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Secure Communications Systems Utilizing Pseudo-Noise Carriers and Sub-carriers

Neff, Rupert

Monterey, California. Naval Postgraduate School

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SECURE COMMUNICATIONS SYSTEMS UTILIZING PSEUDO-NOISE CARRIERS AND SUB-CARRIERS (U)

by

Rupert Theodore Neff

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May 1966

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SECURE COMMUNICATIONS SYSTEMS UTILIZING PSEUDO-NOISE CARRIERS AND SUB-CARRIERS (U)

by

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Submitted in partial fulfillment
for the degree of

MASTER OF SCIENCE IN COMMUNICATIONS ELECTRONICS

from the

UNITED STATES NAVAL POSTGRADUATE SCHOOL

May 1966

Signature of Author [redacted]
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Approved by: [redacted]
Academic Dean

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ABSTRACT

The current literature approaches the subject of communications from the optimum detector point of view, utilizing the principles set forth by Shannon, Davenport, Root, Bode, Peterson, Birdsall, Fox, Weiner, Siebert, Middleton, and many others to carry out detailed analysis of specific detectors, filters, and synchronization processes. This paper approaches communications from a systems point of view, dealing specifically with the family of pseudo-noise systems. The systems discussed are categorized into two groups, the pseudo-noise carrier and the pseudo-noise sub-carrier systems, with emphasis on multiplex techniques. All of the systems discussed are negative dB S/N systems with the exception of a wide-band TV video channel, and a detailed analysis of a representative pseudo-noise sequence of length 15 is given as a background in the auto-correlation functions, power spectral densities, and self-noise spectra which are the characterizing parameters of these waveforms.

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LIST OF ILLUSTRATIONS
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SYMBOLS AND ABBREVIATIONS

X	Product Device
+	Adder
-	Subtractor
Σ	Summer
$\oplus$	Mod 2 Logic
	NOR Logic
90	90° Phase Shift
AM	Amplitude Modulation
AMDT	Amplitude Detector
A_i	Modulating Waveform
B	Double Sided Carrier Bandwidth
BPF	Band Pass Filter
β	FM Modulation Index
CODT	Coherent Detector
dB	$10 \log_{10}$ (power ratio)
DEL	Delay Line
Δf	FM Maximum Frequency Deviation
DISC	Discriminator
ϵ	Base of Natural Log
E_i	Partially Processed Signal
f_o	Clock Frequency
f_{max}	Single Sided Audio Bandwidth
FF	Flip-Flop
FM	Frequency Modulation
GT	Gate

SYMBOLS AND ABBREVIATIONS
(Continued)

HOR	Horizontal Sweep
INV	Inverter
k	Number of Flip-Flop Stages
LIM	Limiter
LPF	Low Pass Filter
M_i	Levels of Sequence i
N	Noise Power
N_i	Length of PN Sequence i
NBF	Narrow Band Filter
PN	Pseudo-Noise
PNG	Pseudo-Noise Generator
P_i	Output of i'th Flip-Flop
P_i^*	Inverted P_i
$P(\omega)$	Power Spectral Density
$R(t_o)$	Autocorrelation Function
ρ	Correlation Coefficient
S	Signal Power
s	Transmitted Signal Waveform
T	Period of Sequence
τ	Basic Pulse Width
VCA	Voltage Controlled Amplifier
VCO	Voltage Controlled Oscillator
VERT	Vertical Sweep
ω_i	$2\pi f_i$
$X(\omega)$	Fourier Transform of x(t)
$\overline{Z}$	Inverted Z

1. Introduction.

In the general communications system the basic limitation is S/N ratio at the output of the receiver. For a desired output S/N of the receiver the required input S/N is determined by the following relationship:

$$(S/N)_{in} = K(B_{info}/B_{carrier})(S/N)_{out}$$

In the conventional receiver, the information channel is modulated directly onto a sinusoidal carrier by AM or FM methods and the transmitted carrier bandwidth is essentially the same as the information bandwidth. Therefore, the receiver processing output S/N is the same order of magnitude as the input S/N, K being a constant with value below 10. If the detection process is coherent, K is smaller by 3 to 5 dB over the non-coherent detection process. However, there is not an appreciable detection gain in either process since the information and carrier bandwidth are essentially equal.

To circumvent the above limitation, one can perform one of two operations on the transmission system. The carrier can be made much wider in bandwidth than the information channel, or a wide-band sub-carrier can be modulated by the information channel, and used in turn to modulate the carrier which also results in a wide-band transmission system. For both of these techniques, the PN, or Maximal Length Shift Register Sequence is ideal, and simple to implement. The following discussions and analyses consider wave-form and system variations all of which are designed to fulfill the one primary objective, bandwidth compression at the receiver to make possible sub-noise reception of the transmitted signal.

The coding of the signal is of secondary importance and shorter sequences in the multiplex systems are, in fact, required.

2. The Single PN Channel.

When considering the general PN channel there are three separate basic systems encountered. There is the PN carrier, AM and FM cases; the PN sub-carrier AM, coherent and non-coherent cases; and the PN sub-carrier FM system.

For short range, line of sight and cable transmissions, the PN carrier can be used and is limited to these applications. The basic AM and FM channel, the variations of which will be discussed in Section 6, is given in Figure 1. Note the wideband signal is compressed in the first stage of detection in both cases. This will also be true in all of the non-coherent systems utilizing the PN code as a sub-carrier. In these systems bandwidth compression must take place prior to the input of the amplitude detector to yield an SN greater than 1 at the detector input. Sub-noise signals cannot be amplitude detected direct without some type of processing gain prior to the detector input.

The basic systems for the PN sub-carrier applications are shown in Figure 2 for the coherent and non-coherent cases. Variations of these will be discussed in Section 7. The information modulation technique is not restricted to AM-DSB-SC or conventional AM but the FM variations are left for the later sections.

In the non-coherent case, the S/N meets the requirement: $(S/N)_{in}(1/\pi)(B_{carrier}/B_{info}) > 1$. Since, for voice and NTDS data systems, the information channel f_{max} is within the 4 KHz range,

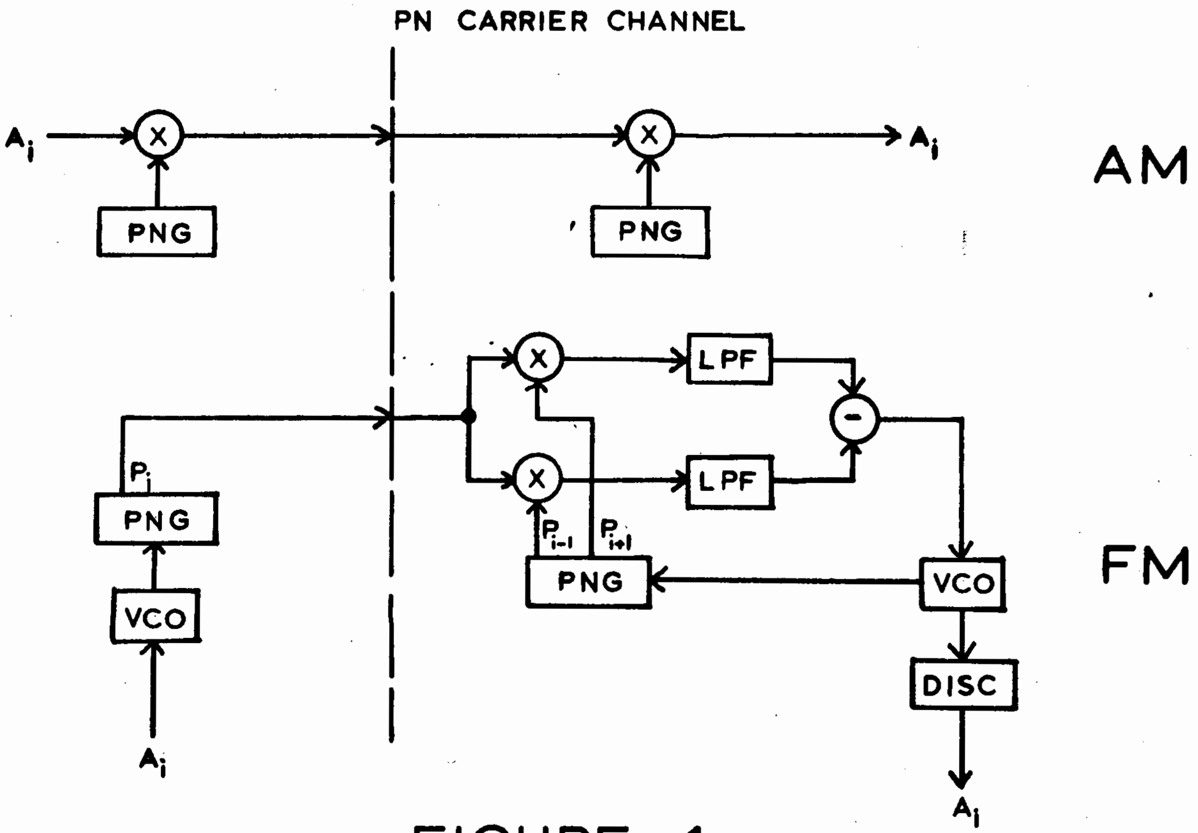

PN CARRIER CHANNEL
A_i
X
PNG
X
PNG
A_i
AM
X
LPF
X
LPF
P_i
PNG
VCO
A_i
P_{i-1}
P_{i+1}
PNG
VCO
DISC
A_i
FM

FIGURE 1

and a typical $B_{carrier}$ is 2 MHz (double sided), $(1/\pi)(10^6/4000)$ = 80.5 or 19 dB. Therefore, with no additive receiver noise, the system processing gain is 19 dB, and the S/N input need only be above -19 dB for amplitude detection of the signal.

The non-coherent system in Figure 2 is also the basic system for PN sub-carrier FM systems as the carrier input at the transmitter can be replaced by an FM signal. The amplitude detector is replaced by a discriminator, and approximately the same processing gain is realized.

The system equations for the coherent cases are the conventional coherent detector equations. With the input, $S(t) = P_i \cos \omega t$, the product device output yields $E_i = P_i \cos^2 \omega t = (1/2)P_i \, (1+\cos 2\omega t)$. Since ω is much larger than $2\pi f_o$, low pass filtering with cutoff at f_o yields the raw PN waveform of a single-sided bandwidth f_o which, upon balanced modulation with P_i and low pass filtering, yields DC or a low frequency information channel. The processing gain in this case is larger by the factor of π since $K = 1$, and no amplitude detection loss is present in the system. A processing gain of 24 dB is thus realized.

The non-coherent system equations for the PN sub-carrier systems are similar for the AM and FM systems. The PN phase shifting in both cases is much faster than the amplitude or frequency variations. The carrier phase need not be known but the PN sub-carrier phase is known by correlation synchronization techniques discussed in Section 5. The transmitted signal may be represented by the conventional expressions. For the AM case, $S(t) = P_i \, (1+mA_i)\cos \omega t$, where m is the AM modulation

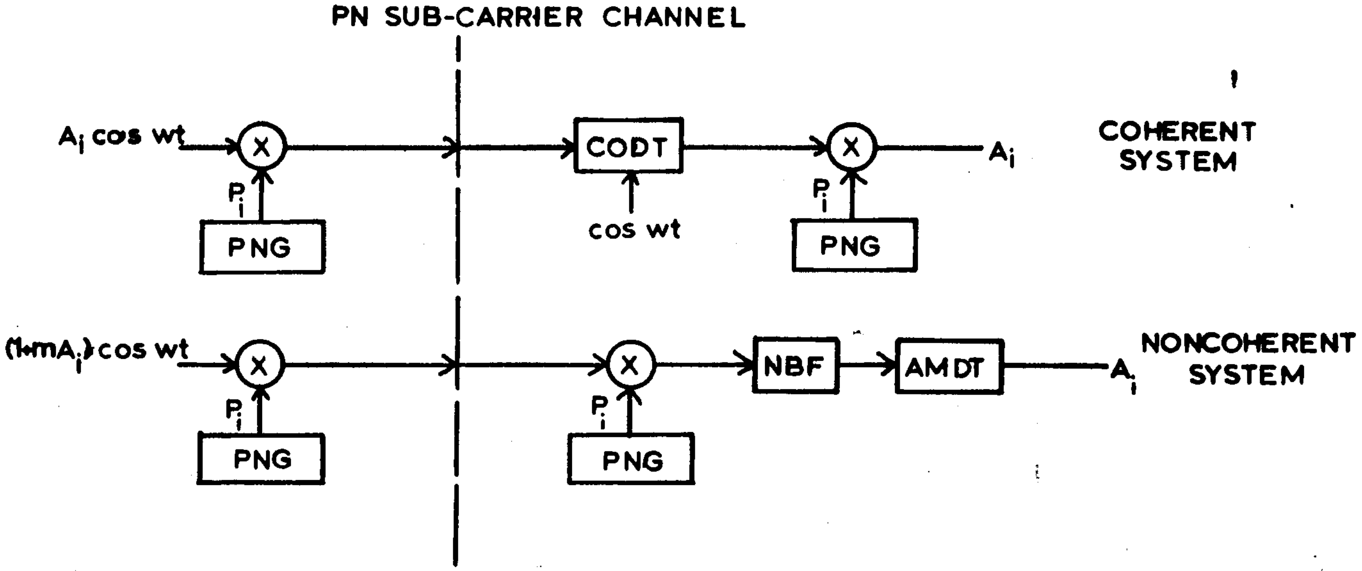


FIGURE 2

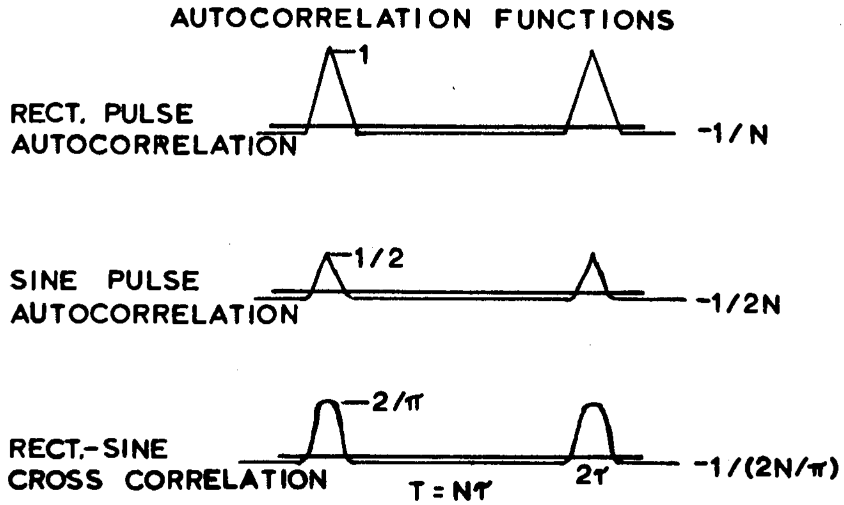


FIGURE 3

index. Since at synchronization $P_i P_i = 1$, the output of the first balanced modulator is a comparatively narrow bandwidth output, $E_i = (1+mA_i)\cos \omega t$ which is amplitude detected. For the FM case the transmitted signal is $S(t) = P_i \cos \omega_A t$, where ω is a function of A, the slowly varying information channel. A more detailed analysis of the information channel frequency restrictions with respect to the PN sequence length and clock frequency will be given in Section 6.

The above equations express the objectives and the final desired relationships expected at synchronization. The detection process in the coherent case performs a bandwidth compression operation on the raw PN waveform, whereas this compression operation is performed on the carrier in the non-coherent case. The cross-coorelation function of the local PNG output and the incoming signal is maximum at synchronization, and optimum matched filter detection results for these systems. However, to acquire and maintain this synchronization is a critical requirement in the PN systems to be discussed. Before delving into system specifics, the characteristics of the waveform to be used must be studied as these will dictate the parameters of the optimum filters to be used, the optimum synchronization techniques, and in general, the receiver will be designed optimally for this waveform in the predetection stages.

3. PN Waveform Analysis.

The Pseudo-Noise sequence, or sometimes called the maximal length shift register code, is one code in the family of four codes called Simplex Codes. The other three, quadratic residue, Hall, and twin prime sequences will not be dealt with in this paper but have the same characteristics as (2) and (3) given below for the PN sequences, where $P = \pm 1$:

(1) $N = 2^k - 1$

(2) $\rho(P_i P_j)_{i \neq j} = (1/N) \sum_{\substack{i=1 \\ i \neq j}}^{N} P_i P_j = \underset{i \neq j}{\text{Average}} \; \rho(P_i P_j) = \begin{cases} -\frac{1}{(N-1)} & N \text{ even} \\ -\frac{1}{N} & N \text{ odd} \end{cases}$

(3) $\rho(P_i P_j)_{i = j} = 1$

Note the correlation coefficient values given above are merely the values of the autocorrelation function at each discrete time shift of the basic sequence. The time shift distances are therefore $n\tau$ where $n = 1, 2, 3, ----N$. A fourth characteristic also bears mentioning and is often applied in practical systems. If the product $P_i P_j^*$ is taken, the resulting sequence will be a time shifted version of the original sequence and the three sequences form a closed triad where any product of one and the conjugate of another yields the third sequence.

The PN code is also near orthogonal if the time shifted cyclic sequences are considered separate words. Since, for the PN sequence $\rho_{i \neq j} = -(1/N)$, the cross-correlation values are small compared with the autocorrelation peak value of 1. This characteristic suggests many applications to multiplex systems.

When considering the autocorrelation function in general under certain circumstances it is much easier to construct an equivalent convolution integral to determine the autocorrelation function. In the development below, if $X^*(\omega) = \pm X(\omega)$, the convolution integral is seen to be equivalent to the autocorrelation integral.

$$R_{conv}(t_o) = \frac{1}{T}\int_{-\frac{T}{2}}^{\frac{T}{2}} x(t)\,x(t_o-t)\,dt = \frac{1}{2\pi T}\int_{-\frac{T}{2}}^{\frac{T}{2}} x(t)\left\{\int_{-\infty}^{\infty} X(\omega)\,\epsilon^{j\omega(t_o-t)}\,d\omega\right\}dt$$

$$= \frac{1}{2\pi T}\int_{-\infty}^{\infty} X(\omega)\,X(\omega)\,\epsilon^{j\omega t_o}\,d\omega$$

$$R_{corr}(t_o) = \frac{1}{T}\int_{-\frac{T}{2}}^{\frac{T}{2}} x(t)\,x(t_o+t)\,dt = \frac{1}{2\pi T}\int_{-\frac{T}{2}}^{\frac{T}{2}} x(t)\left\{\int_{-\infty}^{\infty} X(\omega)\,\epsilon^{j\omega(t+t_o)}\,d\omega\right\}dt$$

$$= \frac{1}{2\pi T}\int_{-\infty}^{\infty} X^*(\omega)\,X(\omega)\,\epsilon^{j\omega t_o}\,d\omega$$

Therefore, the properties of the convolution integral will be used to calculate $R(t_o)$ for the waveforms to be considered. In Figure 3, three correlation functions are shown, the calculations of which are shown below:

(1) Rectangular Pulse PN Waveform

$$R(t_o) = \frac{1}{2\tau}\int_{-\tau}^{\tau}\left[u\left(t+\frac{\tau}{2}\right)-u\left(t-\frac{\tau}{2}\right)\right]\left[u\left(t_o-t-\frac{\tau}{2}\right)-u\left(t_o-t+\frac{\tau}{2}\right)\right]dt = \begin{cases} \dfrac{N-\frac{|t_o|}{\tau}(N+1)}{N} & |t_o|\le\tau \\ 0 & |t_o|>\tau \end{cases}$$

(2) Sine Pulse PN Waveform

$$R(t_o) = \begin{cases} \dfrac{1}{2\tau}\displaystyle\int_{-\tau}^{\tau}\cos\frac{\pi t}{2\tau}\cos\frac{\pi}{2\tau}(t_o-t)\,dt = \dfrac{\tau-|t_o|}{2\tau}\cos\dfrac{\pi|t_o|}{\tau}+\dfrac{1}{2\pi}\sin\pi\left(1-\dfrac{|t_o|}{\tau}\right) & |t_o|\le\tau \\ 0 & |t_o|>\tau \end{cases}$$

(3) Sine Cross-Correlated with Rectangular Pulse

$$R(t_o) = \frac{1}{2\tau}\int_{-\tau}^{\tau}\left[u\left(t+\frac{\tau}{2}\right)-u\left(t-\frac{\tau}{2}\right)\right]\cos\frac{\pi}{2\tau}\left(|t_o|-t\right)dt$$

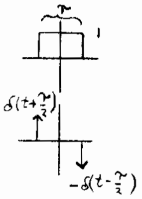
$$= \begin{cases} \dfrac{1}{\pi}\left[1-\sin\dfrac{\pi}{\tau}\left(|t_o|-\dfrac{\tau}{2}\right)\right] & |t_o|\le\tau \\ 0 & |t_o|>\tau \end{cases}$$

As is seen from Figure 3, every complete period shift yields an autocorrelation peak of width 2 cycles and relative height N. These functions are ideal to use in a synchronizing system as a conventional S curve of amplitude versus phase may be constructed over a 4τ interval with a stable null at the center by subtracting or differencing crosscorrelations of the signal with positive and negative time shifted versions of the signal PN.

The method of implementation is discussed in Section 5. However, one can see that such a differencing would yield for case (1), a negative slope at the stable null, but for the other two cases a zero slope results. The latter situation is not desirable at the null as close frequency control with phase variation is required for stability. For cases (2) and (3), this zero slope is eliminated by slightly delaying the correlating waveform associated with the positive peak, and setting the null control voltage slightly above zero for in-synchronization condition. If the sine pulse is normalized to yield DC and low frequency powers equivalent to the rectangular pulse, an amplitude of $\pi/2$ is given to the sine pulses, where the rectangular pulses are of height 1. An advantage of cases (2) and (3) is seen in the rounded peak of normalized height 1, which indicates that the correlation detector will be less sensitive to slight cycle shift than in case 1.

The above discussions have not considered the circuitry involved in sine pulse correlators which are decidedly more difficult to implement. However, the main considerations shown do give the basic relationships which determine receiver detection and synchronization techniques.

For optimum design of filters in the receiving system the signal power spectrum must be known to yield maximum S/N receiver output. The most straightforward method of computing the exact power spectrum for a given PN sequence is by the Fourier Transform. For the simple waveforms considered in the correlation discussion, the transform X(ω) may be calculated by the simple technique as follows:



$$x(t) \Rightarrow X(\omega)$$

$$x'(t) \Rightarrow j\omega X(\omega) = \int_{-\infty}^{\infty} \left[\delta(t+\tfrac{\tau}{2}) - \delta(t-\tfrac{\tau}{2})\right] \epsilon^{-j\omega t}\, dt = \frac{\tau}{j}\left[\frac{\sin\frac{\omega\tau}{2}}{\frac{\omega\tau}{2}}\right]$$

$$P(\omega) = \frac{1}{\tau}\left|X(\omega)\right|^2 = \tau\left[\frac{\sin\frac{\omega\tau}{2}}{\frac{\omega\tau}{2}}\right]^2$$

An extension of this technique yields the envelope of the exact power spectral density of any given PN sequence of basic waveforms such as rectangular or sine pulses. With peak values of ± 1 , the results are given for the PN sequence -1-1-1+1-1-1+1+1-1+1-1+1+1+1+1 .

The rectangular power spectrum is given below:

$$X(\omega) = \frac{1}{j\omega}\left[-1 + 2\epsilon^{-j\omega 3\tau} - 2\epsilon^{-j\omega 4\tau} + 2\epsilon^{-j\omega 6\tau} - 2\epsilon^{-j\omega 8\tau} + 2\epsilon^{-j\omega 9\tau} - 2\epsilon^{-j\omega 10\tau} + 2\epsilon^{-j\omega 11\tau} - 2\epsilon^{-j\omega 15\tau}\right]$$

$$X^*(\omega) = \frac{1}{-j\omega}\left[-1 + 2\epsilon^{j\omega 3\tau} - 2\epsilon^{j\omega 4\tau} + \cdot\;\cdot\;\cdot\;\cdot\;\cdot\;\cdot\;\cdot\;\cdot\;\cdot\; - 2\epsilon^{j\omega 15\tau}\right]$$

$$X(\omega)X^*(\omega) = \frac{1}{\omega^2}\begin{bmatrix} 30 - 32\cos\omega\tau + 4\cos 3\omega\tau - 4\cos 5\omega\tau + 8\cos 6\omega\tau - 12\cos 7\omega\tau \\ + 12\cos 8\omega\tau - 8\cos 9\omega\tau + 4\cos 10\omega\tau - 4\cos 12\omega\tau + 2\cos 15\omega\tau \end{bmatrix}$$

The sine pulse power spectrum is similarly calculated:

$$X(\omega)X^*(\omega) = \left(\frac{\pi}{2}\right)^2\left[\frac{\pi/\tau}{(\pi/\tau)^2 - \omega^2}\right]^2 \begin{bmatrix} 30 + 32\cos\omega\tau + 8\cos 2\omega\tau + 12\cos 3\omega\tau + 8\cos 4\omega\tau \\ + 4\cos 5\omega\tau - 4\cos 7\omega\tau - 4\cos 8\omega\tau - 8\cos 9\omega\tau \\ - 12\cos 10\omega\tau - 16\cos 11\omega\tau - 20\cos 12\omega\tau \\ - 16\cos 13\omega\tau - 8\cos 14\omega\tau - 2\cos 15\omega\tau \end{bmatrix}$$

The spectrum for both cases are plotted in Figures 4 and 5.

Figures 6 and 7 give the relative power of the sidelobes which are considerably greater for the rectangular pulses. The sine pulse spectrum is for the $\pi/2$ peak amplitude and it can be seen in

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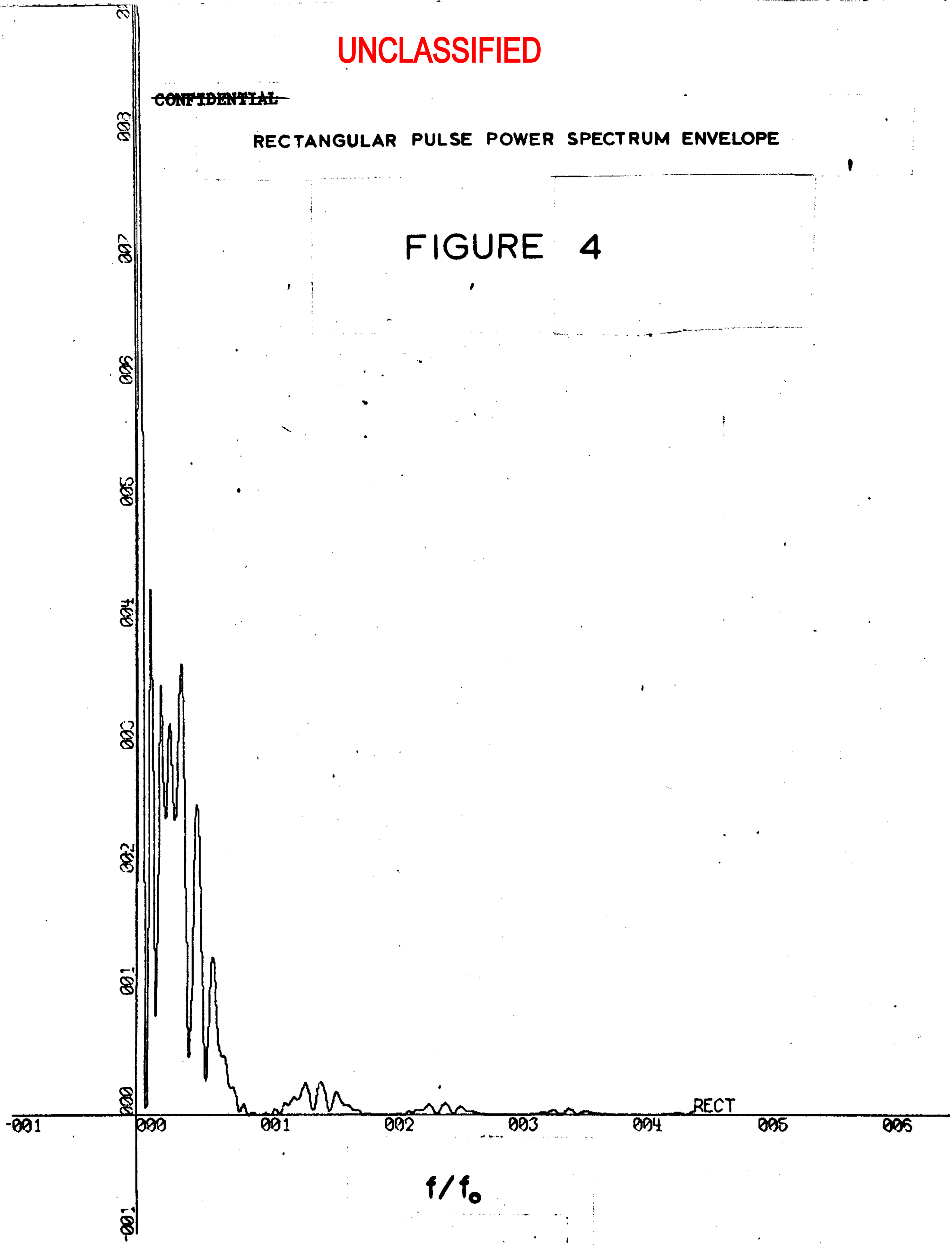
RECTANGULAR PULSE POWER SPECTRUM ENVELOPE

FIGURE 4



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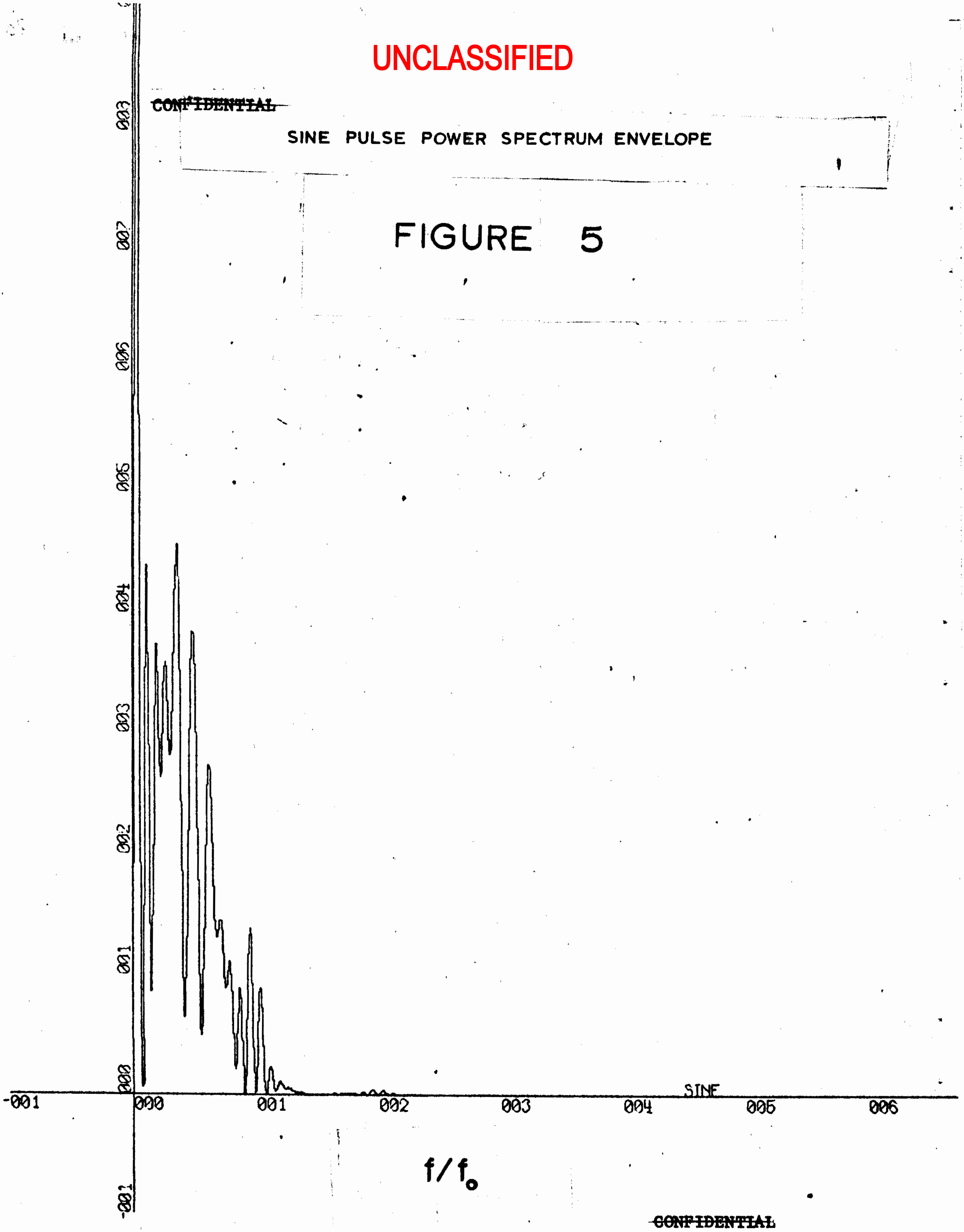
SINE PULSE POWER SPECTRUM ENVELOPE

FIGURE 5



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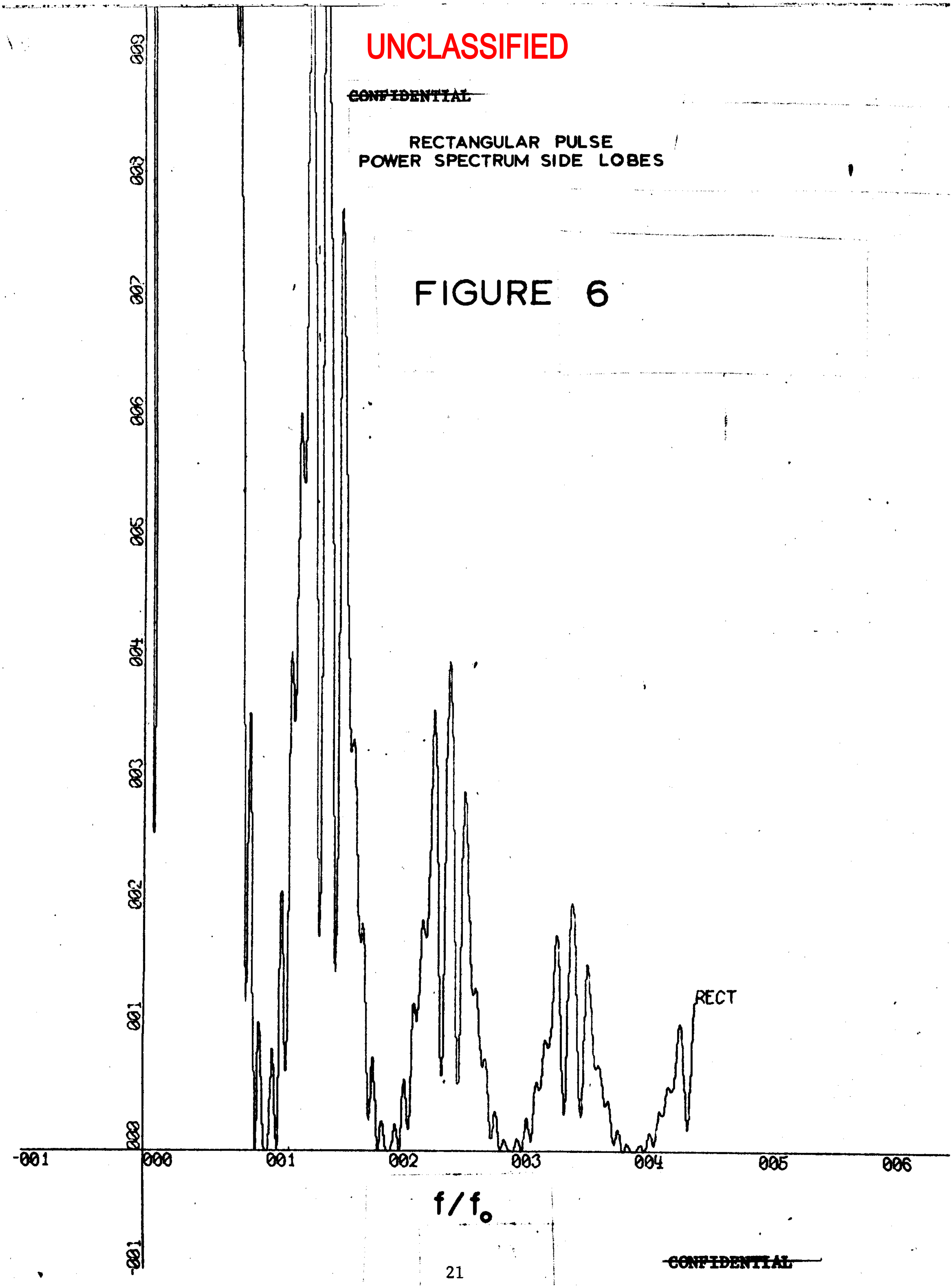
RECTANGULAR PULSE
POWER SPECTRUM SIDE LOBES

FIGURE 6



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SINE PULSE POWER
SPECTRUM SIDE LOBES

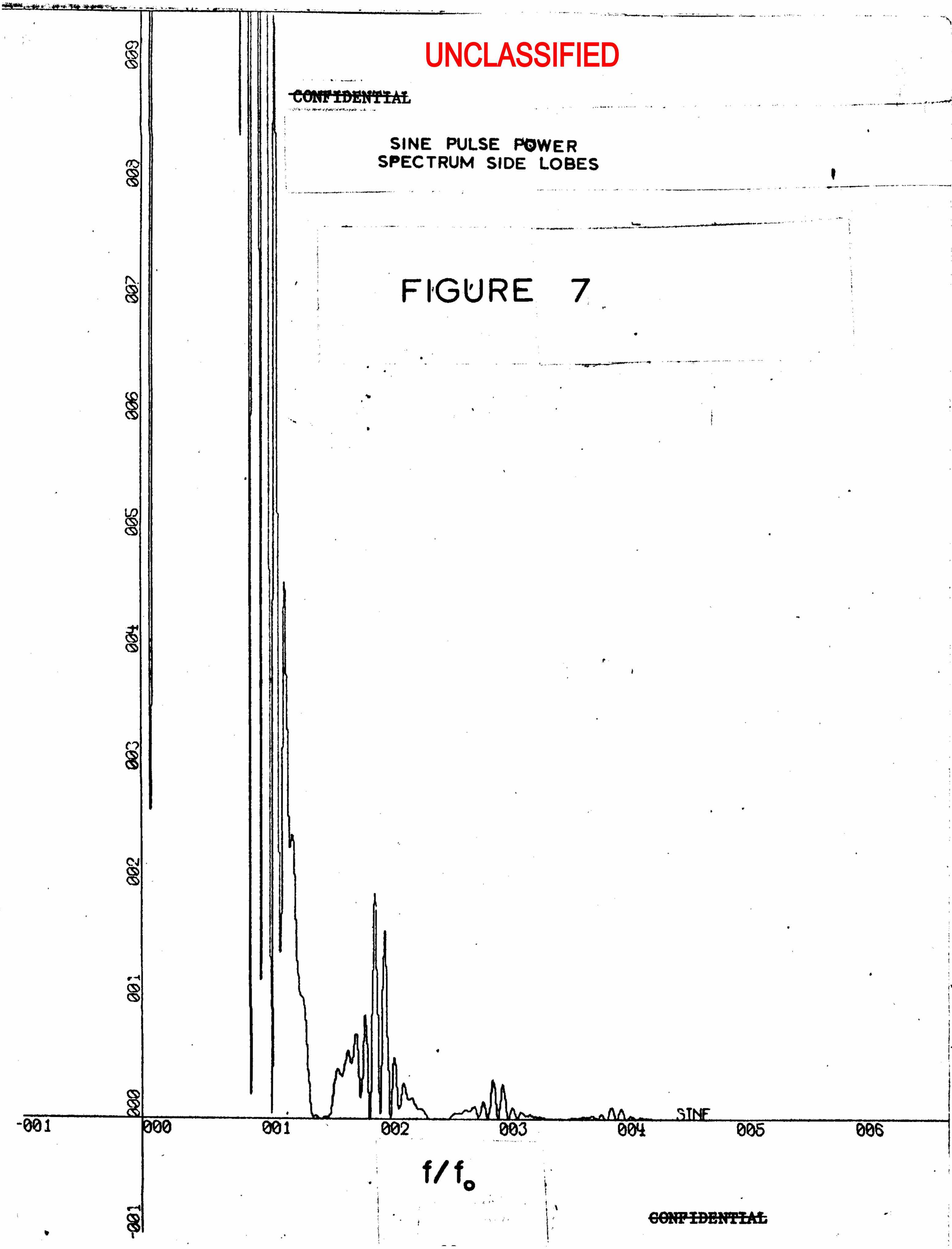

FIGURE 7

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FIGURE 8

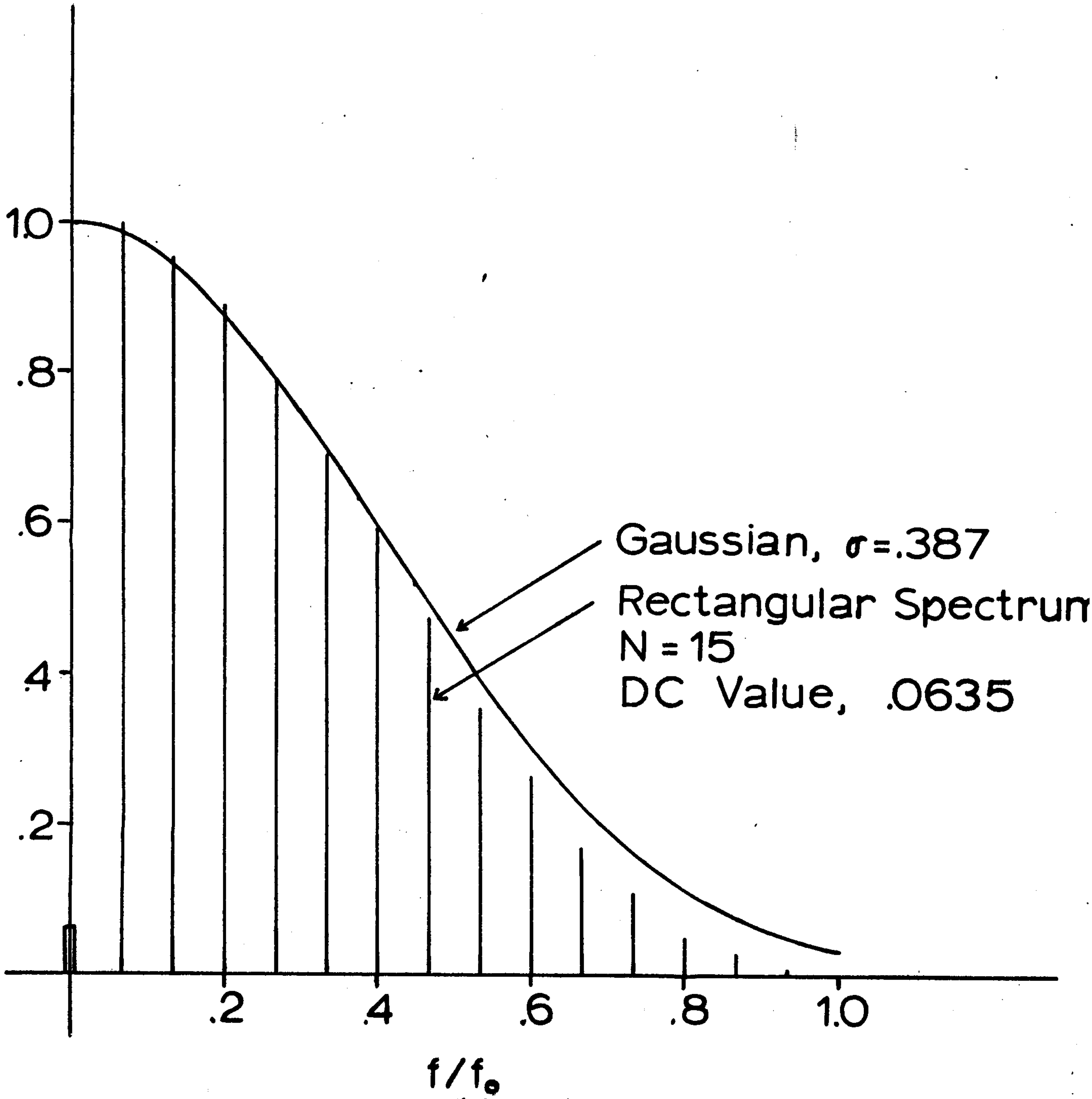

1.0
.8
.6
.4
.2
.2
.4
.6
.8
1.0
f/f₀
Gaussian, σ=.387
Rectangular Spectrum
N = 15
DC Value, .0635

FIGURE 9

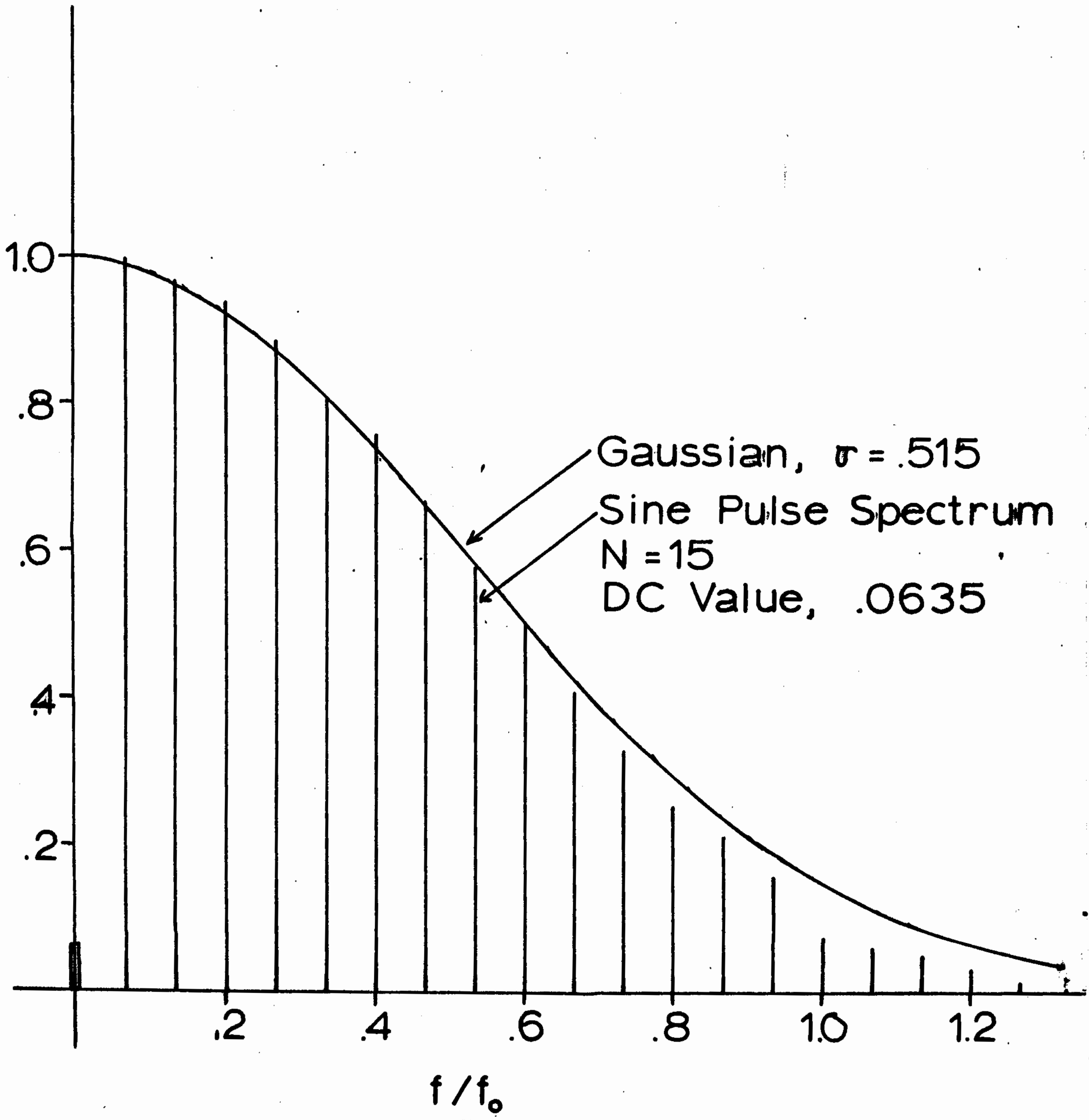

1.0
.8
.6
.4
.2
.2
.4
.6
.8
1.0
1.2
f / f_o
Gaussian, σ = .515
Sine Pulse Spectrum
N = 15
DC Value, .0635

FIGURE 10

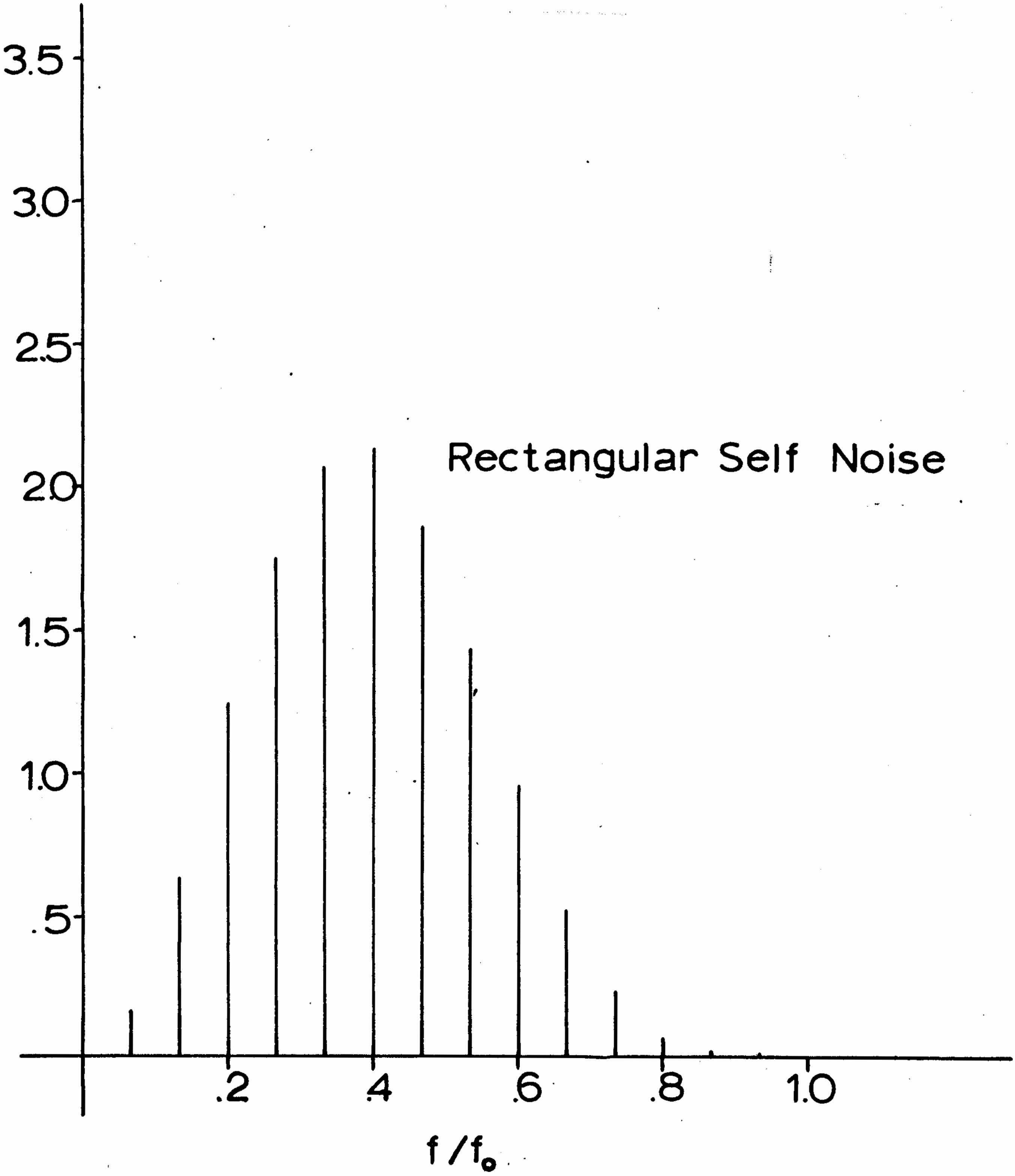
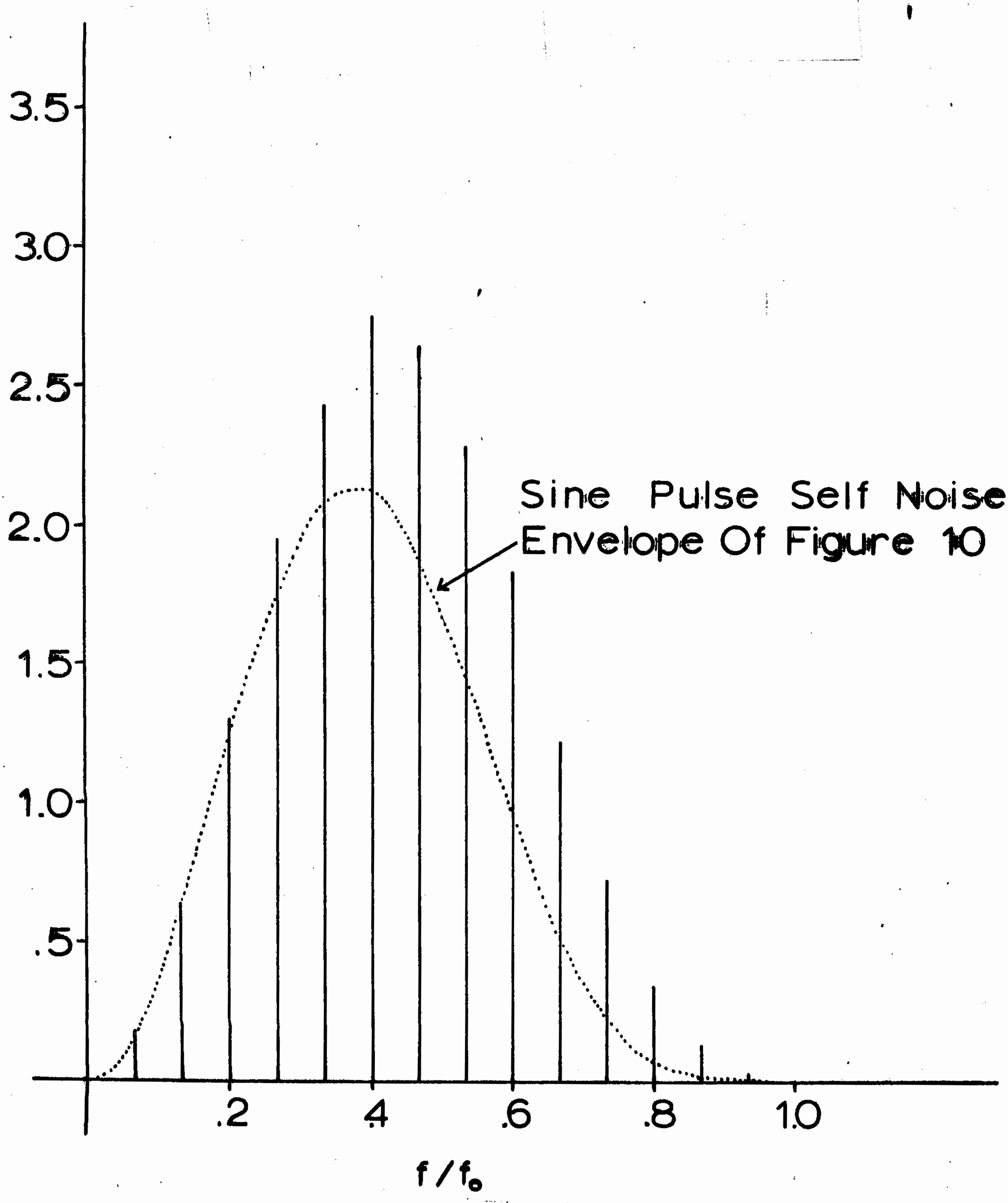


FIGURE 11


3.5
3.0
2.5
2.0
1.5
1.0
.5
Sine Pulse Self Noise
Envelope Of Figure 10
.2
.4
.6
.8
1.0
f / f₀

Figures 4 and 5 that the low frequency values are identical. A few computer values for comparison are given below:

Rectangular		Sine	
f(KHz)	P(ω)	f(KHz)	P(ω)
5.625	.06877	5.625	.06877
11.25	.1924	11.25	.1924
16.88	.3734	16.88	.3735
22.50	.5794	22.50	.5798
67.50	.3647	67.50	.3671

The spectrum for the cyclic sequence of period T is $|X(\omega)|^2 \sum_{n=1}^{\infty} \delta(f-\frac{n}{T})$ where $\delta(f-\frac{n}{T})$ is the Dirac delta function defined by $\int H(f)\,\delta(f-f_0)\,df = H(f_0)$. This is shown in Figures 8 and 9 normalized to the first spike amplitude. Both Gaussian curves were chosen to exactly coincide at zero frequency and 1/3 the first null frequency. In the rectangular case, the resulting error at the null frequency is .036 whereas in the sine case, the error at the null is .0145 thus indicating the sine sequence to be slightly nearer to the Gaussian spectrum. Considering both Gaussian, the approximate ratio of the main lobe powers is merely the inverse ratio of the standard deviations which yields the value of 1.24 dB for sine power to rectangular power ratio.

If the delta spikes are extended to the first sidelobes, the sidelobe maximum level for the rectangular case is -13 dB but for the sine case this maximum is down to -21 dB. The sine pulses are therefore much more efficient as a carrier or sub-carrier for any given PN bandwidth expansion-compression system. In comparing Figure 8 with Figure 4, or Figure 9 with Figure 5, it is seen that even though the envelope is quite erratic, the delta spikes at multiples of 1/T yield a smooth power spectrum envelope. Therefore, not a great deal is gained by going to long sequence lengths except the relative DC spike is reduced proportional to 1/N , and thus crosstalk is also proportionally reduced.

In synchronizing schemes it is necessary to subtract one time shifted correlation from another, both being the same sequence, and the integration times are over one period to yield the controlling S curve. At synchronization, however, this difference has associated with it a power spectrum usually referred to as the sequence self-noise. To yield the S curve, it must be possible to filter out this self-noise and therefore this power spectrum must be known. The self-noise is given in terms of the sequence power spectrum as follows:

$$P(\omega)_{sn} = 2\,P(\omega)\,(1-\cos 2\pi f \tau)$$

This function is plotted for the two cases in Figures 10 and 11. This expression is for sequences shifted by one pulse width. For an n pulse width shift, τn replaces τ in the above expression. This spectrum resembles the Rayleigh density function as should be expected as the joint density function of two orthogonal Gaussian functions is the Rayleigh density function. At low frequencies, the self-noise is equal for the two cases, and therefore, the same optimum filter for removing this self-noise is used.

These three basic characteristics, the power spectrum, correlation function, and self-noise dictate the optimum design objectives for the systems to be considered.

4. Implementation of the PN Generator.

The shift register PN sequences are implemented by either Mod 2 logic which is presently common practice, or by "conditional" logic which is more easily implemented, taking advantage of the JK flip-flop conditioning inputs. The Mod 2 logic block is best expressed by the truth table and a logic diagram as shown in Figure 12. The PN sequence of length $2^k - 1$ is therefore given by the systems in Figure 13, with the reverse sequence and forward sequence implementation shown. Since the reverse sequence has the identical characteristics as the forward sequence, a more simple implementation not requiring Mod 2 logic is shown in the third part of Figure 13. In this system, the input to the first FF is a conditioning input unlike the Mod 2 inputs. This conditioning input gates the clock pulse input to the first JK FF to allow inversion only if the feedback FF value is a 1. Note that if the sequence shown is read upside down, the forward Mod 2 sequence is observed.

Each FF provides a time shifted output of the basic sequence which, due to the orthogonal nature of the code, can be used as separate channel carriers or sub-carriers in a multiplex system. The natural output is the rectangular pulse PN waveform which is the raw FF output. To implement the sine pulse output, additional

MOD 2 IMPLEMENTATION

A	B	(A⊕B)
0	0	0
0	1	1
1	0	1
1	1	0

$A \oplus B = \bar{A}B + A\bar{B} = \overline{(A+\bar{B}) + (\bar{A}+B)}$

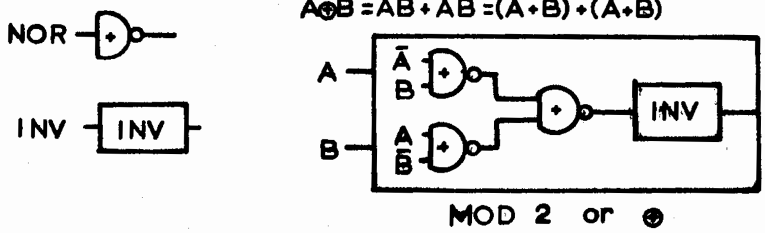


FIGURE 12

PN RECTANGULAR WAVEFORM IMPLEMENTATION

REVERSE SEQUENCE TRUTH TABLE, N=7

A	B	C	CONDITION
1	0	0	RETAIN A
1	1	0	RETAIN A
1	1	1	INV A
0	1	1	INV A
1	0	1	INV A
0	1	0	RETAIN A
0	0	1	INV A

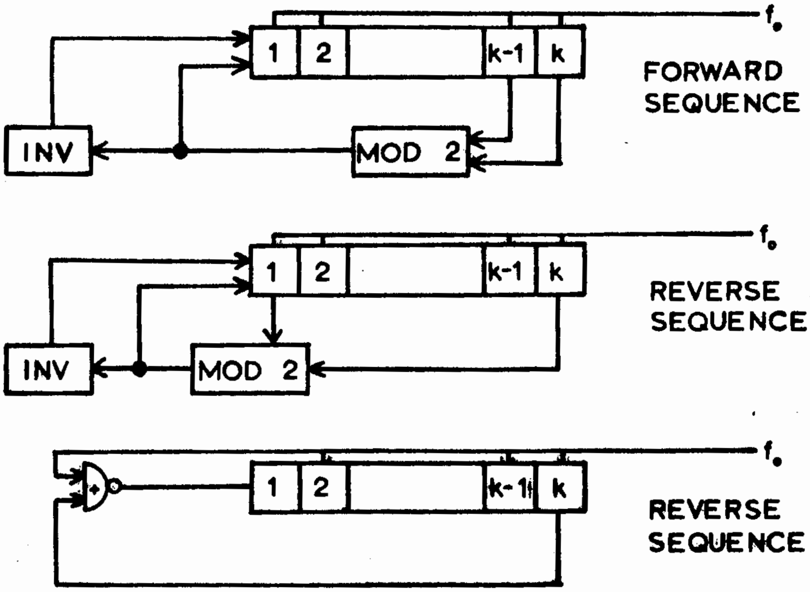


FIGURE 13

circuitry is required. A typical scheme for this is shown in Figure 14, and implemented as shown in Figure 15. Field effect transistors would be ideal for use in the first stage which would allow a high Q circuit and low diode current which would result in a near perfect rectified sine pulse output from the first stage. With some sophistication, the SCS or SCR could be incorporated in the gating system to eliminate the DC bias design complexities. The DC circuitry is desirable in this case to prevent any unwanted phase shifts from distorting the sine pulse cutoff points. In the circuitry shown, the equilibrium state in the gating system is at 1/2 the power supply voltage. The bias and load resistances have been left out for schematic simplicity. With integrated circuit design, the entire unit, with the exception of the inductance, could be incorporated into one module.

The associated basic modulation device for the PN systems is the product device or balanced modulator. Since for a majority of the systems mentioned in this paper, one of the inputs to these devices is a PN sequence, one need only gate an inverter amplifier output with the PN sequence to achieve balanced modulation. This scheme is shown in Figure 16 for the sub-carrier waveform input. This could also be used with the PN carrier input but Mod 2 logic is preferred.

5. Synchronization.

In the latter sections, the basic PN channel, waveform characteristics, and PNG implementation have been discussed. However, when a complete system is considered, many variations utilizing and extending the above techniques are possible. This is especially

PN SINE WAVEFORM IMPLEMENTATION

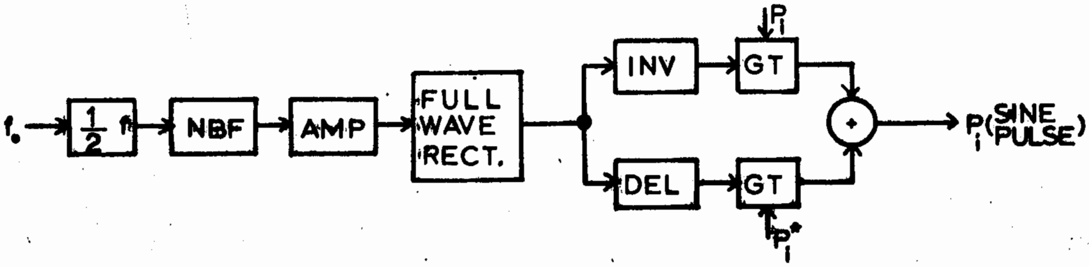


FIGURE 14

TYPICAL CIRCUITRY OF SINE PULSE IMPLEMENTATION

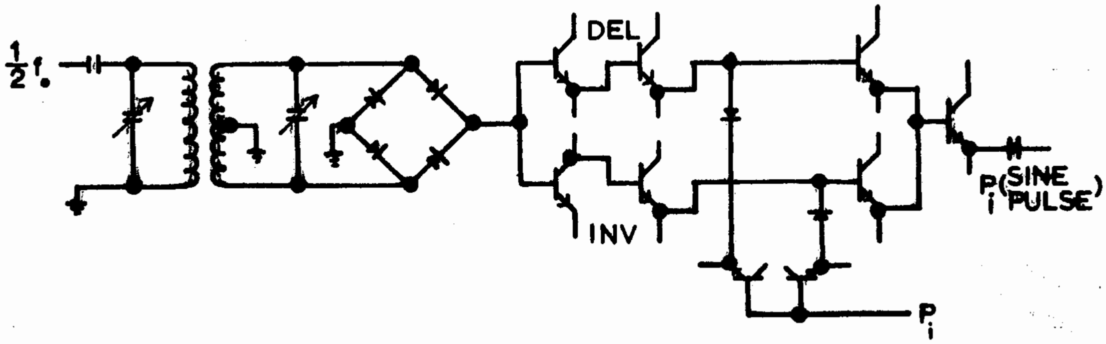


FIGURE 15

PN SUB-CARRIER PRODUCT DEVICE

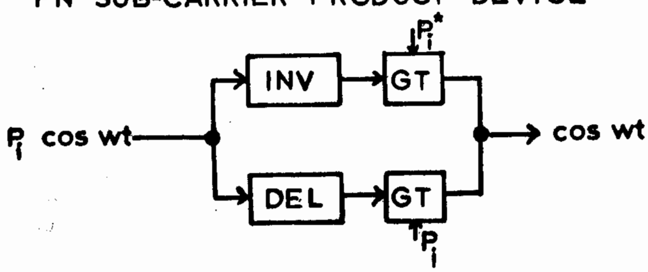
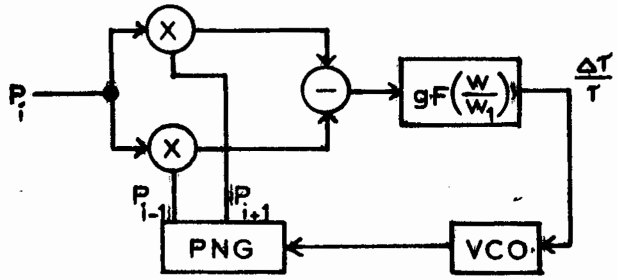


FIGURE 16

true in the area of synchronization of the coherent and non-coherent AM and FM communication links. The primary objective of the synchronization system is the same as that of the common phase or frequency discriminator circuit. An S curve is desired which yields a DC control voltage proportional to the relative phase of the two signals considered. This is achieved, as was mentioned earlier, by subtracting a time shifted version of the correlation output from itself. Many of the earlier synchronizing systems sampled these time shifted correlation outputs by shifting the local PNG back and forth. This produced a desynchronized condition which resulted in 2 or 3 dB loss in signal gain. J. J. Spilker,[14] in 1964, devised a technique which continuously tracks the incoming carrier and provides a true synchronized condition with continuous detection at the peak of the auto-correlation curve. His system is shown in Figure 17 for the new PN waveform but the same system is applicable to the PN sub-carrier systems. A limiter at the input to the correlators makes possible Mod 2 logic balanced modulation utilizing the circuit block shown in Figure 12. A loss of only 1 dB or less results from the limiter action. The loop filter shown is optimum for this waveform with the transfer function, $H(\frac{\omega}{\omega_0}) \stackrel{D}{=} \dfrac{g\,F(\frac{\omega}{\omega_0})}{\frac{\omega}{\omega_0} + g\,F(\frac{\omega}{\omega_0})}$ where $F(\frac{\omega}{\omega_0}) = \dfrac{1 + \sqrt{2}\,\frac{\omega}{\omega_0}}{1 + g\,\frac{\omega}{\omega_0}}$, ω_0 = frequency constant of the filter, g_0 = DC loop gain ($\Delta f_{VCO} = \frac{g_0 \epsilon}{T}$ Hz change), $g_0 \epsilon$ = seconds delay change/second, $g = g_0/\omega_0$. Such a filter is implemented as shown in Figure 18. This filter removes the self-noise and yields the difference of the two correlation functions which is the S curve desired. Without the limiter, sine

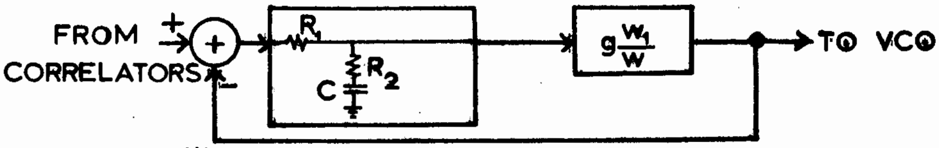
J.R. SPILKER DELAY LOCK SYSTEM



ΔT = DELAY ERROR
$T = 1/f$
w_1 = FILTER FREQ. CONST.

FIGURE 17

OPTIMUM LOOP FILTER IMPLEMENTATION



$$F\left(\frac{w}{w_1}\right) = \frac{1+\sqrt{2}\,\frac{w}{w_1}}{1+g\frac{w}{w_1}}$$

R_2 = ARBITRARY

$R_1 = (g\sqrt{2} - 1)$

$C = \sqrt{2}/(R_2 w_1)$

FIGURE 18

AUTOCORRELATION FUNCTIONS AND RESULTING S CURVES

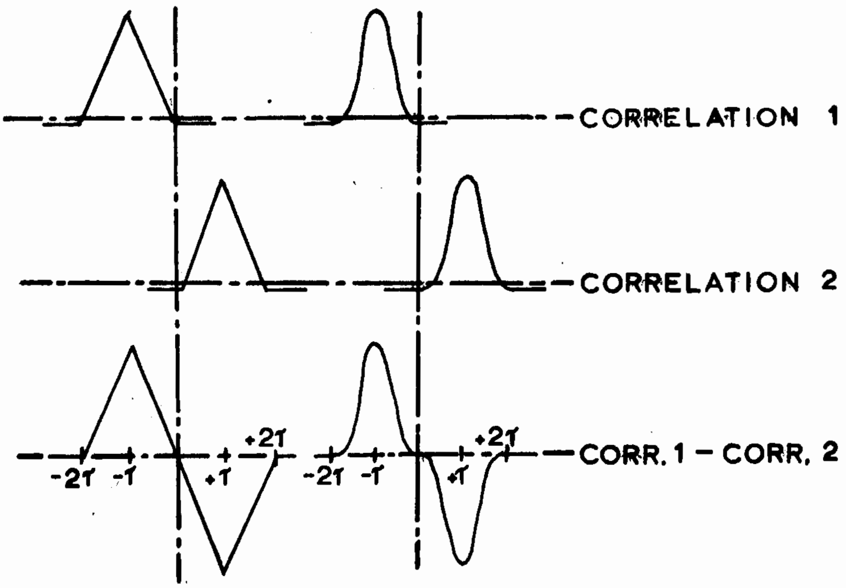


FIGURE 19

COHERENT PN SINE PULSE SUB-CARRIER RECEIVER

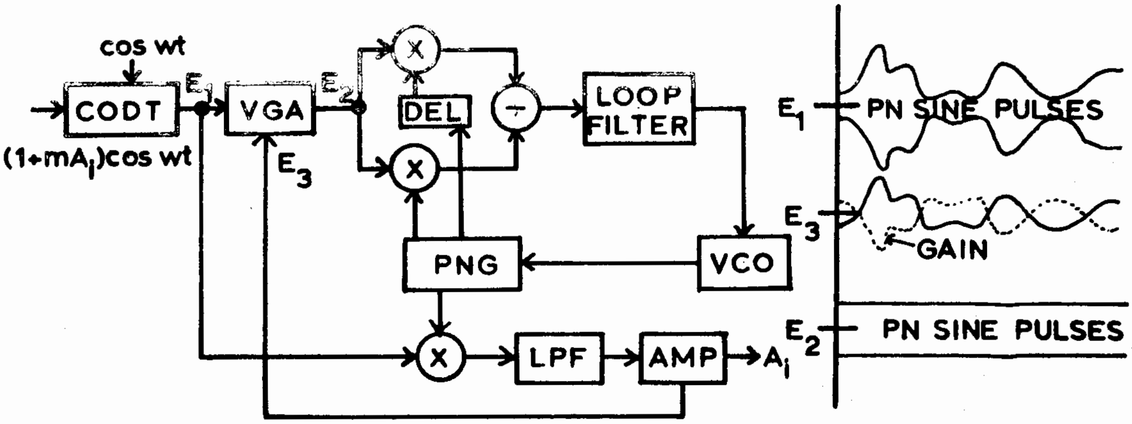


FIGURE 20

J.K. WOLF FILTER

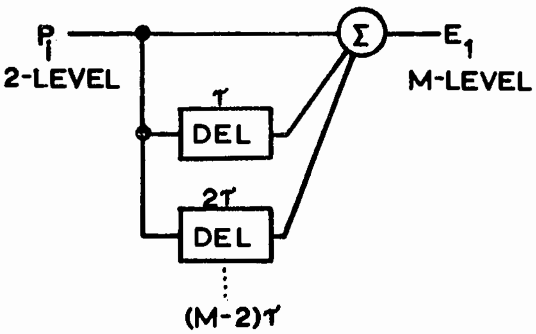


FIGURE 21

impulse input would yield the sine-rectangular cross-correlation functions. The resulting S curves from these two cases are shown in Figure 19. For the two level sequence, synchronization control extends over a relative phase of 2τ and correlation synchronization is lost outside of this range. Shown in Figure 20 is the implementation yielding the S curves given in Figure 19. With the sine pulse input, a slight delay is inserted in the leading FF output line to the top product device to eliminate zero slope at the stable null of the S curve.

Since these systems provide control only over 2 cycles, the tolerance on the frequency is closer than is desirable. Therefore, a synchronization system with a wider control range would literally eliminate the probability of synchronization loss, once lock-on occurred. Such a system is possible and is implemented by constructing a multi-level PN sequence from the incoming 2-level sequence and performing the time shifted correlations on this waveform. The construction of the multi-level PN waveform was shown by J. K. Wolf[17] in 1963. He considered the use of a filter with the transfer function $H(\omega) = \sum_{i}^{M-2} \epsilon^{-j\omega i\tau}$ where M is the number of levels desired at the output with a 2-level input. For this discussion, M will be restricted to even values to always provide a non-return-to-zero (NRZ) waveform, which is more practical for system applications. The implementation of $H(\omega)$ is shown in Figure 21 with P_i being the usual 2-level sequence time function. To simplify the derivation of the auto-correlation function, the approximate power spectrum shall be assumed and the resulting autocorrelation function will be compared with the exact values for a given N. With the power spectrum input, $P(\omega) = \tau\left|\frac{\sin\frac{\omega\tau}{2}}{\frac{\omega\tau}{2}}\right|^2$, the resulting output will be $P_o(\omega) = \tau\left|\sum_{i=0}^{M-2}\epsilon^{-j\omega i\tau}\right|^2\left|\frac{\sin\frac{\omega\tau}{2}}{\frac{\omega\tau}{2}}\right|^2$. But $\sum_{i=0}^{M-2}\epsilon^{-j\omega i\tau} = \frac{1-\epsilon^{-j\omega(M-1)\tau}}{1-\epsilon^{-j\omega\tau}} = \frac{\epsilon^{-j\omega(M-2)\tau/2}\sin\left(\frac{M-1}{2}\right)\omega\tau}{\sin\frac{\omega\tau}{2}}$. Therefore $P_o(\omega) = (M-1)^2\tau\left[\frac{\sin\left(\frac{M-1}{2}\right)\omega\tau}{\left(\frac{M-1}{2}\right)\omega\tau}\right]^2$, where M is still the number of levels of the output sequence. Since the inverse transform of the power spectral density yields the autocorrelation function, the latter function is $R_M(t_o) = \begin{cases}(M-1)\left(1-\frac{|t_o|}{(M-1)\tau}\right) & |t_o| \le \tau(M-1)\\ 0 & |t_o| > \tau(M-1)\end{cases}$. For the PN sequence, the exact expression for the 2-level autocorrelation function is $R_2(t_o) = \begin{cases}\frac{N-\frac{|t_o|}{\tau}(N+1)}{N} & t_o \le \tau\\ -\frac{1}{N} & t_o > \tau\end{cases}$, which for large N yields the two level ideal power spectrum above. The exact expression was found by computing the correlation coefficients for several values of N, and converting these values to the piecewise linear autocorrelation function. For the systems described in this paper, N is not extremely large, and therefore, there is some deviation from the ideal function. If several computations of the correlation coefficient are made for the M level sequences of various lengths, the following exact expression for the autocorrelation function of the M-level, N length sequence is arrived at: $R_{NM}(t_o) = \begin{cases}\left[(M-1)(N-M+2-\frac{|t_o|(N+1)}{(M-1)\tau}\right] & |t_o| \le \tau\\ -(M-1)^2 & |t_o| > \tau\end{cases}$. Some values of the digital calculations are given as follows:

For N = 7, M = 4: $\left[R(t_o)\right]_{t_o=0} = 15$ $\quad\left[R(t_o)\right]_{t_o=(M-1)\tau} = -9$

For N = 15, M = 4: $\left[R(t_o)\right]_{t=0} = 39$ $\quad\left[R(t_o)\right]_{t_o=(M-1)\tau} = -9$

The exact autocorrelation functions are shown for N = 7 and N = 15 in the 4-level case in Figure 22. Since the noncorrelated value,

AUTOCORRELATION OF M-LEVEL PN WAVEFORMS

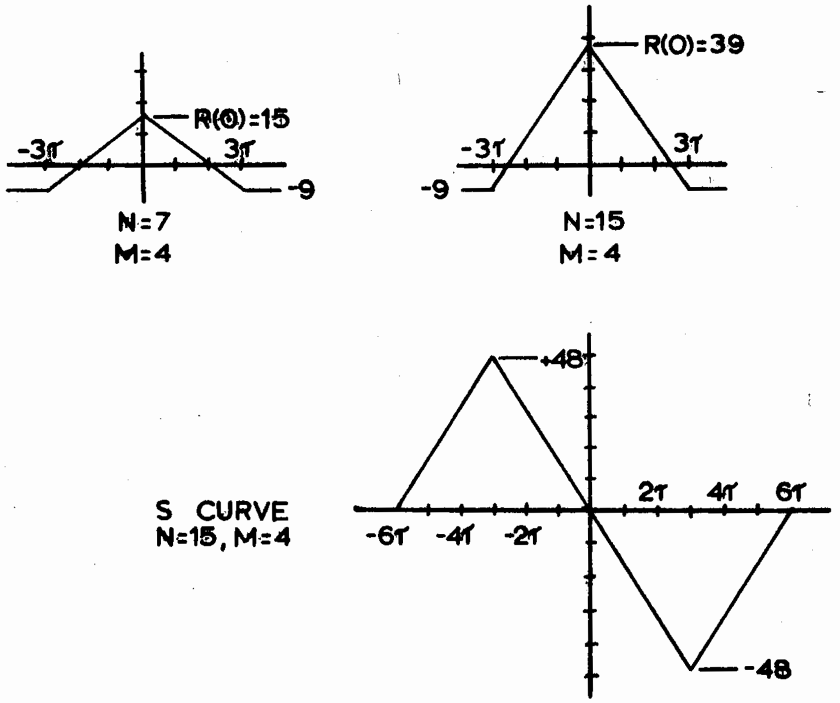


FIGURE 22

FOUR LEVEL SYNCHRONIZATION

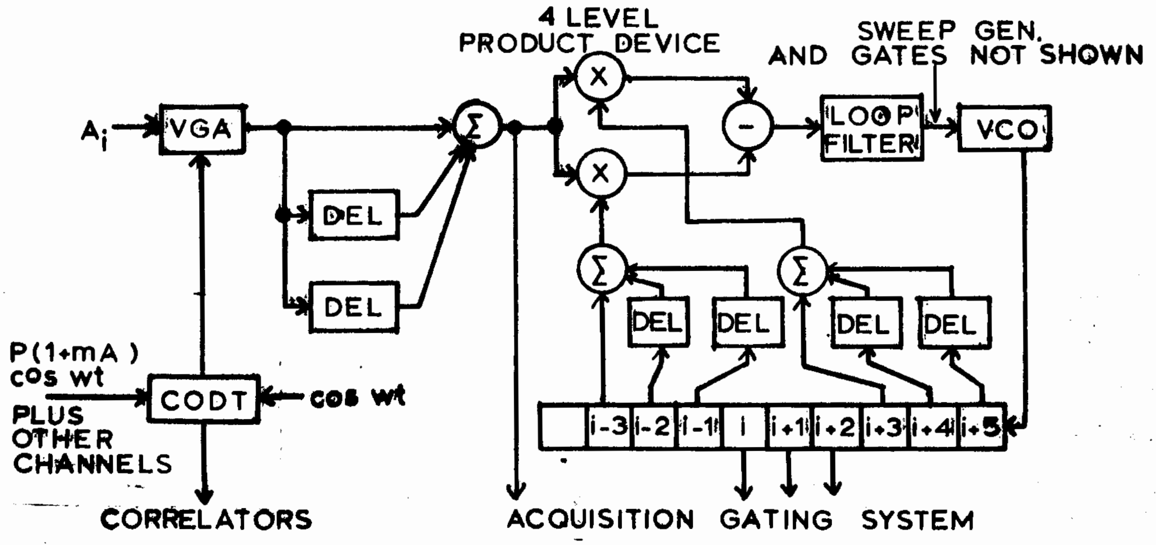


FIGURE 23

normalized to the peak value is $\frac{-(M-1)}{(N-M+2)}$, a direct comparison may now be made with the ideal function derived earlier. For large N, and small M, the true function is seen to approach the ideal function. The negative correlation values are relatively larger but upon shifting and subtracting to form the S curve, a smooth S curve results with zero values outside of the S curve as shown in part 2 of Figure 22. The synchronization control region in this case is 6τ and in general is $2(M-1)\tau$ where M is the number of levels at the output. Also, there is a gain in cross correlation peak value over the 2-level peak which, in general is (M-1)(N+1). For the 2-level case the peak is N+1 which agrees also with this expression.

The above technique yields a wider region of control for the synchronization process utilizing a 2-level input. If $1/T \gg f_{max}$ the information channel could be made relatively insensitive to cycle shifts with M-level synchronization techniques. The information channels need not be M-level, since the only objective of such a technique is to yield a wider phase control region. Therefore, conventional 2-level correlation detection is used for the information channel. Implementation of the M-level synchronization technique is shown in Figure 23. The 4-level balanced modulator may be implemented by gating techniques. The object of such a system is to provide an amplification system with a gain dependent upon the local M-level input. With $P_i = \pm 1$, the gains needed are $\pm 1, \pm 3, \pm 5;$ to $\pm(M-1)$. A DC circuit such as that shown in Figure 24 provides the gating input to the amplifiers, and the

FOUR LEVEL GATING CIRCUIT

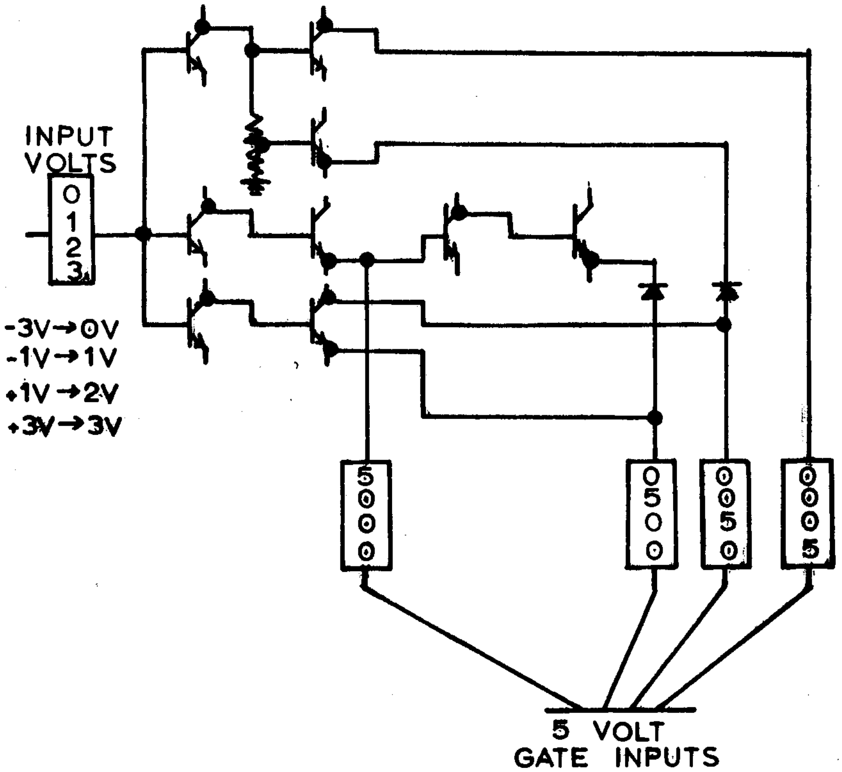


FIGURE 24

FOUR LEVEL PRODUCT DEVICE

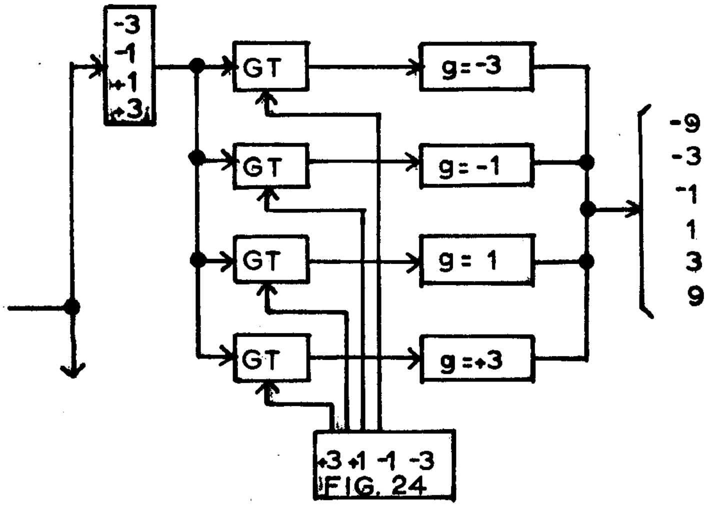


FIGURE 25

complete 4-level balanced modulator implementation is shown in Figure 25. Again, the bias and load resisters are not shown in the circuit schematic for simplicity. The gain amplifiers are 2 transistor stages biased at a mid-voltage range such that a ± 3 volt input does not force the amplifiers into saturation or cutoff prior to providing the ± 9 or ± 3 volt outputs. The transistors are assumed to have a high enough F_{hfe} to pass the rise times of the digital input. This circuit would be an ideal application for the field effect transistor to reduce the noise input of the gates and amplifiers. This implementation is just as valid for the sine pulse inputs, and the cross-correlation outputs yield an S curve similar to the one shown for the rectangular case, but with rounded peaks as shown in Figure 19.

The above techniques are for the coherent detected sub-carrier or for the PN carrier systems. The non-coherent detection process with PN sub-carrier and AM or FM modulation incorporates a second family of synchronization systems. The bandwidth compression process in these systems occurs at the first stage of balanced modulation, yielding a narrow band frequency carrier at the output when synchronization is reached. The system is basically a Spilker system operating on a carrier rather than the raw PN waveform. The block diagram of this technique is shown in Figure 26. The system equations for the correlation branches are given below:

$$E_1(t) = P_i(t)\, P_i(t - \tau_0) \cos \omega t$$

NONCOHERENT SYNCHRONIZATION SYSTEM

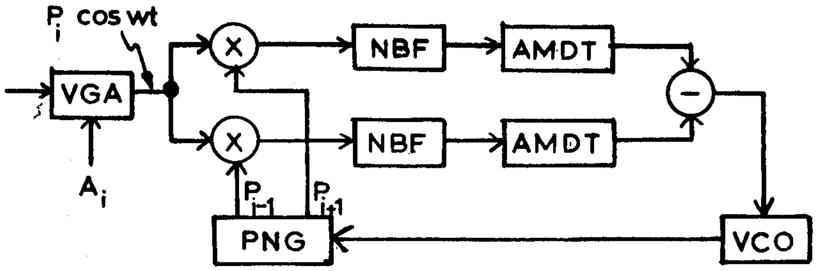


FIGURE 26

EXPANDED S CURVE

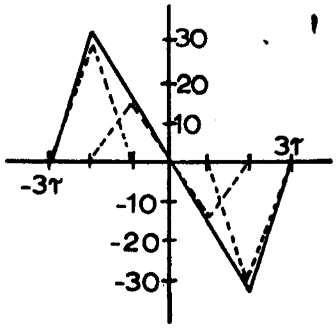


FIGURE 27

NONCOHERENT EXPANDED S SYNCHRONIZING SYSTEM

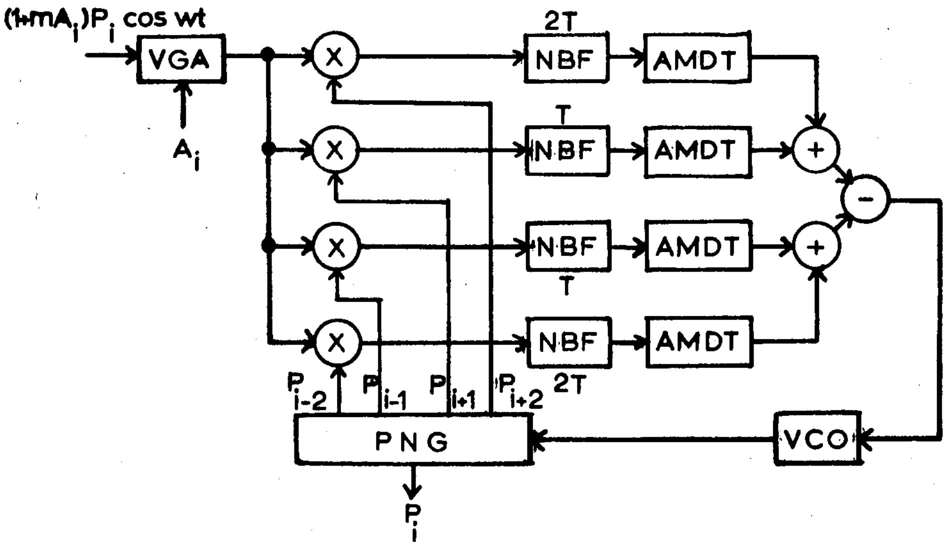


FIGURE 28

FOUR PHASE TRANSMITTER

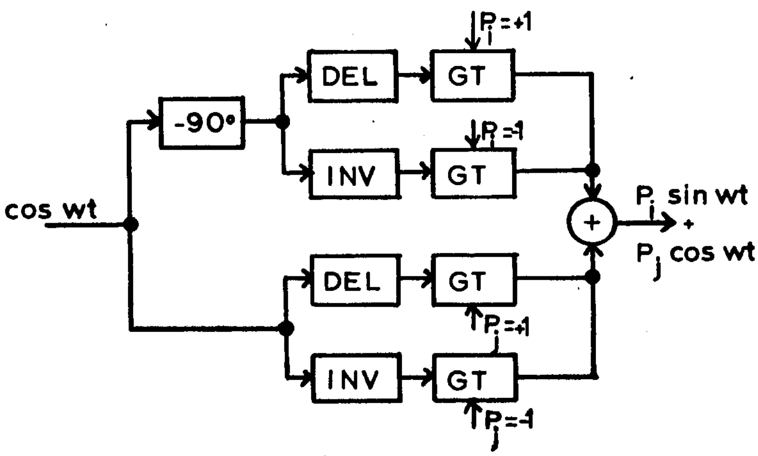


FIGURE 29

$$h(t_1) = \frac{1}{\alpha} f(t_1)\left(\epsilon^{j\omega t_1} + \epsilon^{-j\omega t_1}\right)$$

$f(t_1)$ averages the input over T seconds

By convolution:

$$E_2(t) = \int_0^\infty E_1(t-t_1)\, h(t_1)\, dt_1 = \int_0^\infty P(t-t_1)\, P(t-t_0-t_1)\left[\frac{\epsilon^{j\omega(t-t_1)} + \epsilon^{-j\omega(t-t_1)}}{2}\right]\frac{1}{\alpha} f(t_1)\left(\epsilon^{j\omega t_1} + \epsilon^{-j\omega t_1}\right) dt_1$$

$$= \cos \omega t \left[\frac{1}{\alpha}\int_0^\infty P(t-t_1)\, P(t-t_0-t_1)\, f(t_1)\, dt_1\right] + \frac{1}{\alpha}\int_0^\infty P(t-t_1)\, P(t-t_0-t_1)\, f(t_1)$$
$$\cos \omega(t-2t_1)\, dt_1$$

The latter term is zero as $\cos \omega(t-2t_1)$ varies rapidly in the integral. Therefore, $E_2(t) = (\cos \omega t)\, R_p(t_0)$ $\quad \omega >> \frac{2\pi}{T}$

After amplitude detection,

$$E_3(t) = R_p(t_0)$$

Therefore, the input $E_2(t)$ to the amplitude detector is the carrier modulated by the autocorrelation function. The same S curve results as mentioned in the coherent case. This is mainly a 2-level PN synchronization system. A delay system such as that described for the coherent system would be impractical as the delay tolerances would be too close. However, wider control range is possible by multiple correlation techniques, and the same objective is reached. The basic correlation function yields a peak of N, unnormalized, and a cross-correlation of -1. If the correlation integration times is 2T, a peak of 2N is realized. Therefore, consider a summing process in each branch, of the correlation outputs of $P(t \pm \tau)$ and $P(t \pm 2\tau)$ cross-correlated with the incoming signal, S(t), and the $S(t)\, P(t \pm 2\tau)$ is integrated over 2T. The resulting S curve error voltage occurring at the filter output is shown in Figure 27. Note the control region is over 4τ and with a third correlator a region of 6τ is realized which is equivalent to the coherent system. Implementation of the 4τ control synchronizer is shown in Figure 28. Again, for the sine pulse form, a delay in the leading brance is necessary and a similar control region is attained.

The last synchronization scheme to be discussed is the 4 phase modulation system. In such a system P_i modulates the $\pm 90^\circ$ and P_j modulates the 0 and 180°, and the two modulated carriers are added. The resulting waveform is of the same frequency but is phase shifted in one of four discrete phases. By coherently detecting this waveform at the receiver, the two channels may be reproduced and the usual Spilker synchronizing system may be employed. The basic transmitter is implemented as shown in Figure 29. Since P_i is known relative to P_j, only one synchronization circuit is needed, either about P_i or P_j. A unique method of accomplishing coherent detection utilizing this waveform is shown in Section 7 and therefore there will be no further discussion here.

In the foregoing systems, the primary objective is to yield some type of S curve which has a stable null such that the derivative of the control voltage with respect to phase error is positive over a wide control region, and preferably a linear relationship exists over this region between phase and voltage output such that V_{out} = C(phase error) where C is a positive or negative constant, and the phase error has values from $-N\tau$ to $+N\tau$. The stable null occurs at (phase error) = 0. The realistic system must, in addition to the S curve, have a system which sweeps the local PNG in phase and transfers sweep control to the cross-correlation output when the in-phase control region is reached. This is accomplished by introducing an auxiliary sweep generator which sweeps the VCO clock from a frequency below the transmitted clock frequency to a point above, in a sawtooth or sine wave fashion. By introducing two gates in the circuit, say SCR's or SCS's with rapid gating times, this transfer can be accomplished in more than ample time. The gates are controlled by the cross-correlation value of the zero phase shift register output and the incoming signal. This gating system is shown in Figure 30.

This concludes the discussion of the synchronization techniques which can be used with secure PN systems. Many digital scanning techniques found in current papers have not been discussed due to the complexity of the implementation. In a given communications system, acquisition times are not of critical interest as in the radar case, and therefore this additional complexity would be impractical. Also, acquisition times have not been pursued. J. J. Spilker[14] has dealt with this problem extensively, however, in his paper. Of primary importance in the communications system is the stability of the synchronized state. It has been shown, in this section that the control region can be appreciably extended beyond the 2τ phase error, and that many practical stable coherent and non-coherent systems can be devised utilizing these techniques.

AUTO-SEARCH AND LOCK SYSTEM

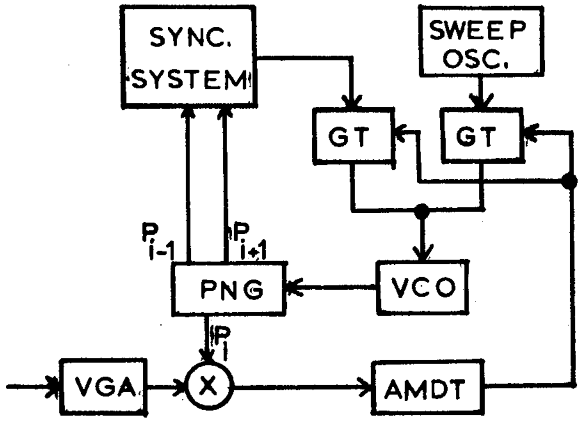


FIGURE 30

PN CARRIER SYSTEM, AM MULTIPLEX

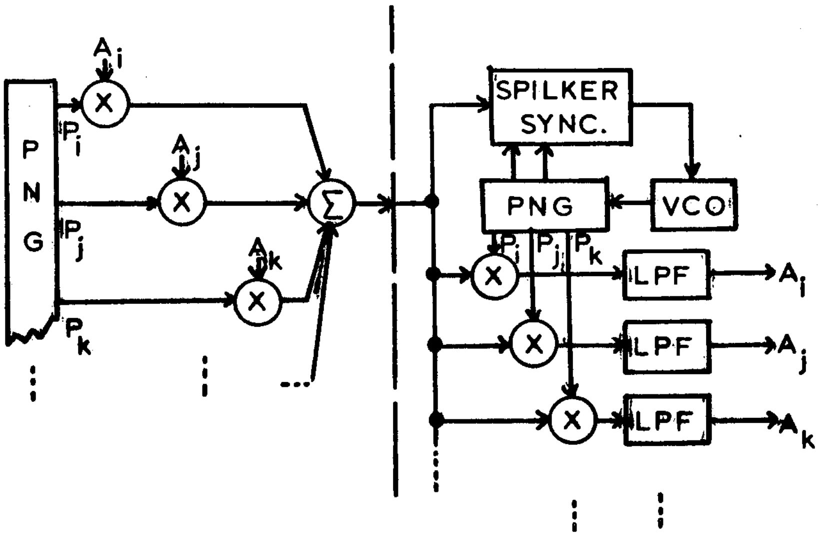


FIGURE 31

CROSS TALK COMPENSATOR

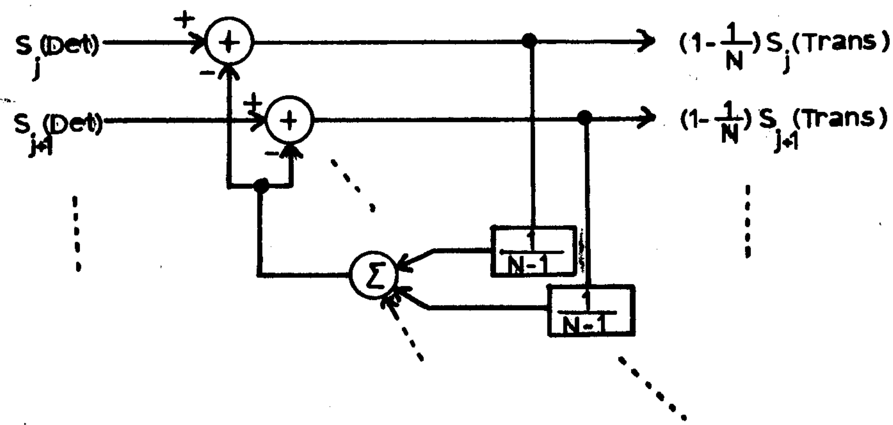


FIGURE 32

6. Multiplex PN Carrier Systems.

These systems are basically DC carrier systems operating in an unconventional form of SSB/SC. At these frequencies, the wideband signal is inefficiently transmitted and received due to the massive antenna requirements, and therefore, only short ranges or cable transmissions are feasible. This type of signal, however, is ideal for multi-channel applications, which can be applied to the sub-carrier systems as well. The information channel, assumed to be narrow band compared with the clock frequency, is balanced modulated with one time shift of the PN sequence, and sent into the transmission medium. With several channels, each FF output has an associated modulation channel, and all channels are linearly summed prior to transmission. The synchronization channel is unmodulated and corresponds to a DC information channel. In Figure 31 this system is shown with A_j corresponding to a DC input or an amplitude modulated DC input with a modulation index of about 1/2 to prevent any phase reversals in the synchronization channel, and to allow this channel to be used for information transmission as well. When in synchronization, each product device in the receiver gives the multiplex signal a weighted value which yields the unique low frequency channel at the output of the LPF. From the spectrum shown in Figure 8, it is seen that the lowest frequency component of the other channels is 1/T with a small DC component which causes crosstalk in the system. The basic system equations for the j'th balanced modulator are as follows:

$$S(t) = A_i(t)P_i \text{ where } B_A = 2\, f_{max} \text{ and } B_P = f_o$$

$$E_j(t) = P_j\, S(t)$$

$$P_jP_j = 1 \quad \text{and } P_jP_i = P_n{}^*$$

For j = i, the resulting product yields $A_j(t)$, and

$$E_j(t) = P_j\ A_i(t)P_i + A_j(t) = A_i(t)P_n{}^* + A_j(t)$$

Low pass filtering yields $A_j(t)$ if the following condition is met:

$N_{max} = f_o/(2f_{max})$ for the general PN multiplex system. An upper bound on the sequence length is thus given for each clock frequency. A sample calculation of N_{max} follows:

i.e. at f_o = 1 MHz , f_{max} = 4 KHz , the resulting N_{max} = 125

Since the primary signal level is below ambient noise, this is not a great restriction.

With many channels, the summation of all the DC components becomes a primary degradation to the system. For the spectrum of Figures 8 and 9, this normalized value is .0635 of the peak power which is near the 1/15 value expected. The reason for the slight discrepancy (.0666 - .0635 = .0031) is the radical behavior of the power spectral density envelope at the 66.7 KHz point which is used for the normalization base. The .0635 was calculated in closed form by noting that the power spectral density expression reduces to the 0/0 form at $\omega = 0$. By De l'Hopital's Rule, one arrives at the expression $|X(\omega)|^2 = \frac{T^2}{2}$ (1944-1942) $= T^2$. Multiplying this by $f_o{}^2$ to normalize, the DC value of 1 results. Then, normalizing this to the peak value at f = 1/T, the value .0635 is arrived at.

The crosstalk magnitude may now be calculated. If $A_j'(t)$ is the detected LPF output at the receiver, and the unprimed function is the actual information transmitted, $A_j'(t) = A_j(t) + (1/N) \quad A_i(t)$. General Electric [13] conceived of a system to eliminate this additive channel energy . Their argument proceeds as follows:

$$A_j'(t) = A_j(t) + \left[\frac{1}{N} \sum_{i=1}^{M} A_i(t)\right] - \frac{1}{N} A_j(t)$$

$$A_j(t)\left(1 - \frac{1}{N}\right) = A_j'(t) - \sum_{i=1}^{M} \frac{\left(1 - \frac{1}{N}\right) A_i(t)}{N-1}$$

The implementation of this argument is shown in Figure 32.

In operation, the wideband system would be band-limited by the antenna system used. For a 2 MHz clock frequency, and a band-limited signal restricted to the broadcast band, 550 KHz to 1.55 MHz, with wavelength ranging from 1790 ft. to 635 ft., 1/4 wavelength matching antennas yield formidable dimensions. Therefore, since the range is short, an ordinary long wire antenna tuned to the mid-frequency would provide the best solution and would be more versatile for possible ECM clock frequency changes.

The above system is merely an extension of the AM basic PN carrier channel described in Section 2. An extension of the second system, the FM system, described in that section is given in Figure 33. The main limitation of this system is the sequence length as for time multiplexing, the sample distance must be $1/2f_{max}$ or less, and the correlation time must be at least one sequence length or $T = N/F_o$. Therefore, for an N length sequence, and M channels, $(N/f_o)M < (1/2f_{max})$, or $M_{max} = f_o/(2Nf_{max})$. For example, at $f_o = 1$ MHz and $N = 15$, $M_{max} = 8$, and similarly for

$N = 31$, $M_{max} = 4$, both assuming a 4 KHz information channel. To circumvent this limitation, f_o may be increased to 5 or 10 MHz. The VCO and discriminator response times must be fast compared with T and $1/2f_{max}$ respectively. To reduce the probability of cycle jump and hence the loss of synchronization, a third and fourth correlator is added at the $\pm 2\tau$ delay FF, as suggested in Section 5. The bandwidth compression in the system with sampling pulse widths of T is approximately $f_oT = N$ which is comparatively low. The post-detection processing to smooth the sampling function outputs is not shown in Figure 33 and consists mainly of a conventional LPF matched to the bandwidth of the information channel. The compression ratio can be increased by increasing the PNG clock frequency and the sampling widths.

The transmitter sampler is synchronized with the PN generator and a possible auxiliary counter depending on the number of channels multiplexed. To eliminate channel ambiguity, a DC bias is put on Channel 1 at the transmitter. At the receiver, the discriminator for Channel 1 is tuned to a slightly higher frequency, out of the range of the other channels, but within the range of the S curve control region. Upon receiving a signal in its pass-band, the Channel 1 discriminator sets the enable FF and at the next timing pulse from the PNG, the sampling stepper is shifted to the next channel input, and the enable FF is returned to its original condition. The sampler proceeds through all of the channels unconditionally stepping every T seconds, and upon returning to

PN CARRIER SYSTEM, FM MULTIPLEX

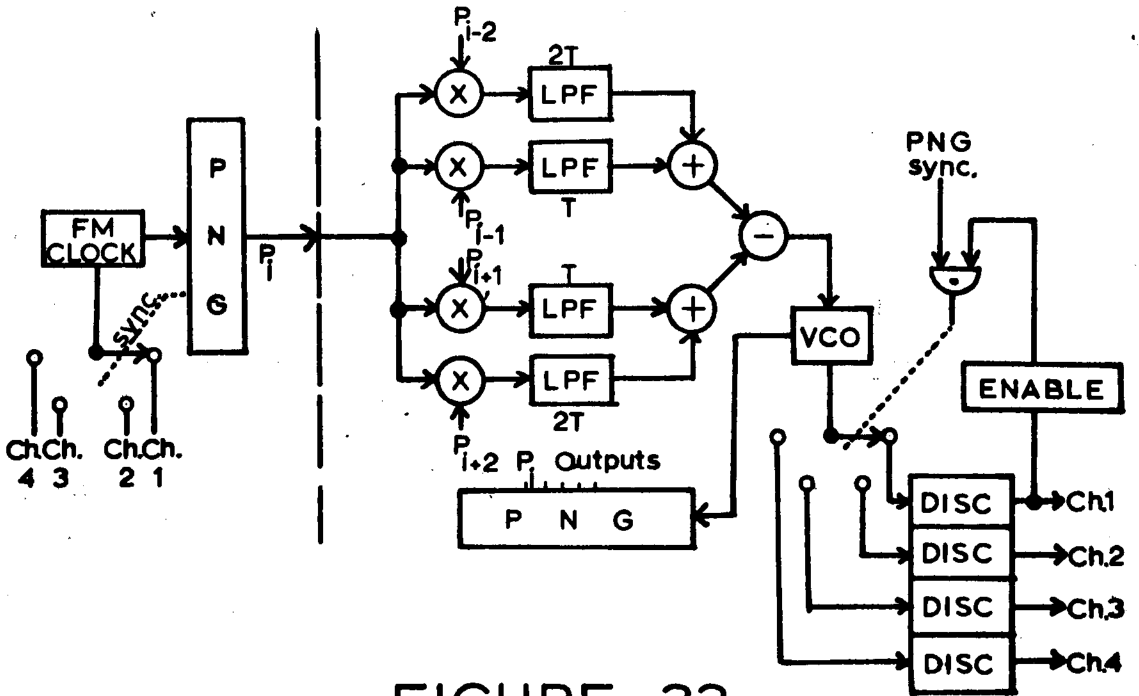


FIGURE 33

PN TRANSMITTED REFERENCE SYNCHRONIZATION SYSTEM

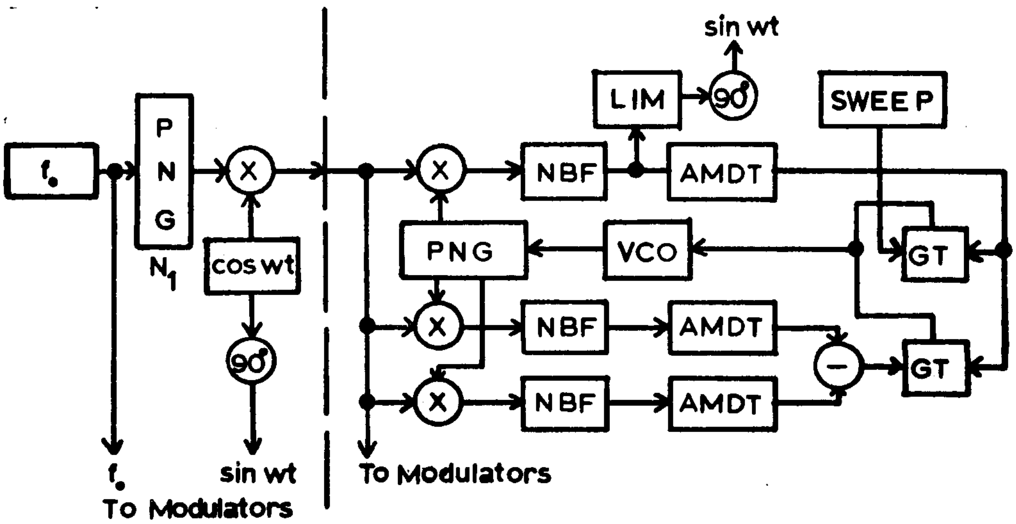


FIGURE 34

Channel 1, cannot proceed further unless a signal is detected to set the enable FF. A continuous check of sampling synchronization is accomplished in this manner. The initial search and lock-on phase is accomplished by the method described in Section 5. The signal, and an amplitude detector with the FM bandwidth yields a gating signal to gate out the sweep oscillator when the in phase condition is reached, or within the control region of the S curve.

This concludes the discussion of the PN carrier systems for short range and cable transmissions. The multiplex detection techniques described here are also applicable to many of the coherent detection multiplex systems described in the following sections.

7. Multiplex PN Sub-Carrier Systems.

The PN carrier systems in the above discussion are bandwidth expansion-compression systems operating directly on the information channel waveform. The PN sub-carrier systems, however, perform the expansion-compression on the information modulated carrier. This carrier, when balanced modulated with a PN sequence, reduces to a random $\pm 90^{\circ}$ phase modulated carrier, sometimes referred to as a bi-phase modulated carrier. The system equations for the signal processing system were derived for the non-coherent case in Section 5, and apply to the systems described here as well with the exception of the 4-phase system. In the non-coherent multiplex system, one need only separate the carrier frequencies transmitted by $2f_{max}$ and utilize the correlation detection process shown in the synchronization system of Figure 26. This is a brute force method but its unsophisticated nature yields a compression ratio of f_o/f_{max} for each channel which is optimum for any of the systems to be discussed. However, there is a 3 to 5 dB loss in this system due to the amplitude detection process.

The coherent detection system offers many advantages over the non-coherent system with respect to multiplexing techniques as well as the 3 to 5 dB processing gain advantage. With coherent detection the multiplexing problem can be reduced to operations on the raw multi-level waveform and the techniques of Section 6 are directly applicable.

When considering the coherent detection problem, one would like to eliminate the conventional carrier phase lock loop and its associated close tolerances, and have a system which yields the carrier phase and PN synchronization in one operation. This can be done by manipulating the orthogonal characteristics of the carrier and PN sequence family in such a way as to make the phase knowledge of one yield the phase knowledge of the other. Since the phase control region of the PN S curve yields more practical timing tolerances, it is more reasonable to choose the PN phase information as the determining factor for the system. This has been done by two methods in this section but undoubtedly many more exist. The first is a type of transmitted reference system where the reference is also a sub-noise transmission. Utilizing the carrier bi-phase modulated waveform with a modulating sequence of different length with respect to the information channels, a reference carrier output may be utilized.

As seen in Figures 8 and 9, the actual spectrum of the modulating sequence consists of 15 delta spikes spaced 1/T in frequency, and the DC spike of value 1/N relative to the envelope height. This leaves the void spaces between for other sequences of other lengths. The PN sequences do not coincide except at the DC value and at f_o which is the first null. Therefore, the carrier phase can be transmitted in quadrature with the information channel utilizing a lower length sequence to place the first delta spike outside of the first information channel spike. The carrier phase is recovered prior to the amplitude detector which provides the in-sinc. gating action. This transmitted reference system is shown in Figure 34. The carrier phase reference system in the receiver cannot share the same clock or VCO with the information channels as the synchronization control voltages in the two cases are independent. The information channel is shown in Figure 35 with $N_2 > N_1$. Once synchronization is attained for the transmitted reference detector and the information detector, the two VCO's can be locked on to each other and the two synchronization systems can be operated in parallel for greater stability. This is possible because the two PNG's in the transmitter are driven by a common clock. Thus, the conventional phase lock loop is eliminated and system security is not compromised by the transmitted reference. The coherent detector output is exactly the PN carrier waveform of M channels described in Section 6. The amplitude detection process in the reference system is extremely narrow band, and therefore, the processing gain prior to the detector

TRANSMITTED REFERENCE AM COHERENT MULTIPLEX SYSTEM

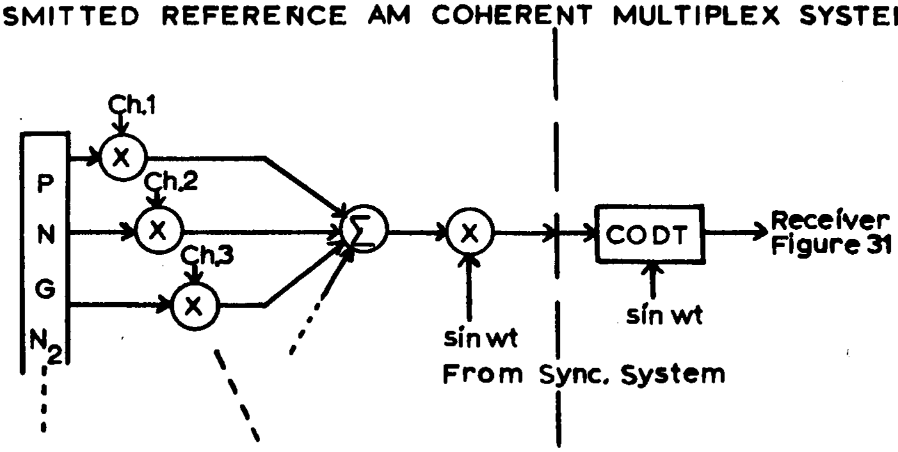


FIGURE 35

can be made as large as for the information channel, which has a larger bandwidth, even with coherent detection. Noise amplitude modulation of the reference carrier can be eliminated by a limiter, and negligible noise phase or frequency modulation is found with respect to the information carrier as both experience the same transmission distortion.

With this system, the carrier can be in the HF, S band or X band and extremely long distances with system security is attainable. Jamming of such a system is extremely difficult since S/N at the receiver to within a few miles of the transmitter is less than one and impossible to detect by a conventional threshold detector.

To give a feel for such a system as that described above an unconventional TV transmission system is proposed below which is in many ways more practical than the present systems used. The video information channel assumes the same resolution requirements as in the present conventional system. Only the synchronizing and audio information is sub-noise and the video channel is continuous and is transmitted conventionally. The system is shown in Figure 36 with the sweep waveform shown. If conventional TV is stripped of retrace and blanking pulses, the actual visible trace is 65.5 microseconds per line with 510 lines per frame and 30 frames per second. Therefore, the basic trace system is a 7.65 KHz system with a two trace period of 131 microseconds. The horizontal sweep waveform is triangular and the vertical sweep waveform is step-triangular with a two sweep period of 1/30 second, and a step period of 65.5

PN AM TELEVISION SYSTEM

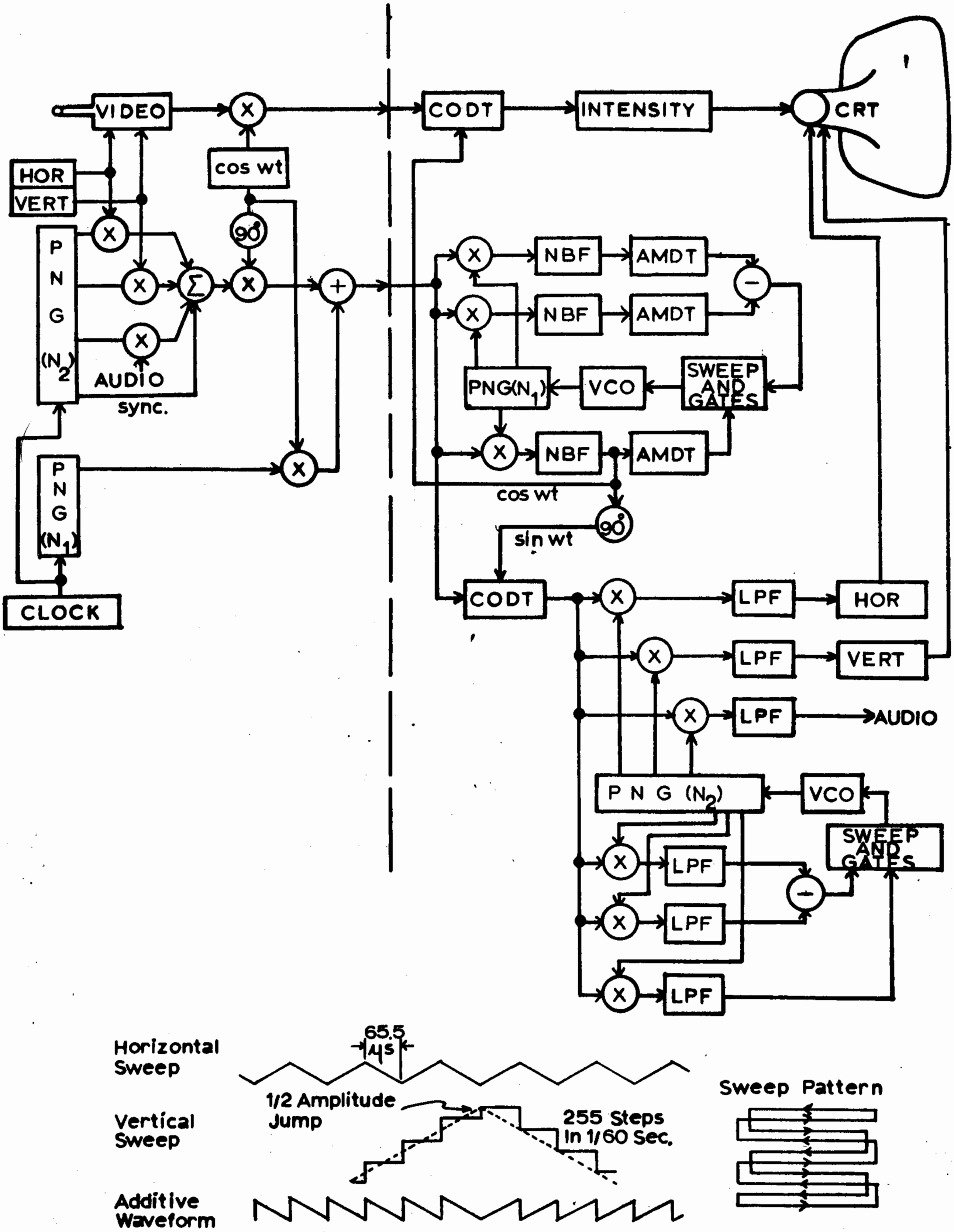


FIGURE 36

microseconds. If the power spectrum of each is analyzed, the triangular waveform yields a $\frac{\sin^2 \frac{\omega\tau}{2}}{(\frac{\omega\tau}{2})^2}$ function whereas the step triangular waveform is of the form $\left[\sum_{n=1}^{M} \frac{\sin \frac{\omega_n\tau}{2}}{\omega_n\tau/2}\right]^2$. Considering the former, the main lobe of the power spectrum is within 15.3 KHz. For the step-triangular function, all of the main lobes are within 15.3 KHz. Therefore, the TV synchronization channel assumes this one sided bandwidth. With the conventional video channel bandwidth of 4.5 MHz , a clock frequency of 4.5 MHz was chosen to match the synchronization bandwidth to the video bandwidth. With this done, a processing gain for the TV synchronization channel is 4.5 MHz/15.3 KHz = 290 or 24.6 dB. Therefore, if the video S/N is 10.6 dB, the synchronization channel can be transmitted 14 dB below ambient noise, neglecting receiver additive noise. With field effect transistors in the early stages of amplification, the receiver additive noise can be made extremely small.

As is seen in Figure 36, the sweep pattern is extremely simple with interlacing simply implemented by the step-triangular waveform. To the 1/30 second triangular waveform is added the sawtooth waveforms to yield the step pattern. At the peak and valley apexes, a gated DC is added and removed respectively to implement the interlacing sweep. The receiver merely utilizes these raw waveforms as direct inputs to the vertical and horizontal deflection plates which greatly simplifies the receiver.

The transmitter and receiver have 2 PN generators and the system is identical to those shown in Figures 34 and 35. The carrier is reconstructed from the transmitted reference and is used to coherently detect the continuous video SSB channel. The horizontal sweep, vertical sweep, and audio channels are balanced modulated by PNG(N_2) and summed along with the synchronization unmodulated channel. This composite signal is then used to balanced modulate the 90^o phase shifted version of the carrier, and transmitted. The transmitted reference is derived from the direct carrier, balanced modulated by a PN sequence of different length, N_1, and transmitted. The processing gain is more than sufficient to detect the sub-noise multiplexed signal through the wide band video and since the two are in quadrature also, the video is further discriminated against.

If it is desired to have the video information also at a sub-noise level, a two sided boundary yields an optimum bandwidth for operation. Considering the sampling of the video channel during the 65.5 microsecond sweep, a 4.5 MHz bandwidth suggests (65.5)(4.5) = 300 revolution calls per line, and a minimum sampling time of 1/9 microsecond, or 9 samples per microsecond, or 500 samples per sweep. For ease of calculation, using 600 samples, 2, 4, or 8 channels could be used to transmit every second, fourth, or eighth sample respectively. The channel samples would be converted to a wide rectangular pulse to fill the gaps, and balanced modulated with a PNG output. Since maximum sequence length is restricted to $N_{max} = f_o/2f_{max}$ and for $2f_{max} = 9$ MHz , N = 15 , the required $f_o = 135$ MHz . This is impractical but reduction is possible by utilizing the above mentioned technique. For N = 15 , 2 channels require a bandwidth of 67.5 MHz , 4 channels require 33.8 MHz , and 8 channels require 17 MHz. If the processing gain is maximized with a given sequence length, for N = 15, the point at which the channels required correspond with the maximum processing gain is 4 channels with a processing gain of 30. and an f_o of 67.5 MHz. Therefore, for a receiver output of 10 dB S/N, the incoming video is at a -4.8 dB level, and the 4.5 MHz synchronization, audio, and deflection information would be at a -14.6 dB, again neglecting' receiver additive noise.

This TV system could be utilized for a secure radar repeater transmission system which could be entirely solid state and would utilize a minimum of parts compared with the conventional TV transmitters and receivers.

A second variation using bi-phase modulation of a transmitted reference is shown in the coherent system of Figure 37. To form a multiplex channel for this case, one cannot balanced modulate the summed multiplex signal at the transmitter as the receiver has no way of detecting this signal without carrier phase information. Therefore, amplitude modulation is used to retain the phase integrity of the carrier and a synchronization channel balance modulates this carrier to yield a wide-band DSB-SC output. With one sideband suppressed, another multiplex channel could be sent over the other sideband, and an effective dual SSB multiplex system results. At the receiver, the carrier is compressed in bandwidth by the synchronization circuit, and the NBF ideally yields the carrier which is

PN AM COHERENT MULTIPLEX SYSTEM

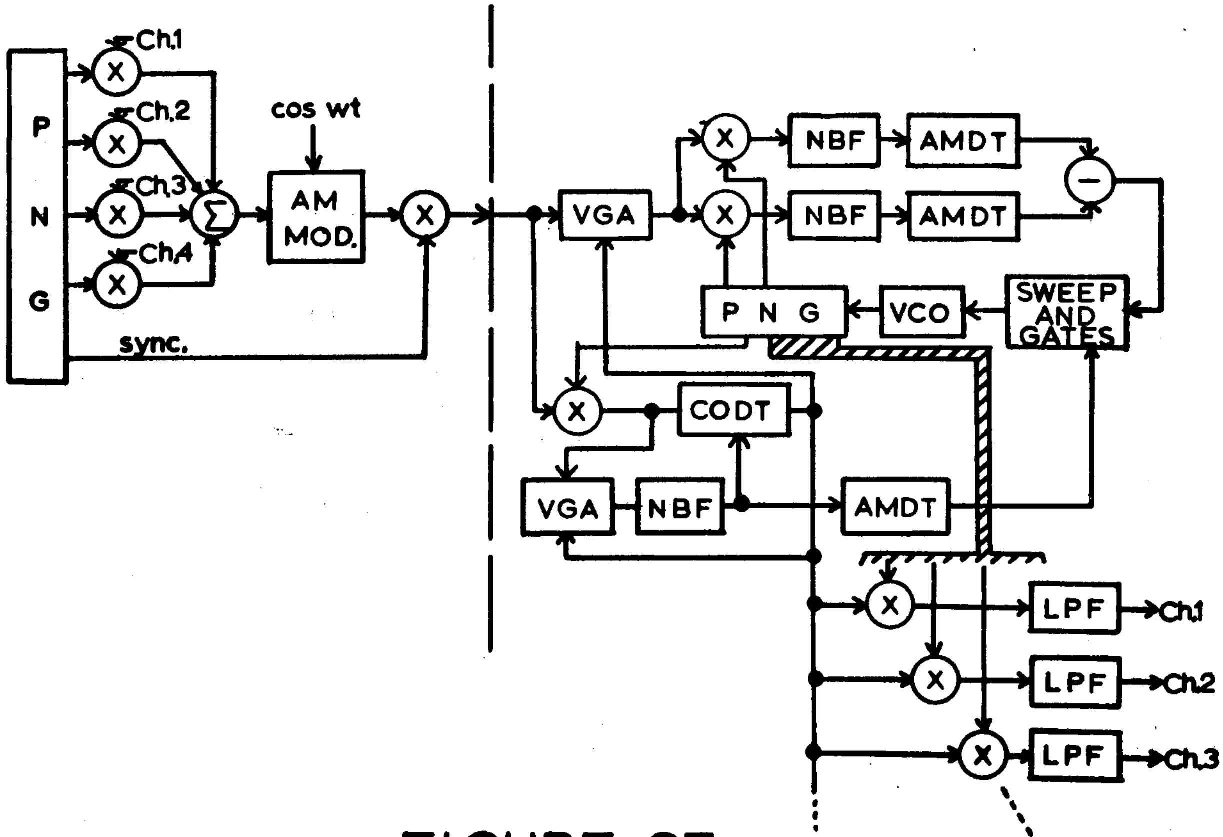


FIGURE 37

utilized in the coherent detector to yield the raw summed multiplex signal. Since synchronization is already obtained, balanced modulation by the unique PN outputs provides the bandwidth compression and the individual channel outputs are realized.

The conventional non-coherent PN system is shown in Figure 38. The channel spacing in frequency need not be the entire bandwidth of the transmitted wideband signal, but need only be twice the information bandwidth. The narrow band filters are matched to the information bandwidth and the bi-phase modulated carrier in Channel 1 is the transmitted reference for the system. This is the brute force method of achieving the multiplex objective and places the burden on the BPF. Sub-noise operation is practical with this system, with the processing gain of 19 dB common to the non-coherent systems.

A more sophisticated system is possible, however, utilizing the shift and add characteristics of the sequences, and the phase quadrature characteristics of the carrier. Consider P_1, P_2, and P_6 as three time shifts of the basic sequence, with the following characteristics: $P(t) = \pm 1$ $\quad P^*(t) = \mp 1$

$$P(t)\,P(t) = 1 \qquad P(t)\,P^*(t) = -1 \qquad P(t) + P^*(t) = 0$$

$$P_1 P_2^* = P_1^* P_2 = P_6$$

$$P_2 P_6^* = P_2^* P_6 = P_1$$

$$P_6 P_1^* = P_6^* P_1 = P_2$$

If the transmitted signal is $S(t) = P_1(t)\cos\omega t + P_2(t)\sin\omega t$, $P_6(t)$ can be recovered at the receiver without phase lock loops or coherent detection in the conventional sense. The following

PN AM NONCOHERENT FREQUENCY DIVISION MULTIPLEX SYSTEM

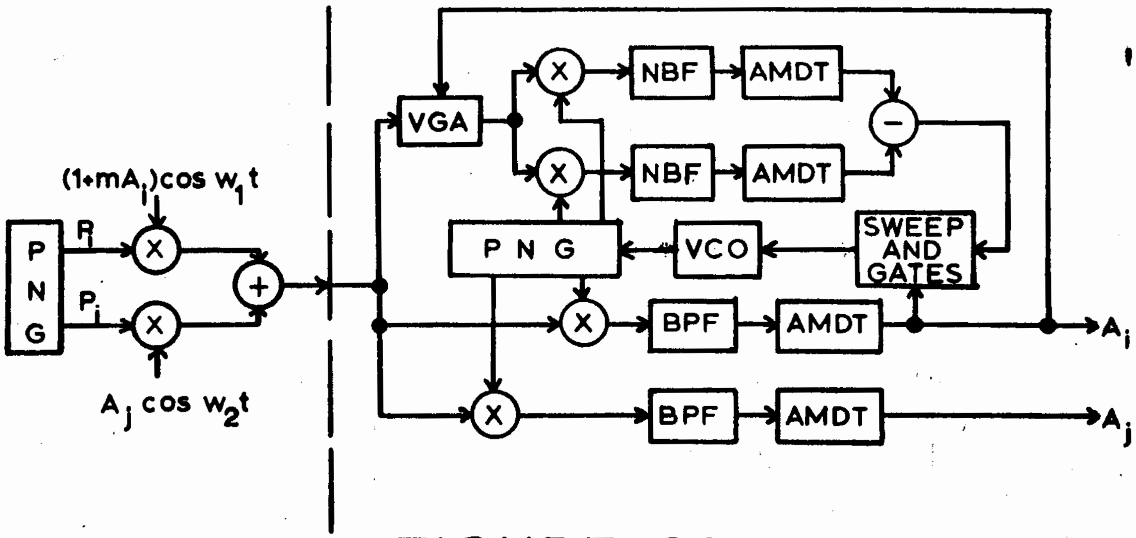


FIGURE 38

PN AM-FM COHERENT MULTIPLEX SYSTEM

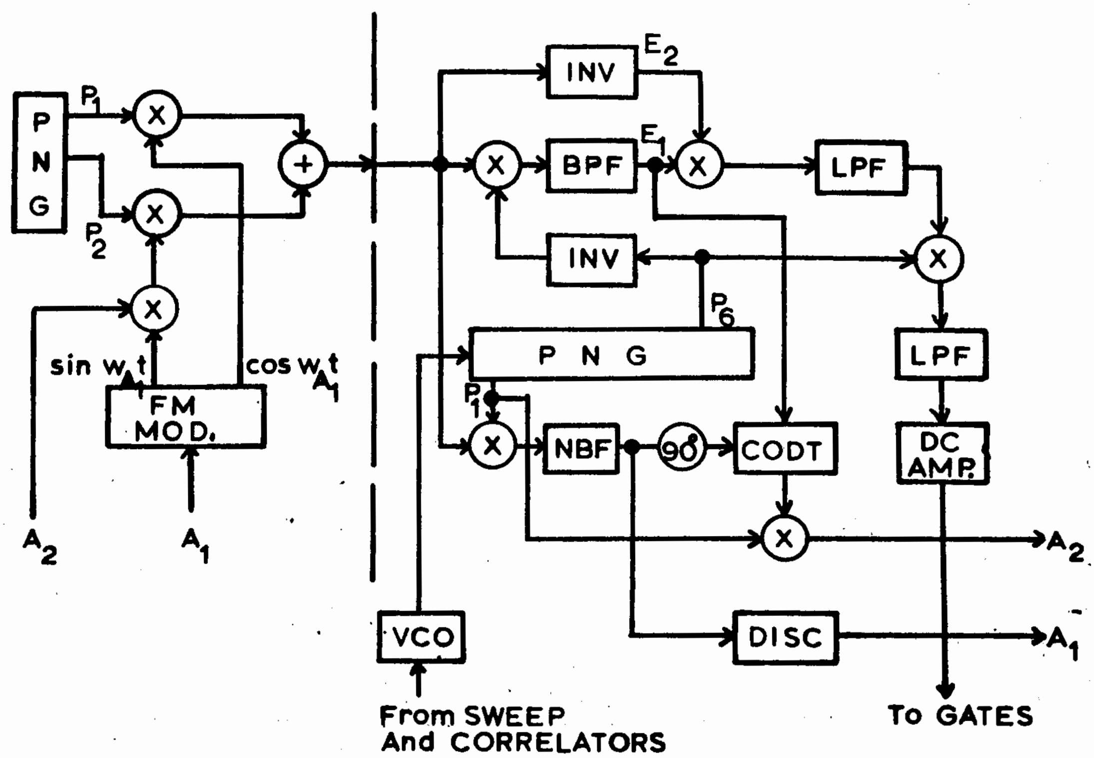


FIGURE 39

detection system equations accomplish this:

$$P_6^*(P_1 \cos \omega t + P_2 \sin \omega t) = P_2 \cos \omega t + P_1 \sin \omega t$$

$$(P_1 \cos \omega t + P_2 \sin \omega t)^*(P_2 \cos \omega t + P_1 \sin \omega t) = P_1^* P_2 \cos^2 \omega t + P_2^* P_1 \sin^2 \omega t$$
$$+ P_1 P_1^* \tfrac{1}{2} \sin 2\omega t + P_2 P_2^* \tfrac{1}{2} \sin 2\omega t$$

With LPF, $E_{out} = P_6 (\cos^2 \omega t + \sin^2 \omega t) = P_6$

Thus, by balanced modulating the output P_6 by P_6 the information or synchronization channel is attained. A unique characteristic of this system is that coherent AM detection is possible concurrently with FM detection utilizing the same frequency. In Figure 39 a dual AM and FM channel is shown which demonstrates this. Since the cosine brance is a constant amplitude FM, this is utilized in the synchronization system to yield the DC control voltage for the VCO. The synchronization system shown in Figure 40 is basically a Spilker system adapted to this transmitted waveform. The PNG shown is the same PNG shown in Figure 39 and the systems are linked to each other as shown. The actual output of the last product device in the synchronization branch of Figure 39 is not precisely P_6 due to the modulation A_2 which is a suppressed carrier modulation. The actual input at this point is $P_6 \left(\frac{1}{2} + \frac{A_2^2}{2}\right)$;(with one branch modulated).
From this it is seen that the P_6 amplitude varies and the DC output of the last product device therefore varies slightly with the A_2 power input. However, all of these variations are additive and will therefore enhance the synchronization stability.

Since there are N triads in a sequence of length N, one might ask if multiplexing by linear addition of the various triad combinations produces a signal which can be separated at the receiver by

SYNCHRONIZATION SYSTEM, AM-FM COHERENT SYSTEM

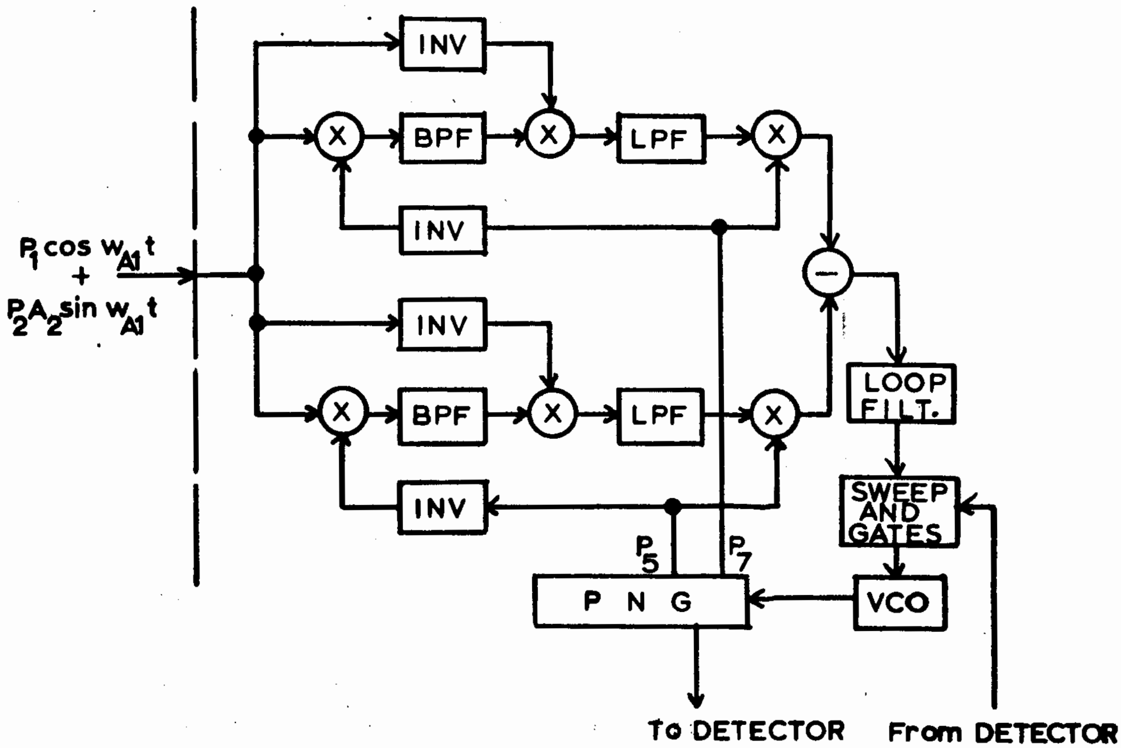


FIGURE 40

using these techniques. The answer is yes but under restricted conditions. Considering a 2 channel system with carriers ω_1 and ω_2 closely spaced in frequency, and each channel modulated like the system of Figures 39 and 40. The system equations are given below:

$$S(t) = P_1 \cos\alpha_1 + A_2 P_2 \sin\alpha_1 + P_2 \cos\alpha_2 + A_3 P_3 \sin\alpha_2 \qquad \text{where } \alpha_i = \omega_i t$$

$$P_6^* S(t) = E_1 = P_2 \cos\alpha_1 + P_1 A_2 \sin\alpha_1 + P_2 P_6^* \cos\alpha_2 + P_3 P_6^* A_3 \sin\alpha_2$$

$$E_2 = P_1^* \cos\alpha_1 + P_1^* A_2 \sin\alpha_1 + P_2^* \cos\alpha_2 + P_3^* A_3 \sin\alpha_2$$

$$\begin{aligned}
E_1 E_2 = {} & P_6 \cos^2\alpha_1 + P_6 A_2^2 \sin^2\alpha_1 - P_6^* \cos^2\alpha_2 - P_6^* A_3^2 \sin^2\alpha_2 \\
& - A_2 \cos\alpha_1 \sin\alpha_1 - \cos\alpha_1 \cos\alpha_2 + P_7 A_3 \cos\alpha_1 \sin\alpha_2 \\
& - A_2 \cos\alpha_1 \sin\alpha_1 + A_2 P_6 \cos\alpha_2 \sin\alpha_1 + P_6 P_7^* A_2 A_3 \sin\alpha_1 \sin\alpha_2 \\
& - \cos\alpha_2 \cos\alpha_1 + P_6 A_2 \cos\alpha_2 \sin\alpha_1 + P_6 P_7^* A_3 \cos\alpha_2 \sin\alpha_2 \\
& + P_7 A_3 \sin\alpha_2 \cos\alpha_1 + P_6^* P_7 A_3 A_2 \sin\alpha_2 \sin\alpha_1 + P_6^* P_7 A_3 \cos\alpha_2 \sin\alpha_2
\end{aligned}$$

$$\begin{aligned}
E_1 E_2 = {} & P_6 \left[(1 - A_2^2)\cos^2\alpha_1 + A_2^2\right] + P_6 \left[(1 - A_3^2)\cos^2\alpha_2 + A_3^2\right] \\
& - 2A_2 \cos\alpha_1 \sin\alpha_1 - 2\cos\alpha_2 \cos\alpha_1 + 2P_6 A_2 \cos\alpha_2 \sin\alpha_1 \\
& + 2P_7 A_3 \cos\alpha_1 \sin\alpha_2 + P_7 P_6^* A_2 A_3 \sin\alpha_2 \sin\alpha_1 + 2P_7 P_6^* A_3 \cos\alpha_2 \sin\alpha_2
\end{aligned}$$

Consider the case $\alpha_1 = \alpha_2$ with LPF_{f_0} at the output.

$$E_1 E_{2_{LPF}} = P_6 \left(1 + \frac{A_2^2}{2} + \frac{A_3^2}{2}\right)$$

This is the synchronizing signal of the system.

Considering $\alpha_1 = \alpha_2$, the first and second terms yield $P_6(1 + \frac{A_2^2}{2} + \frac{A_3^2}{2})$. The third term is eliminated by LPF at cutoff f_o, and the fourth term yields a -1 volt DC. The other terms are eliminated by the LPF which passes the first two terms and the fourth term. Upon balanced modulation by P_6, the DC control voltage is realized with additive stability dependent upon the channel modulation powers. The above system appears to be the most practical with a single FM channel providing the carrier for the AM multiplexed channels.

To add to the discussion of the FM PN systems, a final time multiplexed FM PN sub-carrier system is shown in Figure 41. In this system, the FM can be either wideband such that $\beta >> 1$, and $B = 2 \Delta f$ or narrow band such that $B = 2f_{max}$ for $\beta << 1$, where $\beta = \Delta f / f_{max}$. The non-coherent method of synchronization is used in this case, and processing gain occurs at the first balanced modulator. The processing gain is $f_o/2f_{max}$ for the narrow band FM and $f_o/2\Delta f$ for the wide band FM. The multiplexing is accomplished through a time division multiplex syxtem synchronized with the PNG to eliminate channel ambiguities. In this case, coding is accomplished in two ways. The PN sequence is the natural code in itself, but correct time multiplexing order with relation to the PNG must also be attained before synchronization can be accomplished. As long as the triplet remains adjacent, the sampler can be stepped to any other triplet in a random order as long as the knowledge of the stepping order is known by the receiver. With each order a different and unique sequence is generated, with possible small correlation peaks at one unique switch setting but the next jump would yield a non-correlated output and the sweep generator would continue to sweep the PNG to a sustained correlation peak, at which time lock-on occurs.

No routing gates are required at the discriminator inputs as during the time the central switch of the triplet coincides with the channel desired, the balanced modulator output of that channel is a narrow band FM pulse of length coinciding with the sample length.

PN FM TIME DIVISION MULTIPLEX SYSTEM

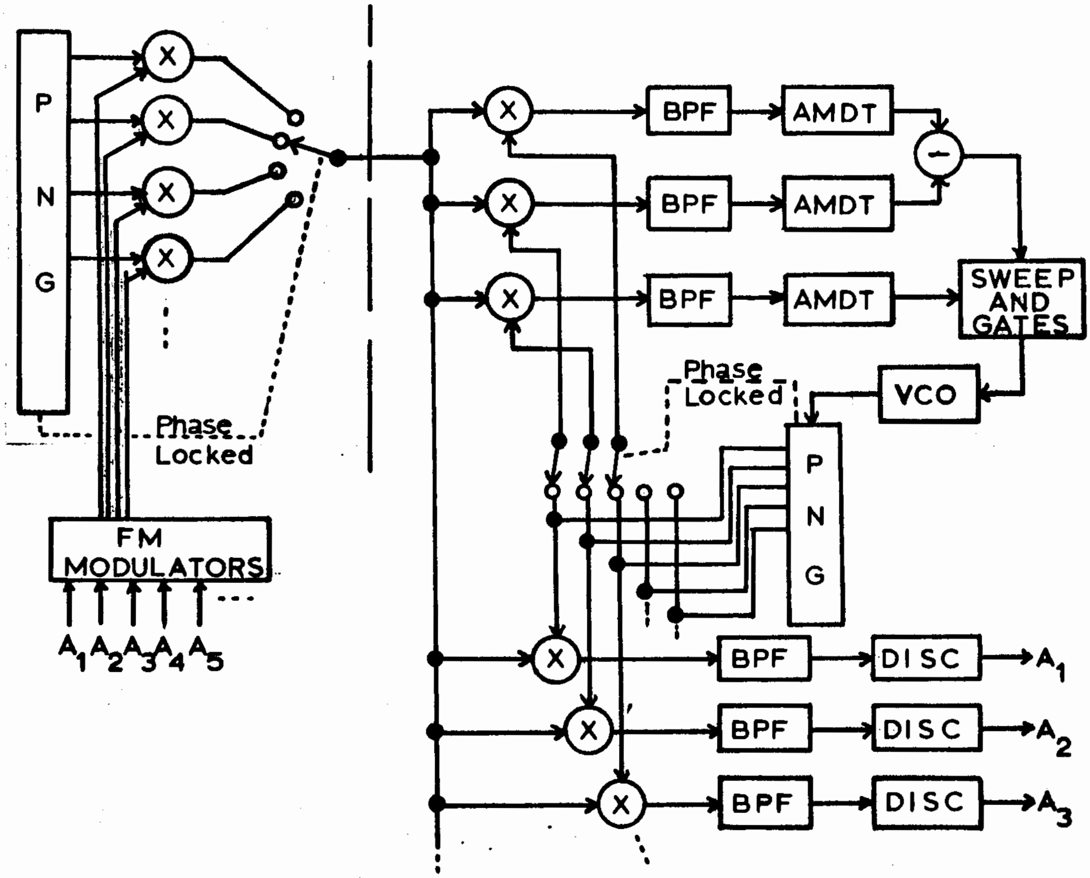


FIGURE 41

When the switch is elsewhere, the balanced modulator output is another wide-band PN modulated FM signal, a small portion of which passes the BPF, but the output signal levels are negligible. Sample lengths in this system can be less than the PN sequence length and the switch positions are set to coincide with specific PNG conditions which remove channel ambiguity.

This concludes the discussion of atmospheric transmissions utilizing PN carriers and sub-carriers. The transmission medium poses few restrictions other than noise effect on the systems discussed. Only at extreme distances do phase shifts and multiple path effects degrade the above systems. These effects are greatly amplified when water is considered a medium of transmission for a wide band signal.

8. Submarine Communications Applications.

When the medium of water is considered for transmitting wide band signals, several characteristics of the medium must be known before the system and the transmission spectrum are chosen. First, interference of underwater shelves and mountains cause serious phase problems. As frequency is increased, say from 1 KHz to 1.5 KHz, the absorption loss for a distant reflected signal is greatly increased whereas the effect on the absorption of the directly transmitted signal is much less. When the propagation loss only is considered, lower frequencies are desirable. If the noise spectrum alone is considered, there is less background noise at the higher frequencies. Finally, if directivity alone is considered, the transducer directivity increases with frequency. A compromise central frequency range therefore may be chosen at 1.0 KHz to 4.0 KHz to yield long range capability with less background noise, and less multipath interference.

Since the processing gain in a given PN sub-carrier system is strictly dependent upon the ratio f_o/f_{max} , the information channel must be extremely narrow with respect to 2.5 KHz. This suggests a slow synchronized teletype channel of 60 words per minute, with a 7 to 5 bit word conversion in the system. With a 5 bit word transmitted in 165 ms a bandwidth of 30 Hz is required. The PN carrier system is highly sensitive to phase shifts, multipath interference, and filtering, all of which occur in the required system. Therefore, the PN sub-carrier system is best adapted to underwater transmissions. Since the sub-carrier must be an order of magnitude higher than the information channel, and an order of magnitude lower than the carrier, and a 10 dB S/N receiver output is required for a .01 % probability of error, a clock frequency of 500 Hz is a good compromise, which yields a processing gain of 12.2 dB. This could be considerably increased by going to a lower bit rate, recording, and playing back at regular teletype speeds. With this clock frequency a severe restriction is imposed on the sequence length with $N_{max} = f_o/2f_{max}$, which dictates a sequence length of 7. If range is compromised, and the transmission frequency is increased to 25 KHz or even 250 KHz, with a 2.5 KHz and a 25 KHz clock frequency respectively, this restriction is increased to $N_{max} = 42$ and $N_{max} = 420$ respectively. The processing gains for these two cases is 82 or 19.1 dB and 820 or 29.1 dB. These are more practical values with the operating values for any one given transmission

SUBMARINE COMMUNICATIONS LINK USING PN AM-FM COHERENT DETECTION

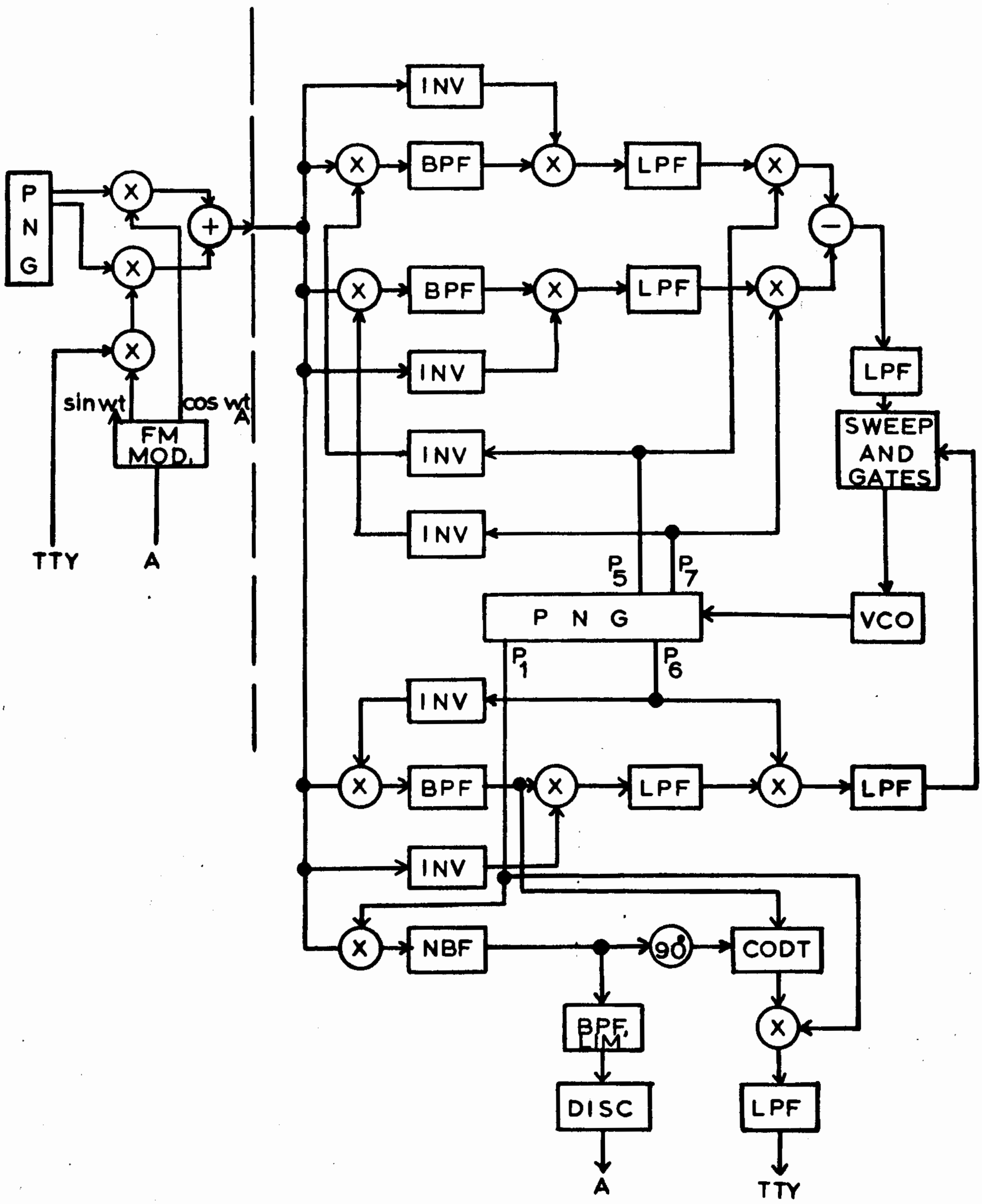


FIGURE 42

dependent upon the communication range, and the communication range to detectable range ratio desired. The detectable range is that which yields an input S/N greater than one, and a threshold detector is capable of determining there is a signal present. This system is shown in Figure 42 utilizing the coherent AM-FM system of Figure 39 and 40. Note that with a teletype signal input as shown, there will be no amplitude variation in the synchronization system. Also several channels of TTY could be transmitted up to the cross-talk limitation dictated by the sequence length. In addition, FSK could be sent over the FM carrier and transmitted reference. This system is also adaptable to voice and NTDS communication links with lower processing gains resulting. The higher carrier frequencies would be required such as 250 KHz and shorter ranges would be experienced.

The versatility of this system in a communication silence environment make it highly desirable for use in any transmission medium. Thus, with such a system the submarine as well is capable of sub-noise communications in all situations.

9. Conclusion.

With the advent of integrated circuitry, the systems discussed in this paper can be easily implemented with a minimum number of discrete parts and at an economical cost. With systems such as these operating at sub-noise levels at predetermined ranges, a complete spectrum of PN transmission frequencies could be superimposed upon the present spectrum without adverse interference, and today's spectrum signal density would be greatly reduced, while military transmission security would be greatly enhanced. The multiplex systems would be ideal for use in an intra fleet communications network and NTDS communications in radio silence conditions, or on a continual basis with frequency and code changing daily.

With code versatility, frequency versatility, and the variety of systems available as shown in this paper, the jamming and ECM immunity becomes extremely high as detection must precede any ECM attempts. The threshold detection range can be made literally within the optical visibility range. Thus, only knowledge of the sequence, frequency, and specific system transmitting technique would lead to compromise. Since many of the components of one variation are interchangeable with those of another, all three of these parameters could be varied daily, or hourly, and the probability of compromise would be extremely small.

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The current literature approaches the subject of communications from the optimum detector point of view, utilizing the principles set forth by Shannon, Davenport, Root, Bode, Peterson, Birdsall, Fox, Weiner, Siebert, Middleton, and many others to carry out detailed analysis of specific detectors, filters, and synchronization processes. This paper approaches communications from a systems point of view, dealing specifically with the family of pseudo-noise systems. The systems discussed are categorized into two groups,the pseudo-noise carrier and the pseudo-noise sub-carrier systems, with emphasis on multiplex techniques. All of the systems discussed are negative dB S/N systems with the exception of a wide-band TV video channel, and a detailed analysis of a representatave pseudo-noise sequence of length 15 is given as a background in the auto-correlation functions, power spectral densities, and self-noise spectra which are the characterizing parameters of these waveforms.

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