plus and minus peak values for auto-correlation for a given sequence is always -1. The reader can verify that this is true for all PN sequences.

The graphs of the correlation data cannot all be included in this report because of the large number. For practical reasons, only the following graph is shown. This one example is fairly typical of all the graphs.

Aperiodic Auto-Correlation n = 31



The curve for a "genuine" random sequence, attained by flipping a coin, is shown to illustrate that the two curves behave in about the same manner. A comparison of these two codes, however, tends to corroborate the results found by DeLong (MIT Report, "Experimental Auto-Correlation of Binary Codes", 1960). He showed that PN sequences in general have lower peak auto-correlation side lobes than sequences selected at random, and in this particular case we get 9 for the random code versus 6 for the PN code.

Although the results for PN sequences for auto-correlation are reasonably good and better than random sequences, it appears that for some mysterious reason, the cross-correlation results for PN sequences are disappointing. One would think that the cross-correlation functions, of two random sequences would be like their auto-correlation functions, but slightly worse only because the auto-correlation functions are symmetric about the origin. The deterioration is due only to the fact that for cross-correlation we have twice as many independent values than for auto-correlation. Therefore, the cross-correlation curve and its peaks might be higher in magnitude by about $\sqrt{2}$. However, the data we have for PN sequences show the cross-correlation peaks to be much higher than this in general.

STUDY PAPER G

POLYPHASE CODES FOR THE AATMS SURVEILLANCE

EDWIN H. FARR

The common coding technique used for range acquisition purposes, such as is proposed for satellite-based AATMS Surveillance Systems, utilizes binary codes. The purpose of this memorandum is to stress that binary codes normally use two-phase signal modulation and that this may be an artificial and needless restriction. Using polyphase or m-phase signal sets $m \geq 2$, may improve surveillance performance significantly enough to justify the additional implementation cost. We develop herein some of the mathematics and some fundamental properties of m-phase codes, and give the computed correlation results of specific codes.

An m-phase code is a complex-valued sequence $\{U_r\}$ of length n, where any element of the sequence is a member of the set

$$\left\{1,\ e^{2\pi i \frac{1}{m}},\ e^{2\pi i \frac{2}{m}}, \ldots, e^{2\pi i \frac{m-1}{m}}\right\}.$$

This set represents signals with phase angles equally spaced in the plane. The aperiodic cross-correlation function $C_{UV}(t)$ of two codes $\{U_r\}$ and $\{V_r\}$ is the Hermitian dot product

$$C_{UV}(t) = \begin{cases} \sum_{r=1}^{n-t} U_r V^*_{r+t}, & t=0,1,\ldots,n-1 \\ \sum_{r=1}^{n+t} U_{r-t} V^*_r, & t=0,-1,\ldots,-(n-1), \end{cases} \quad (1)$$

where Z^* denotes the complex conjugate of Z. Code performance will be determined by the magnitude of Equation 1, namely $|C_{UV}(t)|$.

If $U \equiv V$, we have the usual auto-correlation function. If $|C_{UV}(t)| \leq 1$ for $t \neq 0$, we get the case referred to by Golomb and Scholtz[1] as a "generalized Barker sequence".

Some useful properties of Equation 1 can be obtained. For $t>0$,

$$C^*_{VU}(-t) = \sum_{r=1}^{n-t} V^*_{r+t} U_r$$

$$= C_{UV}(t). \qquad (2)$$

Thus $|C_{VU}(-t)|=|C_{UV}(t)|$ and hence the order of the codes does not matter in determing code performance. Also, setting U = V, we get $C_{UU}(t) = C^*_{UU}(-t)$ and hence one need only run through positive values of t for auto-correlation.

It readily follows from Equation 1 that $C_{U*V*}(t) = C^*_{UV}(t)$ and $C_{UV*}(t) = C^*_{U*V}(t)$, which shows what happens when one or both of the sequences is complemented.

If one sequence is the reverse of the other, i.e. $U_r=V_{n-r+1}$, we have from Equation, $C_{UV}(t) = C^*_{UV}(t)$ and the cross-correlation function is real. For a "palindromic" sequence (one that reads the same forward as backwards) $U_r=U_{n-r+1}$, and hence the autocorrelation function is real. A real correlation function may be of mathematical interest but we have not found any practical value to this whatsoever.

If $\{U_r\}$ and $\{V_r\}$ are both palindromic sequences, by making the proper substitution in Equation 1, we have that $C_{UV}(t) = C^*_{VU}(t)$. From Equation 2 it follows that $C_{UV}(t) = C_{UV}(-t)$ making the cross-correlation function symmetric.

There is a set of transformations that take a set of m-phase codes into another set of codes, not necessarily of m-phase, but with the same correlation functions (in magnitude). For two given codes $\{U_r\}$ and $\{V_r\}$ we define two new codes $\{X_r\}$ and $\{Y_r\}$ of length n by the equations

$$X_r = U_r e^{i(Ar+B)}, \quad Y_r = V_r e^{i(Ar+B)}, \qquad (3)$$

where A and B are real constants. From Equation 1, the cross-correlation function of the new codes is

$$C_{XY}(t) = \begin{cases} \sum_{r=1}^{n-t} X_r Y^*_{r+t} \\ \sum_{r=1}^{n+t} X_{r-t} Y^*_r \end{cases}$$

$$= \begin{cases} \sum_{r=1}^{n-t} U_r e^{i(Ar+B)} V^*_{r+t} e^{-i(Ar+B)-iAt} \\ \sum_{r=1}^{n+t} U_{r-t} e^{i(Ar+B)-iAt} V^*_r e^{-i(Ar+B)} \end{cases}$$

$$= \begin{cases} e^{-iAt} \sum_{r=1}^{n-t} U_r V^*_{r+t} \\ e^{-iAt} \sum_{r=1}^{n+t} U_{r-t} V^*_r \end{cases}$$

$$= e^{-iAt} C_{UV}(t)$$

Thus $|C_{XY}(t)| = |C_{UV}(t)|$ for all t since $|e^{-iAt}| = 1$.

There are many transformations that come from Equation 3 and we shall not discuss all of the implications of this at this time. We just indicate a few simple transformations.

If $A=0$ and $B=2\pi/m$, the m-phase codes are merely shifted in phase one phase angle counter-clockwise. If $B=0$ and $A=2\pi/m$, the m-phase codes undergo "progressively increasing" shifts in phases. For binary codes, setting $B=0$ and $A=\pi$ produces codes whose elements alternate in sign relative to the original codes. Setting $B=0$ and $A=2\pi/m$ transforms binary codes into polyphase codes.

Since so much more is known about binary code performance than for m-phase codes m>2, we must seriously consider the question: Are there m-phase codes m>2, that are significantly better than binary codes? The above analysis shows that there are many m-phase codes that give identical correlation functions to the binary codes. This can be seen first, trivially, by observing that if m is even, the signal elements with phase angles 0 and 180° are members both of the m-phase signal set and the binary set. By restricting transmissions to these two signal elements, we can duplicate the binary transmissions. Secondly, the above transformation in Equation 3 with B=0 and $A=2\pi/m$ transforms, in a non-trivial way, binary codes to m-phase codes if m is even. For m odd, the transformations are not so clear but this point is not important at present.

To sum up then, the best m-phase codes, m even, are at least as good in both auto-correlation and cross-correlation as the best binary codes. And due to the very large number of m-phase codes that exist, it would be reasonable to guess that some are better than the best binary codes. The problem now is to find these better m-phase codes.

It is evident from the literature that very little has been done on the subject of m-phase pulse compression codes, let alone sets of m-phase codes with good aperiodic distinguishability. One early paper on the subject is by DeLong[2] in which he investigated the properties of 3-phase codes only and was able to produce Barker 3-phase codes up to length 9. He did not consider cross-correlation of sets of codes at all.

A paper by Frank[3] gives a constructive method of producing an m-phase code of length m^2. These codes have very high auto-correlation center peak to side peak ratios, at least compared to binary codes of the same length. But to get a code of reasonable length we need a large number of phase angles. For example, for AATMS surveillance a code length around 256 might be used, in which case we would need 16 phase angles. Frank also fails to provide a <u>set</u> of codes with low cross-correlation.

The only sets of polyphase codes known by the writer to exist at present, for a given length and for a "reasonable" number of phase angles, are given by Golomb and Scholtz[1]. The purpose of that paper is to produce all the generalized Barker codes that can be found by computer with a reasonable amount of computation. Since these codes were selected on the basis of their auto-correlation property only, it was not known heretofore what their cross-correlation properties would be like. We have computed all of the cross-correlation functions for the "sextic" Barker sequences and have tabulated the maximum absolute value for each pair. The results are shown below, where the integers designating the code represent phase angles that are multiples of 60°:

n=3	2)
1) 014	1.7
2) 021	

n=4	2)	3)	4)
1) 0131	3.6	2.6	2
2) 1243		3.5	2
3) 5032			2.6
4) 0423			

n=5	2)	3)	4)	5)	6)	7)
1) 01253	4.4	2.6	2.6	3.6	3.6	2.6
2) 12315		2.6	2.6	3	4.4	2.6
3) 43245			4.4	2.6	2.6	2.6
4) 32023				3	2.6	3.6
5) 45254					4.4	3
6) 01305						3
7) 52150						

n=7	2)	3)	4)	5)	6)	7)	8)	9)
1) 0125332	3.6	3.6	4	4.6	4.4	3	3.6	3.6
2) 4021321		2.6	4.4	3.6	3	3.6	2.6	4.6
3) 4311313			3.6	4.6	3	3.3	3	2.6
4) 3400515				5	3.6	3.6	3	3.6
5) 4515154					4	4.4	4.6	4.4
6) 3423305						3.5	3.6	6.1
7) 1021250							6.2	3.6
8) 5410145								4
9) 1251032								

n=8	2)	3)	4)	5)	6)	7)	8)	9)	10)	11)
1) 01231152	4.6	4	4	3.6	4.4	3.6	3.6	5.6	4.6	4.4
2) 54314001		3	4.6	7	3	4.4	4.4	5	3.6	4
3) 23403054			3	3	6.1	4.6	3.6	3.6	4.4	4
4) 32142401				4.6	3.6	3.6	4	3.5	4	3.5
5) 21030223					3.6	3.5	4.4	5	3.6	3.6
6) 50142005						4.6	4.4	3.6	4.6	4
7) 53543350							4.6	3.5	4	3.5
8) 25322245								5.2	4.4	3.6
9) 03550532									5.6	3.6
10) 34122143										4.6
11) 01503254										

n=9	2)	3)	4)	5)	6)	7)	8)	9)	10)	11)	12)	13)	14)	15)	16)	17)	18)
1) 012303054	4.6	3.6	3.6	6.2	3.5	7.9	4.4	7	5.3	3.6	3.5	4.4	4.4	4.6	4	4	4.4
2) 543244040		4	4.4	5.3	5	5.3	4.6	3.6	6.6	4	5.6	4	6.1	3.6	4.4	4.6	5.6
3) 141335432			5.6	3.6	5.6	5	4.6	4.6	3.6	5.2	5.3	5.6	3.6	4.6	4	4.6	4.4
4) 321525123				3.6	7	3.6	6.6	3.6	4.6	4.6	5	4.6	4	4	4.4	3.6	4.6
5) 450252005					3.6	5.2	3.5	5.3	4.6	3.6	4.6	4.6	4.6	5.3	4	5.6	4.4
6) 105352440						4	6.9	4.4	4	5.6	4.4	4.4	5	4.4	4.4	4.6	3.5
7) 133524210							3.6	6.1	5.6	4.4	4.6	4.4	4	4.6	4.6	4.6	5.6
8) 211530345								5	3.6	4.4	3.6	6	4.4	3.6	4	4.4	4
9) 234152005									3.6	4	5.6	4	3.6	4.4	4.6	3.6	4.6
10)320011302										4	4.4	3.5	6.1	3.6	4.4	3.6	5
11)025112154											3.6	5	4	6.9	4.6	4.6	3.6
12)430443401												4	5.3	3.6	5.6	3.6	4.4
13)013433152													3.6	5.3	3.6	5.6	3.6
14)542122352														4	3.6	3.6	5.6
15)014050532															4	4.4	3.6
16)430454023																3.6	7.8
17)451322143																	4
18)103232301																	

n=10	2)	3)	4)	5)	6)	7)	8)	9)	10)	11)
1) 0123030432	5.3	5.6	3.6	6	3.6	8.7	3.5	6.1	4.4	5.3
2) 5431122413		4.6	4.4	4	5.3	5.6	4.4	4.6	6.2	4.4
3) 3403540432			4.6	5	4.4	5.3	3.6	7.2	3.6	7.2
4) 2015105123				4.6	4.6	4.6	5.3	5.6	4.4	5.2
5) 4503434205					4.4	5.3	4.6	5.3	4	7.2
6) 1043130112						4	5.2	4.6	5.3	4.6
7) 5001514310							4.6	6	4.4	5.3
8) 4314231245								4.4	7	3.6
9) 2350453143									3.6	6.6
10)3205051401										4
11)1241213154										

n=11	2)	3)	4)	5)	6)
1) 01230151532	4	5.2	5	4.4	4.6
2) 54304045023		5.2	4.6	4	5.2
3) 50402231532			5.3	6.1	4
4) 05153324523				3.6	6.1
5) 45240004254					4.4
6) 10315551201					

n=12	2)	3)
1) 013331252152	7.6	4.6
2) 124545353154		6
3) 431534342401		

These codes are all distinct, in the sense that no one code can be transformed into another by one of the above transformations. The codes are different, but equivalent by a transformation, to those in Reference 1, where they are identified by the numerically lowest sequence. The identification in Reference 1 produces a lot of leading zeros and this has the undesirable effect of causing the cross-correlation values to be high with zero shift. To correct this, we transformed the codes in Reference 1 so as to make them appear as random-looking as possible.

The codes only go up through length 12 and therefore the length may be short for our purpose. However, these results are still more pessimistic than we had hoped. Perhaps another direction to go at this time is to construct polyphase PN sequences, analogous to the binary PN sequences, and determine their aperiodic correlation properties. Some of the mathematical foundation of these generalized codes is given in a paper by Zierler[4].

REFERENCES

1. S.W. Golomb and R.A. Scholtz, "Generalized Barker Sequences", IEEE Trans. on IT-11, October 1965, p. 533-537.
2. D.F. De Long, "Three-Phase Codes", MIT Lincoln Lab. Group Report, 47-28, July 1959.
3. R.L. Frank, "Polyphase Codes with Good Nonperiodic Correlation Properties", IEEE Trans. on IT, January 1963, p. 43-45.
4. N. Zierler, "Linear Recurring Sequences", Journal of SIAM, Vol. 7, p. 45, March 1959.

STUDY PAPER H

CONSTRAIN ON THE LENGTH OF PR CODES DUE TO DOPPLER EFFECTS IN THE AATMS SURVEILLANCE

B.S. GOLDSTEIN

In the AATMS surveillance bi-phase pseudo-random codes are utilized. A long code is desired to maximize the processing gain. The length of the code, however, is constrained by doppler and other frequency off-sets. In the receiver the filter processing gain G as given by TRW[1] is

$$G = L \left[\frac{\sin(\pi L \Delta f \tau_{ch})}{\pi\ L\ \Delta f\ \tau_{ch}} \right]^2 \qquad (1)$$

where L is the number of code elements, γch is the compressed pulse width and it is the frequency off-set. To maximize the gain G, the following relationships are given by TRW

$$L_{opt} = \frac{0.372}{\Delta f\ \tau_{ch}} \qquad (2)$$

$$G_{opt} = \frac{0.23}{\Delta f\ \tau_{ch}}$$

These relationships are derived in the attached appendix. In Figure H-1 is plotted the normalized gain G/L versus $L\tau_{ch}$ with the frequency off-set of as a parameter. From these curves the proper choice of L can be determined for the postulated restrictions. For example, if

$$\Delta f = 4 \times 10^3 \text{Hz}$$

$$G/L = 0.9 \text{ (Allowable degradation)}$$

then, the following condition should be satisfied

$$L\ \tau_{ch} \leq 4 \times 10^{-5}$$

Thus, if τ_{ch}=0.1μsec., the number of code elements should be constrained by L≤400elements. If τch=0.2μsec., then L≤ 200 elements.

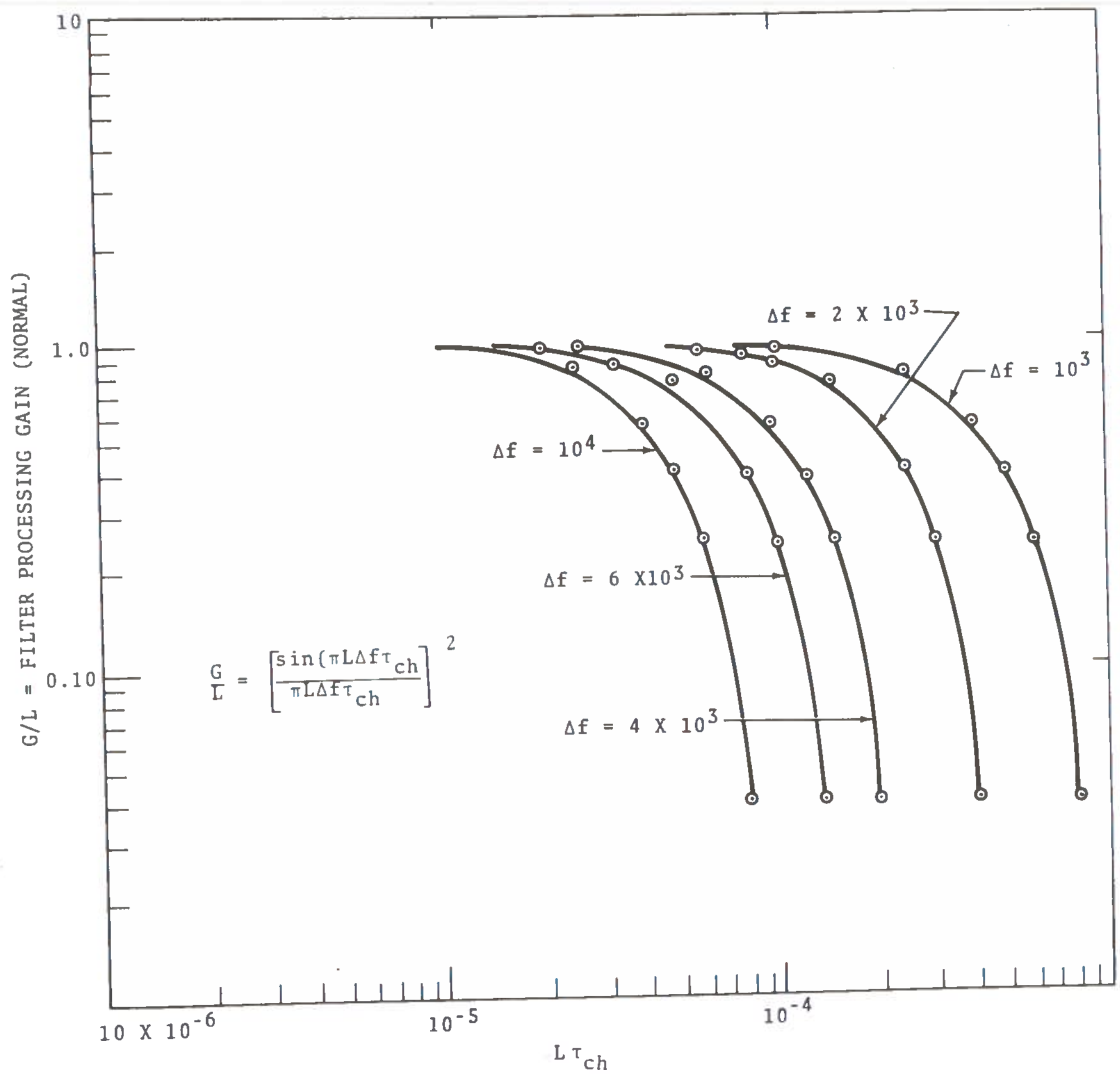


Figure H-1 Normalized Filter Processing Gain Versus $L\tau_{ch}$ with the Frequency Off-Set as a Parameter

REFERENCE

1. "Communication, Navigation and Surveillance for Aircraft Operating Over the Contiguous United States," TRW Report, Dec. 1970, Vol. I, prepared under Contract NAS S-21535.

APPENDIX A

The filter processing gain is given by

$$G = L\left[\frac{\sin(\pi L \Delta f \tau_{ch})}{\pi\ L\ \Delta f\ \tau_{ch}}\right]^2$$

where

L = bi-phase code length in bits

Δf = total frequency off-set

τ_{ch} = compressed pulse width

Let

$$a = \pi \Delta f \tau_{ch}$$

$$L = x$$

$$G = \frac{\sin^2 ax}{a^2\ x}$$

$$\frac{\alpha G}{\alpha x} = \frac{1}{a^2}\left[\frac{1}{x}\, x 2a \text{ Sin } ax \text{Cos} ax\ \frac{-1}{x^2} \text{ Sin}^2 ax\right] = 0$$

or $2ax = \tan ax$

Solving $\quad ax \approx 1.16 \approx \pi \Delta f \tau_{ch} L$

$$L = \frac{1.16}{\pi \Delta f \tau_{ch}}, \quad G = \frac{0.82}{(1.16)\pi \Delta f \tau_{ch}}$$

STUDY PAPER I

I. FALSE ALARMS, PROBABILITY OF DETECTION AND THE EFFECTS OF SIDELOBES ON THE PROPOSED AATMS SURVEILLANCE

B.S. GOLDSTEIN

The performance of a surveillance system is determined by the number of false alarms the system generates and its capability to detect aircraft. These performance criteria are directly related to the system signal-to-noise ratio requirements. In this study paper, the following is determined:

1. The false alarm rate and probability of detection as a function of signal-to-noise ratio in a single receiver channel in the presence of receiver noise only, from which signal energy requirements can be established.
2. The number of alarms generated by sidelobes when auto-correlating or cross-correlating pseudo random codes in the receiver; it is shown that with the codes so far available, the system would generate a high alarm rate.

FALSE ALARMS, PROBABILITY OF DETECTION, AND SIGNAL ENERGY REQUIREMENTS.

In the proposed satellite aided surveillance system, pseudo random codes are transmitted via satellites to receiving ground stations. In the receiver the waveforms are matched filter detected, compressed and envelope detected. If we assume that this corresponds to envelope detection of a narrow band Gaussian process, then the envelope statistics are Raleigh distributed. For such a process, the probability of false alarm P(FA) and the probability of detection P(D) can be evaluated from either Rice's expressions or the Marcum Q-function. According to Rice, the P(D) is given by

$$P(D) = \int_{V}^{\infty} \frac{R}{N_o} \exp\left(-\frac{R^2+S^2}{2N_o}\right) I_o\left(\frac{RS}{N_o}\right) dR \qquad (1)$$

where

R = amplitude (random variable)

N_o = rms noise

V = threshold voltage

S = peak amplitude (sinusoid)

I_o = Bessel function of zero order

The P(FA) is given by setting the signal amplitude S=0. That is, with no signal present the P(FA) is,

$$P(FA) = \int_v^{\infty} \frac{R}{N_o} \exp\left(\frac{-R^2}{2N_o}\right) dR = e^{-\frac{V^2}{2N_o}} \qquad (2)$$

In Figure I-1, the P(FA) is plotted as $\log_{10}$ P(FA) versus v^2/No in dB's, from which the alarm rate can be readily assessed.

In the proposed surveillance system, the P(FA) should be small to prevent system saturation. The number of false alarms per second, n, is proportional to system bandwidth and is given by approximately

$$n = W \exp\left(-\frac{V^2}{2N}\right) \approx W x P\ (FA) \qquad (3)$$

Where W is the system bandwidth.

For example, if the processing system can tolerate 10 alarms/sec., and W = 10 MHz as contemplated for the LIT type system, the threshold v^2/No must be set so that

$$P(FA) = \frac{n}{W} \leq 10^{-6} \qquad (4)$$

The expression for the P(D) as a function of the various parameters was given in Equation 1 above. The computational values are drawn in Figure I-2 as a function of signal-to-noise ratio with P(FA) as a parameter. The required signal-to-noise for a given P(D) and acceptable number of alarms per second can be readily evaluated from Figure I-2 and Equation 4. For example,

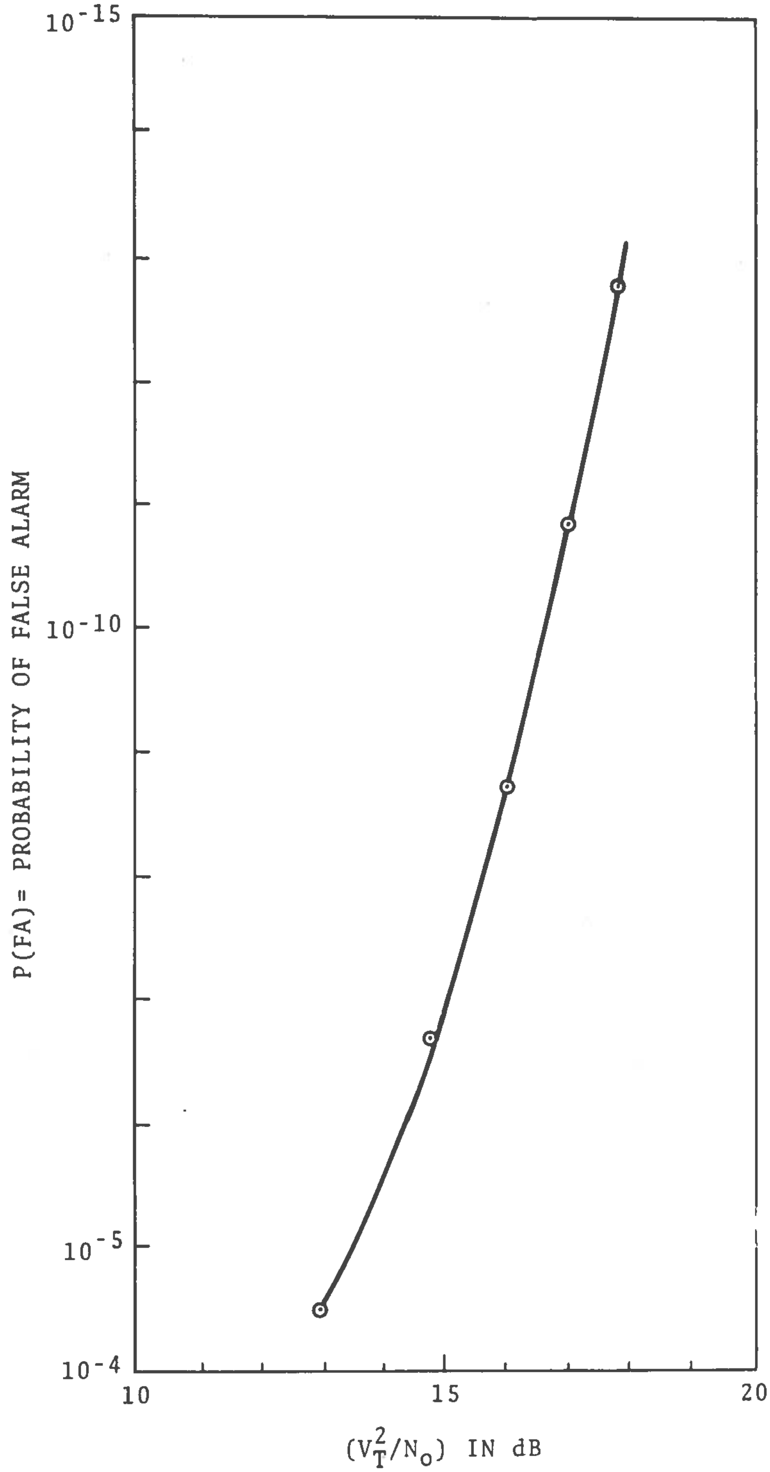


Figure I-1. Probability of False Alarm vs. Normalized Threshold V_T^2/No (V_T = Threshold voltage)

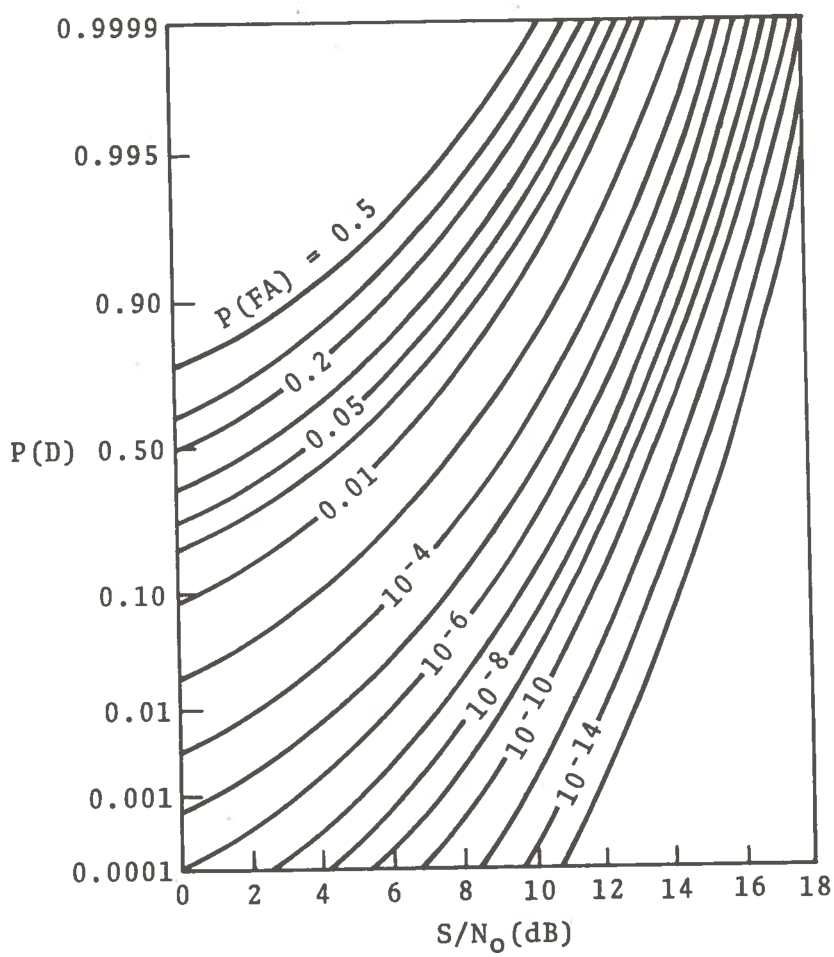


Figure I-2. Signal To Noise Ratio Versus Probability of Detection with P(FA) as a Parameter

IF:	n	= 10 alarms/sec.
	B	= 10MHz
then,	P(FA)	= 10^{-6} (Equation 4)

From Figure 2, it can be seen that

For:	P(FA)	= 10^{-6}
and,	P(D)	= 99.99%
	the (SNR)	16dB

For lower P(D), less signal power would be required, However, in the surveillance system under consideration, a high P(D) is required because of the large anticipated number of aircraft in the system. If we assume 100,000 aircraft in Conus transmitting once per second, then even with a P(D) of 99.99%, each second an average of 100 aircraft would not be detected per receiving station. These aircraft would have to be picked up on the next transmitting cycle.

THE EFFECT OF SIDELOBES ON THE FALSE ALARM RATE IN 4TH GATCS SURVEILLANCE

In a typical surveillance system, a threshold level is set for a specified probability of false alarms. In the presence of a signal the threshold is exceeded and the target is detected. If the sidelobe level is high enough, the threshold may trigger even though no target is present, thus generating a false alarm. To evaluate the false alarm rate due to sidelobes, we set the following initial system design characteristics:

P(FA) = probability of false alarm = 10^{-6}

P(D) = probability of false alarm = 10^{-6}

P(D) = probability of detection = 99.99%

For the specified initial conditions, the required signal-to-noise ratio SNR=16 dB.

The sidelobe level may be considered as signal energy in a gaussian noise which generates a false alarm whenever the set threshold is exceeded. If we denote the sidelobe level to be X dB below the main peak signal value, then from Rice's curves we can construct the following table.

Sidelobe level X dB below peak	3	6	10	13	16
$P(FA)_s$ due to sidelobe	0.9	0.25	0.01	8×10^{-4}	10^{-4}

where $P(FA)_s$ is the probability of false alarm due to sidelobes with a level of X dB below peak signal value.

The total number of false alarms/sec. due to sidelobes is given by

$$\text{False alarms/sec. due to sidelobes} = \left(\frac{mNa}{Tn}\right) P(FA)_s \qquad (5)$$

where

m = no. of transmitted pulses/T

Na = no. of aircraft

T = up-date rate

n = no. of independent channels

Suppose now that $Na = 10^5$; T = 2 sec.; m = 2; n = 1. Then, the number of false alarms/sec. due to sidelobes is given as follows:

X dB below peak	10 dB	13 dB	16 dB
False alarms/sec.	1000	70	10

The above false alarm rate is given for one sidelobe only. If there are N sidelobes, the number of false alarms is given more generally by

$$\text{Total false alarms/sec. per channel} = \left(\frac{mNa}{Tn}\right) \sum_{n=1}^{N} \left[P(FA)_s\right] \qquad (6)$$

where N is the number of sidelobes generated per pulse, X dB below peak value.

For a truncated pseudo random code of length L, the number of sidelobes generated per correlation per pulse is

$$N = 2[L - 1] \qquad (7)$$

but, only the significant sidelobes need to be considered.

STUDY PAPER J

SYSTEM FALSE ALARM OF MULTI-PULSE WAVEFORMS PROPOSED FOR THE AATMS SURVEILLANCE

B.S. GOLDSTEIN

In the proposed satellite aided surveillance system multi-pulse waveforms are being considered for aircraft transmissions. In the original LIT type system proposed by TRW, each aircraft transmits a single bi-phase pseudo-random modulated waveform per up-date interval. In a multi-pulse transmission, it is hoped that:

1. The aircraft identification problem can be alleviated;

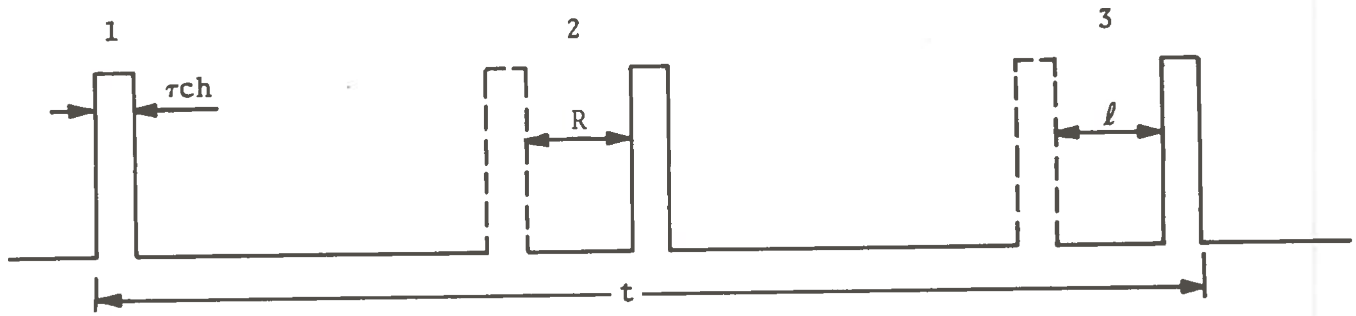
2. The processing complexity can be reduced;

3. The reliability of the system can be increased

In Figure J-1 is shown, for example, a video detected three pulse waveform proposed by Autonetics. The aircraft identity is determined from the unique pseudo-random codes, frequency channels and the relative position of the pulses with respect to each other. If a pulse from another aircraft occurs within these pulse repetition intervals a false identification may result. Because of the large number of transmitting aircrafts, the resulting false alarm rate may saturate the system.

Both Autonetics and RCA calculated the alarm rates for the respective recommended waveforms. However, since a large number of waveforms, configurations, and system parameters can be postulated, it was felt to be advisable to express these calculations in generalized formulas in order to see how system performance is affected by the choice of different waveforms and parameters. These expressions should aid in the evaluation of competitive waveforms. The method of analysis carried out is simple minded and should be considered only as a first attempt to obtain a quantitative measure of system reliability.

The number of pulses present in a channel due to aircraft transmissions is



τch = PULSE WIDTH

$\ell R/\delta$= PRI = NO. OF PULSE REPETITION INTERVALS

δ= WIDTH OF ONE PRI

t = WAVEFORM DURATION

Figure J-1. Detected Compressed Address Pulses from an Aircraft

$$\text{no. of pulses/sec.} = \left(\frac{m}{T}\right)\left(\frac{Na}{nF}\right) \qquad (1)$$

where

m = no. of pulses/transmission

T = up-date rate

N_a = no. of aircraft

n = no. of PR codes

F = no. of frequency channels

If the basic pulse width interval is δ, then the probability of a pulse within such an interval due to interfering pulses is

$$\begin{array}{l}\text{no. of interfering}\\ \text{pulses}/\delta\end{array} = \left(\frac{m}{T}\right)\left(\frac{Na}{nF}\right)\delta \qquad (2)$$

It should be pointed out that the above expression is an average number. If we assume that the pulses are Poisson distributed, then the probability of more than one pulse occuring within such an interval can be readily evaluated. A more thorough analysis would have to take this into account. The problem of finding the first pulse needs also to be resolved. However, the approach taken by the contractors will be followed here to obtain the generalized expressions.

An identity is registered when no pulses occur in intervals corresponding to a true address. The probability of such an occurrence is given by

$$\begin{array}{l}\text{probability of}\\ \text{false address}\end{array} = \left[\left(\frac{m}{T}\right)\left(\frac{Na}{nF}\right)\delta\right]^{m \times PRI} \qquad (3)$$

where PRI is the number of pulse repetition intervals or number of positions in which an pulses register as an address.

The slave stations carry out similar identification procedures. However, the same false identity can occur only within a time span of about 48 milliseconds which is the maximum delay of signal arrival from several satellites. Hence the probability of some false identity at a slave station is given by

$$\text{probability of false address at slave station} = \left[\left(\frac{m}{T}\right)\left(\frac{Na}{nF}\right)\delta\right]^m x\ \frac{48}{\delta} \qquad (4)$$

If three, time of difference, measurements are correlated to register an identity, the system false alarm per channel per, SFA is

$$SFA = \text{system false alarm/channel}/\delta = \qquad (5)$$

$$\left\{\left[\left(\frac{m}{T}\right)\left(\frac{Na}{nF}\right)\delta\right]^m x\ \frac{48}{8}\right\}^3 x \left\{\left[\left(\frac{m}{T}\right)\left(\frac{Na}{nF}\right)\delta\right\}^m x\ PRI$$

and the total system false alarm per second TSEA/sec. is

$$TSFA/sec. = \frac{(SFA)\ X\ nF}{\delta} \qquad (6)$$

From the above expressions the impact of various parameters on system performance can be readily assessed. Since the up-date rate and the number of aircraft is determined by system requirement, the significant quantity to be assessed is $\left(m\delta/nF\right)^m$ which has the greatest impact on system reliability. The quantity δ should be considered as a tolerance with which aircraft positions or pulse arrival can be measured, particularly when correlating between different receiving stations. A limited but more complex analysis was carried out by Lincoln Laboratory for specific stated conditions. In all probability better statistical models and simulations will be required to determine system reliability with a high degree of confidence.

REFERENCES

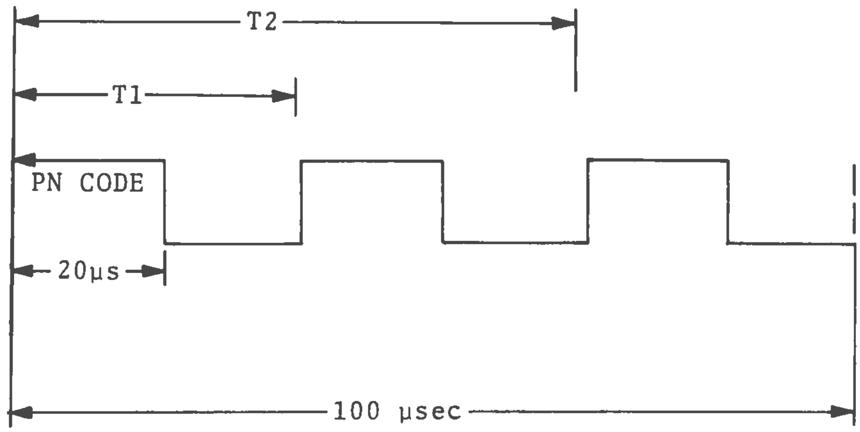
1. I.G. Stiglitz et. al., "Concept Formulation Studies of the Surveillance Aspects of the Fourth Generation Air Traffic Control System" Lincoln Laboratory Report ATC-7, September 1971, prepared under Contract DOT/TSC-241

STUDY PAPER K

THE EFFECT OF FALSE ALARMS ON DATA PROCESSING IN THE AUTONETICS 3-PULSE SYSTEM

J. DUMANIAN

Assuming a transmission rate per aircraft of 1.0 + 1.2 = 2.2 codes per second, Autonetics has calculated false alarms. The codes are actually transmitted in a 3 pulse code train where each 20 usec pulse is actually 200 chips (.1 us. per chip) of a PN pseudo-random sequence; see the figure below.



Ten transmission frequencies and 10 PN codes are used to partition a million aircraft population to 10^4; 100 possible locations or times T_1 for the second pulse and 100 possible locations or times T_2 for the third pulse complete the identity. One-tenth microsecond precision is used for location of these pulses.

Assuming that all the pulse trains emitted for surveillance purposes at 1.2 codes per second or 3 codes per 2.5 seconds and all the pulse trains emitted for communications enter the position determining computer; then according to Autonetics (see page 7-109), in the July, 1971 Final Report, the probability that a false position is calculated is 1.76×10^{-12}.

The above analysis does not consider"overwriters" i.e. when a genuine signal has been overwritten by noise or other signal and results in multiple possibilities, some correct, and some are false alarms.

If it is assumed that on single overwrites, both possible identities are passed on to the position determining computer then the following analysis, items 9-28, apply. It is reasonable to assume that two identities can be transferred to the position determining computer in 100 microseconds using conventional integrated circuit logic.

9. Probability a random pulse arrives at a particular time on a particular channel of the master satellite in an acceptable T_1 or T_2 position = item 3, on page 7-109,* x the total possible T_1 and T_2 positions = $3.2 \times 10^{-4} \times 200 = .064$.

10. The average number of opportunities per second per channel for a random pulse to cause an overwrite on a user identity =

$$\frac{10^5 \text{ transmitting aircraft}}{10 \text{ (freq's)} \times 10 \text{ (PN codes)} \times 2.5 \text{ (secs.)}} = 400 \text{ opp/}_{\text{sec}}$$

11. The average frequency of occurrence of the above event = item 9.x item 10 = $.064 \times 400 = 25.6$ times/sec

This phenomenon can be repeated in a slave on the same valid ID with probability approaching 1 or on some other valid ID.

12. The average number of opportunities of this second kind = 200 (possible positions) x $\frac{1}{10}$ (aircraft in flight) x

$$\frac{.048 \text{(secs)}}{2.5 \text{ (secs)}} = .385$$

13. The total number of opportunities for an extra pulse to produce the same false ID in a slave for each opportunity in the master = 1 + .385 = 1.385.

14. The probability of producing the same false ID in the same channel (of a slave) during a 48 ms. period centered about the time of the false ID in the master = item 3 x item 13 =

$$3.2 \times 10^{-4} \times 1.385 = 4.43 \times 10^{-4}$$

*Autonetics Final Report "Study and Concept Formulation For A Fourth Generation Air Traffic Control", July 1971.

15. The probability of the above event occurring in all three slaves =

$(\text{item } 14)^3 = 8.71 \times 10^{-11}$

16. The false alarm rate per channel due to single overwriters is item 15 x item 11 =

$8.71 \times 10^{-11} \times 25.6 = 2.24 \times 10^{-9}$

17. The system false alarm rate due to two causes; the combination of noise and user pulses and the single overwrite on valid ID's = item 8 + 100 item 16 =

$1.76 \times 10^{-12} + 100 \times 2.24 \times 10^{-9} = 2.24 \times 10^{-7}$

The acceptance of two possible identities during single overwrites increases the false alarm rate (over a system which disallows overwrites to be considered) but reduces the target dismissal rate.

18. The time between false alarms =

$$\frac{1}{2.24 \times 10^{-7}} = 4.46 \times 10^{6} \text{ secs.}$$

The receiver signal to thermal noise ratio is assumed to equal 16 dB. The mismatched pulse interference at the input to the pulse compression filter has peak power equal to the desired signal peak power at that point = $\frac{S}{200^2}$ where S is the peak post-compression signal power.

On the average, there will be $10^4 \times 2.2$ pulses per second at the input of a bank of 10 filters. Each pulse will be 200 x .1 usec long. The combined interference power at the input of each filter bank will be

$\frac{S}{200^2} \times 10^4 \times 2.2 \times 200 \times 10^{-7} = .0022 \times \frac{S}{200}$. If treated like random noise the average output interference power from each pulse compression filter will be 200 times the above or $.0022S = \frac{S}{450}$.

The signal to <u>total</u> noise ratio will be

$$\frac{S}{\frac{S}{40} + \frac{S}{450}} = 27.2, \text{ or } 14.35 \text{ dB.}$$

It was assumed in the false alarm discussion that the probability of noise crossing the threshold was .0001

Thus for P_n = .0001, $\frac{S}{N}$ = 14.35 dB, Rice's curves give

Pd = 99.9%

In order to detect a user, all 3 x 4 (sat's) = 12 pulses must cross thresholds.

19. Prob. (12 crossings) = Prob $\left[\text{(one crossing)}\right]^{12}$ =

$$\left[1-.001\right]^{12} = 1 - 12 \times .001 = .988$$

20. Prob (failure to detect a valid user ID due to failing to cross a threshold) = .012

Failure to detect due to more than one overwrite pulse occurs when 2 or more valid user ID's coincide and when random pulses occur.

21. The probability of 2 users coinciding on one channel of one satellite receiver =

$$\frac{10^3 \times 100}{2.5 \text{ sec}} \text{ ms} = \frac{10^{-4}}{2.5}$$

22. The prob. of the above in at least one of 4 sat receivers = 4 x (item 21) = .00016

23. The prob. of more than 1 overwrite of a valid ID in the master receiver = 1- prob (no overwrites) - prob (exactly one overwrite)

24. The probability of no overwrites = $\left[\text{1-prob (pulse at particular position)}\right]^{200} = \left[1- 3.2 \times 10^{-4}\right]$ = .936

25. The probability of exactly one overwrite =

$$\binom{200}{1}\left(3.2 \times 10^{-4}\right)^{1}\left(1-3.2 \times 10^{-4}\right)^{199} = 200 \times 3.2 \times 10^{-4}$$

$$\times .936 = .06$$

26. The probability of more than one random overwrite of a valid ID = 1 - item 24 - item 25 = 1 - .936 - .06 = .004

27. The probability of the above event occurring in at least one of the 4 satellites = 4 x item 26 = .016

28. Target dismissal = item 22, (coincidence of users) + item 27, (more than one overwrite) + item 20, (failure to cross thresholds = 2.8%

29. If it is assumed that all overwrites are detected and dropped out, the probability that a false position is calculated is as determined by Autonetics, 1.76×10^{-12}, but the target dismissal rate is increased to 4 (1 - item 24) + item 20 = 26.8%. While this false alarm rate will not tax the position determining computer, the 6% overwrites (1-item 24) will necessitate a small increase in the capacity of the preprocessor or associative memory.

STUDY PAPER L

THE EFFECT OF FALSE ALARMS ON DATA PROCESSING USING A MODIFIED AUTONETICS 2-PULSE SYSTEM

J. DUMANIAN

Autonetics has proposed a 2-pulse aircraft identity transmission system where aircraft will transmit to satellite receivers; L-band emissions, which will be relayed to ground stations. This change from a 3-pulse system to a 2-pulse system is aimed at reducing multi-access code noise. The first pulse will be on one of ten frequencies and using one of 10 pseudo-random codes. The codes will be 400 chips long, .1 us per chip; thus, each pulse will be 40 usecs long. The second pulse will be spaced 70 to 80 usecs from the first and its code will be from another set of 10 400-chip codes. There will, as with the first pulse, be a choice of ten frequencies. Thus, in a 160 usec period, 10^6 possible aircraft can be uniquely coded.

A Pulse			B Pulse				
Freq.		Codes	Freq.		Codes	possible A-B pulse spacings	
10	X	10	10	X	10	100	$= 10^6$ A/C ID's

On detection, each aircraft could be listed in a 23-bit listing. This comes from allowing 4 bits each to code frequencies and PN codes for each of the A and B pulses and 7 bits for pulse spacing.

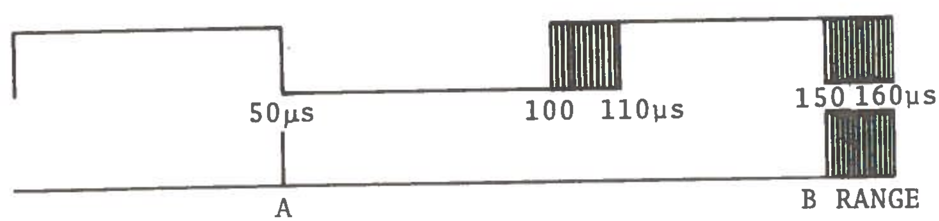
From each satellite, there will be individual streams of 2-pulse trains. Assuming a 2-pulse train is emitted from each of 100,000 aircraft every 2 secs for surveillance purposes and 99 pulses of communication data every 100 seconds, then there will be 1+1 = 2 pulses per second being transmitted per satellite to ground installations for processing. For this analysis, it is assumed that these pulses are 50 microseconds long, coded with a 500-chip truncated PN code. The spacing between the two pulses is between 50 and 60 microseconds, i.e., at .1 usec spacings. The leading pulse will be coded from a group of 10 500-bit codes which were truncated from 511-bit PN codes; the trailing pulse will be coded from another group of ten. Codes other than PN codes can be considered. For communication, a separate set of 10 codes will be used for the leading pulse and a different set of 10 codes for the trailing pulse; this makes 40 codes in all. There are 48 available 511-bit PN codes to select from. Using 40 codes simplifies the surveillance and communication message sorting and reduces the false alarm rate. An interesting comparison is that with this system, 100% of the baud time will be filled at the input to the bank of analog matched filters; with the 3-pulse Autonetics system only 44% were filled, but with the RCA 5-pulse system, 400% were filled.

To implement a detection system for surveillance, it is proposed that 10 counter register channels be used to service each satellite channel. These counter and register channels will appear as shown below:

COUNTDOWN COUNTER	COUNT-UP COUNTER
1000 COUNT	100 COUNT

A FREQ REG.	A CODE REG.	B FREQ REG.	B CODE REG.	POSITION REG.	TOA REG.
4 BITS	4 BITS	4 BITS	4 BITS	7 BITS	20 BITS

Since there are 100,000 airborne aircraft emitting 2 pulses per 2 seconds (one A, one B), the average rate of detection of A pulses is one per 20 usecs. The format of the A and B pulse train is shown below:



The A pulse becomes quantized as a .1 us pulse on passage through an analog matched filter, the B pulse becomes quantized as a .1 usec pulse in a similar manner 100 to 110 microseconds later. As soon as an A pulse is quantized, a selected available counter-register channel, which has been waiting for this event, is contacted and the frequency number (1-10) is stored in the A pulse frequency register. The A pulse PN code number is stored in the A Code Register and a busy bit is indicated. At the same time, the channel 1000-count countdown counter is allowed to count down .1 usec clock pulses. The next available counter register channel is given ready status. When the countdown counter reaches zero, the count-up register is started and another busy bit is raised to the "1" value. The count-up counter allows for the acceptance of all B pulses during 10 microseconds following 100 microseconds after the A pulse. During the time when the count-up counter is operating, incoming B pulses result in frequency numbers, B code numbers and time of arrival data being stored in the appropriate registers of all the counter-register channels which are alerted to B pulses. The A and B pulse information are then fed to an associative memory which will match data from four satellite channels. As soon as the A and B registers and time-of-arrival data are transferred, the counter-register channel is cleared and placed on available status. Ten stages of counting are used in the countdown counter and 7 in the count-up counter. A central "clock" counter supplies the time-of-arrival bits whenever needed. The central clock counter counts 100 MHz clock pulses and requires about 42 stages.

The determination of 10 counter--register channels is based on the average arrival rate of A pulses, 1 every 20 microseconds and the usage time per counter-register channel of 110 microseconds, which specifies a need for 5 1/2 counter-register channels. Ten was selected to allow for bursts of data. This simple configuration has been assumed to form a basis for making an error analysis.

The false alarm calculations are as follows:

1. The problem of a B pulse occurring in a particular 100 nanosecond interval for 1 frequency (but allowing for any of 10 codes) is

$$\frac{10^5 \text{ (A/C)}}{100 \text{ (positions)} \times 10 \text{ (freq)}} \times \frac{1}{2 \text{ (secs)}} \times 10^{-7} \text{ (secs)} = 5 \times 10^{-6}$$

2. Assume that the probability of noise appearing like a B pulse under the same set of circumstances, i.e., from 10 channels is 10^{-5}.

3. The combined probabilities of 1 and 2, 1.5×10^{-5} represent the false representation of a B pulse in a .1 usec interval of a freq-code channel.

4. The probability of an A pulse appearing in a .1 usec interval of a particular freq-code channel is

$$\frac{10^5 \text{ A/C}}{100 \text{ (positions)}} \times \frac{1}{2 \text{ (secs)}} \times 10^{-7} \text{ (secs)} = 5 \times 10^{-5}$$

5. The probability of an incorrectly outputted A-B pair in a freq-code channel is

$$\underset{\text{A pulse and noise}}{(5 \times 10^{-5} + 10^{-5})} \quad \underset{\text{B pulse and noise}}{(1.5 \times 10^{-5})} \quad \underset{\text{positions}}{(100)} = 9 \times 10^{-8}$$

6. The probability of outputting the same false target in a master channel and 3 slave channels (from 3 other satellites) during 48 milliseconds (the maximum time for all 4 TOA's to arrive at the ground) is

$$9 \times 10^{-8} \left[9 \times 10^{-8} \times \frac{58 \times 10^{-3}}{10^{-7}} \right] = 7.29 \times 10^{-12}$$

7. The probability of the system outputting a false target is

$$100 \times 7.29 \times 10^{-12} = 7.29 \times 10^{-10}$$

8. The system False Alarm Rate is

$$10^7 \times 7.29 \times 10^{-10} = 7.29 \times 10^{-3} \text{ FA/sec}$$

or, the expected time between false alarms is approximately 2 minutes.

9. In 48 milliseconds, each freq-code channel will produce 24 actual targets; obtained from $\frac{48 \text{ msec}}{.1 \mu\text{sec}}$ X item 4. In a corresponding period, only .00432 false targets will be generated; obtained from $\frac{48 \text{ msec}}{.1 \mu\text{sec}}$ X item 5. After the 48 millisecond interval, the preprocessor or associative memory transfers the sorted aircraft ID's to the position determining computer. Neither the preprocessor nor the position determining computer is taxed by the false data presented to them.

STUDY PAPER M

JAMMING SATANS AND OTHER ATC SATELLITE SYSTEMS AND THE EFFECT ON DATA PROCESSING

J. DUMANIAN

It must first be observed that for large budget jammers, e.g., major nations, all projected systems for air traffic control are vulnerable. Even back-up systems can be overburdened by action from formidable opponents. This type of action which would be certainly destructive of human lives would only be initiated when an act of war is entertained.

Consideration should be made, however, of the low budget jammer; for example, a student on drugs, a disgruntled ex-controller, etc. Ways should be devised to keep the action of the low budget jammer from over-taxing the ATC system.

Consider an Air/Satellite/Ground Time-of-Arrival Surveillance System, e.g., SATANS. At peak times, 100,000 aircraft will transmit their ID code for 100 usec at 2.5 sec intervals. This means that 10 secs of data are electrically summed during a 2.5 sec time period on satellite receiving antennas and the resultant signals are rebroadcast to the ground station. Here the 400% workload is divided among 10 channel filters; 40% per filter. Subsequent compression or resolution results in reducing the 20 usec pulses (five of which make up the 100 usecs mentioned earlier) to .1 usec pulses. This is a 200 to 1 factor. This allows the interlacing of data on 10 channels to be easily processed by a single computer. In other words, in 10 successive .1 usec intervals, 10 aircraft ID codes with time tags can be stuffed in a buffer register memory, prior to the matching and computational operations.

Now assume that a low budget jammer has built an L-band transmitter and is transmitting signals from the ground with an average power which is comparable to that of the peak power of the aircraft transmitter, 3.5 KW. Or, he could have n (a small number of) transmitters each randomly transmitting at a 1/n X 100% duty cycle, with the same total power as before. That is, in each 20 usec band he is sending signals to the satellite. The question is: What effect will this have on the total system? First, to partially qualify the term low budget, it is required to quote a rule-of-thumb. It is that a transmitter of this type can be purchased and/or modified for a dollar per average watt of power. With the jammer filling each baud with signals, the effect at the point of summation on the satellite receiver antenna of 12.5 secs of data (instead of 10.0 previously noted) during each 2.5 sec interval; this is an increase similar to that of a sudden addition of 25,000 aircraft. It must be assumed that there exists non-limiting circuitry in the satellite and ground equipment to accommodate this new business.

When referred to page 23 of the SATANS Report, Preliminary Analysis of Performance and Cost by Elliot Austein, RCA, the effect of the additional aircraft will be such that the s/50 term, the interference caused by crosstalk among ID signals from 100,000 aircraft, in the denominator of the signal to total noise ratio, will be increased by 25%. The signal to <u>total</u> noise ratio will be reduced to 12.9 dB and the probability of detection (using data from 5 satellites) will become 97%. This means that 3% of the aircraft at any instant in time will not be detected. However, this will be on a random basis and should not visibly affect air traffic control. A more clever maneuver would be a jamming technique which would continually interrogate one of the 10 channels. This would cause a complete loss of 10% of the flying aircraft, i.e., the same aircraft would appear lost continuously. To accomplish this would require knowledge of the pseudo-random coded ID pulse trains, and actually the workings of the SATANS System. The jamming power has been fixed at 3.5 KW for this discussion. However, if the power is increased, the amplifiers in the system can be driven into non-linear regions and the probability of detection of aircraft can be severely reduced.

A satellite system can be designed which shows more immunity to jamming, e.g., a synchronized S/A/G TOA surveillance system. Coded messages identifying each of five satellites are synchronously transmitted by the satellites. The receiving aircraft then transmits the satellite code plus their own identification code to the ground station. This system is superior to the previous system since only aircraft in view of the jammer can be jammed. Thus, the location of those aircraft in view of the jammer would be lost to the ground stations, which may amount to only a few percent.

Among the disadvantages to this scheme are:

1. The need to have synchronized systems on-board five satellites. There is a small cost factor here, but worse still is the need for occasional resynchronization or error correction by signals from the ground station. This opens the door again for the jammer. Actually, jamming the satellite controlling mechamisms, i.e., the control thrusters, can cause harm. A narrow time window and a carefully guarded code can probably assure against a noise jammer interfering with the control mechanisms and also the resynchronization circuits.

2. The bunching of returns to the ground station. The longest segment of the satellite-to-aircraft-to-ground path is the satellite-to-aircraft journey and the greatest variation of this is due to the elliptical satellite path which varies between 16,000 miles and 25,000 miles above the earth in the region of the aircraft. This will cause a bunching of signals to the aircraft between .1 secs and .15 secs after transmittal from the satellites. The variation in slant ranges in the aircraft-to-ground portion of the transmission merely adds a 10% dispersive factor to the above times. To reduce the bunching, the satellite signals can be spaced 1/5 sec apart, i.e., each satellite transmission is staggered with respect to the others. In addition, agile multibeam receiving arrays on the ground can be used to pick up the signals.

STUDY PAPER N

A COMPARISON OF DATA PROCESSING REQUIREMENTS IN FUTURE ATC SYSTEMS

The MIPS rates estimated for System I - Boeings proposed satellite system, System II - Autonetics prpposed hybrid system, System III - version of the upgraded third generation ATC system which uses a satellite-Dabs concept are listed below and an attempt is made to correlate the results.

SYSTEM I DATA PROCESSING RATE ESTIMATES

Function	MIPS
Surveillance	190.52
Satellite Tracking & Control	16.00
A/G Communications	19.71
Flight Plan Processing & Flow Control	36.40
Flight Service Station Terminals	157.40
Weather & Facilities	13.48
Aircraft Control	92.88
Collision Avoidance	130.32
Display Terminals	38.34
	681.57

SYSTEM II DATA PROCESSING RATE ESTIMATES

Function	MIPS rate/cntr	No. of cntrs	Total MIPS
Surveillance Processing			96
Communications Processing			15
Control Processing			212
			323

The 96 MIPS for Surveillance Processing is accounted for by .80 MIPS at the Enroute Centers and 3 MIPS at the Major Hubs and .66 MIPS at the Minor Hubs.

Some of the items which would normally be expected to consume data processing power, e.g., displays, are included but others are not; thus, this list is incomplete.

SYSTEM III DATA PROCESSING RATE ESTIMATES

The analysis of this system is based on the ATCAC data processing estimates.

The L.A. basin estimates as given in the article by Blake and Nelson in the IEEE Journal of March 1970 are extrapolated for the sizing given below which was suggested by Neal Blake of the FAA.

In the L.A. example, 8,000 instantaneous airborne aircraft are divided into 3,800 enroute controlled and 4,200 terminal controlled with 42.7 MIPS required in the former case for all enroute ATC functions and 48.5 in the latter case for all terminal ATC functions. Assuming that the 100,000 instantaneous airborne A/C for demand level 3c are divided into 3,800/8,000 X 100,000 = 47,500 enrouted controlled aircraft and 4,200/8,000 X 100,000 = 52,500 terminal controlled. The total MIPS rate then equals 42.7 X 47,500/3,800 + 48.5 X 52,500/ 4,200 = 1,141 MIPS. This figure includes a factor of 2 increase for instruction mix degradation, a factor of 2 increase for executive overhead and a factor of 1-1/2 increase for high-level language inefficiency, for a total increase factor of 6.

Under similar ground rules, Autonetics (actually UNIVAC) estimated approximately the same number of MIPS for an earlier proposed satellite system and Boeing had 3,600 MIPS in earlier estimates for their satellite system. Since then, both Autonetics and Boeing have reduced their estimates, mainly by assuming that the programs for the most part would not be written in high-level languages.

OTHER OBSERVATIONS

Boeing actually scales the "raw" or non-degraded individual program MIPS rate given in the article Blake and Nelson: Future ATC Data Processing Requirements, p. 394-399, March 1970, IEEE Proceedings e.g., CAS, .25 MIPS in the enroute area, etc. Boeing uses data processing estimates from this source and scales up by a factor of 2, 4, or 6; the reasoning behind this is that often used programs can be written in machine language.

Autonetics, in Vol. III pp. D-192, D-193, D-194 of their final report on Fourth Generation Air Traffic Control Study, Nov. 1971, tries to equate the ATCAC data processing estimates for 1990 time period with their own. They used 2 examples. In one, their MIPS rate was 73% of the ATCAC figures; in the other, their MIPS rate was 230% of the ATCAC figures.

See the chart below for a comparison of the data processing estimates for the surveillance and control functions for System I and II.

COMPARISON OF DATA PROCESSING IN SYSTEM I & SYSTEM II

ATC FUNCTION	System I Rate MIPS	System II Rate MIPS
Surveillance	190.52	96
Control (including flow, CA)	269.60	212
Communications	19.71	15

Obviously, there are differences. However, the differences are only partially accounted for. For example, the Autonetics system is designed to have instantaneous acquisition, i.e., a tracking history does not have to be built up to declare targets; thus, data processing is not required for this function. Boeing has proposed a hardware tracker, but their signal design is such that longer queues of data have to be sorted to match signals from different satellites, which does require more data processing.

CONCLUSIONS

The data processing estimates for Systems I, II and III are different, but so are the systems. Found in System I and System II are some features aimed at reducing the data processing workloads. These bear confirmation from analytical and experimental bases. Failure to initiate common ground rules at the time of the initial design of Systems I and II has reduced the credibility of post-comparisons.