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MEASUREMENT OF THICKNESS OF LOSSY LAYERED MEDIA

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THE DESIGN OF A HIGH DYNAMIC-RANGE
CORRELATOR RADAR FOR THE MEASUREMENT OF
THICKNESS OF LOSSY LAYERED MEDIA

by

Hafez Mohamed Hafez, B.Sc.Engg., M.Sc. Engg.

A thesis submitted to the Faculty of Graduate Studies
and Research in partial fulfillment of the requirements
for the degree of

Doctor of Philosophy

Faculty of Engineering

Carleton University

Ottawa, Ontario

Canada

December 1978

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ABSTRACT

The measurement of lossy medium thickness using modulated RF signals is often limited by the interference between the sidelobes of a strong reflection from one surface and the weak reflection from the other. The theory and design of a novel VHF radar system which has the capability of cancelling all the time sidelobes of all the radar echos is presented in this thesis. The new system employs a digitally modulated transmitted waveform (Pseudorandom-MSK signal) and a correlator receiver which correlates the received signal with a reference signal forming two quadrature channels. The receiver outputs are then processed to suppress the time sidelobes of the two quadrature signals.

A potential application of the system is the measurement of the depth of fresh water in remote lakes and rivers. A detailed study of the electromagnetic characteristics of fresh water is given, with a description of a novel broadband technique for the measurement of the attenuation rate and the dispersion characteristics of fresh water.

Finally, the implementation of the novel radar system is presented with the proper choice of the system parameters to suit the particular problem of measuring the depth of fresh water. The experimental performance of the system was tested under laboratory conditions, and a significant

reduction of the sidelobe levels was demonstrated. The system described in this thesis provides at least 12 dB improvement in mainlobe/sidelobe ratio relative to conventional radar systems.

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LIST OF IMPORTANT SYMBOLS

$\bar{\epsilon}$	complex permittivity
ϵ	relative complex permittivity
ϵ'	real part of ϵ
ϵ''	the negative of the imaginary part of ϵ
ϵ_0	permittivity of the high frequency end of the dipolar dispersion region
ϵ_s	static relative permittivity ($\omega=0$)
ϵ_v	permittivity of space
τ	dipole relaxation time
γ	complex propagation constant
α	real part of γ
β	imaginary part of γ
σ_i	ionic conductivity
σ_p	equivalent dipole conductivity
δ	dielectric loss angle
σ_0	radar cross-section per unit length
τ_0	time delay per unit length
ρ_0	time resolution
B_n	noise bandwidth
H	thickness of the lossy medium
L_1	losses at air-water interface
L_2	losses at bottom-water interface
$\theta(t)$	complex phase of the modulated signal

T	clock period
u_k	bipolar data
u_I, u_Q	in-phase and quadrature bipolar data
N	sequence length
t_1	delay time axis
$I(t_1), Q(t_1)$	in-phase and quadrature correlation functions
G_I, G_Q	in-phase and quadrature correlator outputs
γ	receiver output
ψ	unknown phase of the received signal
f_c	clock frequency
Δf_c	offset of the clock frequency
T_o	code duration

CHAPTER 1

INTRODUCTION

1.1 INTRODUCTION

In recent years, increased emphasis has been placed on remote sensing as a technique for understanding and evaluating natural resources. Radar sensors are now being used to explore vast areas of the earth. For example, radar sensors are used to study oceans and their currents and for applications in geophysics, mineral exploration, topographic mapping, agriculture and forestry [1][2],[3][4]. Radar maps are providing an inventory of potential mineral resources, new transportation routes and sites for agriculture [1]. The possible applications of radar for major economic and social improvement seem almost limitless.

Fresh water is one of the most important natural resources. It is an essential component for the agricultural industry, and is also important as an alternative means for commercial transportation. Therefore, the development of an airborne radar sensor to explore and evaluate fresh water supplies should be one of the most potentially interesting and rewarding applications of radar remote sensing techniques.

The evaluation of fresh water supplies in remote lakes and rivers requires the development of an airborne

radar sensor capable of estimating the water depth at different locations. The operation of such a radar is based on measuring the differential time delay between the two radar echos from air-water and water-bottom interfaces (see Fig. 1.1a). However, because of the inevitable losses of the electromagnetic waves while propagating through the water and at the interfaces, the reflection from the bottom interface will be much weaker than the reflection from the top interface. This will result in a self-interference problem between the weak echo and the time side lobe of the strong echo as illustrated in Fig. 1.1b.

In this thesis, a novel solution for the self-interference problem is presented. The solution provides a new approach to the design of the transmitted waveform in order to remove self-interference between radar echos. In addition, novel experimental and analytical models of the electromagnetic properties of fresh water at VHF frequencies are presented.

In Section 1.2, the background and objective of the thesis are presented in greater details, while the thesis organization is presented in Section 1.3.

(a)

(b)

Fig. 1.1. Measurement of the fresh water depth using an air-borne radar sensor.
 (a) Model
 (b) The two echo responses at the receiver output.

1.2 BACKGROUND AND OBJECTIVE

Although radar sensing techniques have been developed to a high degree of sophistication, the state-of-the-art for terrain mapping and exploration is still advancing at a rapid rate, and some problems are still without satisfactory solutions.

One of the major limitations on the performance of radar sensors is their limited dynamic range in comparison with the wide dynamic range of the natural terrain reflectivities. This is particularly important in applications that require electromagnetic probing of the terrain surface in order to explore the layered structure of the ground, or to measure the thickness of lossy layered media.

An important example of such a problem is the estimation of the fresh water depth using an air-borne radar sensor. The importance of this applications stems from the fact that an air-borne surveillance sensor can help evaluate the amount of fresh water in remote areas. It can also help estimate the suitability of remote lakes and rivers as alternative transportation routes for industrial and commercial purposes.

Three techniques may be considered for the measurement of the depth of fresh water, namely: acoustic, optical, and electromagnetic.. Acoustic systems have been used for underwater communications and detections and for monitoring

the thermal structure of the lower atmosphere [5]. However, very little information is available about the use of acoustic air-borne sounders for sensing the depth of water. Optical methods, on the other hand, can be used to provide information as to the location and depth of shallow sea water features. However, the light absorption by a coloured water represents a doubtful problem [6]. The electromagnetic technique is the third technique which can be used to measure the depth of fresh water. It has potential economical advantages over the optical system and is the technique studied in this thesis. The EM technique is based on measuring the differential time delay between the reflected waves from the air-water and the water-bottom interfaces. A possible measurement method is shown in Fig. 1.1.

In the consideration of developing an air-borne radar system capable of measuring the depth of fresh water, an adequate knowledge of the attenuation characteristics of the water is necessary. The lack of such information, particularly in the VHF band of frequencies, has suggested that theoretical and experimental studies of the fresh water characteristics must be made. A simple novel technique is described in the first part of this thesis for broadband measurement of the attenuation rate of EM energy in fresh water. The technique employs a network analyzer to measure the S-parameters of a broadband coaxial transmission line filled with a sample of fresh water. The

experimental results presented in this thesis indicate that the attenuation rate at low frequencies is surprisingly low (less than 2 dB/m) and relatively independent of frequency up to a frequency of approximately 100 MHz. For frequencies higher than 100 MHz, the attenuation rate increases rapidly. For water conductivity of $.01 \text{ } \Omega\text{-meter}^{-1}$ (typical for fresh water) and at 100 MHz, the attenuation rate is only 2 dB/m. This low value suggests that it should be possible to develop a VHF radar sensor capable of sensing the depth of relatively shallow fresh water.

In spite of these encouraging results, the detection capability of the system is still seriously limited by the slowly decaying time sidelobes of the strong reflection of the air-water interface. This residual response can be an order of magnitude higher than the weak reflection from the bottom terrain, resulting in a severe self-interference problem as illustrated in Fig. 1.1b.

This problem cannot be solved by increasing the transmitted power or improving the system resolution. It can only be solved by careful design of the transmitted waveform and by improving the receiver filtering process [7].

A novel solution to this problem is presented in the second part of the thesis. The solution is based on two points. First, a special waveform is designed to allow for

sidelobe cancellation. Second, a CW waveform is utilized in order to transfer the critical filtering into the audio band, where wideband filters with flat responses are easy to construct. The transmitted waveform is a digitally modulated VHF signal, where a cyclic pseudorandom binary sequence (PRBS) is used to modulate a VHF carrier using the relatively new modulation technique known as Minimum (Frequency) Shift Keying. It was observed that the autocorrelation function of the MSK modulated signal has a peak at the origin and sinusoidal sidelobes. By forming two quadrature channels in the receiver, these sidelobes can be virtually cancelled by a simple delay and addition of the two quadrature channels. The correlation technique used in the system utilizes a slight offset of the PRBS's clocks of the transmitted and reference signal. This slight offset of the clock frequencies allows for an integration over a large number of PRBS periods, and in effect, the autocorrelation output will have a spectrum at a much lower frequency than the range.

new system is, in effect, a reflectometer with a wide dynamic range. It can be used to solve a wide variety of remote sensing problems, where the dynamic range of the natural terrain reflectivities vary widely.

In developing this thesis, we attempt to maintain a balance between theoretical and experimental work; and also

to maintain a balance between the academic value of work and the practical and potential economical value.

1.3 THESIS ORGANIZATION

This thesis is divided into two parts. The first deals with the definition of the radar problem and includes theoretical and experimental studies of the VHF electromagnetic characteristics of fresh water, the nature of the air-water interface reflection characteristics, and the water-bottom interface characteristics. The radar system requirements such as bandwidth, dynamic range, transmitted power, etc. are included as well as a survey of the classical radar systems and their dynamic range limitations. The second part of the thesis contains the new approach of solving the self-interference problem and the proposed technique to cancel the sidelobes of the strong reflection. Also, a detailed description of the practical implementation of the system and the achieved results are included.

In Chapter 2, the theoretical and experimental studies of the electromagnetic energy losses in fresh water are presented. The results for the measured attenuation rate in samples of fresh water obtained using a novel broadband measurement technique at frequencies between 33 and 363 MHz are presented. The remainder of the Chapter deals with the characteristics of the top and bottom reflections

and a statement of the necessary system requirements. In Chapter 3, an evaluation is made of the suitability of classical radar systems (e.g. short pulse radar, FM-CW, etc.) for this particular application. The assessment of each system is based on its ability to solve the interference problem described in the previous section. The novel correlator receiver for sidelobe suppression is presented in Chapter 4. This Chapter also includes the error analysis and the effects of errors on the system performance. A detailed description of the system hardware is given in Chapter 5, together with the results obtained under laboratory conditions by simulating the targets by means of transmission lines. The system performance, advantages and disadvantages are summarized in Chapter 6, with final conclusions and recommendations for further work.

CHAPTER 2

ATTENUATION RATE IN FRESH WATER AND STATEMENT OF SYSTEM PARAMETERS

2.1 INTRODUCTION

The design of a radar sensor for fresh water depth measurements requires an adequate knowledge of the electrical characteristics of the water. It also requires knowledge of the air-water and water-bottom interfaces. This information will determine the system design requirements, such as the operating frequency, system bandwidth, etc. In this Chapter, these three points are discussed.

In Section 2.2, theoretical and experimental studies of the attenuation rate in fresh water are given. First, the dispersion theory in polar liquids is reviewed, and the attenuation rate in pure water is determined based on this theory. The effect of added ionic solutions is considered next, and the attenuation mechanism associated with these solutions is explained. This is followed by a summary of the results of previous measurements. In the last part of the section, a novel measurement method for the attenuation rate is introduced, together with the practical results of laboratory measurements.

Section 2.3 discusses the general characteristics of the air-water and water-bottom interfaces. The nature of

wave reflection at the two interfaces is described, and the reflectivity dependence on surface roughness, frequency and wave polarization is discussed.

The system parameters and constraints are discussed in Section 2.4. The operating frequency, bandwidth, dynamic range and antenna height are determined based on the analysis given in Sections 2.2 and 2.3. The Chapter is concluded in Section 2.5.

2.2 ATTENUATION RATE IN FRESH WATER

2.2.1 Introduction

The original theory of dispersion in polar liquids (such as water) is due to Debye [8]. The complex permittivity of water as predicted by Debye's theory has been confirmed by extensive measurements at microwave frequencies [9]-[15]. C.H. Collie, J.B. Hasted and D.M. Riston [9] have described three different methods to measure the complex permittivity of water. These are:

- 1) The direct measurement of the absorption coefficient in a rectangular waveguide containing water;
- 2) The measurement of the absorption coefficient in two cylindrical waveguides containing water, the two diameters being selected to be just above and just below the cut-off value for the waveguide used;

- 3) Measurement of the change in Q of an H_{01} resonator on the introduction of an axial capillary containing water.

The measurements on water were carried out for wavelengths of 10, 3 and 1.25 cm. T.J. Buchanan and E.H. Grant [10] have described another method for wavelengths between 4 and 50 cm. In their measurement technique (which utilizes a balanced bridge), the transmitted signal is divided between a water absorption cell and a piston attenuator, subsequently being combined at a magic T junction and fed to a receiver. When the signals are equal and in anti-phase, the receiver shows no response. The same technique has been used by J.B. Hasted and H.M. El-Sabeh [11] for wavelengths of 1.26, 3.28 and 9.22 cm. J.A. Soxton and J.A. Lane [12] reported the use of superhetrodyne receivers for .62, 1.24 and 3.21 cm wavelengths; their basic technique was to maintain constant output power from the receiver by adjusting the attenuation in a piston attenuator with and without the absorption cell. E.H. Grant [13] used a short-circuited line, and by measuring the position of particular points in the standing wave, it was possible to determine the absorption coefficient of the liquid. G.A. Vilulich and D.F. Evans [14] described a low-frequency balanced bridge to measure the permittivity of water at frequencies between .5-100 kHz.

Although the behaviour of water has been studied extensively at microwave frequencies and low frequencies (.5-100 kHz), very little has been reported concerning the electromagnetic properties of water, and particularly fresh water, in the VHF frequency range (30-300 MHz).

The attenuation of the electromagnetic waves in fresh water occurs because of two different mechanisms. At high frequencies, the waves are attenuated mainly because of the dipole relaxation phenomenon, while at low frequencies the ionic conductivity of the dissolved salts is the main reason for attenuation [15]. It has been recognized [15] that the transition region in the case of fresh water lies in the VHF range. In this band, the effects of both the dipole relaxation and the ionic conductivity should be taken into account.

The theory of anomalous dispersion in polar liquids is presented in Section 2.2.2. The attenuation rate in pure water and the effect of ionic conductivity are discussed in Sections 2.2.3 and 2.2.4, respectively. Section 2.2.5 summarizes the significant results reported to date by others. A broadband S-parameter transmission measurement technique for determining the attenuation rate in fresh water is described in Section 2.2.6. The technique employs an HP 8410A network analyzer and a broadband transmission cavity (33-363 MHz range) to provide attenuation results which agree well with theory.

2.2.2 Theory of Anomalous Dispersion in Polar Liquids

In its original form, Debye's theory accounts satisfactorily for the observed behaviour of many simple solutions of polar substances in non-polar solvents.

It has been necessary, however, to modify the theory somewhat in the consideration of pure polar liquids such as pure water, because of the strong interaction fields between the molecular dipoles in such liquids, which are neglected in Debye's theory. The necessary modification of Debye's theory has been introduced by Onsager's [16] and Frohlich and Sack [17].

The complex permittivity of the pure water, ϵ , according to the modified Debye theory may be expressed as

$$\epsilon = \epsilon' - j\epsilon'' = \epsilon_0 + \frac{\epsilon_s - \epsilon_0}{1 + j\omega\tau} \quad (2.1)$$

where ϵ_s = static permittivity

ϵ_0 = permittivity of the high frequency end of the dipolar dispersion region

τ = relaxation time of the dipole

and ω = radian frequency.

From Eqn. (2.1), we may derive the following relations:

$$\epsilon' = \frac{\epsilon_s + \epsilon_0 x^2}{1 + x^2} \quad (2.2)$$

and

$$\epsilon'' = \frac{x(\epsilon_s - \epsilon_0)}{1+x^2} \quad (2.3)$$

where $x = \omega\tau$. ϵ_0 has a constant value of 4.9 which is independent of temperature [12], while ϵ_s and τ are strong functions of temperature. Table I gives the commonly accepted values of ϵ_s and τ as functions of temperature [18].

Table I. ϵ_s and τ vs temperature [12].

TEMP °C	ϵ_s	τ Picosec
0	88.2	18.7
10	84.2	13.6
20	80.4	10.1
30	76.7	7.5
40	73.1	5.9
50	69.8	4.7

Using Eqns. (2.2) and (2.3) together with the values of ϵ_s and τ given in Table I, one can compute the values of ϵ' and ϵ'' as functions of frequency over the dipole dispersion region at a certain temperature.

Figs. 2.1 and 2.2 show such relations for three different temperatures.

Fig. 2.1. The real part of the complex permittivity.
(Pure water, $\epsilon_0 = 4.9$)

Fig. 2.2. The imaginary part of the complex permittivity.
(Pure water, $\epsilon_0 = 4.9$)

2.2.3 Attenuation Rate in Pure Water

Consider a homogeneous medium and a uniform planar wave. The wave equation may be written in the form of the Helmholtz equation as:

$$\nabla^2 E - \gamma^2 E = 0 \quad (2.4)$$

where

$$\gamma^2 = j \omega \mu_r (\sigma_i + j\omega\bar{\epsilon}) \quad (2.5)$$

$\bar{\epsilon}$ is the complex permittivity of the medium given by

$$\bar{\epsilon} = \epsilon_r (\epsilon' - j\epsilon'') \quad (2.6)$$

and σ_i = medium conductivity in Ω -meter⁻¹

μ_r and ϵ_r are the permeability and permittivity of space, respectively. If the wave travels in the y-direction, one possible solution of Eqn. (2.4) is:

$$\tilde{E}(y,t) = \text{Re}\{E_0 \cdot \exp[j(\omega t - \beta y)]\} \cdot e^{-\alpha y} \quad (2.7)$$

where E_0 is the wave amplitude at $y=0$ and α and β are the real and imaginary parts of the propagation constant γ , respectively. Then,

$$\alpha = \text{Re}\{[j \omega \mu_r (\sigma_i + j\omega\bar{\epsilon})]^{1/2}\} \quad (2.8)$$

and the attenuation rate (AR) in dB/m is (-8.686α) . For pure water, the ionic conductivity σ_i is zero, and Eqn. (2.8) can be simplified to:

$$\alpha = \text{Re}\{[j \omega \mu_r \epsilon_r (j\omega\epsilon' + \omega\epsilon'')]^{\frac{1}{2}}\} \quad (2.9)$$

or

$$\alpha = \frac{\omega}{c} [\epsilon' \sec(\delta)]^{\frac{1}{2}} \sin(\delta/2) \quad (2.10)$$

where c is the velocity of light and $\delta = \tan^{-1}(\epsilon''/\epsilon')$.

Finally, the attenuation rate in pure water can be written as

$$(\text{AR})_{\text{pure water}} = -8.686 \frac{\omega}{c} [\epsilon' \sec(\delta)]^{\frac{1}{2}} \sin(\delta/2) \quad (2.11)$$

Using Eqns. (2.11) and (2.2), the attenuation rate of pure water is calculated and plotted in Fig. 2.3 [18], for different values of temperature. Fig. 2.3 shows that, for frequencies less than 50 GHz approximately, the attenuation rate falls rapidly at a rate of approximately 100 dB/m/decade.

2.2.4 Effect of Ionic Solution in Water

In the case of fresh water, there are dissolved salts in the water. The presence of such ions in the water will lead to ionic conductivity, σ_i , and consequently α is now defined by Eqn. (2.8) rather than Eqn. (2.9).

In Eqn. (2.8), the term containing σ_i will be added to the imaginary part of the complex permittivity to yield

$$\epsilon''_{\text{eff}} = \epsilon'' + \frac{\sigma_i}{\omega \epsilon_r} \quad (2.12)$$

Fig. 2.3. Rate of attenuation in pure water.

Although the mechanism of energy absorption arising from dipole relaxation is different from that for ionic conductivity, their effects are the same as far as the attenuation rate is concerned, hence justifying Eqn. (2.12).

The question of whether or not the parameters of pure water (ϵ_s and τ) will be affected by the presence of ions has been addressed by Hasted, Riston and Collie [19]. In their work, they concluded that the presence of appreciable quantities of an electrolyte in water causes a change in the dipole relaxation time and a decrease in the dielectric constant, and in strong solutions, such as some sea water, this is a factor which cannot be neglected. However, in the case of fresh water, the influence of salt on the dipole relaxation time and dielectric constant can be neglected. The total effective conductivity can be given by

$$\sigma_{\text{total}}(\omega, t) = \sigma_i(t) + \sigma_p(\omega, t) \quad (2.13)$$

where t represents temperature and $\sigma_p(\omega, t)$ is the equivalent conductivity of the dipole relaxation. Accordingly, ϵ''_{eff} for fresh water can be written as

$$\epsilon''_{\text{eff}} = \omega \epsilon_r \sigma_{\text{total}}(\omega, t) \quad (2.14)$$

Now, we should be able to use Eqn. (2.11) to calculate the attenuation rate in fresh water by replacing ϵ'' by ϵ''_{eff} in the expression for δ . Using the parameters given in Table I,

the attenuation rate in fresh water has been calculated as a function of frequency and with σ_i and temperature as parameters. The results are plotted in Fig. 2.4.

It can be seen from Fig. 2.4 that, for any particular σ_i , the frequency range can be divided roughly into three regions; in the lower band, the effect of the dipole relaxation is very small compared to the ohmic losses; in the middle range, both σ_i and σ_p must be considered, while in the upper band, the effect of σ_p predominates. The effect of temperature is very small in the lower band while it is more pronounced in the middle and upper bands.

An important and surprising result of Fig. 2.4 is that the attenuation rate for $\sigma_i = .01$ $\Omega^{-1}\text{-meter}^{-1}$ is only 2 dB/m at 100 MHz. This value of σ_i is considered by many references (see for example [20]) as a typical value of fresh water conductivity.

2.2.5 Summary of the Results of Previous Measurements

Extensive measurements have been made to study the dielectric properties of water, and to confirm Debye's theory for water and other polar liquids. Some of the important experiments were due to Lane and Saxton [12]; Collie, Hasted and Riston [9]; and Hasted and El-Sabeh [4]. These results are summarized and tabulated in Tables II, III and IV. The index of refraction n , and the absorption

Fig. 2.4 Attenuation rate in water (effect of ionic conductivity).

coefficient k in Tables II and IV are related to the complex permittivity by the following relation:

$$(n-jk)^2 = \epsilon' - j\epsilon'' \quad (2.45)$$

Table II. Dielectric properties of water (Lane and Saxton).

TEMP °C	$\lambda=6.2\text{mm}$		$\lambda=1.24\text{cm}$			$\lambda=3.21\text{cm}$	
	k	n	k	n	n	k	n
	meas.	calc.	meas.	meas.	calc.	meas.	calc.
0	2.04	3.45	2.77	4.75	4.70	2.89	7.14
10	2.37	3.94	2.90	5.45	5.42	2.44	7.80
20	2.59	4.44	2.86	6.12	6.10	2.00	8.14
30	2.70	4.90	2.67	6.70	6.62	1.60	8.24
40	2.70	5.35	2.41	7.10	7.02	1.29	8.22
50	2.63	5.73	2.13	7.30	7.20	1.08	8.13

Table III. Complex permittivity of water (Hasted and El-Sabeh).

TEMP °C	$\lambda=1.267\text{cm}$				$\lambda=3.282\text{cm}$			
	ϵ'	ϵ'	ϵ''	ϵ''	ϵ'	ϵ''	ϵ''	ϵ''
	cal.	meas.	cal.	meas.	cal.	meas.	cal.	meas.
0	15.1	14.5	27.4	27.5	45.8	46.0	41.7	41.6
10	22.0	22.0	32.4	33.0	56.2	56.0	37.8	37.0
20	30.5	31.0	35.6	35.0	63.3	63.0	31.5	31.5
30	38.6	38.7	35.5	35.5	66.2	65.8	25.2	26.0
40	44.0	43.5	33.5	34.0	66.3	66.3	20.9	20.5
50	48.2	48.0	30.5	30.5	65.2	65.5	16.5	16.5
60	50.9	50.5	26.9	26.5	63.5	63.5	13.4	13.5

TEMP °C	$\lambda=9.22\text{cm}$			
	ϵ'	ϵ'	ϵ''	ϵ''
	cal.	meas.	cal.	meas.
0	78.2	80.0	26.5	26.5
10	79.0	79.3	19.5	19.4
20	77.3	77.8	13.8	13.9
30	75.2	75.8	10.3	10.2
40	72.1	73.0	8.1	8.4
50	69.5	69.7	6.3	6.5
60	66.6	66.0	4.7	5.0

Table IV. Dielectric properties of water (Collie, Hasted and Riston).

TEMP °C	ϵ_s	$\lambda=1.27\text{cm}$				$\lambda=3.21\text{cm}$			
		n cal.	n exp.	k cal.	k exp.	n cal.	n exp.	k cal.	k exp.
0	882	4.92	4.94	2.81	2.86	7.31	7.28	2.83	2.86
10	842	5.60	5.55	2.92	2.91	7.88	7.73	2.39	2.43
20	803	6.20	6.25	3.82	2.86	8.16	8.08	1.96	1.97
30	767	6.71	6.75	2.64	2.67	8.25	8.11	1.58	1.57
40	731	7.02	7.00	2.39	2.40	8.21	8.20	1.28	1.29
50	698	7.23	7.26	2.10	2.11	8.12	8.15	1.04	1.05
60	666	7.34	7.30	1.83	1.87	8.01	7.99	.856	.865
75	621	7.34	7.35	1.50	1.52	7.80	7.82	.663	.672

It should be noted that most of the reported measurements were at microwave frequencies with the exception of Cooper's work [21] which was conducted in the frequency ranges of (.93-13 MHz) and (690-4320 MHz). Also, the studies of the effect of salts on the properties of water have been made on strong solutions, the minimum concentration of salts being 2 or 3%. This is more useful in characterizing sea water than fresh water.

2.2.6 Attenuation Rate Measurements of Fresh Water Samples

The sample holder shown in Fig. 2.5 was constructed in order to perform attenuation rate measurements on fresh water samples. The holder is a section of coaxial line 20" in length, with diameters of the inner and the outer conductors equal to .25" and .75", respectively. The space between the two conductors is filled with water when under

Fig. 2.5. Sample holder.

test and the characteristic impedance of the loaded line is about 7.3 ohms. The sample holder resonates (maximum power transfer at resonance) when the electrical length is an integer number of half wavelengths. For the dimensions given in Fig. 2.5, the resonant frequencies in MHz are given by:

$$f = 33 n \text{ MHz; } n=1,2,3,\dots \quad (2.16)$$

The sample holder is useful over a wide range of frequencies since measurements may be made for $n=1$ to 10 or even higher. It is of interest to note that other techniques described to date are, in general, narrowband in nature.

Fig. 2.6 illustrates the measurement setup. In this arrangement, a standard sweep oscillator is employed to supply a frequency swept signal to the HP 8410A S-parameters test set which is adjusted to measure the S_{21} parameter. The sample holder is connected between the two ports, 1 and 2. An oscilloscope is used to display the variation of $|S_{21}|^2$ with frequency.

For a coaxial line of normalized impedance R (the ratio between the loaded line impedance and the impedance of the measuring system) and length l inserted between ports 1 and 2 of the S-parameters set, the relationship between incident, transmitted and reflected powers ($|a_1|^2$, $|b_2|^2$ and $|b_1|^2$ respectively) is given by [22]:

Fig. 2.6. Measurement setup.

$$\begin{bmatrix} b_1 \\ a_1 \end{bmatrix} = \frac{1}{4} \begin{bmatrix} \sqrt{R} + \frac{1}{\sqrt{R}} & \sqrt{R} - \frac{1}{\sqrt{R}} \\ \sqrt{R} - \frac{1}{\sqrt{R}} & \sqrt{R} + \frac{1}{\sqrt{R}} \end{bmatrix} \begin{bmatrix} (\frac{1}{\sqrt{R}} - \sqrt{R}) e^{-\gamma \ell} b_2 \\ (\frac{1}{\sqrt{R}} + \sqrt{R}) e^{\gamma \ell} b_2 \end{bmatrix} \quad (2.17)$$

Then

$$a_1 = \frac{(R+1)^2}{4R} [e^{\gamma \ell} - \left\{ \frac{R-1}{R+1} \right\}^2 e^{-\gamma \ell}] b_2 \quad (2.18)$$

and the S_{21} parameter is then given by:

$$S_{21} = \frac{b_2}{a_1} = \frac{4R/(1+R)^2}{e^{\gamma \ell} - \left[(1-R)/(1+R) \right]^2 e^{-\gamma \ell}} \quad (2.19)$$

and

$$|S_{21}|^2 = \frac{[16R^2/(1+R)^4] e^{-2\alpha \ell}}{1 + \left(\frac{1-R}{1+R} e^{-\alpha \ell} \right)^4 - 2 \left(\frac{1-R}{1+R} e^{-\alpha \ell} \right)^2 \cos 2\beta \ell} \quad (2.20)$$

where $\gamma = \alpha + j\beta$ is the complex propagation constant of the propagation medium. $|S_{21}|^2$ has its maxima when $\cos 2\beta \ell = 1$, then

$$|S_{21}|_{\max}^2 = \frac{[16R^2/(1+R)^4] e^{-2\alpha \ell}}{[1 - \left(\frac{1-R}{1+R} e^{-\alpha \ell} \right)^2]^2} \quad (2.21)$$

Solving Eqn. (2.21) for the attenuation constant α gives:

$$\alpha = \frac{1}{2\ell} \ln [x + \sqrt{x^2 - y}] \quad \text{neper/m} \quad (2.22)$$

where

$$x = \left(\frac{1-R}{1+R} \right)^2 + \frac{8R^2}{|S_{21}|_{\max}^2 (1+R)^4} \quad (2.23)$$

and

$$y = \left(\frac{1-R}{1+R} \right)^4 \quad (2.24)$$

The attenuation rate (AR) in dB/m is then

$$AR = 8.686 \alpha \quad \text{dB/m} \quad (2.25)$$

Samples of fresh water were obtained from lakes in the Gatineau area near Ottawa, Ontario, Canada. The DC ionic conductivities of the samples were measured and found to be between .03 and .01 Ω -meter⁻¹. The DC conductivity was used in the calculation of α since it is known, [21], that the conductivity of water is relatively independent of frequency for frequencies less than 1 GHz (see Fig. 2.7).

The sample holder was loaded with sample of fresh water, and the impedance of the loaded line was measured using the Time Domain Reflectometer (TDR). For all the samples, the measured impedance was found to be 7.3 ohms.

For the normalized impedance $R = 7.3/50$, and $\ell = 20''$ the expression of the attenuation rate can be written as:

$$AR = 8.549 \ln \left[0.555 + \frac{0.0989}{|S_{21}|_{\max}^2} + \frac{0.1977}{|S_{21}|_{\max}^2} + \frac{0.0098}{|S_{21}|_{\max}^4} \right]^{\frac{1}{2}} \quad (2.26)$$

Measurements were made on the transmission parameter $|S_{21}|^2$ over the frequency range (33-363) MHz. The attenuation rate was then calculated using Eqn. (2.26), and the results

Fig. 2.7. Variation of conductivity with frequency.
21°C, 3% NaCl solution.
a) Experimental curve
b) Measured value at .95-13 MHz/region

are plotted in Fig. 2.8.

The theoretical and measured data in Fig. 2.8 are in good agreement. Furthermore, the measurements indicate that the attenuation rate below 100 MHz is dominated by the ionic conductivity and is therefore independent of frequency. Increasing the frequency above 100 MHz results in a rapid increase of the attenuation rate because of the dipole relaxation phenomenon.

2.3 AIR-WATER AND WATER-BOTTOM INTERFACES

2.3.1 Introduction

Radar echos are directly related to the nature of the terrain illuminated by the electromagnetic waves. The relationship is complicated by the numerous terrain parameters that affect these echos. The most important parameters that affect radar echos are: the detailed composition of the illuminated area, the angle of incidence, the wave polarization and radar frequency. Because of this inherent difficulty, the following discussion should be regarded as an illustration of the general trends of the relative echo strengths rather than a firm description of the phenomenon. The purpose of the discussion given below is to impose a set of constraints on the experiment conditions such as the height of the antenna, the antenna orientation and the state of the water surface.

Fig. 2.8. Measurement results, $T=20^{\circ}\text{C}$, $\epsilon'=80.4$.

— calculated values
 pure water $\square$
 $\sigma=0.01$ $\circ$
 $\sigma=0.03$ $\triangle$

2.3.2 General Characteristics of Terrain Echos

Terrain reflectivities are usually described in terms of the radar cross-section per unit illuminated area σ° . Several experiments [7], [23], have showed that the variation of σ° with the grazing angle, θ , takes the general form shown in Fig. 2.9. In the figure, three distinct regions can be identified. The first region (near vertical incidence) is called the quasi-specular region, σ° in this region is fairly large compared to the other two regions. This is apparently due to specular scattering from normally oriented surfaces. Below some grazing angle, θ_0 , there is little likelihood of significant specular return, and the reflectivity is much less. This second region is called the plateau region. The boundary between the quasi-specular region and the plateau region is called the transition angle, θ_0 , and is not sharply defined, particularly for smooth surfaces. At very low grazing angle, σ° decreases rapidly with decreasing angle due to the destructive effect of the interference between the direct and scattered waves. This third region is called the interference region. The transition between the plateau and interference regions is called the critical angle, θ_c .

2.3.3 Effect of Surface Irregularities

All natural terrains have certain degrees of irregularities that affect their reflectivities. Terrain

Fig. 2.9. General dependence of σ^0 on grazing angle.

roughness depends not only on the physical construction of the terrain, but also on the wavelength and the angle of incidence (grazing angle). A terrain may be considered rough at one frequency and smooth at another. Moreover, at the same frequency, a terrain could change its status from rough to smooth at different grazing angles.

Lord Rayleigh suggested a way to distinguish between the extremes of smooth and rough surfaces. Rayleigh's criterion for a surface to be considered smooth is usually given as [7]:

$$\Delta h \sin \theta < \lambda/8 \quad (2.27)$$

where Δh is the RMS height of surface irregularities, θ is the grazing angle and λ is the wavelength.

Available experimental data indicate that for small values of the parameter $(\frac{\Delta h \sin \theta}{\lambda})$, the reflection coefficient increases with increases in Δh . However, for large values $(\frac{\Delta h \sin \theta}{\lambda} \gg 1/8)$, it appears that reflection coefficient is independent of Δh . [23]

2.3.4 σ^0 Dependence on Frequency

Most experimental results indicate that σ^0 varies inversely with the wavelength, λ . The wavelength dependence of σ^0 is frequently expressed in terms of λ^{-n} , where n is normally between zero and one [23]. Moreover, the critical

grazing angle, θ_c , is frequency dependent, according to M.W. Long [24], and may be calculated from the following relationship:

$$\sin \theta_c = \frac{\lambda}{4\pi\Delta h} \quad (2.28)$$

which indicates that θ_c increases with the wavelength.

In the case of water surfaces and at VHF frequencies, the curve of σ° vis θ is expected to be similar to the corresponding curve at microwaves, except that the critical angle θ_c is much larger, and the value of σ° in the plateau region is lower, as shown in Fig. 2.10.

2.3.5 σ° Dependence on Polarization

Based on several theoretical models and experimental results, the effect of the transmit and receive polarizations is more pronounced in cases of slightly rough rather than rough surfaces. According to Barrik and Peak [25], the average back scattering cross-section per unit area, σ° , can be expressed as:

$$\sigma^\circ_{ij} = 4\pi(2\pi/\lambda)^4 \sin^4\theta |T_{ij}|^2 W \quad (2.29)$$

where i and j refer to the incident and scattered polarizations, respectively, and W is a parameter which depends on the surface roughness. T_{ij} is the polarization dependent term given as:

Fig. 2.10. σ° dependence on frequency.

$$T_{HH} = \frac{(\mu_r - 1)[(\mu_r - 1)\sin^2\theta + \epsilon_r \mu_r] - \mu_r^2(\epsilon_r - 1)}{[\mu_r \cos\theta + \sqrt{\epsilon_r \mu_r - \sin^2\theta}]^2} \quad (2.30)$$

and

$$T_{VV} = \frac{(\epsilon_r - 1)[(\epsilon_r - 1)\sin^2\theta + \epsilon_r \mu_r] - \epsilon_r^2(\mu_r - 1)}{[\epsilon_r \cos\theta + \sqrt{\epsilon_r \mu_r - \sin^2\theta}]^2} \quad (2.31)$$

and

$$T_{HV} = T_{VH} = 0 \quad (2.31)$$

where H and V denote horizontal and vertical polarizations, respectively, and ϵ_r and μ_r are the relative permittivity and permeability, respectively.

Based on this model, J.W. Wright [26] conducted several experiments which demonstrated that under the same conditions, the surface echos in the case of vertical polarization are substantially higher than those for horizontal polarization. Several other experiments reported in [23] and [24] confirmed Wright's work. The difference between the two cases as suggested by Maurice [24] is shown in Fig. 2.11, and the results of Wright's experiment are shown in Fig. 2.12.

2.3.6 Summary

From the discussion presented in the above subsections, the following can be concluded.

Fig. 2.11. σ° dependence on polarization.

Fig. 2.12. Example of σ° dependence on polarization.

1. For an operating frequency of 100 MHz ($\lambda=3$ meter), the Rayleigh criterion given in Eqn. (2.23) indicates that the water surface can be considered smooth over the entire range of grazing angle up to a water wave height of 37.5 cm. Since most fresh water surfaces are within this range, it is reasonable to assume that the air-water interface is a smooth surface. On the other hand, the wavelength of the electromagnetic waves in water is only 33 cm at 100 MHz, and since the bottom terrain irregularities are of the same value (33 cm) or more, then the water-bottom interface can be considered rough in general.
2. For the air-water interface, and for water wave height less than 24 cm, Eqn. (2.24) indicates that the plateau region will entirely disappear at 100 MHz, which means that the surface reflectivity decreases rapidly for grazing angles less than 90° . This is in contrast with the case of the bottom terrain where the reflectivity decreases at a much slower rate.
3. The discussion presented in Section 2.3.5 indicates that a vertically polarized antenna provides a much higher reflectivity in the plateau region in comparison with a horizontally polarized

antenna. Since we are interested only in the specular return near ($\theta=90^\circ$), the antenna should be horizontally polarized.

4. It is important to maintain the antenna height as low as possible for two reasons. First, a low antenna provides a fairly strong echo from the air-water interface and this will ease the problem of transmitter/receiver isolation for the case of a CW radar and will also reduce the required transmitted power; second, a low antenna reduces the water surface clutter in the required resolution cell. In other words, the echo strength from the air-water interface decays rapidly with time.

2.4 STATEMENT OF SYSTEM PARAMETERS

The received power from the air-water interface, P_1 , depends on the height of the antenna R_0 , the reflection coefficient at the interface Γ , the wavelength of the transmitted signal λ and the gain of the transmit and receive antenna G . For a perfectly smooth and infinite surface, the radar equation can be written as:

$$P_1 = \frac{P_t G^2 |\Gamma|^2 \lambda^2}{(4\pi)^2 R_0^2} \quad (2.32)$$

where P_1 is the received power and P_t is the transmitted power. An alternative form of Eqn. (2.32) is:

$$P_1(\text{dB}) = P_t(\text{dB}) + 2G(\text{dB}) + 20 \log_{10} \left(\frac{|\Gamma|\lambda}{4\pi R_0} \right) - S \quad (2.33)$$

where S is a safety factor included to account for the uncertainty of the radar equation prediction.

The received power from the bottom terrain, P_2 , can be written as:

$$P_2(\text{dB}) = P_1(\text{dB}) - (2L_1 + L_2 + 2A.H) \quad (2.34)$$

where L_1 and L_2 are the losses at the top and bottom interfaces, respectively, A is the attenuation rate in water and H is the water depth.

The operating frequency of the system is chosen to be 100 MHz. This choice is a reasonable compromise between the attenuation rate in water and the size of the antenna. At 100 MHz, an antenna gain of 4.5 dB requires a total physical area of approximately 2 m^2 . The antenna height should be maintained as low as practically possible; in our analysis, R_0 is taken as 10 meters. A reasonable value of the safety factor S is 10 dB and $|\Gamma|$ for the air-water interface is equal to 0.8. Substituting these values in Eqn. (2.33) yields:

$$P_1 = P_t - 35.4 \text{ dB} \quad (2.35)$$

At 100 MHz, the attenuation rate in fresh water, A , is approximately equal to 2 dB/m, while the loss at the air-water interface, L_1 , is 8.9 dB. Eqn. (2.31) can then be written as:

$$P_2(\text{dB}) = (P_t - 35.4) - (17.75 + L_2 + 4H)$$

or

$$P_2 = P_t - (53.2 + 4H + L_2) \text{ dB} \quad (2.36)$$

Using Eqns. (2.35) and (2.36), the difference in power level between the top and bottom reflections is:

$$P_1 - P_2 = 17.8 + 4H + L_2 \quad \text{dB} \quad (2.37)$$

where L_2 is the loss at the bottom terrain interface. It varies widely from one terrain to the other and may take any value between 10 and 30 dB.

The minimum detectable signal is related to the noise bandwidth, B_n , by the well-known relation:

$$\text{MDS} \approx k T_0 F_n B_n \quad (2.38)$$

where k is Boltzmann constant, T_0 is the receiver temperature in degrees Kelvin and F_n is the receiver noise figure.

For $B_n = 15 \text{ MHz}$ and $F_n = 10 \text{ dB}$, Eqn. (2.38) yields:

$$\text{MDS} = -92.2 \text{ dBm} \quad (2.39)$$

If the maximum water depth to be measured is 5 meters, and

L_2 is taken as the worst case of approximately 30 dB, Eqns. (2.36) and (2.39) yield the minimum transmitted power as:

$$P_{t_{\min}} \approx + 11 \text{ dBm} \quad (2.40)$$

The time delay per unit length in the water is given by the following relationship:

$$\tau_0 = \frac{\sqrt{\epsilon'}}{c} \quad \text{sec/m} \quad (2.41)$$

where ϵ' is the real part of the complex permittivity of water and c is the speed of light. For fresh water at 20°C and 100 MHz, $\epsilon' = 80$, and

$$\tau_0 \approx 30 \quad \text{nsec/m} \quad (2.42)$$

The propagation delay in the water of depth H meters (round trip) is then:

$$\tau = 60 H \quad \text{nsec} \quad (2.43)$$

The theoretical time resolution of the system, ρ_0 , is approximately equal to the inverse of the transmitted signal bandwidth B , [27], i.e.

$$\rho_0 \approx \frac{1}{B} \quad \text{sec} \quad (2.44)$$

Combining Eqns. (2.44) and (2.43) results in a minimum water depth of:

$$H_{\min} = \frac{10^8}{6B} \quad \text{m} \quad (2.45)$$

Substituting $B \approx 15$ MHz in Eqn. (2.45) gives:

$$H_{\min} = 1.11 \text{ meters} \quad (2.46)$$

For the minimum depth $H_{\min}$ (= 1.11 meter), and for $L_2 = 20$ dB, the amplitude of the received signal from the bottom terrain is approximately 42 dB below the received signal from the air-water interface. Therefore, the time sidelobes of the top reflection should be maintained well below 42 dB at time offset from the main peak of $1/B$ sec. Furthermore, for larger values of H , the bottom reflection is further reduced, and at least a similar reduction in sidelobe level of the strong reflection is required.

2.5 SUMMARY

In this Chapter, a detailed theoretical study of the electromagnetic characteristics of fresh water was given, with a description of a novel broadband technique for the measurement of attenuation rate and the dispersion characteristics of fresh water. Measurement results have indicated that the attenuation rate is reasonably low at low frequencies and up to about 100 MHz, hence this region was chosen for our system. A theoretical review of the nature of wave reflections at the air-water and water-bottom interfaces has indicated that the air-water surface can be considered smooth at VHF bands while the water-

bottom interface is rough in general. Also, it was concluded that the antenna should be as low as practically possible and horizontally polarized, in order to prevent strong reflections from grazing angles less than 90° . A statement of the system requirement was given in Section 2.4, in which the important parameters of the system and the system requirements were determined. In Chapter 3, different radar systems are analyzed in order to evaluate their suitability for the problem of water depth measurement.

CHAPTER 3

DYNAMIC RANGE LIMITATION ON DIFFERENT RADAR SYSTEMS

3.1 INTRODUCTION

The problem of measuring the depth of shallow fresh water places a severe constraint on the design of a radar system. The interference between the residual response of the top surface echo and the weak response from the bottom requires the radar system to have high resolution and wide dynamic range, simultaneously.

Most of the classical radar systems achieve a high dynamic range by using variable gain receiver techniques [7]. However, such methods are appropriate only for detecting targets sufficiently separated in distance. Other systems are designed with ultra-high resolution capability without much attention to the dynamic range; such systems are used to resolve targets of approximately equal strengths.

These two approaches cannot be used in our case, and a new technique must be developed, with particular attention to enhancing the dynamic range of the receiver without using a variable gain control, while maintaining the basic resolution requirements. One should note that the dynamic range of the receiver is not limited by the thermal noise or clutter, but rather is limited by the residual response from a strong reflection.

In general, radar systems can be divided into two main categories: pulse radars and continuous wave (CW) radars and each class has its advantages and disadvantages. Pulse radars are usually power limited because the pulse repetition frequency (PRF) is chosen through the traditional trade-off between large unambiguous range and high average power [7]. Pulse radars typically have low sensitivities and wide bandwidths. Nevertheless, they are widely used because of the well-established technology, and because of their capability to detect a moving target with minimum sensitivity to Doppler frequency shift.

CW radars, on the other hand, have very high sensitivities, their bandwidths are usually much smaller than the pulse radars, and they are also less expensive. But, CW radars have the problem of leakage between the transmitter and receiver which limits their maximum range [7].

Pulse radars can be further divided into short pulse and pulse compression techniques. The former consists of a train of short CW bursts, while the latter consists of a train of longer modulated pulses. The CW radars can also be divided into several forms, among them only two serve the purpose of our discussion. These are the FM-CW radar [28] and the phase-coded waveform radars [29]. A third class of radars which may be included in the discussion is the homodyne coherent radar [30]. This latter system achieves

a high resolution capability, while maintaining the system bandwidth at minimum level, but as illustrated later suffer from limited unambiguous range capabilities.

In this Chapter, we shall review the main features of these systems, with a special emphasis on their potential for reducing the residual response of a strong reflection within the required resolution time. In Section 3.2, the short pulse radar is discussed, with a mathematical analysis of the theoretical limit of the sidelobe level. This is followed by a discussion of the pulse compression techniques in Section 3.3. In Section 3.4, a detailed analysis of the theoretical and practical limitations of the sidelobe level in FM-CW systems is given. The homodyne coherent radar is discussed in Section 3.5. The Chapter is then concluded in Section 3.6.

3.2 SHORT PULSE RADAR

3.2.1 Introduction

The short pulse radar is conceptually the simplest of all radar systems. A train of short bursts of CW is transmitted, with the pulse repetition period greater than the round trip delay of the largest range of interest. The pulse envelope may be a rectangular, triangular, trapezoidal, or any other shape; however, a rectangular envelope has the highest average power. The received pulses are

first down-converted into the IF stage, and then fed to the detector which might be an ~~envelope detector~~, a square wave device or a coherent detector. All of these three detectors achieve the same signal-to-noise ratio when the received power is above a certain limit [29].

In the following subsections, attention is directed to the sidelobe level at the output of different receiving filters and different pulse envelope shapes. It is concluded that the Gaussian filter-Gaussian pulse represents the optimum shape for sidelobe reduction. In Section 3.2.2, the impulse response and the bandpass characteristics of different filters are presented. The special case of a Gaussian filter-Gaussian pulse is further discussed in Section 3.2.3, and concluding comments are given in Section 3.2.4.

3.2.1. Characteristics of Different Receive Filters

The received signal from a point target in free space is a replica of the transmitted signal, greatly attenuated, delayed in time and shifted in frequency by the Doppler effect. We assume here that the Doppler shift is negligible or can be removed, along with the carrier frequency, which leaves the low-frequency modulated function which describes the signal envelope and phase.

We will consider a single, non-repetitive waveform of finite energy $E_1 \gg N_0$, where N_0 is the spectral density

of the input noise. Fig. 3.1 shows the mathematical model of the receiving filter, where the filter response is described by a transfer function, $H(f)$, in the frequency domain and an impulse response, $h(t)$, in the time domain. The input and output pulses are described by $x(t)$ and $y(t)$ in the time domain and $X(f)$ and $Y(f)$ in the frequency domain, respectively. We will reference the filter functions in order to place the maxima at $f=0$ and $t=0$, although these reference points may be shifted to other values by appropriate circuit elements.

The optimum matched filter has a frequency response which is the complex conjugate of the received signal spectrum, i.e.

$$H(f) = X^*(f) \quad (3.1)$$

Therefore, the output pulse, $Y(t)$, is:

$$Y(t) = \int_{-\infty}^{\infty} |H(f)|^2 e^{j2\pi ft} df \quad (3.2)$$

The output of a matched filter may also be expressed in terms of the time response of the filter as follows:

$$Y(t) = \int_{-\infty}^{\infty} h(\lambda)h(t-\lambda)d\lambda \quad (3.3)$$

Fig. 3.2 illustrates the impulse characteristics of different types of filters, and the corresponding bandpass characteristics are shown in Fig. 3.3 [7].

Fig. 3.1. Mathematical model of the receiving filter.

Fig. 3.2. Impulse characteristics of different filters.

- Rectangular
- Two-tuned stage
- . - . - Five-pole Gauuian
- Infinite-pole Gaussian

Fig. 3.3. Deviation from center frequency per 6db bandwidth
Bandpass characteristics of different filters.

- Rectangular
- . - . Two-tuned stage
- - - - Five-pole Gaussian
- Infinite-pole Gaussian

Two factors are usually predominant in the choice of the receive filter. These are the receiver sensitivity and the effective duration of the received pulse. The receiver sensitivity is determined by the noise bandwidth of the receiver filter, B_n , which is defined as: [31]

$$B_n = \int_{B^0} |H(f)|^2 df \quad (3.4)$$

where B^0 is the bandwidth which contains an amount of energy arbitrary close to the total energy of $H(f)$. Similarly, the effective pulse duration, τ_n , is defined as: [31]

$$\tau_n = \int_{\tau^0} |h(t)|^2 dt \quad (3.5)$$

where τ^0 is the time duration which contains an amount of energy arbitrarily close to the total energy of $h(t)$.

In the case of two targets of vastly different reflectivities, the pulse duration as defined in Eqn. (3.5) is not adequate since the residual response of the strong target might mask the response of the weak target at a time offset greater than τ_n . In this case, it is essential to design the receive filter such that the residual response of $Y(t)$ decays rapidly at a time t removed from the pulse peak at $t=0$. Fig. 3.2 shows that the Gaussian filter has the best impulse response as far as the time sidelobes are concerned. In the following section, the case of a

Gaussian matched filter shall be treated in more detail.

3.2.3 The Case of a Gaussian Pulse-Gaussian Filter Receiver

The transfer function of an idealized Gaussian filter can be written as: [31]

$$H(f) = \exp(-f^2/2\sigma_f^2) \quad (3.6)$$

where σ_f is the variance of the frequency response. The corresponding impulse function, $h(t)$, is: [32]

$$h(t) = [2\pi\sigma_f^2]^{-1/2} \exp(-2\pi^2\sigma_f^2 t^2) \quad (3.7)$$

$$= [2\pi\sigma_t^2]^{-1/2} \exp(-t^2/2\sigma_t^2) \quad (3.8)$$

where the time variance σ_t is related to the frequency variance σ_f by:

$$\sigma_t = 1/2\pi\sigma_f \quad (3.9)$$

The noise bandwidth, B_n , of the Gaussian filter can be calculated from Eqn. (3.4) by setting the integration limits to $\pm \infty$ as follows:

$$B_n = \int_{-\infty}^{+\infty} \exp(-f^2/\sigma_f^2) df \quad (3.10)$$

This gives:

$$B_n = \sqrt{\pi} \sigma_f \quad (3.11)$$

Substituting Eqn. (3.11) into Eqn. (3.7) gives:

$$h(t) = (2 B_n^2)^{1/2} \exp(-2\pi B_n^2 t^2) \quad (3.12)$$

The received pulse, $x(t)$, has an expression similar to that of the filter impulse response with a normalized amplitude, equal to A , i.e.

$$x(t) = A \exp(-2\pi B_n^2 t^2) \quad (3.13)$$

The output pulse $y(t)$, is the convolution integral of $x(t)$ and $h(t)$. Using Eqns. (3.12) and (3.13), we get:

$$y(t) = A[2B_n^2]^{1/2} \int_{-\infty}^{\infty} \exp\{-2\pi B_n^2 [\lambda^2 + (t-\lambda)^2]\} d\lambda \quad (3.14)$$

A straightforward evaluation of Eqn. (3.14) leads to:

$$y(t) = A/\sqrt{2} \exp(-\pi B_n^2 t^2) \quad (3.15)$$

Therefore, the output response at time t may be expressed as:

$$y(t) = 20 \log_{10} (y(t)/y(0)) \text{ dB}$$

or

$$y(t) \approx -27.29(B_n t)^2 \text{ dB} \quad (3.16)$$

A plot of Eqn. (3.16) is shown in Fig. 3.4. The figure shows that at time $t=1.5/B_n$ the output pulse is 61.4 dB below the peak value at $t=0$.

To complete the analysis, we should also consider the received signal-to-noise ratio for the case of a train

Fig. 3.4. Filter output for the idealized case of Gaussian filter-Gaussian pulse.

of Gaussian pulses. In this case, $y(t)$ can be rewritten as:

$$y(t) = \frac{A}{\sqrt{2}} \sum_{n=-\infty}^{+\infty} \exp[-\pi B_n^2 (t - nT_0)^2] \quad (3.17)$$

where T_0 is the pulse repetition period. The average signal power at the output of the filter is then:

$$P_{av} = \frac{A^2}{2T_0} \int_{-T_0/2}^{T_0/2} \exp(-2\pi B_n^2 t^2) dt \quad (3.18)$$

When T_0 is much greater than the time variance of the Gaussian pulse, the integration limits of Eqn. (3.18) can be set to $\pm \infty$, and then

$$P_{av} = \frac{A^2 \cdot \text{PRF}}{2\sqrt{2} B_n} \quad (3.19)$$

where PRF is the pulse repetition frequency which is equal to $1/T_0$. The signal-to-noise ratio at the filter output is defined as:

$$S/N = P_{av} / N_0 B_n \quad (3.20)$$

where N_0 = power spectral density of the noise

$$= k T_0 F \quad (3.21)$$

where k is Boltzmann's constant (1.38×10^{-23}), T_0 is the noise temperature (290 °K) and F is the receiver noise figure. Combining Eqns. (3.19), (3.20) and (3.21), we obtain:

$$S/N = \frac{A^2 \cdot PRF}{2\sqrt{2} k T F B_n^2} \quad (3.22)$$

A more useful format of Eqn. (3.22) is:

$$S/N = 1.767 \times 10^{20} \frac{P_p \cdot PRF}{F B_n^2} \quad (3.23)$$

where P_p is the peak power of the received signal.

The case of a Gaussian pulse and a Gaussian matched filter represents the theoretical limit on the sidelobe level, and the performance of other waveforms should be compared with this idealized case. This ideal case cannot be realized in practice, since the Gaussian pulse is not a time-limited function. In practice, however, the transmitted pulse has a finite length, and only an approximation of the Gaussian pulse can be achieved.

3.2.3. Practical Limitations of Short Pulse Radars

The discussion given above illustrates that both the shape and the bandwidth of the receive filter contribute to the sidelobe level of the output response. Generally, the bandwidth of a radar receiver should be wide enough to reduce the ringing of the output pulse; moreover, a rectangular bandpass should be avoided, and the lowest sidelobes are exhibited by networks which approximate Gaussian characteristics.

Thus far, we have considered the filtering process to take place only at the IF stage. In doing so, we neglected the contribution of the receive antenna to the whole impulse response of the receiver. This approximation is valid only in a narrowband system, where the bandwidth of the receiver antenna can be made approximately flat over the desired IF bandwidth. In VHF systems, however, this is not true, since the antenna response cannot be made flat over 20 or 30% bandwidth. In fact, in such systems, the antenna response may dominate the shape of the receive filter response. Therefore, an idealized Gaussian filter cannot be realized in a wideband system of interest, and the sidelobes will certainly be much higher than the value predicted for an ideal Gaussian filter. Nathanson [32] has predicted -32.4 dB for the time sidelobe level as a practical limit while the theoretical result for a Gaussian filter as illustrated in Fig. 3.4 suggests a value of ≈ 61 dB at $t = (1.5/B_n)$.

3.2.4 Summary

The short pulse radar was shown to suffer from a ringing problem due to the imperfection of the antenna response and the IF filter over the large bandwidth required. It was also shown that not only a large bandwidth of the receive filter is required to reduce the sidelobe level, but even more important is the shape of the filter response.

and the shape of the transmitted pulse. Gaussian filter-Gaussian pulse was shown to provide the lowest sidelobe level (61.4 dB at $t=1.5/B_n$). However, this ideal case was shown to be unrealizable. The sidelobe level in the case of short pulse radar is limited to approximately 34 dB.

3.3 PULSE COMPRESSION TECHNIQUES

3.3.1 Introduction

Pulse compression involves the transmission of a long coded pulse and the processing of the received echo to obtain a relatively narrow pulse. The increased detection capability of a long-pulse radar system is achieved while retaining the range resolution capability of a narrow pulse system [33]. Several advantages are obtained. Transmission of a long pulse permits a more efficient use of the average-power capability of the radar. Generation of high-peak-power signals is avoided. The average power of the radar may be increased without increasing the pulse repetition frequency and hence decreasing the radar's unambiguous range. In addition, the radar is less vulnerable to interfering signals.

A pulse compression radar is a practical implementation of a matched filter system. The coded signal may be generated either as a frequency response $H(f)$ or as a time response $h(t)$ of a coding filter. In Fig. 3.5a, the coded

Fig. 3.5. Pulse compression radar using:

- a) Conjugate filter
- b) Time inversion
- c) Correlation

signal is obtained by exciting the coding filter $H(f)$ with a unit impulse. The received signal is fed to the matched filter whose frequency response is the complex conjugate $H^*(f)$ of the coding filter. The output of the matched filter section is the compressed pulse, which is given by the inverse Fourier transform of the product of the signal spectrum $H(f)$ and the matched filter response $H^*(f)$.

$$y(t) = \int_{-\infty}^{\infty} |H(f)|^2 e^{j2\pi ft} df \quad (3.24)$$

The filter is also matched to a signal if the signal is the time inverse of the filter's response to a unit impulse. This is achieved by applying the time inverse of the received signal to the compression filter, as shown in Fig. 3.5b. Identical filters may be used for both expansion and compression, or the same filter may be used for both expansion and compression with appropriate switching between the transmitting and receiving functions.

In practice, multiple delays and correlators are used to cover the total range interval of interest. Hence, correlation processing as shown in Fig. 3.5c is equivalent to matched filtering.

The output of the matched filter consists of the compressed pulse accompanied by range sidelobes. Spectrum weighting of the output signal's may be used to reduce the

time sidelobes; this is discussed in the following subsection.

3.3.2 Weighting Techniques

The primary objective of weighting in the frequency or time domain is to reduce the sidelobes in the other domain. Since we are interested in reducing the time sidelobes, frequency weighting functions are considered.

Consider the case when a rectangular pulse is received first by a matched filter and then fed to a frequency weighting circuit. The performance achieved with various frequency-weighting functions is summarized in Table V.

Pedestal height H is defined in the case of Hamming weighting function amplitude at the band edge ($f = \pm B/2$) when the function has been normalized to unit amplitude at the band center ($f = 0$). A normalized logarithmic plot of the output pulse magnitude as a function of time is shown in Fig. 3.6. [7].

The uniform weighting (curve A) has the worst time sidelobes performance; the first lobe is only -13.2 dB below the main lobe, while the following lobes fall off at a rate of 6 dB/octave. Taylor weighting has a performance intermediate between the uniform case and Hamming weighting which represents the best case. Hamming weighting results in a wide main lobe (double the width of the uniform weighting), but the sidelobes are low. The first lobe is -42.8 dB below the main lobe, decaying at a rate of 6 dB/octave.

Table V. Different weighting functions.

Weighting Function, $u(f)$	Pedestal Height $H, \%$	Mainlobe Width -3 dB	Peak Sidelobe Level dB	Far Sidelobe Fall-off
<u>Uniform Weighting</u>				
$u(f) = \text{constant}$	100	$.886/B$	-13.2	6 dB/octave
<u>Hamming</u>				
$u(f) = H + \frac{(1-H)}{\cos^2(\pi f/B)}$	8	$1.33/B$	-42.8	6 dB/octave
<u>(Cosine)ⁿ</u>				
$u(f) = \cos^2(\pi f/B)$	0	$1.46/B$	-31.7	18 dB/octave
$u(f) = \cos^3(\pi f/B)$	0	$1.66/B$	-39.1	24 dB/octave
$u(f) = \cos^4(\pi f/B)$	0	$1.94/B$	-47.0	30 dB/octave
<u>Taylor</u>				
$u(f)$ $= W_0 \left[1 + 2 \sum_{m=1}^{n-1} F_m \cos\left(\frac{2\pi m f}{B}\right) \right]$				
$n = 8$	11	$1.25/B$	-40.0	6 dB/octave

Fig. 3.6. Comparison of compressed pulse shapes for three frequency weightings.

- Uniform weighting
- Hamming weighting
- · - · - Taylor weighting ($\bar{n}=6$)

3.3.3 Summary

Pulse compression techniques with different weighting functions were considered. It was shown that Hamming weighting function has the lowest sidelobe level (42.8 dB). However, this level is not adequate for water depth measurement, since the echos from the water-bottom interface could be weaker than this value. In the following section, a detailed analysis of the third classical radar system (FM-CW) is given.

3.4 FM-CW RADAR

3.4.1 Basic Concept

A simplified block diagram of a FM-CW radar is shown in Fig. 3.7a, and the system operation can be explained as follows. The carrier frequency, f_0 , is modulated by a periodic low frequency signal producing the instantaneous frequency shown in Fig. 3.7b. The modulating signal may be triangular, sinusoidal or any other periodic signal; however, in a multi-target environment, the instantaneous frequency should be a linear function of time, and therefore only triangular or saw-tooth waveforms are used to modulate the carrier. In Fig. 3.7b, a triangular waveform is used with a frequency ($f_1 = 1/T$). The RF frequency deviation is Δf_0 Hz around the center frequency, f_0 . The received signal is a delayed version of the transmitted one with some amplitude scaling factor; hence, the instantaneous frequency

Fig. 3:7. FM-CW radar.

- a) Basic configuration
- b) Instantaneous transmitted frequency, f_T ; received frequency, f_R ; beat frequency f_B .

of the received signal is as shown in Fig. 3.7b. The time delay, τ , is related to the target range, R , through the following relation:

$$\tau = \frac{2R}{v} \quad (3.25)$$

where v is the propagation velocity in the medium of interest. A continuous mixing of the received and transmitted signals will result in a beat frequency, f_B , as shown in Fig. 3.7b. The beat frequency can be calculated from the geometry of Fig. 3.7b, and is equal to

$$f_B = \frac{2\tau\Delta f_o}{T} \quad (3.26)$$

Using Eqn. (3.25) and (3.26), we get:

$$f_B = \frac{4\Delta f_o}{vT} R \quad (3.27)$$

Therefore, the beat frequency is proportional to the target range, R , and a frequency counter or a spectrum analyzer can be used to measure the range. In multi-targets environment, however, the spectrum of the beat signal contains several spectral lines, each line corresponding to a target. The distance between any two targets can be determined by measuring the frequency difference between the corresponding spectral lines. In this latter case, a counter cannot be used and the spectrum analyzer is an essential part of the system.

3.4.2 Spectrum of the Beat Signal

In a FM-CW system, the range measurement takes place in the frequency domain; therefore, the analysis of the beat signal spectrum is of prime interest. The exact analysis of the beat spectrum is very complicated and therefore some justified assumptions should be made to simplify the analysis. We shall first assume a perfectly linear swept frequency and a constant amplitude; the effect of sweep non-linearity and amplitude modulation will be taken into account in Sections 3.4.4 and 3.4.5.

In general, the beat signal can be expressed as follows:

$$x(t) = a \cos(2\pi \int_{-\infty}^t f_b(t) dt) \quad (3.28)$$

where $f_b(t)$ is the instantaneous beat frequency and a is the amplitude. In case of a triangular modulating signal, $f_b(t)$ takes the form shown in Fig. 3.8a and it can be expressed as the sum of a d-c term, f_{B0} , and a periodic term, $f'_b(t)$ with zero average value.

$$f_b(t) = f_{B0} + f'_b(t) \quad (3.29)$$

$f'_b(t)$ is shown in Fig. 3.8b while f_{B0} is equal to:

$$f_{B0} = f_B - \frac{T}{T} f_B \approx f_B \quad (3.30)$$

Fig. 3.8. The instantaneous frequency and phase of the beat signal.

The approximation in Eqn. (3.30) is valid since $\frac{\tau}{T}$ is much smaller than unity. Eqn. (3.28) can now be written as:

$$x(t) = a \cos(2\pi f_B t + \phi(t)) \quad (3.31)$$

where

$$\phi(t) = 2\pi \int_{-\infty}^t f'_b(t) dt \quad (3.32)$$

$\phi(t)$ is shown in Fig. 3.8c. From the geometry of the figure, the maximum value of $\phi(t)$ can be calculated as:

$$|\phi(t)|_{\max} \approx \frac{\pi}{2} \tau f_B \quad (3.33)$$

Using Eqn. (3.26), Eqn. (3.33) is rewritten as:

$$|\phi(t)|_{\max} \approx \pi \left(\frac{\tau}{T}\right) (\tau \Delta f_0) \quad (3.34)$$

According to Woodward [27], the time resolution of a radar system is approximately equal to the inverse of the bandwidth of the transmitted waveform ($\frac{1}{\Delta f_0}$ in our case), which means that $(\tau \Delta f_0)$ is in the order of unity. Since $\frac{\tau}{T}$ is usually much smaller than one, then:

$$|\phi(t)|_{\max} \ll 1 \quad (3.35)$$

Consequently, the waveform given in Eqn. (3.31) can be regarded as a narrow band FM signal (NBFM). However, this assumption is valid only under the condition:

$$\tau^2 \ll \frac{T}{\pi \Delta f_0} \quad (3.36)$$

Assume $|\phi(t)| \ll 1$, so that $\cos \phi(t) \approx 1$ and $\sin \phi(t) \approx \phi(t)$.

Recall that $\cos(A+B) = \cos A \cos B - \sin A \sin B$

so that,

$$x(t) \approx a \cos(2\pi f_B t) - a \phi(t) \sin(2\pi f_B t) \quad (3.37)$$

The Fourier transform of $x(t)$ is then:

$$\begin{aligned} X(\omega) &= \frac{a}{2} \{ \delta(\omega + \omega_B) + \delta(\omega - \omega_B) \} \\ &\quad - \frac{a}{2} \{ \Phi(\omega + \omega_B) - \Phi(\omega - \omega_B) \} \end{aligned} \quad (3.38)$$

where $\delta(\cdot)$ is the Dirac-delta function, and $\Phi(\omega)$ is the Fourier transform of the variable phase $\phi(t)$.

In the frequency domain, the spectral line at $f = f_B$ represents the desired target indication where f_B is proportional to the target range. The second term in Eqn. (3.38), however, represents the undesired sidebands, which might mask the spectral lines of other weaker targets. Therefore, the dynamic range of the received signals is determined by the relative amplitudes of the first and second terms in Eqn. (3.38).

The Fourier transform of the variable phase $\phi(t)$ is related to the Fourier transform of the function $f'_b(t)$ by the following relation:

$$\Phi(f) = \frac{1}{jf} F'_b(f) \quad (3.39)$$

where $F'_b(f)$ is the Fourier transform of $f'_b(t)$. Moreover, $f'_b(t)$ is a periodic function of time, with a period of $T/2$.

seconds, and it can be represented by its Fourier series:

$$f'_b(t) = \sum_{n=-\infty}^{+\infty} C\left(\frac{2n}{T}\right) e^{j4\pi nt/T} \quad (3.40)$$

where $C\left(\frac{2n}{T}\right)$ is the coefficient of the n th harmonic which is given as:

$$C\left(\frac{2n}{T}\right) = \frac{2}{T} \int_{-T/4}^{T/4} f'_b(t) e^{-j4\pi nt/T} dt \quad (3.41)$$

A straightforward evaluation of Eqn. (3.41) yields:

$$\begin{aligned} C\left(\frac{2n}{T}\right) &= \left(\frac{f_B \tau}{T}\right) \text{sinc}^2(\pi n \tau / T); \quad n \neq 0 \\ &= 0; \quad n = 0 \end{aligned} \quad (3.42)$$

Substituting Eqn. (3.42) into (3.40) gives:

$$f'_b(t) = \frac{2f_B \tau}{T} \sum_{n=1}^{\infty} \text{sinc}^2(\pi n \tau / T) \cos(4\pi nt / T) \quad (3.43)$$

Using Eqns. (3.39) and (3.43), the n th component of the spectrum $\Phi(f)$ is:

$$|\Phi_n| = \frac{f_B \tau}{2n} \text{sinc}^2(\pi n \tau / T) \quad (3.44)$$

occurring at frequency offset of $\frac{2n}{T}$ away from the beat frequency f_B . Therefore, the sidelobe power at a frequency offset from the beat frequency, f , is:

$$P_s(f) = \left(\frac{f_B \tau}{T}\right)^2 \frac{\text{sinc}^4(\pi \tau f / 2)}{f^2} \quad (3.45)$$

The frequency sweeping rate $\left(\frac{1}{T}\right)$, and the RF frequency deviation (Δf_0) can be selected to provide a very low level sidelobe (below 60 dB). However, other considerations should

be taken into account, such as the relation between the beat frequency and the sweeping rate, also the relation between the target range and the RF bandwidth of the system. These considerations are discussed below.

3.4.3 RF Bandwidth and Sweeping Rate

Eqn. (3.45) is applicable only in the case of a single tone signal interrupted periodically with a small phase variation. The period between interruptions, $T/2$, should be much larger than the cycle period of f_B , i.e.

$$f_B \gg \frac{1}{T} \quad (3.46)$$

Also, for the phase variation (at the interruption instant) to be small, τ should be much smaller than the cycle period of f_B , i.e.

$$\frac{1}{\tau} \gg f_B \quad (3.47)$$

The two inequalities (3.46) and (3.47) set the limit on the beat frequency. A violation of any of these two conditions will result in a rapid degradation of the sidelobe level.

The beat frequency, f_B , is determined by the round trip delay, τ , the RF frequency deviation Δf_0 and the sweeping rate $\frac{1}{T}$. The relation between these parameters is rewritten here as:

$$f_B T = \tau \Delta f_0 \quad (3.48)$$

Since $f_B \gg \frac{1}{T}$, then $\tau \Delta f_0$ is also much greater than one, i.e.

$$\Delta f_0 \gg \frac{1}{T} \quad (3.49)$$

For a triangularly swept frequency, the RF bandwidth is approximately equal to the frequency deviation Δf_0 . Therefore, the RF bandwidth should be much larger than the inverse of the nearest target delay. This point is particularly important in the application of fresh water depth measurement, since the antenna height above the water surface should be as small as practically possible for the reasons discussed in Chapter 2. This requires a large RF bandwidth. For example, if the antenna height is 30 meters, the RF bandwidth should be more than 30 MHz to allow a reasonable number of cycles between phase interruptions. This is one of the disadvantages of FM-CW systems for our particular problem. The second major disadvantage is the amplitude fluctuation of the received signal which is discussed below.

3.4.4 Effect of Amplitude Modulation

The amplitude modulation of the received signal is caused by the frequency dependent response of different components within the radar system such as the antenna, filters, amplifiers, etc. and by the frequency dependence of the reflectivity of the target. Since the FM-CW system detects the targets in the frequency domain, the spectrum of an amplitude modulated signal may be mistaken as a target or it may mask the response of a weak target. This problem

is particularly important in our application for two reasons; first, the transmitted waveform is a wideband signal with a low carrier frequency and this will enhance the problem of controlling the response of the system components over the entire bandwidth; secondly, in a multi-target environment, limiters cannot be used to stabilize the amplitude of the received signal.

To illustrate this problem quantitatively, let us assume that the received signal was amplitude modulated by a random signal. The rms value of the amplitude variation must be below 0.015 dB in order to obtain spurious sidelobe levels below -55 dB. However, an amplitude variation of less than 0.015 dB is impossible to achieve without the use of a limiter. For an optimistic estimation of an amplitude variation of 0.1 dB, the sidelobe level is -38.8 dB. A weak target response may be below this limit.

3.4.5 Effect of Sweep Non-Linearity

Another problem which affects the dynamic range of the FM-CW system is the frequency sweep linearity. The deviation from a linear relationship between the instantaneous transmitted frequency and time is caused by either the non-linearity of the triangular modulated signal or the non-linearity of the frequency modulator or both.

For an idealized linearly swept frequency, the instantaneous transmitted and received frequencies can be

written as:

$$f_t^o = f_o + k_o t \quad (3.50)$$

$$f_r^o = f_o + k_o (t - \tau) \quad (3.51)$$

where τ is the round trip delay of the target, and k_o ($= 2\Delta f_o / T$) is the sweeping rate. The beat frequency, f_B^o , is the difference between the transmitted and received frequencies:

$$f_B^o = f_t^o - f_r^o = k_o \tau \quad (3.52)$$

In general, the relationship between f_t and time is not linear because of the non-linearity of the control voltage, and the characteristic of the V.C.O. itself. Therefore, more general expressions of f_t and f_r are:

$$f_t = f_o + k_o (1 + e(t)) t \quad (3.53)$$

and

$$f_r = f_o + k_o (1 + e(t - \tau)) (t - \tau) \quad (3.54)$$

The corresponding beat frequency is then:

$$f_B = \{k_o (1 + e(t)) t\} - \{k_o (1 + e(t - \tau)) (t - \tau)\} \quad (3.55)$$

We shall assume that the error function $e(t)$ is a slowly varying function of time, and since the round trip delay time is small compared to the sweeping time, then:

$$e(t - \tau) \approx e(t) \quad (3.56)$$

Therefore, the expression of the beat frequency, f_B , can

2

be simplified to:

$$f_B \approx k_0 \tau (1 + e(t)) \quad (3.57)$$

or

$$f_B = f_B^0 + \delta f_B \quad (3.58)$$

where δf_B is the error in the beat frequency caused by the sweep error and it is:

$$\delta f_B = k_0 \tau e(t) \quad (3.59)$$

The beat signal can now be written as:

$$x(t) = A \cos(2\pi f_B t + 2\pi k_0 \tau \int_{-\infty}^t e(t) dt) \quad (3.60)$$

Again, if we assume that the phase error then the sidebands of the beat signal spectrum are:

$$|S_2|^2 = f_B^2 \left| \frac{E(f - f_B^0)}{(f - f_B^0)} \right|^2 \quad (3.61)$$

where $|E(f)|^2$ is the power spectral density of the error function $e(t)$. For example, if $e(t)$ is a sinusoidal function with maximum value $|E_m|$, and periodic time equals the sweeping time of the sweep generator, a linearity better than 0.4% is required to keep the sidelobes below 60 dB. However, this value is achievable using a YIG V.C.O. if the control voltage is generated digitally.

3.4.6 Summary

The FM-CW radars have two major disadvantages which make them unattractive for our application. First, the

amplitude modulation due to the circuit components imperfections and other spurious effects makes it almost impossible to keep the sidelobes below approximately 38 dB. This sidelobe level is not adequate for our application. Secondly, the FM-CW radars require RF bandwidth which is approximately 30% more than other systems to achieve the same resolution. This is undesirable, particularly in the case of VHF systems like the one in our interest.

The last system to be considered is the bi-frequency homodyne receiver which is discussed below.

3.5 MULTIPLE FREQUENCY HOMODYNE RADAR

A schematic diagram representing a bi-frequency homodyne radar is shown in Fig. 3.9 [30]. In the transmitter, the second harmonic of the fundamental is generated in a non-linear $\times 2$ frequency multiplier. The two frequencies are then directed to a target consisting of two dielectric interfaces. The return signals are power detected in the receiver to find the dispersive-loss introduced by the target. Also, the fundamental is frequency doubled and compared with the second harmonic in a coherent phase detector to determine the dispersive phase of the target.

The dynamic range limitations of this system are obvious, since the measurement of the dispersive phase and the dispersive amplitude will be difficult when the return from one interface is weak. An estimated dynamic range of

Fig. 3.9. Frequency homodyne radar.

this system may be less than 25 dB. The other major disadvantage of the bi-frequency homodyne radar is the severely limited unambiguous range. The range of this system is limited to half the wavelength of the highest frequency. For frequency of 100 MHz, the unambiguous range is determined by half the wavelength of the second harmonic (300 MHz); this range is only 8.3 cm in water.

3.6 SUMMARY AND CONCLUSIONS

In this Chapter, we have discussed several types of classical radar systems. The evaluation of each system was based on its ability to solve the problem of self-interference between radar echos in multi-target environments. The short pulse radar was shown to suffer from a ringing problem due to the imperfection of the antenna response over a relatively large bandwidth. This problem will probably limit the sidelobe level to about 34 dB which is not adequate for our application. Pulse compression techniques with different weighting functions were also considered. Hamming weighting functions were shown to have the lowest sidelobe level (42.8 dB), but the level was not adequate for our application. FM-CW radars have sidelobe levels which are limited to approximately 39 dB because of the spurious amplitude modulation of the received signal. Finally, the bi-frequency homodyne radar was considered. This system was shown to suffer from a very limited

unambiguous range and very limited dynamic range.

We have also indicated that the case of a short pulse radar with Gaussian filter and Gaussian pulse represents the ideal, although non-realizable, case. In this case, the sidelobe level is approximately 61 dB at $t = 1.5/B_n$, removed from the center pulse, where B_n is the noise bandwidth of the system. This theoretical value shall be taken as a reference and the performance of the novel system proposed in Chapter 4 of this thesis will be compared to this value at the end of Chapter 5.

CHAPTER 4

CORRELATOR RECEIVER WITH PSEUDORANDOM SEQUENCES AND SIDELobe SUPPRESSION

4.1 INTRODUCTION

In Chapter 2, we have described the problem of interference between the radar echos of two targets with vastly different reflectivities. The ability of the classical radar systems to solve this problem has been examined in Chapter 3, and it was concluded that these systems will not provide a satisfactory solution for the problem. It was also stated in Chapter 2 that a solution for this problem may be achieved by a special design of the radar transmitted waveform, with particular attention to the suppression of the time sidelobes at the receiver output.

In this Chapter, we present a new radar waveform which has the potential of time sidelobe suppression. The waveform design makes use of the recent advances in the digital modulation techniques being used in modern communication systems, and is based on obtaining the autocorrelation function of the modulated signals which have desirable sidelobe construction.

In Section 4.2, the basis of this new approach is discussed and a detailed derivation of the new waveform design is given together with an explanation of the sidelobe

suppression mechanism. The detailed error analysis is then given in Section 4.3.

4.2 THEORY OF SUPPRESSED SIDELOBE CORRELATOR

4.2.1 Introduction

The ability of a radar to resolve stationary targets in range depends on the autocorrelation function of the transmitted waveform [27]. Ideally, the autocorrelation function should have a single lobe peaking at the origin and extending over a resolution time of $\pm T$ seconds, and have zero value elsewhere. In theory, this can be realized by a short pulse of duration T seconds. However, it has been shown in Chapter 3 that in practice, the correlation function of a short pulse will have high sidelobes that extend well beyond the resolution time T . An alternative waveform is a repetitive CW signal; the autocorrelation of such a signal will always have certain sidelobes which are higher than the required level (about 50 dB below the peak value). An excellent waveform for modulating the transmitter is the Pseudo Random Binary Sequence (PRBS) codes [34], [35]. The autocorrelation function of the code has a single peak for each code length and sidelobe level which is inversely proportional to the code length. However, longer codes require long integration time in the correlator receiver and real time measurements become excessively difficult.

It will be shown in this Chapter that the new technique presented in this thesis achieves very low sidelobe levels using relatively short PRBS codes.

Numerous other waveforms have been proposed by different workers, and a comprehensive discussion of the waveform design is given by Skolnik [7]. Skolnik has concluded that it is not always possible to design a waveform to have a particular autocorrelation function. He has also indicated that the ideal case of an autocorrelation function with a single spike at the origin and zero value elsewhere is unrealizable.

The new approach for waveform design presented in this thesis is based on obtaining two autocorrelation functions, and then cancelling the sidelobes by combining these two functions. The peaks of the two autocorrelation functions should alternate every T_0 seconds, where T_0 is the maximum unambiguous round trip delay, and the sidelobes should be equal in magnitude and opposite in polarity at any instant. The realization of such an arrangement requires coherency between the two autocorrelation functions; it also requires the relative amplitude and shape of the sidelobes to be independent of the target range.

Recent developments in communication theory have resulted in new multi-phase modulation methods, where the modulated signal is presented by the sum of two or more

orthogonal signals, and the autocorrelation function of the modulated signal depends on the modulation technique as well as the modulating signal. Among these modulation methods, the class of four-phase modulation is of particular interest, since the modulated signal of that class can be separated into two quadrature signals which can then be processed separately to obtain two quadrature autocorrelation signals. The quadrature autocorrelations can be made to satisfy the requirements of the sidelobe cancellation if the modulating signal is a pseudorandom binary sequence (PRBS). However, a proper encoding of the PRBS is required. One particular modulation technique, Minimum (Frequency) Shift Keying, MSK, is found to satisfy these requirements with no further encoding process of the pseudorandom code. It also requires only 75% of the bandwidth which is required for the more conventional phase shift keying, PSK, modulation [36].

In the following subsections, the properties of the pseudorandom codes are first summarized in Section 4.2.2.

In Section 4.2.3, a brief discussion of the MSK modulation is given, followed by the detailed mathematical derivation of the autocorrelation function of the MSK modulated signals in Section 4.2.4. Finally, the technique used to cancel the sidelobes is presented in Section 4.2.5.

4.2.2 Pseudorandom Binary Sequences

Pseudorandom Binary Sequences (PRBS) have structures similar to that of random sequences and therefore possess desirable autocorrelation properties. The characteristics of these sequences are discussed in a number of references [29], [34], [35]. Here, we shall briefly summarize some of the important characteristics that serve the purpose of our discussion.

The length of a PRBS code is $N = 2^{(n-1)}$, where n is the number of stages in the shift register code generator. The length N is equal to the number of subpulses in the sequence.

Periodic codes may be obtained when the shift register generator is left in continuous operation. Aperiodic codes are obtained when the generator output is truncated after one complete sequence. The autocorrelation functions for these two cases differ with respect to the sidelobes structure. Fig. 4.1 illustrates the autocorrelation function for the periodic and aperiodic cases together. This figure was obtained using the Digital Processor Oscilloscope (Tektronix Model WP 1210) by cross-correlating two complete sequences against one complete sequence of a 15 element code. The sidelobe level in the middle section represents the periodic case; it is uniform at a level of (-1) , while the two end sections represent the aperiodic case.

Fig. 4.1. The cross-correlation of two complete sequences and one complete sequence of 15-element pseudorandom code.

The autocorrelation function is repetitive with a period of NT seconds where T is the subpulse duration, and with a peak value equal to N . Hence, the peak to sidelobes ratio for the periodic case is equal to N . For the aperiodic case, the sidelobe structure of each half of the autocorrelation function has odd symmetry around an average value of $(-\frac{1}{2})$ [7].

Other important characteristics of the PRBS codes are:

1. The difference between the number of $(+1)$ subpulses and (-1) subpulses is exactly one.
2. The multiplication of the code by a delayed version of itself results in the same code with a new delay.
3. Two similar consecutive pulses (e.g. $(+1)$ followed by $(+1)$ or (-1) followed by (-1)) occur $(\frac{N-1}{2})$ times within the code, while two different consecutive pulses occur $(\frac{N+1}{2})$ times.

There are other characteristics of the PRBS codes; however, they are of no immediate interest.

4.2.3 Minimum (Frequency) Shift Keying

Minimum Shift Keying (MSK) is a type of digital modulation which can be viewed either as a form of continuous phase Frequency Shift Keyed (FSK) modulation or as a special case of Offset-Keyed Quaternary Phase Shift Keyed (OK-QPSK) modulation [36], [37]. In the first FSK view, the MSK waveform can be written as: [36]

$$Y(t) = \cos(\omega_c + \frac{\pi u_k}{2T} + x_k); (k-1)T \leq t \leq kT \quad (4.1)$$

where ω_c is the carrier radian frequency, $u_k = \pm 1$ is the bipolar data being transmitted at a rate $R = 1/T$, and x_k is a phase constant which is valid over the k th binary data interval. Fig. 4.2a illustrates the FSK nature of the MSK waveform, with a radian frequency $(\omega_c + \frac{\pi}{2T})$ being transmitted for $u_k = 1$ and a radian frequency $(\omega_c - \frac{\pi}{2T})$ being transmitted for $u_k = -1$. The tone spacing in MSK is one-half that employed in conventional orthogonal FSK modulation giving rise to the name "minimum" shift keying. During each T second data interval, the value of x_k is a constant determined by the requirement that the phase of the waveform be continuous at the bit transition instants $t = kT$. Applying this requirement to the argument of Eqn. (4.1) results in the recursive phase constraint:

$$x_k = x_{k-1} + (u_{k-1} - u_k) \frac{\pi(k-1)}{2} \quad (4.2)$$

Fig. 4.2b illustrates the continuous phase, constant amplitude MSK waveform. For coherent detection, a reference value of x_k , say x_0 , can be set to zero without loss of generality. This assumption will be used in all that follows, with the result, using Eqn. (4.2), that $x_k = 0$ or π modulo 2π .

(a) Tone spacing in MSK.

(b) Continuous phase nature of MSK.

(c) Example of excess phase.

(d) MSK excess phase trellis.

Fig. 4.2. Characteristics of MSK.

We can define $\theta(t)$ as:

$$\theta(t) = x_k + \left(\frac{\pi u_k}{2T}\right)t \quad (4.3)$$

such that $\theta(t)$ is a piecewise-linear phase function of the MSK waveform in excess of the carrier term (which linearly increases in phase). Using the recursive phase constraint, $\theta(t)$ is plotted in Fig. 4.2c for a particular data sequence u_k . The phase function in Fig. 4.2c is a particular path along the phase trellis of possible paths shown in Fig. 4.2d. Fig. 4.2d illustrates that over each T second interval, the phase of the MSK waveform is advanced or retarded precisely 90° with respect to the carrier phase, depending upon whether the data for that interval is $+1$ or -1 , respectively.

The MSK modulated signal can also be expressed as an OK-QPSK waveform as follows: [37]

$$y(t) = u_Q \cos(\omega_c t) \cos\left(\frac{\pi t}{2T}\right) + u_I \sin(\omega_c t) \sin\left(\frac{\pi t}{2T}\right) \quad (4.4)$$

where u_Q is related to the original data $u(t)$ by the following relationship:

$$u_Q(m) = u(n) \oplus u_Q(m-1) \quad (4.5)$$

where $\oplus$ denotes modulo-2 addition, with a similar expression for u_I .

An expression for the power spectra of the MSK modulated signal is given by [36] as:

$$G_{\text{MSK}}(f) = \frac{8 P_c T (1 - \cos 4\pi f T)}{\pi^2 (1 - 16 T^2 f^2)^2} \quad (4.6)$$

where P_c is the power in modulated waveform. This relation is shown in Fig. 4.3. The autocorrelation function of the MSK modulated signal is presented in the following subsection.

4.2.4 The Autocorrelation Function of a MSK Modulated Signal

The waveform of an MSK modulated signal can be written in complex form as:

$$f(t) = \text{Re}\{A_c \exp[j(\omega_c t + \theta(t))]\} \quad (4.7)$$

where A_c and ω_c are the carrier amplitude and radian frequency, respectively, and

$$\theta(t) = \frac{\pi}{2T} \int_0^t u(t) dt \quad (4.8)$$

where T is the clock period of the code, and

$$u(t) = u_k = \pm 1; (k-1)T \leq t \leq kT \quad (4.9)$$

u_k is the periodic PRBS $\{u_1, u_2, \dots, u_N\}$.

In Eqn. (4.8), the reference angle $\theta(0)$ has been chosen arbitrarily to equal zero.

The autocorrelation function of Eqn. (4.7) can be written as:

Fig. 4.3. Power spectral density for MSK signal.

$$\begin{aligned}
 R(t_1) &= \frac{1}{NT} \int_{NT} \cos(\omega_c t + \theta(t)) \cos(\omega_c t + \theta(t+t_1)) dt \\
 &= \frac{1}{2NT} \int_{NT} \cos[2\omega_c t + \theta(t) + \theta(t+t_1)] dt \\
 &\quad + \frac{1}{2NT} \int_{NT} \cos[\theta(t+t_1) - \theta(t)] dt \quad (4.10)
 \end{aligned}$$

The first term of Eqn. (4.10) has spectrum around $2\omega_c$ and its average value is zero. Therefore, Eqn. (4.10) can be written as:

$$R(t_1) = \frac{1}{2NT} \int_{NT} \cos[\theta(t+t_1) - \theta(t)] dt \quad (4.11)$$

Eqn. (4.11) may be written in a complex form as:

$$\begin{aligned}
 R(t_1) &= \frac{1}{NT} \int_0^{NT} \exp[j\theta(t, t_1)] dt \\
 &= I(t_1) + j Q(t_1) \quad (4.12)
 \end{aligned}$$

where t_1 is the autocorrelation parameter and

$$\theta(t, t_1) = \theta(t+t_1) - \theta(t) \quad (4.13)$$

In Eqn. (4.12), a scaling factor of no immediate interest has been dropped. The value of the integration in Eqn. (4.12) depends on the characteristics of both the PRBS code and the modulation technique.

For a differential time delay $t_1 = nT + \tau_0$ and at any instant of time $t = mT + \tau$, where m and n are integer

numbers, and $0 < \tau_0, \tau' < T$, the phase difference $\theta(t, t_1)$ can be written as:

$$\theta(t, t_1) = \theta(mT + \tau', nT) + \frac{\pi \tau_0}{2T} u_{m+n+2} \quad (4.14)$$

It follows that:

$$\begin{aligned} \exp[j\theta(t, t_1)] &= \cos\left(\frac{\pi \tau_0}{2T}\right) \exp[j\theta(mT + \tau', nT)] \\ &\quad + j u_{m+n+2} \sin\left(\frac{\pi \tau_0}{2T}\right) \exp[j\theta(mT + \tau', nT)] \end{aligned} \quad (4.15)$$

Using Eqns. (4.12) and (4.15), gives:

$$R(t_1) = \frac{1}{NT} \left[\cos\left(\frac{\pi \tau_0}{2T}\right) x_1 + j \sin\left(\frac{\pi \tau_0}{2T}\right) x_2 \right] \quad (4.16)$$

where

$$x_1 = \int_0^{NT} \exp[j\theta(mT + \tau', nT)] dt \quad (4.17)$$

and

$$x_2 = \int_0^{NT} u_{m+n+2} \exp[j\theta(mT + \tau', nT)] dt \quad (4.18)$$

For an MSK modulated signal, the phase variation over the clock period T is $\pm \frac{\pi}{2}$; i.e.

$$\begin{aligned} \theta((k+1)T) - \theta(kT) &= \frac{\pi}{2} ; u_k = +1 \\ &= -\frac{\pi}{2} ; u_k = -1 \end{aligned} \quad (4.19)$$

Using Eqn. (4.19) and knowing that the phase of an MSK modulated signal is continuous, one gets:

$$\theta(mT + \tau', nT) = \frac{\pi}{2} \left[\sum_{k=m+1}^{m+n} u_k \right] + \frac{\pi \tau'}{2T} [u_{m+n+1} - u_{m+1}] \quad (4.20)$$

The first term of Eqn. (4.20) gives $\pm \frac{\pi}{2} n$, while the second term is a function of τ , and it can be written as:

$$\begin{aligned}
 z &= \frac{\pi \tau}{2T} [u_{m+n+1} - u_{m+1}] = 0 & ; u_{m+n+1} &= u_{m+1} \\
 &= \frac{\pi \tau}{T} & ; u_{m+n+1} &= 1 \\
 & & ; u_{m+1} &= -1 \\
 &= -\frac{\pi \tau}{T} & ; u_{m+n+1} &= -1 \\
 & & ; u_{m+1} &= +1
 \end{aligned} \quad (4.21)$$

Moreover, the phase continuity allows us to replace the integrations in Eqns. (4.17) and (4.18) by summations as follows:

$$x_1 = \sum_{m=0}^{N-1} e^{\pm j \frac{\pi n}{2}} \int_{(m-1)T}^{mT} e^{jz} d\tau \quad (4.22)$$

and

$$x_2 = \sum_{m=0}^{N-1} u_{m+n+1} e^{\pm j \frac{\pi n}{2}} \int_{(m-1)T}^{mT} e^{jz} d\tau \quad (4.23)$$

In case of $n \neq rN$, where r is an integer, and if the number of (+1)'s in the code is greater than the number of (-1)'s by exactly one, it can be shown (see Appendix A) that x_1 and x_2 take the following simple forms:

$$x_1 = -T e^{-j \frac{\pi n}{2}} \quad (4.24)$$

and

$$x_2 = T e^{-j \frac{\pi n}{2}} \quad (4.25)$$

Using Eqns. (4.16), (4.24) and (4.25), one gets:

$$R(t_1) = -\frac{1}{N} \left\{ \cos\left(\frac{\pi \tau_0}{2T}\right) - j \sin\left(\frac{\pi \tau_0}{2T}\right) \right\} e^{-j \frac{\pi n}{2}} \quad (4.26)$$

Since $t_1 = nT + \tau_0$, then $R(t_1)$ can be written as:

$$R(t_1) = -\frac{1}{N} e^{-j \left(\frac{\pi t_1}{2T} \right)} \quad (4.27)$$

Finally, the quadrature autocorrelation functions $I(t_1)$ and $Q(t_1)$ can be written as:

$$\begin{aligned} I(t_1) &= \text{Re} \left\{ -\frac{1}{N} e^{-j \left(\frac{\pi t_1}{2T} \right)} \right\} \\ &= -\frac{1}{N} \cos\left(\frac{\pi t_1}{2T}\right) \end{aligned} \quad (4.28)$$

and

$$\begin{aligned} Q(t_1) &= \text{Im} \left\{ -\frac{1}{N} e^{-j \left(\frac{\pi t_1}{2T} \right)} \right\} \\ &= \frac{1}{N} \sin\left(\frac{\pi t_1}{2T}\right) \end{aligned} \quad (4.29)$$

Eqns. (4.28) and (4.29) represent the time sidelobes of the quadrature correlation functions $I(t_1)$ and $Q(t_1)$. They have simple sinusoidal shapes and equal magnitudes.

When the differential delay between the reference and received codes is a multiple number of the code duration, the phase of the I and Q terms will be constant over the integration time, and it can take one of four values: 0, $\pi/2$, π or $3\pi/2$. As a result, one of the terms will have a peak value while the other will disappear. The phase of the MSK

modulated signal advances by $\pi/2$ for every positive pulse and is retarded by $\pi/2$ for every negative pulse. Since the PRBS have an odd number of subpulses, and with the assumption that the number of positive pulses is greater than the number of negative pulses by one, it is evident that:

$$\theta(kNT) - \theta((k-1)NT) = \frac{\pi}{2} \quad (4.30)$$

Eqn. (4.30) indicates that the peaks of the I and Q channels will alternate every NT seconds. The correlation peaks extend over a time period of 2T seconds centered at time instants of r NT, where r is any integer number. In this interval, the phase difference $\theta(t, t_1)$ can be written as:

$$\theta^\circ(t, t_1) = \frac{n\pi}{2} + \frac{\pi\tau_0}{2T} \quad ; \quad -T \leq \tau_0 \leq T \quad (4.31)$$

$$; n = rN$$

Substituting Eqn. (4.31) into (4.12) yields :

$$R^\circ(t_1) = \frac{1}{NT} \int_0^{NT} \exp(j \frac{\pi}{2} (n + \frac{\tau_0}{T})) dt \quad (4.32)$$

The I and Q terms are then given as:

$$I^\circ(t_1) = \cos(\frac{\pi\tau_0}{2T}) \cos(\frac{\pi n}{2}) \quad ; \quad n = 2rN \quad (4.33)$$

$$-T \leq \tau_0 \leq T$$

and

$$Q^\circ(t_1) = \cos(\frac{\pi\tau_0}{2T}) \sin(\frac{\pi n}{2}) \quad ; \quad n = (2r+1)N$$

$$-T \leq \tau_0 \leq T \quad (4.34)$$

Combining Eqns. (4.28), (4.29), (4.33) and (4.34) yields the complete description of the terms I and Q as follows:

$$\begin{aligned}
 I(t_1) &= -\frac{1}{N} \cos\left[\frac{\pi}{2} \left(n + \frac{\tau_0}{T}\right)\right] ; [(2rN+1)] \leq n \leq [2(r+1)N-2] \\
 &\quad ; 0 \leq \tau_0 \leq T \\
 &= \cos\left(\frac{\pi n}{2}\right) \cos\left(\frac{\pi \tau_0}{2T}\right) ; n = 2rN \quad (4.35) \\
 &\quad ; -T \leq \tau_0 \leq T
 \end{aligned}$$

and

$$\begin{aligned}
 Q(t_1) &= \frac{1}{N} \sin\left[\frac{\pi}{2} \left(n + \frac{\tau_0}{T}\right)\right] ; [(2r+1)N+1] \leq n \leq [(2r+3)N-2] \\
 &\quad ; 0 \leq \tau_0 \leq T \\
 &= \sin\left(\frac{\pi n}{2}\right) \cos\left(\frac{\pi \tau_0}{2T}\right) ; n = (2r+1)N \\
 &\quad ; -T \leq \tau_0 \leq T \quad (4.36)
 \end{aligned}$$

Fig. 4.4 illustrates the waveforms of I and Q terms for pseudorandom sequence of $N = 7$. The time sidelobes of the I and Q terms vary sinusoidally along the time axis with a period of $4T$ seconds; they are 90° out-of-phase with an amplitude of $1/N$ of that of peak value.

The expression for the autocorrelation of an MSK modulated signal as presented above is the basis of the suppressed sidelobes correlator receiver. The desirable structure of the I and Q terms of the autocorrelation function can be used to suppress the sidelobes at the receiver output as explained in the following section,

Fig. 4.4. The I and Q waveforms for $N = 7$.

4.2.5 Receiver Configuration

The functional block diagram of the receiver is shown in Fig. 4.5. The transmitted signal is an MSK modulated signal which can be expressed as:

$$x_t(t) = A_c \exp\{j(\omega_c t + \theta(t))\} \quad (4.37)$$

The received signal is a delayed and attenuated version of the transmitted one and it can be written as:

$$x_r(t) = \alpha A_c \exp\{j(\omega_c t + \theta(t - t_0) + \Psi)\} \quad (4.38)$$

where α is an attenuation factor, t_0 is the round trip delay time and $\Psi (= \omega_c t_0)$ is a target range dependent phase. The reference signal is a typical version of the transmitted one, with the only difference being that the modulating PRBS is continuously delayed in time at a slow rate. The reference signal can be written as:

$$y_r(t) = B_c \exp\{j(\omega_c t + \theta(t - \tau))\} \quad (4.39)$$

where B_c is a constant and τ is the reference code delay with respect to the transmitted one.

Upon reception the received signal is first split into two in-phase channels and then multiplied by the two quadrature components of the reference signal. The output of the two multipliers are then integrated in two identical low pass filters, yielding the signals:

Fig. 4.5. Receiver block diagram.

$$G_I(t_1) = \operatorname{Re}\left\{\frac{1}{NT} \int_0^{NT} \exp\{j(\theta_d(t, t_1) - \Psi)\} dt\right\} \quad (4.40)$$

and

$$G_Q(t_1) = \operatorname{Im}\left\{\frac{1}{NT} \int_0^{NT} \exp\{j(\theta_d(t, t_1) - \Psi)\} dt\right\} \quad (4.41)$$

where

$$\theta_d(t, t_1) = \theta(t - \tau) - \theta(t - t_0), \quad (4.42)$$

and

$$t_1 = t_0 - \tau \quad (4.43)$$

Using Eqns. (4.12), (4.40) and (4.41), the two integrator outputs can be written as:

$$G_I(t_1) = \cos \Psi I(t_1) + \sin \Psi Q(t_1) \quad (4.44)$$

and

$$G_Q(t_1) = \cos \Psi Q(t_1) - \sin \Psi I(t_1) \quad (4.45)$$

where $I(t_1)$ and $Q(t_1)$ are given by Eqns. (4.35) and (4.36), respectively. The two signals G_I and G_Q are then delayed by a time delay equivalent to one clock period T seconds, the delayed and undelayed versions of G_I and G_Q are then combined in sum and difference circuits, as shown in Fig. 4.5, resulting in the two signals γ_I and γ_Q .

$$\gamma_I = G_I(t_1 - T) + G_Q(t_1) \quad (4.46)$$

$$\gamma_Q = G_I(t_1) - G_Q(t_1 - T) \quad (4.47)$$

Using Eqns. (4.44) - (4.47) gives:

$$\gamma_I(t_1) = \cos \Psi [I(t_1 - T) + Q(t_1)] + \sin \Psi [Q(t_1 - T) - I(t_1)] \quad (4.48)$$

and

$$\gamma_Q(t_1) = \cos \Psi [I(t_1) - Q(t_1 - T)] + \sin \Psi [Q(t_1) + I(t_1 - T)] \quad (4.49)$$

The terms $[I(t_1 - T) + Q(t_1)]$ and $[I(t_1) - Q(t_1 - T)]$ can be evaluated using Eqns. (4.35) and (4.36):

$$\begin{aligned} [I(t_1 - T) + Q(t_1)] &= \sin\left(\frac{\pi t_1}{2T}\right) ; 2rNT \leq t_1 \leq (2rN+2)T \\ &= -\sin\left(\frac{\pi t_1}{2T}\right) ; [(2r+1)N-1]T \leq t_1 \leq [(2r+1)N+1]T \\ &= 0 ; \text{elsewhere} \quad (4.50) \end{aligned}$$

and

$$\begin{aligned} [I(t_1) - Q(t_1 - T)] &= \cos\left(\frac{\pi t_1}{2T}\right) ; (2rN-1)T \leq t_1 \leq (2rN+1)T \\ &= -\cos\left(\frac{\pi t_1}{2T}\right) ; (2r+1)NT \leq t_1 \leq [(2r+1)N+2]T \\ &= 0 ; \text{elsewhere} \quad (4.51) \end{aligned}$$

Substituting Eqns. (4.51) and (4.50) into (4.49) and (4.48) gives the complete expressions for the terms $\gamma_I(t_1)$ and $\gamma_Q(t_1)$ as follows.

$$\underline{(2rN-1)T \leq t_1 \leq 2rNT}$$

$$\gamma_I(t_1) = -\sin \Psi \cos\left(\frac{\pi t_1}{2T}\right)$$

$$\gamma_Q(t_1) = \cos \Psi \cos\left(\frac{\pi t_1}{2T}\right)$$

$$\underline{[(2r+1)N-1]T \leq t_1 \leq (2r+1)NT}$$

$$\gamma_I(t_1) = -\cos \Psi \sin\left(\frac{\pi t_1}{2T}\right)$$

$$\gamma_Q(t_1) = -\sin \Psi \sin\left(\frac{\pi t_1}{2T}\right)$$

$$\underline{2rNT \leq t_1 \leq (2rN+1)T}$$

$$\gamma_I(t_1) = \left[\cos \Psi \sin\left(\frac{\pi t_1}{2T}\right) - \sin \Psi \cos\left(\frac{\pi t_1}{2T}\right) \right]$$

$$\gamma_Q(t_1) = \left[\sin \Psi \sin\left(\frac{\pi t_1}{2T}\right) + \cos \Psi \cos\left(\frac{\pi t_1}{2T}\right) \right]$$

$$\underline{(2r+1)NT \leq t_1 \leq [(2r+1)N+1]T}$$

$$\gamma_I(t_1) = \left[\sin \Psi \cos\left(\frac{\pi t_1}{2T}\right) - \cos \Psi \sin\left(\frac{\pi t_1}{2T}\right) \right]$$

$$\gamma_Q(t_1) = -\left[\cos \Psi \cos\left(\frac{\pi t_1}{2T}\right) + \sin \Psi \sin\left(\frac{\pi t_1}{2T}\right) \right]$$

$$\underline{(2rN+1)T \leq t_1 \leq (2rN+2)T}$$

$$\gamma_I(t_1) = \cos \Psi \sin\left(\frac{\pi t_1}{2T}\right)$$

$$\gamma_Q(t_1) = \sin \Psi \sin\left(\frac{\pi t_1}{2T}\right)$$

$$\underline{[(2r+1)N+1]T \leq t_1 \leq [(2r+1)N+2]T}$$

$$\gamma_I(t_1) = \sin \Psi \cos\left(\frac{\pi t_1}{2T}\right)$$

$$\gamma_Q(t_1) = -\cos \Psi \cos\left(\frac{\pi t_1}{2T}\right)$$

$$\gamma_I(t_1) = \gamma_Q(t_1) = 0 ; \text{ elsewhere} \quad (4.52)$$

In Fig. 4.6, the waveforms of γ_I and γ_Q are schematically illustrated for different values of the ambiguous

Fig. 4.6a. The correlator output waveforms when $\psi = 0$.

Fig. 4.6b. The correlator output waveforms when $\psi = 45^\circ$.

Fig. 4.6c. The correlator output waveforms when $\omega = 90^\circ$.

phase Ψ . The waveshape dependency on the phase Ψ can be removed by using a square-and-add circuit at the final stage of the receiver as illustrated in Fig. 4.5. Then, the final output of the correlator is:

$$\gamma(t_1) = \gamma_I^2(t_1) + \gamma_Q^2(t_1) \quad (4.53)$$

Using Eqn. (4.52) yields:

$$\begin{aligned} \gamma(t_1) &= \left[1 + \cos\left(\frac{\pi t_1}{T}\right) \right] & ; (rN-1)T \leq t_1 \leq rNT \\ &= 2 & ; rNT \leq t_1 \leq (rN+1)T \\ &= \left[1 - \cos\left(\frac{\pi t_1}{T}\right) \right] & ; (rN+1)T \leq t_1 \leq (rN+2)T \\ &= 0 \text{ elsewhere} & \end{aligned} \quad (4.54)$$

The output waveform is illustrated schematically in Fig. 4.7. Eqn. (4.54) and Fig. 4.7 illustrate that the time sidelobes are removed at the receiver output. However, the mainlobes are widened to become $3T$ seconds instead of $2T$ seconds, which means that the system basic resolution is degraded from T seconds to $1.5T$ seconds. We should note, however, that in any system the reduction of the sidelobes is always accompanied by an increase in the mainlobe width.

4.3 SYSTEMATIC AND RANDOM ERRORS

Thus far, we have ignored the system noise together with some other effects such as the error in the delay circuit or the mismatch between the different components within the system. The sole purpose of Section 4.2 was to

Fig. 4.7. Correlator output.

establish an expression for the correlator output, and to indicate that, theoretically at least, the sidelobes are cancelled. For a practical system, however, the sidelobes level is determined by the errors at different parts of the circuits. These errors are of two kinds: systematic and random. The systematic errors are, in general, recognizable and can be corrected or minimized by careful design, high quality circuit components and system calibration while the random errors are not related to correctable components since they are produced by the thermal noise and clutter.

In the following subsections, we shall analyze the effects of these errors and evaluate the corresponding sidelobe levels.

4.3.1 Random Errors

First, we shall consider the case of random amplitude and phase variations on the sidelobe levels. A schematic diagram of the mathematical model is shown in Fig. 4.8.

The signal $x(t)$ at the multiplier input is the sum of the desired signal $S^\circ(t-t_0)$ and a noise term $n(t)$.

$$x(t) = S^\circ(t-t_0) + n(t) \quad (4.55)$$

where t_0 is the signal delay. The noise term is assumed to be an additive white Gaussian random process with zero mean.

A perfect noise-free modulated signal $S(t)$ represents the reference signal at the L.O. port of the double balanced

Fig. 4.8. Mathematical model for noise calculations.

mixer multiplier. The output signal is proportional to the multiplication of the two input signals; then,

$$y(t) = x(t) \cdot S(t) \quad (4.56)$$

The autocorrelation function of the mixer output is [38]:

$$\begin{aligned} R_y(t_1, t_2) &= E(y_1 y_2) \\ &= E\{x_1 x_2 S_1 S_2\} \\ &= E\{(S_1^o + n_1)(S_2^o + n_2) S_1 S_2\} \\ &= E(S_1^o S_2^o S_1 S_2) + E(n_1) E(S_1 S_2 S_2^o) \\ &\quad + E(n_2) E(S_1 S_2 S_1^o) + E(n_1 n_2) E(S_1 S_2) \quad (4.57) \end{aligned}$$

where $E(\cdot)$ is the expectation function, the subscripts 1 and 2 denote the two instants t_1, t_2 and $S^o \triangleq S(t-t_0)$. Since the signals and the noise are independent, then:

$$R_y(t_1, t_2) = E(S_1^o S_2^o S_1 S_2) + E(n_1 n_2) E(S_1 S_2) \quad (4.58)$$

By setting $\tau = t_1 - t_2$, and for a stationary process,

$$R_y(\tau) = R_{SS^o}(\tau) + R_S(\tau) R_n(\tau) \quad (4.59)$$

where $R_{SS^o}(\tau)$ is the cross-correlation between the transmitted and received signal, $R_S(\tau)$ is the autocorrelation of the reference waveform and $R_n(\tau)$ is the autocorrelation of the noise term. Let us first consider the term $R_S(\tau)$. If the reference signal is given by:

$$S(t) = A \cos[\omega_c t + \theta(t)] \quad (4.60)$$

where $\theta(t)$ contains the code information. Then,

$$R_S(\tau) = A^2 E\{\cos[\omega_c t_1 + \theta(t_1)] \cdot \cos[\omega_c t_2 + \theta(t_2)]\} \quad (4.61)$$

Apart from a scaling factor, the expectation function in Eqn. (4.61) represents the autocorrelation function of the reference waveform, which has been evaluated in Section 4.3, and its value is given by the 1 term in Eqns. (4.32) and (4.35). This waveform can be regarded as a multiplication of two waveforms: $a(t)$ and $\cos(\frac{\pi t}{2T})$. This idea is illustrated in Fig. 4.9. The power spectral density of $R_S(\tau)$ can be derived as follows:

If the Fourier transform of the function $a(\tau)$ is $A(\omega)$, then the Fourier transform of $R_S(\tau)$ can be written as:

$$\begin{aligned} G(\omega) &= F\{R_S(\tau)\} \\ &= \frac{1}{2} [A(\omega + \omega_0) + A(\omega - \omega_0)] \end{aligned} \quad (4.62)$$

where $\omega_0 = \frac{\pi}{2T}$. The function $a(\tau)$ is a periodic function having the following Fourier transform:

$$A(\omega) = \left\{ \frac{A^2}{N} \text{sinc}(\omega T) \sum_{n=-\infty}^{+\infty} \delta(\omega - n\omega_1) \right\} - \frac{A^2}{N} \delta(0) \quad (4.63)$$

where $\omega_1 = \frac{2\pi}{2NT}$, and $\text{sinc}(\cdot) = \frac{\sin(\cdot)}{(\cdot)}$. Substituting Eqn. (4.63) into (4.62), we obtain:

Fig. 4.9. The correlation function of the transmitted waveform as a multiplication of two waves.

$$\begin{aligned}
 G(\omega) = & \frac{A^2}{N} \left\{ \text{sinc}[(\omega + \omega_0)T] \sum_{n=-\infty}^{+\infty} \delta(\omega + \omega_0 - n\omega_1) \right. \\
 & \left. + \text{sinc}[(\omega - \omega_0)T] \sum_{n=-\infty}^{+\infty} \delta(\omega - \omega_0 - n\omega_1) \right\} \\
 & - \frac{A^2}{N} [\delta(\omega_0) + \delta(-\omega_0)] \quad (4.64)
 \end{aligned}$$

We can use the relation between ω_0 and ω_1 (i.e. $\frac{\omega_0}{\omega_1} = \frac{N}{2}$) to modify Eqn. (4.64) and rewrite it as:

$$\begin{aligned}
 G(\omega) = & \frac{A^2}{2N} \sum_{n=-\infty}^{+\infty} \left\{ \text{sinc}\left(\frac{\pi(n+N)}{N}\right) + \text{sinc}\left(\frac{\pi n}{N}\right) \right\} \delta\left(\omega - \frac{(N+2n)}{4NT}\right) \\
 & - \frac{A^2}{2N} \{ \delta(\omega - \omega_0) + \delta(\omega + \omega_0) \} \quad (4.65)
 \end{aligned}$$

The spectrum of the received signal consists of a number of spectral lines spaced by a frequency space equal to $1/2NT$.

The maximum value at the origin is:

$$G(0) = \frac{2A^2}{\pi N} \quad (4.66)$$

The noise bandwidth, B_n , is given by:

$$B_n = \frac{1}{G(0)} \int_{-\infty}^{+\infty} G(\omega) d\omega \quad (4.67)$$

$$= M \cdot \frac{1}{2NT} \quad (4.68)$$

where M is the number of spectral lines within the noise bandwidth which can be determined by equating the total power in time and frequency domain, i.e.

$$M \cdot G(0) = \frac{A^2}{2} \quad (4.69)$$

Then,

$$M = \frac{\pi N}{4} \quad (4.70)$$

Now, the noise is assumed to be white Gaussian noise and limited to a bandwidth B_n around f_0 , and with power spectral density η . Fig. 4.10 shows the power spectra for both the reference signal and the noise.

The noise term of the signal $y(t)$ is the convolution integral of the power spectra $G(f)$ and $N(f)$. This spectrum has a staircase shape with the top width equal to $1/NT$ and the bottom width equal to $2B_n$ as shown in Fig. 4.10.

The second term of Eqn. (4.59) is $R_{SS^*}(\tau)$ which represents the cross-correlation between the received and reference signals. Both signals have spectrums similar to the one given in Eqn. (4.65) and approximated as shown in Fig. 4.10. The spacing between the spectral lines, however, is not the same in the two spectra due to the clock frequency offset Δf_c . The two spectra can be written as:

$$G(f) = \frac{2A^2}{\pi N} \sum_{n=1}^{M/2} \delta(f \pm n\Delta_1) \quad (4.71)$$

and

$$G^*(f) = \frac{2a^2}{\pi N} \sum_{n=1}^{M/2} \delta(f \pm n\Delta_2) \quad (4.72)$$

where 'a' is the amplitude of the received signal and Δ_1 and Δ_2 are the frequency spacings between the spectral lines

Fig. 4.10. Power spectral densities of the signal, noise, and output noise.

in $G(f)$ and $G^o(f)$, respectively, and are given by:

$$\Delta_1 = \frac{f_c}{2N} \quad (4.73)$$

$$\Delta_2 = \frac{f_c + \Delta f_c}{2N} \quad (4.74)$$

Cross-correlating $G(f)$ and $G^o(f)$ results in the signal power spectrum at the output of the multiplier, $Z(f)$. Fig. 4.11 shows the power spectra $G(f)$, $G^o(f)$ and $Z(f)$. The bandwidth of the signal spectrum $Z(f)$ as shown in Fig. 4.11 is $M\epsilon$, where ϵ is the difference between Δ_2 and Δ_1 :

$$\begin{aligned} \epsilon &= \Delta_2 - \Delta_1 \ll \Delta_1 \\ &\ll \Delta_2 \end{aligned} \quad (4.75)$$

or

$$\epsilon = \frac{\Delta f_c}{2N} \quad (4.76)$$

At the output of the low pass filter, the total signal power is:

$$P_S = \frac{M}{2} \left(\frac{2A^2}{\pi N} \right) \cdot \left(\frac{2a^2}{\pi N} \right) \quad (4.77)$$

and the noise power is

$$P_n = (2N\epsilon) \cdot \frac{A^2 \eta}{4} \quad (4.78)$$

Using Eqns. (4.70), (4.75), (4.76), (4.77) and (4.78), the signal-to-noise ratio is:

Fig. 4.11. Spectral densities of the signals $S(t)$ and $S^o(t)$ with rectangular approximation and the filtered spectrum of $Z(f)$.

$$\frac{S}{N} = \frac{2a^2}{\pi N \Delta f_c \eta} \quad (4.79)$$

or

$$\frac{S}{N} = \frac{4P_c}{\pi N \Delta f_c \eta} \quad (4.80)$$

where P_c is the received power. For $N = 63$, $\Delta f_c = 11.73$ kHz and $\eta = k T_o$, where k is Boltzmann's constant and T_o is the noise temperature, Eqn. (4.80) gives a receiver sensitivity of -116 dBm. In practice, however, a noise figure of approximately 10 dB should be taken into account and the expected sensitivity is then -106 dBm.

Now, consider a random amplitude variation with an RMS value of δ and white spectrum over the noise bandwidth B_n . Neglecting the thermal noise, η in Eqn. (4.79) can be replaced by $\left(\frac{\delta^2}{B_n}\right)$. Then, from Eqn. (4.78), the noise power at the receiver output is:

$$P_n = 2N\epsilon \cdot \frac{A^2}{4} \cdot \frac{\delta^2}{B_n} \quad (4.81)$$

Using Eqns. (4.76), (4.70) and (4.68), we obtain:

$$P_n = \frac{2\delta^2 A^2 \Delta f_c}{\pi f_c} \quad (4.81)$$

Then, the r.m.s. value of the sidelobe level is:

$$\sigma_1 = \sqrt{P_n} = \sqrt{2} \delta A \sqrt{\frac{\Delta f_c}{\pi f_c}} \quad (4.81)$$

The peak of the main lobe is equal to aA , then:

$$(\text{Mainlobe/Sidelobe})_{\text{voltage ratio}} = \frac{a}{\sqrt{2\delta}} \sqrt{\frac{\pi f_c}{\Delta f_c}} \quad (4.82)$$

For amplitude variation of 0.1 dB, $f_c = 15$ MHz and $\Delta f_c = 11.73$ kHz. Equation (4.82) yields mainlobe/sidelobe equal to 68.75 dB.

4.3.2 Systematic Errors

In order to cancel the sidelobes at the receiver output, the differential delay between the two reference channels must be exactly T seconds. A small error in that delay might result in rapid degradation of the sidelobe level. Assuming a small error in the delay time, ΔT , and by combining Eqns. (4.35), (4.36) and (4.44)-(4.47), one gets:

$$\begin{aligned} \gamma_I = & \cos \Psi \left\{ \frac{-1}{N} \cos\left(\frac{\pi t_1}{2T} - \frac{\pi}{2} + \Delta\theta\right) + \frac{1}{N} \sin\left(\frac{\pi t_1}{2T}\right) \right\} \\ & + \sin \Psi \left\{ \frac{-1}{N} \cos\left(\frac{\pi t_1}{2T}\right) - \frac{1}{N} \sin\left(\frac{\pi t_1}{2T} - \frac{\pi}{2} + \Delta\theta\right) \right\} \end{aligned} \quad (4.83)$$

and

$$\begin{aligned} \gamma_Q = & \sin \Psi \left\{ \frac{-1}{N} \cos\left(\frac{\pi t_1}{2T} - \frac{\pi}{2} + \Delta\theta\right) + \frac{1}{N} \sin\left(\frac{\pi t_1}{2T}\right) \right\} \\ & + \cos \Psi \left\{ \frac{-1}{N} \cos\left(\frac{\pi t_1}{2T}\right) - \frac{1}{N} \sin\left(\frac{\pi t_1}{2T} - \frac{\pi}{2} + \Delta\theta\right) \right\} \end{aligned} \quad (4.84)$$

where

$$\Delta\theta = \frac{-\pi\Delta T}{2T} \quad (4.85)$$

Eqns. (4.83) and (4.84) can be written as:

$$\begin{aligned} P_1 = & \frac{1}{N} \left\{ -\cos \Psi \sin\left(\frac{\pi t_1}{2T} + \Delta\theta\right) + \cos \Psi \sin\left(\frac{\pi t_1}{2T}\right) \right. \\ & \left. - \sin \Psi \cos\left(\frac{\pi t_1}{2T}\right) - \sin \Psi \cos\left(\frac{\pi t_1}{2T} + \Delta\theta\right) \right\} \end{aligned} \quad (4.86)$$

and

$$P_2 = \frac{1}{N} \left\{ -\sin \Psi \sin\left(\frac{\pi t_1}{2T} + \Delta\theta\right) + \sin \Psi \sin\left(\frac{\pi t_1}{2T}\right) \right. \\ \left. - \cos \Psi \cos\left(\frac{\pi t_1}{2T}\right) + \cos \Psi \cos\left(\frac{\pi t_1}{2T} + \Delta\theta\right) \right\} \quad (4.87)$$

or

$$P_1 = \frac{1}{N} \left\{ \sin\left[\Psi - \frac{\pi t_1}{2T} - \Delta\theta\right] - \sin\left[\Psi - \frac{\pi t_1}{2T}\right] \right\} \quad (4.88)$$

$$P_2 = \frac{1}{N} \left\{ \cos\left[\Psi - \frac{\pi t_1}{2T}\right] - \cos\left[\Psi - \frac{\pi t_1}{2T} - \Delta\theta\right] \right\} \quad (4.89)$$

Then, for small values of $\Delta\theta$,

$$P_1 \approx \frac{\Delta\theta}{N} \cos\left(\Psi - \frac{\pi t_1}{2T}\right) \quad (4.90)$$

and

$$P_2 \approx \frac{\Delta\theta}{N} \sin\left(\Psi - \frac{\pi t_1}{2T}\right) \quad (4.92)$$

Then

$$|P_1| = |P_2| = \frac{\Delta\theta}{N} \quad (4.92)$$

Consequently, the maximum error due to imperfect sidelobe cancellation is:

$$\sigma_\theta = \frac{\sqrt{2}\Delta\theta}{N} \text{ aA} \quad (4.93)$$

The r.m.s. values in Eqns. (4.81) and (4.93) are independent and the maximum sidelobe level is:

$$\sigma_{S.L.} = \sqrt{\sigma_\theta^2 + \sigma_1^2} \quad (4.94)$$

For $N = 63$, $f_c = 15 \text{ MHz}$, $\Delta f_c = 11.73 \text{ kHz}$, $\Delta\theta = 1^\circ$ and $\delta = .1 \text{ dB}$,

and using Eqns. (4.81), (4.93) and (4.94), the sidelobe level is approximately equal to 67 dB.

4.4 CONCLUSION

A novel technique for sidelobe cancellation of a CW radar signal modulated with a pseudorandom sequence was presented in this Chapter. A closed form expression representing the correlator output was derived. The analysis shows that the cancellation of the sidelobes will result in increasing the mainlobe width from two to three clock periods, hence reducing the resolution by 30 %. The error analysis is also presented, and it was shown that the overall sensitivity of the system is approximately -106 dBm. An expression for the amplitude of residual sidelobes due to phase and amplitude error is also given. It was estimated that a random amplitude error of .1 dB and phase error of 1° will produce a residual sidelobe of approximately 67 dB below the mainlobe. The detailed design of the system and the experimental results are discussed in the following chapter.

CHAPTER 5

SYSTEM DESIGN AND IMPLEMENTATION

5.1 INTRODUCTION

In Chapter 4, we presented a mathematical analysis of the suppressed sidelobes correlator receiver, with particular reference to the suitability of such a system for the simultaneous detection of targets with vastly different reflectivities. In this Chapter, we shall discuss the details of the practical design and implementation of the system proposed in Chapter 4. The system design is directed to solve the particular problem of measuring the depth of fresh water, and the different parameters of the system are chosen to suit this application. The choice of the parameters is based on the discussion given in Chapter 2. The system has been tested under laboratory conditions by means of simulating the multiple targets using long coaxial cables, variable attenuators and power dividers.

In Section 5.2, the block diagram of the system is given, with a brief discussion of the function of each block and the interactions between the different parts of the circuit. The operation of each block is then discussed in greater detail in Sections 5.3 through 5.7. For each block, the practical and theoretical considerations involved in the design are discussed, and the results of the practical

testing are presented. In Section 5.8, the results of the laboratory simulation for one, two and three targets are presented and discussed. A comparison between the system performance and the predicted performance of other classical radar systems is presented in Section 5.9 where the Chapter is concluded by giving a summary and final comments on the system operation and performance.

5.2 SYSTEM BLOCK DIAGRAM

The transmitter and receiver systems consist of five main subsystems namely the clock processor, the code generators, the RF modulator, the RF signal processor and the detection circuit. A functional block diagram of the system is shown in Fig. 5.1.

The input to the clock processor (the first subsystem) is a sinusoidal signal of frequency $f_c = 15$ MHz. This signal is used to generate two clock signals at f_c and $f_c + \Delta f_c$, where Δf_c is a small frequency offset. The two clock signals are then used to clock two identical binary code generators in the code generator circuit. The two clock signals f_c and $f_c + \Delta f_c$ are further processed to produce two pairs of quadrature (clock /4) sinusoidal signals. One pair is fed to each of the transmitter modulators and to the receiver modulator.

The second subsystem is the code generator circuit. This block contains two identical PRBS generators, which

generate two identical PRBS codes at two different frequencies f_c and $f_c + \Delta f_c$. The codes are further processed to produce two pairs of odd and even bit streams, and one pair is fed to each of the RF modulators. The RF modulator circuits contain two MSK modulators operating at the same carrier frequency, $f_o = 100$ MHz. One of the modulators produces the transmitted signal, while the other produces the reference signal. Each modulator receives four signals, two from the clock processor and two from the code generator circuit, which are used to MSK modulate the 100 MHz carrier.

The fourth subsystem in the block diagram shown in Fig. 5.1 is the RF signal processor. In this part of the circuit, the received signal and the reference signal are split in order to form the two quadrature channels, I and Q. In each of these channels, the received and reference signals are first multiplied using a mixer and then the products are integrated in two identical low pass filters. The outputs from the filters are fed to the detection circuit. In the detection circuit, the I and Q signals are further processed to cancel the sidelobes and to shape the main lobe of the correlation function. The output from the detection circuit is then displayed using an oscilloscope.

In the following sections, the detailed design and operation of each subsystem is further discussed.

Fig. 5.1. System block diagram.

5.3 THE CLOCK PROCESSOR

The clock processor circuit has four functions:

1. To offset the input sinusoidal signal f_c by a small amount Δf_c producing two frequencies f_c and $(f_c + \Delta f_c)$.
2. To generate two clock signals from the two frequency signals f_c and $f_c + \Delta f_c$ which are then fed to the code generator circuits.
3. To generate two pairs of quadrature signals, $\{\cos(2\pi f_c t/4)$ and $\sin(2\pi f_c t/4)\}$ and $\{\cos(2\pi(f_c + \Delta f_c)t/4)$ and $\{\sin(2\pi(f_c + \Delta f_c)t/4)\}$, and to supply a pair to each of the RF modulators.
4. To generate a clock signal at a frequency equal to Δf_c to trigger the display oscilloscope.

In the following subsection, we shall discuss the implementation of these four functions in some detail.

5.3.1 - Frequency Offset

We have indicated above that the transmitted and reference codes are generated at slightly different frequencies. The frequency offset Δf_c is an important design parameter, since it will determine the noise bandwidth and hence the sensitivity of the receiver. We shall first calculate the appropriate value of Δf_c .

Consider that the transmitted code is generated at the clock frequency f_c , while the reference code is generated at $(f_c + \Delta f_c)$, then the transmitted code repetition time, T_t , is:

$$T_t = \frac{N}{f_c} \quad (5.1)$$

where N is the number of subpulses in the PRBS code. The reference code repetition time, T_r , is:

$$T_r = \frac{N}{f_c + \Delta f_c} \quad (5.2)$$

After N clock pulses, the differential delay between the transmitted and reference codes is:

$$\begin{aligned} b &= T_t - T_r \\ &= N \left(\frac{1}{f_c} - \frac{1}{f_c + \Delta f_c} \right) \end{aligned} \quad (5.3)$$

The cross-correlation between the received and reference signals requires an integration of the product signal over a period of time $T = T_t \approx T_r$ and during this time the differential delay must be nearly constant. In order to satisfy this condition, b should be much less than one clock pulse duration, i.e.

$$b \ll \frac{1}{f_c} \quad (5.4)$$

Using Eqns. (5.3) and (5.4), we have:

$$N \left(\frac{1}{f_c} - \frac{1}{f_c + \Delta f_c} \right) \ll \frac{1}{f_c} \quad (5.5)$$

Choosing $b_{\max} = \frac{1}{10f_c}$, Eqn. (5.5) gives:

$$(\Delta f_c)_{\max} = \frac{f_c}{(10N-1)} \quad (5.6)$$

For $f_c = 15$ MHz and $N = 63$, the maximum value of the frequency offset is $(\Delta f_c)_{\max} \approx 23.8$ kHz.

To ensure a linear relation between the correlator output and the range R , the differential delay between the transmitted and reference codes should be a linear function of time. This requires Δf_c to be kept constant with time. This can be achieved by locking the two clock frequencies f_c and $f_c + \Delta f_c$ together, such that any drift in one of the clock frequencies causes a similar drift in the other leaving Δf_c constant.

The clock processor block diagram is shown in Fig.

5.2. A phase locked loop (PLL) is used to offset the frequency f_c by Δf_c . The operation of the PLL can be explained as follows. The free running frequency of the voltage controlled oscillator (V.C.O.) is set initially as close as possible to f_c by adjusting the timing capacitor c . The output of the V.C.O. (f_R) is then buffered and mixed with the frequency (f_c) and the mixer output is low-pass-filtered to obtain the frequency difference ($f_R - f_c$). A voltage comparator is used to generate a square wave at the frequency difference ($f_R - f_c$) and the n th harmonic of this signal is isolated out using a narrow BPF. The harmonic signal, $n(f_R - f_c)$, is then applied to port (2) of the phase

Fig. 5.2. Clock processor.

comparator. The input frequency f_c is divided by m using a binary counter then fed to port (1) of the phase comparator. When the loop is in locking condition, the two frequencies at ports (1) and (2) are equal, i.e.

$$\frac{f_c}{m} = n(f_R - f_c) = n \Delta f_c \quad (5.7)$$

or

$$\Delta f_c = \frac{f_c}{mn} \quad (5.8)$$

In the operation described above, the PLL acts as a multiplier with a relatively large multiplication factor ($f_R/n\Delta f_c$). In such applications, it is important to take into account that the phase comparator is actually a mixer, and that its output contains the sum and difference frequency components of the input signals. The difference component is a dc signal (the error signal) which controls the frequency of the V.C.O. to keep the loop in locking condition. The sum frequency components, if not well filtered, will induce incidental FM at the V.C.O. output. This occurs because the V.C.O. is running at many times the frequency of the input signal and the sum frequency component which appears at the control voltage of the V.C.O. causes a periodic variation of its frequency. Therefore, it is generally necessary to filter the error signal quite heavily to remove the sum frequency components. The consequence of this filtering is a reduced capture range and an under-damped loop transient response.

In order to offset a 15 MHz frequency by about 20 kHz, a multiplication factor of 750 is needed. This factor is quite large, and incidental FM on the V.C.O. output is difficult to avoid. Therefore, it is necessary to lock the loop to one of the harmonics of the frequency difference Δf_c . A limitation on the ~~use~~ of harmonics, however, is that the locking range decreases with successively higher and weaker harmonics. In practice, it is difficult to use the 10th order and higher harmonics.

The values of $m = 256$ and $n = 5$ are chosen to offset the clock frequency $f_c = 15$ MHz by

$$\Delta f_c = \frac{f_c}{mn} = \frac{15 \text{ MHz}}{256 \times 5} = 11.72 \text{ kHz} \quad (5.9)$$

The low pass filter of the phase-locked loop requires a large time constant, τ , in order to adequately filter the sum frequency component. This leads to another practical problem, that of loop stability. The stability problem can be solved by using a lag-lead type of filter as shown in Fig. 5.3a.

For the PLL 562, the locking range ($2f_L$) and the capture range ($2f_{cap}$) are given by the following empirical relationships:

$$2f_L = 2k_V \quad (5.10)$$

and

$$2f_{cap} = 2 \frac{k_V}{\tau_1} \quad (5.11)$$

(b) Transfer function

Fig. 5.3. The lag-lead low-pass filter of the phase locked loop and its transfer function.

where k_V = the total open loop gain = $k_1 k_2$

and k_1 = V.C.O. conversion gain = f_o

k_2 = the phase comparator and amplifier gain

$$k_2 = \frac{.04V_s}{\sqrt{1+(V_s/40)^2}} \quad (5.12)$$

where V_s is the input signal (at port (2)) in mv rms.

For $V_s = 10$ mV rms and $f_o = 15$ MHz

$$k_V = 5.82 \times 10^6 \quad (5.13)$$

and the locking range around f_o is:

$$f_2 = \pm 5.82 \text{ MHz} \quad (5.14)$$

In order to guarantee a minimum attenuation of A dB at the radian frequency $\omega_2 = \frac{1}{\tau_2}$ in Fig. 5.3, the ratio between R_1 and R_2 can be calculated from the following relation:

$$20 \log\left(\frac{R_2}{R_1 + R_2}\right) = -A \text{ dB} \quad (5.15)$$

where R_1 is an internal resistance of the PLL 562 with a typical value of 6 k Ω . Therefore, if we choose A to be 50 dB, then from Eqn. (5.15), we obtain:

$$R_2 = 19 \Omega \approx 22 \quad (5.16)$$

Fig. 5.3b shows the transfer function of the LPF. The frequency $f_2 = \frac{1}{2\pi\tau_2}$ is chosen to be 60 kHz (approximately

one octave below the sum frequency). Then,

$$\tau_2 = 2.65 \times 10^{-6} \text{ sec} \quad (5.17)$$

and

$$C = \frac{\tau}{R_2} = 1.16 \times 10^{-7} \approx .1 \text{ } \mu\text{f} \quad (5.18)$$

It follows that:

$$\tau_1 = (R_1 + R_2)C \approx 6 \times 10^{-4} \text{ sec} \quad (5.19)$$

Substituting Eqn. (5.19) into Eqn. (5.11) gives the capture range around f_0 as:

$$f_{\text{cap}} = \pm 98.5 \text{ kHz} \quad (5.20)$$

The offset method presented above will ensure a fixed value for the clock frequency difference Δf_c , and hence allow for a linear relationship between the radar range and the correlator differential time delay.

5.3.2 Generation of the Quadrature (Clock/4) Signals

The generation of the two quadrature signals $\cos\left(\frac{2\pi f_c t}{4}\right)$ and $\sin\left(\frac{2\pi f_c t}{4}\right)$ takes place in the circuit shown in Fig. 5.4. The clock frequency, f_c , is first divided by two in a J-K flip-flop (1). The two outputs Q and $\bar{Q}$ are further divided by two in two other flip-flops (2) and (3). The outputs of these flip-flops are then low-pass filtered and a-c coupled to the RF modulator. A similar circuit is used for the other clock frequency ($f_c + \Delta f_c$).

Fig. 5.4. Divide by 4 and quadrature signals forming circuit.

5.3.4 Trigger Signal

The trigger circuit is used to generate a trigger signal at a periodic time equivalent to one complete code length (i.e. $\frac{N}{\Delta f_c}$). A binary counter is used to generate the trigger signal as shown in Fig. 5.5. The clock signal of the counter is the Δf_c signal which is taken directly from the frequency offset circuit as shown in Fig. 5.2. The counter counts up to 63 and is then reset to the initial state. The clear pulse of the counter is the trigger signal used to trigger the display oscilloscope.

5.4 THE CODE GENERATOR CIRCUIT

Linear feedback shift register generators provide a simple and convenient method for generating pseudorandom sequences. A typical shift register generator for code length $N = 63$ is shown in Fig. 5.6a and the sequence is shown in Fig. 5.6b/.

The n stages of the shift register are initially set to all 1's. The outputs from specific individual stages of the shift register are summed by a modulo-2 adder to form the input. The shift register is pulsed in its normal manner at the clock frequency rate. The output of any stage is then a pseudorandom binary sequence. When the feedback connections are properly chosen, the output is a sequence of maximal length with the desired properties explained in Chapter 4.

Fig. 5.5. Trigger circuit.

Fig. 5.6a. PRBS generator for $N = 63$.

Fig. 5.6b. PRBS for $N = 63$.

Only two components are needed (except from the clock generator), to generate a maximal length pseudorandom binary sequence. These are a modulo-2 adder and a shift register in which the various stages are accessible. Modulo-2 addition is equivalent to the logical EX-OR function.

One problem that is encountered in using an m-sequence generator is that of setting the shift register starting state. This is particularly important when two identical sequences are to be generated. The difficulty arises since most shift registers have only the capability of being reset to all zero states simultaneously, a condition which would lead to a trivial all-zeros sequence. A simple solution to avoid this difficulty is to generate a PRBS using EX-NOR feedback, obtaining a complementary code in which case the all zeros state is permitted as a starting state.

The output data y_k from the shift register generator must be processed to provide bipolar odd and even streams of the encoded data. The data processor for the transmitted signal is shown in Fig. 5.7 with a similar circuit for the reference signal.

First, the sequence is fed to a J-K flip-flop to provide the required differential encoding [39]. The relation between the input and output sequences of the first J-K flip-flop is:

$$u_k = y_k \oplus y_{k-1} \quad (5.26)$$

Fig. 5.7. Sequence processor.

where y_k is the input sequence, u_k is the output sequence and $\oplus$ denotes the modulo-2 addition.

The sequence u_k is fed simultaneously to two D-type flip-flops. The $(\text{clock}/2)$ and $(\overline{\text{clock}}/2)$ signals (generated by the second J-K flip-flop) are used to sequentially clock the two D flip-flops. Thus, two parallel sequences u_Q and u_I are formed, and then a-c coupled to the modulator.

5.5 THE MODULATOR

An MSK signal can be viewed as a quadriphase signal with offset keying and sinusoidal symbol shaping. In this visualization, the modulated waveform can be written as [39]:

$$m(t) = u_Q \cos(\omega_0 t) \cos\left(\frac{\pi t}{2T}\right) + u_I \sin(\omega_0 t) \sin\left(\frac{\pi t}{2T}\right) \quad (5.27)$$

where ω_0 is the carrier radian frequency, T is the clock period and u_Q and u_I are the quadrature and in-phase symbols (± 1). Each symbol is extended to $2T$ seconds.

Eqn. (5.27) represents the basic approach of the modulator implementation as shown schematically in Fig. 5.8.

The carrier frequency (100 MHz) is generated by means of a stable, temperature compensated, crystal controlled oscillator. The quadrature carriers are then formed by means of a 90° hybrid. The two bipolar sequences u_I and u_Q are low-pass filtered (using an RC coupler) to limit the base-band spectrum and are then used to amplitude modulate the

Fig. 5.8. MSK modulator.

two quadrature carriers $\cos(\omega_0 t)$ and $\sin(\omega_0 t)$ in mixers 1 and 2. The amplitude modulated carriers are next multiplied by the (clock/4) sinusoidal waveforms in mixers 3 and 4 resulting in the terms $\{u_Q \cos(\omega_0 t) \cos(\frac{2\pi f_c}{4} t)\}$ and $\{u_I \sin(\omega_0 t) \sin(\frac{2\pi f_c}{4} t)\}$. Finally, these two signals are summed together forming the MSK modulated signal $m(t)$.

The normalized power spectral density of the MSK modulated signal as given by S.A. Gronemeyer et al [36] is:

$$G(f) = \frac{8(1+\cos 4\pi fT)}{\pi^2(1-16T^2f^2)^2} \quad (5.28)$$

where f is the frequency offset from the carrier frequency. This equation has been derived for an infinite random sequence.

The normalized power spectral density is plotted in Fig. 4.3 with the magnitude taken as 0 dB at $f = 0$. The MSK spectrum falls off at a rate proportional to f^{-4} for relatively large values of (f/f_c) , while the main lobe of the double sided spectrum extends over $(1.5 \times f_c)$ Hz. For a clock frequency of 15 MHz, the required RF bandwidth is 22.5 MHz in contrast with the 30 MHz required for a short pulse radar having the same resolution or the approximately 40 MHz required in the case of FM-CW radars.

Fig. 5.9 shows the observed baseband spectrum of a pseudorandom binary code of $N = 63$ while the spectrum of the modulated carrier is shown in Fig. 5.10. The latter figure shows a close agreement between the measured spectrum and the theoretical spectrum shown in Fig. 4.3.

Fig. 5.9. Measured spectrum of the PRBS code.
10 dB/cm, 15 MHz/cm

Fig. 5.10. Spectrum of the MSK modulated signal.
10 dB/cm, 15 MHz/cm

5.6 THE RF SIGNAL PROCESSOR

The block diagram of the RF signal processor is shown in Fig. 5.11. The power level of the reference signal at the input (Point A) is -9 dBm. This level is raised to +3 dBm (at Point B) using a combination of a wideband amplifier (22 dB gain) and an attenuator (-10 dB). The reference signal is then divided in a wideband 90° hybrid. The received signal, on the other hand, is divided into two in-phase signals using a power divider. The received and reference signals are then multiplied in two identical mixers and the outputs are low-pass filtered producing the I and Q correlation functions.

The function of the low-pass filters is to average the mixer outputs over a time interval equal to $(\frac{N}{\Delta f})$ sec, and their identical transfer functions are chosen to match the spectrum of the correlation signal derived in Chapter 4.

The spectrum envelope of the correlator output is similar to that shown in Fig. 4.3, but the frequency axis should be normalized to Δf_c rather than f_c and the center frequency is set to zero. Therefore, the cut-off frequency of the LPF is chosen to be:

$$f_{c.o.} = 1.5 \Delta f_c \approx 17.58 \text{ kHz} \quad (5.31)$$

A second order, active low-pass filter is implemented with cut-off frequency equal to 17.58 kHz. Such a filter will pass 99.5% of the power in the desired signal.

Fig. 5.11. The RF signal processor.

5.7 THE DETECTION CIRCUIT

The block diagram of the detection circuit is shown in Fig. 5.12, while the circuit diagram is shown in Fig.

5.13. The two signals $I(t)$ and $Q(t)$ are first amplified by 40 dB. The amplified signals are then individually delayed by $\frac{1}{\Delta f_c}$ ($\approx 85.32 \mu\text{sec}$) in a dual channel CCD analog delay line. The sum and difference circuits produce the terms $[I(t) - Q(t - \frac{1}{\Delta f_c})]$ and $[Q(t) + I(t - \frac{1}{\Delta f_c})]$ in which the time side-lobes are cancelled leaving only the peaks of the correlation functions.

As has been indicated in Chapter 4, the delay and addition operation results in a distorted shape of the correlation peaks (see Fig. 4.4) depending on the propagation phase ambiguity Ψ . The square-and-add circuit removes the effect of the angle Ψ , and establishes a shape which is independent of the propagation delay (see Fig. 4.5).

The output stage of the detection circuit consists of a logarithmic amplifier and a threshold circuit. The logarithmic amplifier is used to compress the output signal for display purposes, and the threshold voltage is set to suppress the residual sidelobe fluctuations.

The circuit was implemented using $\mu\text{a} 747$ dual operational amplifiers, and each amplifier was d-c offset using a $10 \text{ k}\Omega$ potentiometer. The low-pass filter at the outputs of the CCD delay lines are similar to the filters used in

Fig. 5.12. The block diagram of the detection circuit.

Fig. 5.13. Detection circuit.

the RF signal processor circuit shown in Fig. 5.10. The RCA 3040A operational amplifiers were used to perform the square operation.

A calibration curve, which relates the received signal power level to the voltage output of the detection circuit, is given in Section 5.8.2, where the measurement procedure and the achieved results are discussed.

In order to obtain the two terms $\{I(t) + Q(t - \frac{1}{\Delta f_c})\}$ and $\{Q(t) - I(t - \frac{1}{\Delta f_c})\}$ the outputs of the two channels $I(t)$ and $Q(t)$ must be equally delayed by precisely $\frac{1}{\Delta f_c}$ seconds ($\approx 85.32 \mu\text{sec}$). Such precision and equality can be obtained by a dual CCD analog shift register.

The Fairchild CCD321M delay module was used for this purpose. The CCD321M is a complete delay module intended for use in signal processors where precisely controlled delay of analog information is required. The module is a printed circuit board containing a Fairchild CCD321 dual 455-bit analog shift register, together with the input and output signal processing circuitries. The delay time for analog signals through the CCD321M is determined by the frequency of the clock signal which can be provided by an external source or by an internal V.C.O. oscillator. The CCD321M can be used as a 910 bits single channel delay line or as two 455 bits lines for delaying two independent analog signals. The latter mode is used for delaying $I(t)$ and $Q(t)$ separately as shown in Fig. 5.12.

The delay of each channel, D , can be calculated as follows:

$$D = \frac{455}{f_{\text{clock}}} = \frac{1}{\Delta f_c} \quad (5.32)$$

Substituting $\Delta f_c = 11.72 \text{ kHz}$ in Eqn. (5.32) gives

$$\text{delay} = D = 85.32 \text{ } \mu\text{sec}$$

Therefore, the clock frequency of the delay line is

$$f_{\text{clock}} = 5.33 \text{ MHz.}$$

Fig. 5.14 illustrates schematically the CCD delay line with the conditions of the input and output signals identified. The input signals are gain controlled by individual potentiometers and a.c. coupled through 22 μf capacitors into 100 $\text{k}\Omega$ loads at the device inputs. Two emitter followers provide the sampled and hold output waveforms at a 75 Ω source impedance level. The two external LPF's at the outputs have bandwidths of 17.58 kHz and the output waveforms are delayed replicas of the input waveforms.

5.8 LABORATORY SIMULATION

The test of the system performance was carried out by means of simulating the radar targets under laboratory conditions. The test of the system sensitivity and dynamic range involves the simulation of one, two and three targets with different amplitudes and differential delays.

Fig. 5.14. Dual channel CCD delay line.

In the following subsections, we shall first describe the techniques used for target simulation. This is followed by a description of the experimental procedure and a discussion of the results obtained.

5.8.1 Target Simulation

Long coaxial cables, hybrid junctions and variable attenuators were used to simulate the targets. In the case of a single target, only a variable attenuator is used to connect the transmitter output to the receiver input, as shown in Fig. 5.15a. The power level of the received signal can be changed by simply varying the setting of the variable attenuator.

The case of two targets at different ranges is simulated using a hybrid junction and two coaxial cables of different lengths, as shown in Fig. 5.15b. The input signal from the transmitter at port (1) is divided equally at ports (2) and (3). Port (2) of the hybrid junction represents the near target. It is connected to a variable attenuator of attenuation A_1 , and a coaxial cable terminated by a short circuit at the far end, the round trip delay provided by the cable is D_1 second. Port (3), on the other hand, represents the far target. It also has a variable attenuator of attenuation A_2 , and a longer cable which is also terminated by a short circuit. The delay provided by the cable is D_2 .

(a) Case of one target.

(b) Case of two targets.

(c) Case of three targets.

Fig. 5.15. Arrangements for target simulation.

where $D_2 > D_1$. The signals from the short circuit terminations are reflected back to the hybrid junction, where they are combined at port (4) which connects to the receiver input.

The coaxial cables used have an attenuation constant of $a = .12$ dB/m, and a delay constant of $b = 4.5$ nsec/m. The characteristics of the hybrid junction used are shown in Fig. 5.16. The isolation between ports (1) and (4) of the hybrid junction is more than 50 dB over the frequency range of 1-500 MHz.

The case of three targets was simulated as shown in Fig. 5.15c. The arrangement in this case is similar to the case of two targets, except that the signal at port (3) of the hybrid is further divided, using a power splitter/combiner, forming two branches. Three cables of different lengths, and three variable attenuators are used to control the delay and attenuation in each branch. Therefore, the receiver receives three echos from the three short circuit terminations. The three echos arrive at the receiver input at different instants and with different amplitudes.

The analogy between the simulation setup and the problem of fresh water depth measurement may be made using the following parameters:

R_0 = the antenna height above the water surface = 10 m

λ = wavelength of the carrier frequency = 3 m

$|\Gamma|$ = reflection coefficient at the air-water interface = .8

Fig. 5.16. Characteristics of the hybrid used in target simulation

G = gain of the transmit/receive antenna = 4.5 dB

H = depth of water

$2L_1$ = losses at the air-water interface = 17.75 dB

L_2 = losses at the bottom terrain

Using Eqn. (2.29) in Chapter 2, the total attenuation of the air-water interface reflection is 35.4 dB, and the corresponding round trip delay is 66.67 nsec. Using Eqn. (2.33), the total attenuation of the signal received from the bottom terrain is:

$$P_2 = -(53.2 + 4H + L_1) \text{ dB} \quad (5.33)$$

while the corresponding delay, using Eqn. (2.40), is:

$$\tau_2 = 66.67 + 60 H \text{ nsec} \quad (5.34)$$

Consider the simulation setup for two targets shown in Fig. 5.17b. Port (2) represents the air-water interface, while port (3) represents the bottom terrain. Then,

$$D_1 = 2bl_1 = 66.67 \text{ nsec} \quad (5.35)$$

$$D_2 = 2bl_2 = 66.67 + 60 H \text{ nsec} \quad (5.36)$$

$$2A_1 + 2al_1 = 35.4 \text{ dB} \quad (5.37)$$

$$2A_2 + 2al_2 = 53.2 + 4H + L_2 \text{ dB} \quad (5.38)$$

where A_1 and A_2 are the variable attenuator values, and l_1 and l_2 are the lengths of the cables at ports (2) and (3)

respectively, a is the attenuation constant of the coaxial cable and b is the delay constant. For a certain value of H and L_2 , the cable lengths and the attenuation values can be determined using Eqns. (5.35) and (5.36). For example, for $H = 4$ meters and $L_2 = 20$ dB, the simulation parameters are: $\ell_1 = 7.4$ m, $\ell_2 = 34$ m, $A_1 = 17$ dB and $A_2 = 40$ dB.

In the following section, the experimental procedure is described and the results achieved are discussed.

5.8.2 Measurements and Results

Throughout the experiment, the output power of the carrier oscillator was kept constant at 6 dBm. Since the same carrier is shared by the transmit and reference modulators, each modulator receives a carrier power of 3 dBm. Furthermore, in each modulator, the carrier power is further divided in a 3 dB quadrature hybrid. Consequently, the carrier power at the inputs of mixers (1) and (2) in Fig. 5.8 is 0 dBm, which is the recommended value for the bi-phase switching application. The power level at the modulator output was measured and found to be -9 dBm. The reference signal is amplified using a 22 dB gain wideband amplifier followed by a 10 dB attenuator before the quadrature hybrid divider. Therefore, the power level of the reference signal at the input of each correlator is 0 dBm, while the power level of the received signal varies depending on the attenuation in the target simulation circuit.

The case of a single target was first examined in order to confirm the theory given in Chapter 4. The single target simulation arrangement shown in Fig. 5.15a was used, with the attenuator set at 30 dB. The waveforms at the input of the delay line are shown in the double trace photo in Fig. 5.17. The measurement shows that the output waveforms of the I and Q correlators agree with the shapes predicted by theory. The sidelobes are nearly sinusoidal and the peaks appear every $\frac{63}{\Delta f_c}$ seconds as expected. Also, the peaks have two different amplitudes because of the propagation phase Ψ . To further examine the sidelobes shape, a dual trace magnified photo of the $I(t)$ and $Q(t - \frac{1}{\Delta f_c})$ terms at the input of the summing amplifier is shown in Fig. 5.18. This figure shows clearly the nearly sinusoidal shape of the sidelobes. The mainlobe/sidelobe ratio was measured and found to be 33 dB, which is 3 dB below the theoretical value of 36 dB. The ambiguous phase, Ψ , is partially responsible for this difference. When Ψ is adjusted to be nearly equal to zero, a mainlobe/sidelobe ratio of 35 dB was measured. Figs. 5.19 and 5.20 illustrate the output of the sum amplifier with errored and perfect delays, respectively. When the delay is not exactly $\frac{1}{\Delta f_c}$, the residual sidelobes are visible in Fig. 5.19 while the sidelobes are nearly cancelled in the case of nearly perfect delay in Fig. 5.20. We should also note the distorted shape of the correlation peak after the

Fig. 5.17. $I(t)$ and $Q(t)$ at the inputs of the delay line.
Case of single target with $A_1 = 30$ dB.
30 mV/div and 2 msec/div

Fig. 5.18. $I(t)$ and $Q(t - \frac{1}{\Delta f_c})$ at the inputs of the
summing c amplifier. Case of single
target with $A_1 = 30$ dB.

Fig. 5.19. The waveform at the output of the sum amplifier. Imperfect delay case of one target. $A_1 = 30$ dB.
10 mV/cm, .5 msec/div.

Fig. 5.20. The waveform at the output of the sum amplifier near perfect delay. One target. $A_1 = 30$ dB.
10 mV/cm, .5 msec/div.

addition of terms I and Q. Such distortion is more pronounced when the peaks of the two channels are further pushed apart in order to cancel the sidelobes.

The receiver sensitivity was measured using the one target simulation setup. The attenuation between the transmitter and receiver was increased in steps and the output from the detection circuitry was observed on the oscilloscope. An overall sensitivity of approximately -95 dBm was measured which is comparable to the -106 dBm theoretical value predicted in Chapter 4. The calibration curve of the output voltage as a function of the input signal level is shown in Fig. 5.21, where the abscissa represents the attenuation A_1 which is introduced between the transmitter and receiver, while the ordinate represents the output peak in volts. The threshold voltage was taken as .1 volt.

The case of two targets was evaluated next. The first delay D_1 was set to zero while A_1 was maintained at 30 dB. A long cable was used to simulate the delay of the second target. The delay D_2 was measured using the Time Domain Reflectometer (TDR) and found to be approximately 240 nsec (round trip) which corresponds to about 34 meters of cable length. The total attenuation of the second target was 46 dB (including the attenuation introduced by the long cable). Therefore, the amplitude difference between the first and second target is 16 dB. Figs. 5.22 and 5.23 show the output

Fig. 5.21. Output voltage vs received signal attenuation.

Fig. 5.22. Case of two targets, $A_1 = 30$ dB and $A_2 = 46$ dB, $D_1 = 0$, $D_2 = 240$ nsec. Imperfect delay. 1 V/div and .5 msec/div.

Fig. 5.23. Case of two targets $A_1 = 30$ dB, $A_2 = 46$ dB. $D_1 = 0$ and $D_2 = 240$ nsec. Perfect delay. 1 V/div. and .5 msec/div.

of the receiver when the delay is imperfect and perfect, respectively. When the CCD delay differs from $\frac{1}{\Delta f_c}$, the residual sidelobes rise to about 35 dB below the strongest target level. Adjusting the delay to be nearly perfect (i.e. $\approx \frac{1}{\Delta f_c}$) results in reducing the sidelobes below the threshold level as shown in Fig. 5.23. Figs. 5.24 to 5.27 represent the case of two targets with the CCD correctly adjusted to $\frac{1}{\Delta f_c}$. The level of one target is kept constant at 30 dB, while the level of the other target is changed in steps from 45 dB to 82 dB. The second target was detected when its level was 52 dB below the first target as shown in Fig. 5.27 and this value represents the signal dynamic range at the receiver input. We should also note that the sidelobes of the strong target are kept 54 dB below the mainlobe. Finally, the case of three targets is simulated as shown in Fig. 5.15c. The three targets are separated by $D_{12} = 240$ nsec and $D_{13} = 390$ nsec; the first target attenuation was 30 dB, the second was 36 dB and the third was 46 dB and 78 dB in Figs. 5.28 and 5.29, respectively.

5.9 CONCLUSION

The experimental results presented above demonstrates the advantage of the suppressed sidelobes correlator system in reducing the time sidelobe level. The delay-and-addition process in the two channels was shown to reduce the sidelobe level from 36 dB in a single channel to 54 dB;

Fig. 5.24. Case of two targets $A_1 = 30$ dB, $A_2 = 45$ dB,
 $D_1 = 0$, $D_2 = 390$ msec.
 1 V/div., .5 msec/div.

Fig. 5.25. Case of two targets: $A_1 = 30$ dB, $A_2 = 58$ dB.
 $D_1 = 0$, $D_2 = 390$ nsec.
 1 V/div., .5 msec/div.

Fig. 5.26. Case of two targets: $A_1 = 30$ dB, $A_2 = 74$ dB.
 $D_1 = 0$, $D_2 = 390$ nsec.
 1^1 V/div., 2^2 .5 msec/div.

Fig. 5.27. Case of two targets: $A_1 = 30$ dB, $A_2 = 82$ dB.
 $D_1 = 0$, $D_2 = 390$ nsec.
 1^1 V/div., 2^2 .5 msec/div.

Fig. 5.28. Case of three targets: $A_1 = 20$ dB, $A_2 = 36$ dB,
 $A_3 = 46$ dB
 $D_1 = 0$, $D_2 = 240$ nsec, $D_3 = 390$ nsec.
 1^1 V/div., 2 .2 msec/div.

Fig. 5.29. Case of three targets: $A_1 = 30$ dB, $A_2 = 36$ dB,
 $A_3 = 78$ dB.
 $D_1 = 0$, $D_2 = 240$ nsec, $D_3 = 390$ nsec.
 1^1 V/div., 2 .2 msec/div.

that is an 18 dB improvement. Relative to classical radar systems, the novel system presented in this thesis provides a time sidelobe level which is 16 dB better than the FM-CW radar, and about 20 dB better than the short pulse system. In Table VI, the time sidelobe level of the suppressed sidelobe correlator receiver system is compared with the predicted results of several classical radar systems, and it is also compared with the idealized (but unrealizable) case of the Gaussian filter-Gaussian pulse described in Chapter 3. This ideal case is taken as a reference (zero degradation from optimum) in Table VI.

It is noted that the suppressed sidelobe correlator receiver has the closest performance to the ideal system, and its performance is suitable for water depth measurement. It should be noted also from the results obtained with three targets that the system suppresses the sidelobe of each target automatically regardless of the number of targets.

Although the proposed system is suitable for the water depth measurement, it exhibits the following disadvantages or limitations which must be taken into account when considering other applications. First, the mainlobe of the correlation function extends over three clock periods due to the delay-and-addition process. This is in contrast with a mainlobe width of only two clock periods observed before the delay-and-addition circuit. Therefore, the basic

Table VI. Sidelobe level of different radar systems.

SYSTEM	SIDELobe LEVEL, dB	DEGRADATION FROM OPTIMUM, dB
Suppressed sidelobe correlator receiver	54	7
Short pulse radar	34	27
FM-CW	38	23
Pulse compression with Hamming weighting function	42.8	18.2
Gaussian pulse-Gaussian filter	61	0

3 3

OF/DE

Resolution Test Chart

resolution capability of the overall system is reduced by approximately 30 % compared with the resolution of a single channel to allow for sidelobe cancellation modulators, working at slightly offset clock frequencies. Secondly, the proposed system exhibits a relatively low sensitivity (about -106 dBm) compared to other CW systems (e.g. FM-CW). This is due to the fact that a modulated reference signal is used to detect the code echo, which results in coloured output noise with the peak of the noise spectrum occurring in the band of the received filter (see Fig. 4.9). The third disadvantage of the system is its sensitivity to Doppler frequency shift, which limits the system applications to stationary targets or nearly so. Finally, the time required for measurement may represent another disadvantage in some applications. When $\Delta f_c = 11.73 \text{ kHz}$ and $N = 63$, the time required to scan a complete code length is approximately 5.4 msec.

CHAPTER 6

CONCLUSION AND RECOMMENDATIONS

6.1 SUMMARY

The development of a novel radar sensor capable of estimating the depth of fresh water has been described in this thesis. This new approach to radar waveform design can be used in a wide variety of remote sensing applications in layered lossy media, where the probing electromagnetic energy encounters severe attenuation. In addition, a detailed study of the electromagnetic characteristics of fresh water is given, with a description of a novel broadband technique for the measurement of the attenuation rate and the dispersion characteristics of fresh water.

The thesis is divided into two parts. The first part includes a theoretical and experimental study of the propagation characteristics in fresh water, and the nature of wave reflections at the air-water and the water-bottom interfaces. It also includes a statement of the system parameters and a survey of classical radar systems. The second part includes the theory of the novel approach for radar waveform design as well as the details of the system design, implementation and the obtained results.

In Chapter 2, a theoretical study of the EM propagation characteristics in fresh water is given. The study

summarizes the theories which describe the mechanisms of EM attenuation and dispersion in water. It was concluded that in fresh water there are two mechanisms that are responsible for EM attenuation. The first mechanism, called the dipole relaxation phenomenon, dominates the attenuation characteristics at high frequencies. At low frequencies, however, ionic conductivity dominates the attenuation characteristics. For fresh water, it was established that the intermediate range of frequencies occurs in the VHF band. In this frequency range, both mechanisms should be taken into account. Because of the lack of adequate information concerning the attenuation characteristics of water in the VHF band, it was necessary to conduct experimental measurements of the attenuation rate. A novel broadband measurement technique is described in Chapter 2. The technique employs a network analyzer to measure the S-parameters of a broadband coaxial transmission line filled with a sample of fresh water. The measurement was conducted in the frequency range 33-363 MHz and the results agree well with the theory. The attenuation rate at frequencies around 100 MHz is found to be reasonably low (2 dB/m), and hence this frequency range was chosen for the radar system.

The characteristics of the EM reflection from the air-water interface and the bottom terrain interface was also studied in Chapter 2. It was concluded that at VHF frequencies

the air-water surface can be considered perfectly smooth. Consequently, the EM reflection will be mainly specular. The water-bottom interface, on the other hand, was shown to be a rough surface, and the EM reflection will be mainly a scattered return. The losses at the bottom interface vary widely depending on the nature of the terrain, and may range from 10 to 30 dB. Based on these studies, it was concluded that the amplitudes of the top and bottom echos may differ by 40 dB or more.

The statement of system requirements given at the end of Chapter 2 indicates the necessity of using a radar system which provides a substantially reduced time sidelobe at the receiver output in order to detect the weak reflected wave from the bottom terrain.

In Chapter 3, an evaluation of the performance of different classical radar systems is given. The evaluation is based on the ability of each system to reduce the sidelobe level at the receiver output. It was concluded that the short pulse radar exhibits a sidelobe level that is limited to a minimum of 34 dB because of the critical filtering required at the IF stage and because of the imperfect antenna response over a wide band of frequencies. This value, however, is not sufficient for our application. The sidelobe levels of the FM-CW radar was shown to be limited to a minimum of approximately 38 dB. The limiting

factor in this case is the spurious amplitude modulation introduced by the target dispersion characteristics and by the non-flat response of the antenna over the large bandwidth required for this system. The conclusion of Chapter 3 indicates the necessity of developing a new technique for the transmitted waveform design, with particular attention to the reduction of the sidelobe level.

In Chapter 4, a novel approach for the design of the transmitted waveform is presented. The technique is based on obtaining two correlation functions at the receiver output, such that by a simple delay and addition process the sidelobes can be cancelled entirely. The technique utilizes the relatively new modulation method known as Minimum (Frequency) Shift Keying, MSK, which was found to have desirable autocorrelation characteristics. A detailed analysis of the new technique is presented in Chapter 4. This analysis demonstrates that the time sidelobes at the receiver output are limited only by the random noise and the systematic errors in the receiver circuit. Sidelobe levels of 67 dB for an amplitude error of .1 dB and a phase error of 1° were predicted by the analysis.

In Chapter 5, the detailed description of the system design and implementation is given, together with the achieved results obtained by simulating the target under laboratory conditions. The measurement shows an achievable

sidelobe reduction of 54 dB which is at least 12 dB better than the predicted values for classical radar systems.

Although the proposed system is suitable for the water depth measurement, it exhibits the following disadvantages, or limitations. First, it was shown in Chapters 4 and 5 that the basic spatial resolution of the system was reduced by about 30 % to allow for sidelobe cancellation. This degradation in resolution is a direct result of the delay-and-addition process. Secondly, the overall system sensitivity is relatively low (-106 dBm) compared with other CW systems. This reduced sensitivity is due to the fact that a modulated reference signal is used to detect the target echo. The use of a modulated reference signal results in coloured output noise with the peak of the noise spectrum occurring in the band of the receive filter. The third disadvantage of the system is its sensitivity to Doppler frequency shift. Although in the case of measuring the water depth using an airborne system, the Doppler effect is a minor problem, other applications may require the detection of moving targets, and in such cases the system will not perform satisfactorily. Finally, the time required for measurement may be another disadvantage in some applications. In our case, the time required for a single measurement is 5.4 msec ($N/\Delta f_c$); this time can be reduced by either increasing the frequency offset Δf_c or reducing the code length, N .

6.2 FUTURE RESEARCH WORK

Continued research work in many areas of the system presented in this thesis is required on both analytical and technological levels.

On the technological side, an additional subsystem should be incorporated to track and cancel the reflection from the antenna at the outputs of the I and Q correlators. In addition, antenna studies should be undertaken to reduce the size of the VHF antenna. Microstrip antennas might be particularly suitable for air-borne radar systems.

On the analytical side, the autocorrelation function of an MSK signal modulated by a single PRBS code should be analyzed. A successful sidelobe cancellation in this case will allow the use of the system in a pulsed mode, hence solving the problem of transmitter/receiver isolation. In addition, a method of searching a small portion of the code length should be investigated in order to reduce the measurement time.

In Chapter 4, we have indicated that recent developments in communication theory have resulted in new multi-phase modulation techniques, in which the modulated signal is the sum of a number of orthogonal signals. The possibility of adopting these recent advances in communication engineering to radar waveform design should be considered. The multi-phase received radar signals could be resolved into their

orthogonal components, and by processing these components in a microprocessor or real time digital computer, additional information about the target could possibly be obtained.

This procedure may be particularly interesting in the case of Synthetic Aperture Radar (SAR), where a two-dimensional picture is obtained by processing the received signals.

Four phase or eight phase PSK or M-ary MSK modulations could be used to modulate the carrier forming new waveforms for the SAR applications, which may have advantages over the presently used FM pulse. In addition, the possibility of adopting the technique presented in this thesis to the field of echo-canceller receivers in satellite communication systems should be investigated, particularly when the communication channel uses MSK as a digital modulation technique.

Finally, concerning potential applications of the suppressed sidelobe correlator receivers, many studies should be undertaken to adopt the system to applications such as deep space measurements, soil moisture content profiling and wide dynamic range TDR.

REFERENCES

1. Bauer, M.E., "Technological Basis and Applications of Remote Sensing of the Earth's Resources", IEEE Trans. on Geoscience Electronics, Volume GE-14, pp. 3-9, January 1976.
2. Smith, P.L., Jr., K.R. Hardy, and K.M. Glover, "Applications of Radar to Meteorological Operations and Research", Proceedings of the IEEE, Vol. 62, pp. 724-45, 1974.
3. Ulaby, F.T., "Radar Measurement of Soil Moisture Content", IEEE Transactions on Antenna and Propagations, Vol. AP-22, pp. 257-65, March 1974.
4. Special Issue on Radio Oceanography, IEEE Transactions on Antenna and Propagation, Vol. AP-25, January 1977.
5. Parry, H.D. and M.J. Sanders, Jr., "The Design and Operation of an Acoustics Radar", IEEE Transactions of Geoscience Electronics, Vol. GE-10, No. 1, pp. 58-61, January 1972.
6. Hogg, J.E., A.R. Spadaro and M. Felscher, "The Airborne Pulsed Neon Blue-Green Laser. A New Oceanographic Remote Sensing Device", Proc. of The 6th Int'l Symp. on Remote Sensing, Vol. II, pp. 1061-1074, October 1969.
7. Skolnik, M.I., "Radar Handbook", McGraw-Hill Book Company, 1970.
8. Debye, P., "Polar Molecules", The Chemical Catalogue Co. Inc., New York, 1929.
9. Collie, C.H., J.B. Hasted, and D.M. Riston, "The Dielectric Properties of Water and Heavy Water", Proc. Phys. Soc. 73, pp. 145-160, 1948.
10. Buchanan, T.J., and E.H. Grant, "Phase and Amplitude Balance Methods for Permittivity Measurements Between 4 and 50 cm", The British Journal of Applied Physics, Vol. 49, pp. 64-66, February 1955.
11. Hasted, J.B., and H.M. El-Sabeh, "The Dielectric Properties of Water in Solutions", Transactions of the Faraday Society, Vol. 49, pp. 1003-1011, 1953.

12. Lane, J.A. and J.A. Saxton, "Dielectric Dispersion in Pure Polar Liquids at Very High Radio Frequencies. I. Measurements on Water, Methyl and Ethyl Alcohols", Proc. of Royal Society, 213, pp. 400-408, 1953.
13. Grant, E.H., "A Simple and Rapid Method of Measuring the Complex Permittivity of a Liquid", British Journal of Applied Physics, Vol. 6, p. 181, May 1955.
14. Vidulich, G.A., D.F. Evans, and R.L. Key, "The Dielectric Constant of Water and Heavy Water Between 0 and 40°", Journal of Physical Chemistry, Vol. 71, No. 3, pp. 656-662, February 1967.
15. Saxton, J.S., "Electrical Properties of Water", Wireless Engineer, Vol. 26, pp. 288-292, September 1949.
16. Onsager, L., "Electric Moments of Molecules in Liquids", J. Amer. Chem. Soc. 58, pp. 1486-1493, 1936.
17. Frohlich, H., and R. Sack, "Dielectric Properties of Dipolar Substances", Proc. Roy. Soc. A, 182, pp. 338-403, 1944.
18. Royer, G.M., "The Dielectric Properties of Ice, Snow and Water at Microwave Frequencies and the Measurement of the Thickness of Ice and Snow Layers with Radar", CRC Report No. 1242, June 1973.
19. Hasted, J.B., D.M. Riston and C.H. Collie, "Dielectric Properties of Aqueous Ionic Solutions. Parts I and II.", The Journal of Chemical Physics, pp. 1-21, January 1948.
20. Harvey, A.F., "Microwave Engineering", Academic Press, London and New York, 1963.
21. Cooper, R., "The Electrical Properties of Salt-Water Solutions Over the Frequency Range 1-4000 MC/S", The Journal of the Institute of Electrical Engineers, pp. 69-75, March 1946.
22. Kerns, D.M., and R.W. Betty, "Basic Theory of Waveguide Junctions and Introductory Microwave Network Analysis", Pergamon Press, 1962.
23. Long, M.W., "Radar Reflectivity of Land and Sea", Lexington Books, 1975.

24. Long, M.W., "On the Polarization and the Wavelength Dependence of Sea Echo", IEEE Transactions on Antennas and Propagation, Vol. AP-13, pp. 749-754, September 1965.
25. Barrick, D.E., and W.H. Peake, "Scattering from Surfaces with Different Roughness Scales: Analysis and Interpretation", Battelle Memorial Institute Research Report, Nov. 1, 1967.
26. Wright, J.W., "Backscattering from Capillary Waves with Application to Sea Clutter", IEEE Transactions on Antennas and Propagation, Vol. AP-14, pp. 749-754, November 1966.
27. Woodward, P.M., "Probability and Information Theory with Applications to Radars", Pergamon Press, 1953.
28. Wheeler, G.J., "Radar Fundamentals", Prentice-Hall Inc., 1967.
29. Berkowitz, R.S., "Modern Radar", John Wiley and Sons Inc., 1965.
30. Wight, J.S., "Coherency and its Application in the Design of Ultra-High Resolution Radar for the Measurement of Thicknesses of Remote Layered Media", Ph.D. Thesis, Department of Electronics, Carleton University, June 1976.
31. Barton, D.K., "Handbook of Radar Measurements", Prentice-Hall, Inc., 1969.
32. Nathanson, F.E., "Radar Design Principles", McGraw-Hill Book Company, 1969.
33. Barton, D.K., "Radars. Vol. 3. Pulse Compression", Raytheon Company, 1975.
34. Reid, M.S., "A Millimeter Wave Pseudorandom Coded Meteorological Radar", IEEE Transactions on Geoscience Electronics, Vol. GE-7, pp. 146-156, July 1969.
35. Craig, S.E., W. Fishbein and O.E. Rittenbach, "Continuous Wave Radar with High Range Resolution and Unambiguous Velocity Determination", IRE Transactions on Military Electronics, pp. 153-161, 1962.
36. Gronemeyer, S. and A. McBride, "MSK and Offset QPSK Modulation", IEEE Transactions on Communications, Vol. COM-24, pp. 809-820, August 1976.

37. Mathwich, H.R., J.F. Balcewiz, and M. Hecht, "The Effect of Tandem Band and Amplitude Limiting on the E_c/E_b Performance of Minimum (Frequency) Shift Keying (MSK)", IEEE Transactions on Communications, Vol. COM-22, pp. 1525-1539, October 1974.
38. Davenport, W.B., and W.L. Root, "An Introduction to the Theory of Random Signals and Noise", McGraw-Hill Book Company Inc., 1958.
39. Taylor, D.P., S.T. Ogletree and S.S. Hoykin, "A Prototype 60 Mb/s Fast Frequency Shift Keying Modem", ICC Conference, pp. (S1-26)-(S1-30), June 1977.

APPENDIX A

The derivations of Eqns. (4.17) and (4.18) in Section 4.2.4 are best understood through an example. Consider the transmitted data $u(t)$ as antipolar pseudorandom cyclic sequence of $N = 15$ as shown in Fig. A.1a. The corresponding phase of the MSK modulated signal $\theta(t)$ is shown in (c). (the solid line) $\theta(t)$ is a continuous function in time. It increases by 90° for $u(t) = +1$ and decreases by 90° for $u(t) = -1$. Fig. A.1b illustrates the same code delayed by three clock periods, and the dotted line in (c) shows the corresponding delayed phase $\theta(t-3T)$ ($n=3$). When the two MSK modulated signals are multiplied in a non-linear device, the low frequency component at the output of the multiplier will have the phase difference $\theta_d(t, 3T)$ ($=\theta(t)-\theta(t-3T)$) as shown in (d). The cosine and sine waveforms are shown in (e) and (f), respectively. Now, consider the time interval $\{(m-1)T \leq t \leq mT\}$; in this interval, the output waveforms depend on the data $u(t)$ and $u(t-3T)$. There are two possibilities:

1. For $u(t) = u(t-3T)$; $(m-1)T \leq t \leq mT$

$$\theta_d(t, 3T) = \pm \frac{\pi}{2}$$

$$\cos \theta_d = 0$$

$$\sin \theta_d = 1$$

Fig. A.1. I and Q terms at the input of the integrator for a delay of $3T$.

- a) pseudorandom sequence ($N = 15$) = $u(t)$;
 b) $u(t-3T)$;
 c) the phase of the MSK modulate signals $\theta(t)$ and $\theta(t-3T)$;
 d) the phase difference $\theta_d(t, 3T) = \theta(t) - \theta(t-3T)$;
 e) $\cos(\theta_d)$; f) $\sin(\theta_d)$

Therefore,

$$\int_{(m-1)T}^{mT} \cos(\theta_d) dt = 0$$

and

$$\int_{(m-1)T}^{mT} \sin(\theta_d) dt = \pm T$$

2. For $u(t) \neq u(t-3T)$; $(m-1)T \leq t \leq mT$

$$\theta_d(t, 3T) = \pm\left(\frac{\pi}{2} - m + \frac{2\pi}{T} t\right)$$

$$\int_{(m-1)T}^{mT} \cos(\theta_d) dt = \pm \frac{T}{\pi}$$

and

$$\int_{(m-1)T}^{mT} \sin(\theta_d) dt = 0$$

We should note that the waveform of the phase difference $\theta_d(t, 3T)$ resembles the original code but it is inverted and delayed and the transition from $-\pi/2$ to $+\pi/2$ is linear. However, this waveform possesses the unique characteristic of the pseudorandom code. For instance, the number of transitions from $-\pi/2$ to $+\pi/2$ is equal to the number of transitions from $+\pi/2$ to $-\pi/2$. This leads to:

$$\sum_{m=1}^N \int_{(m-1)T}^{mT} \cos(\theta_d) dt = 0 \quad (A.1)$$

Second, the waveform of the sine term is the same pseudo-random sequence with sinusoidal pulse shaping; then

$$\sum_{m=1}^N \int_{(m-1)T}^{mT} \sin(\theta_d) dt = -T \quad (A.2)$$

Combining (A.1) and (A.2), then

$$\int_0^{NT} e^{j\theta_d(t,3T)} dt = 0 - jT \quad (A.3)$$

For $n = 5$, there will be an added phase difference of π which inverts the sine term and yields:

$$\int_0^{NT} e^{j\theta_d(t,5T)} dt = 0 + jT \quad (A.4)$$

In general, for n odd,

$$\int_0^{NT} e^{j\theta_d(t,nT)} dt = 0 + jT \sin \frac{n\pi}{2} ; n \text{ odd} \quad (A.5)$$

Fig. A.2 illustrates the case of $n = \text{even number} = 4$, following the same procedure as before:

$$\int_0^{NT} e^{j\theta_d(t,nT)} dt = -T \cos \frac{n\pi}{2} + j 0 ; n \text{ even} \quad (A.6)$$

Combining (A.5) and (A.6) yields:

$$x_1 = \int_0^{NT} e^{j\theta_d(t,nT)} dt = -T e^{-j \pi n/2} \quad (A.7)$$

An important feature of the pseudorandom code is that if the code is multiplied by itself delayed the result is the same code inverted, and the integration of the new code

Fig. A.2. I and Q terms at the input of the integrator for a delay of $4T$.

a) $u(t)$	d) $\theta_d = \theta(t) - \theta(t-4T)$
b) $u(t-4T)$	e) $\cos(\theta_d)$; f) $\sin(\theta_d)$
c) $\theta(t)$ and $\theta(t-4T)$	

always gives (-1) . This fact could be applied directly to evaluation x_2 :

$$\begin{aligned}
 x_2 &= \int_0^{NT} u_{m+n+1}(t) e^{j\theta_d(t, nT)} dt = -\{-T e^{-j n\pi/2}\} \\
 &= T e^{-j n\pi/2} \quad (A.8)
 \end{aligned}$$

END

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