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COHERENCY AND ITS APPLICATION IN THE DESIGN OF
ULTRA-HIGH RESOLUTION RADAR FOR THE MEASUREMENT OF
THICKNESSES OF REMOTE LAYERED MEDIA

by

James Stuart Wight

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

Doctor of Philosophy

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ABSTRACT

Radars used for the determination of the thicknesses of remote layered media often encounter significant problems due to conflicts between desired resolution requirements and realizable bandwidth capabilities. The design and construction of a novel radar system which provides at least an order of magnitude improvement in resolution is described in this thesis. This system is an application of a new field of harmonic interferometry based on the novel concept of harmonic coherency which makes use of multiple frequency signals.

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CHAPTER 1
INTRODUCTION

A. Ultra-High Resolution Radar Design

Radars or active sensing systems, applied to the determination of the thicknesses of ice layers or industrial films and sheets, encounter significant problems arising from conflicts between desired resolution requirements and realizable bandwidth capabilities.

Experimental electromagnetic measurements for sea ice indicate that losses are minimum over the HF and VHF spectrum⁽¹⁻⁴⁾. Consequently, a successful remote sensing system for sea ice should operate in these frequency bands. However, the maximum resolution achieved for a modulated carrier radar operating in the VHF band, say under 200MHz, is of the order of 1ft. This is unsatisfactory for many important applications ranging from the clearing of ice blockage in shipping lanes to the determination of the thickness of ice landing-strips and roadways for Arctic exploration⁽⁵⁾.

A similar problem exists in the remote, rapid and accurate, production line measurements of industrial film thicknesses such as oil, and industrial sheet thicknesses such as plexiglass or building materials, using modulated carrier radars. The large bandwidths and high centre frequencies required by these systems for satisfactory resolution are prohibitive due to high cost and stringent component requirements⁽⁶⁾. A reduction in centre frequency and bandwidth would alleviate

this problem; however, the resolution attainable would become unsatisfactory.

It is apparent that a reduction in centre frequency and bandwidth is required while maintaining the high resolution accuracy provided by conventional (modulated carrier) radar systems for both these applications. At least an order of magnitude improvement in resolution is shown in this thesis to be realizable using a novel radar system based on the coherent phase relationship which can exist between harmonically diverse electromagnetic signals.

B. Coherency

In the broadest sense of the term, the principle of coherency refers to the presence of an analytic relationship between the phases of diverse waves at different points in time^(7,8). Only through this relationship can resonance phenomena, spatially distributed wavefronts, and standing wave interference patterns exist. The operation of all homodyne measurement systems⁽⁹⁻¹³⁾, interferometers⁽¹⁴⁻¹⁶⁾, holographic systems^(17,18) and synthetic-aperture radars⁽¹⁹⁻²¹⁾ is dependent upon this principle. In addition, coherency is the basis for all passive electromagnetic components.

In its generally encountered form, that of spatial coherency, a phase reference is *conserved* between several signals of the same carrier frequency, each propagating along separate paths. A second and novel

type of coherency, called harmonic coherency, can be developed by use of a space-frequency transformation presented in this thesis. As a result of this transformation, a phase reference can be conserved between several signals propagating along one path, each with a different carrier frequency. With this concept as a basis, a new field of harmonic interferometry can be introduced and applied to the design of ultra-high resolution radar for remote sensing of layered media.

C. Organization

The familiar principle of spatial coherency is presented in the following chapter and two of its fundamental properties discussed. A space-frequency transform is then presented which introduces the concept of harmonic coherency and develops the new field of harmonic interferometry. As a result, the phase reference which is maintained in a multipath, single-frequency interferometric system can also be conserved in a single-path, multi-frequency system. This new system readily lends itself to remote phase measurements, since its channel paths always maintain equal length for non-dispersive propagation.

In Chapter 3, harmonic interferometry is applied to the radar measurement of the bulk characteristics of remote layered media. Due to properties borrowed through the duality relationship with spatial interferometry, a bi-frequency system, used as a radar to determine the physical separation of two dielectric interfaces, yields at least an order of magnitude improvement in resolution over a modulated carrier

radar occupying the same frequency band. Consequently, at least an order of magnitude reduction in centre frequency and bandwidth can be employed without loss of resolution performance. The construction of two systems, one for the measurement of ice thickness and one for industrial film and sheet thicknesses, is described in this Chapter and the experimental data for these systems are presented.

Finally, conclusions and a brief discussion on related areas is given in Chapter 4.

CHAPTER 2

THE PRINCIPLE OF COHERENCY

A. Spatial Coherency

Spatial coherency refers to the conservation of a phase reference between several spatially distributed and differentially delayed signals of the same frequency. The two fundamental properties of this principle, the maximum phase resolution obtainable in the presence of noise and the maximum differential time delay tolerable between the signals, determine two important theoretical constraints for spatial interferometric applications.

The maximum phase resolution which can be achieved between two spatially isolated signals can be readily derived for a typical phase measurement system in which a low power, noise embedded signal is received from a target and is compared with a high power, noise free transmitter signal. Assuming a Gaussian noise of bandwidth, BW, the noise superimposed on the low power CW signal fluctuates at the same average frequency as the CW signal giving rise to the waveform:

$$z(t) = \text{Re} \left\{ [A + x(t) + jy(t)] e^{j\omega_0 t} \right\} \quad (1)$$

where A is the signal amplitude and x(t) and y(t) are the in-phase and phase quadrature components of the noise. Assuming the signal is large compared to the noise, the noise will manifest itself as a phase shift as shown in Fig. 1.

Fig. 1. Noise induced phase shift of a CW signal.

$$\phi(t) = \tan^{-1} \phi(t) = y(t)/A \quad (2)$$

Consequently, the statistical variance of the time suppressed phase, σ_ϕ^2 becomes:

$$\sigma_\phi^2 = \frac{1}{A^2} \sigma_y^2 \quad (3)$$

where σ_y^2 is the variance of the phase quadrature noise component.

Using Rice's theorem (22,23), $\sigma_x^2 = \sigma_y^2 = \sigma_N^2$. Eqn. (3) becomes:

$$\sigma_\phi^2 = \frac{1}{A^2} \sigma_N^2 \quad (4)$$

The average power of the narrow band noise is:

$$\sigma_N^2 = \int_0^\infty N_o |\Psi(f)|^2 df \quad (5)$$

$$= N_o BW \quad (6)$$

where N_o is the one-sided power spectral density of the noise, $\Psi(f)$ is the filter transfer function and BW is the effective bandwidth. Consequently, the variance of the phase in Eqn. (4) becomes:

$$\sigma_\phi^2 = \frac{N_o BW}{A^2} \quad (7)$$

Since N is Gaussian with zero mean, the probability density function for the phase error must also be Gaussian with zero mean:

$$f(\phi) = \frac{1}{\sqrt{2\pi} \sigma_\phi} e^{-\phi^2 / 2\sigma_\phi^2} \quad (8)$$

A phase error which ~~is~~ exceeded only 1% of the time can be determined by:

$$2 \int_0^{\phi_0} f(\phi) d\phi = .99 \quad (9)$$

and equals

$$\phi_0 = 1.6\sigma_\phi \quad (10)$$

Using Eqn. (7), Eqn. (10) becomes:

$$\phi_0 = \frac{1.6}{\sqrt{A^2 / N_0 BW}} \quad (11)$$

This equation, which relates the signal to noise ratio necessary to achieve a desired error in phase, describes an important theoretical constraint for interferometric systems. For example, a received signal to noise ratio of 30dB is required by all systems in order to maintain the phase uncertainty below 3 electrical degrees for 99% of the time.

The second fundamental property of coherency which is important in interferometric applications, is the maximum differential time delay which can be tolerated between the separate signals without disturbing their analytical phase relationship. This delay is governed by the effective bandwidth of the AM and FM noise associated with the signals, as becomes evident from a consideration of an oscillator

represented in complex notation by:

$$z_1(t) = \text{Re} \left\{ n(t) e^{j\theta} e^{j\omega_o t} \right\} \quad (12)$$

Here, the envelope consists of the modulating noise, $n(t)$ of bandwidth BW and a random phase angle θ uniformly distributed from 0 to 2π . If this signal is split and differentially delayed, the resulting two signals will remain coherently related for any delay less than the autocorrelation interval of the noise⁽²⁴⁾:

$$R_n(\tau) = \frac{1}{BW} \quad (13)$$

Since this interval is inversely related to the bandwidth of the modulating noise, a large difference in signal path lengths can be tolerated for noise-free oscillating sources. This maximum time interval then becomes the second important design parameter, especially for large path-length interferometric systems.

B. Harmonic Coherency

A second and novel type of coherency called *harmonic coherency* is introduced in this thesis by the application of a space-frequency duality relationship illustrated in Fig. 2. Physically discrete, single frequency channels generated in the power splitter of Fig. 2a are transformed by this relationship into harmonically discrete, single path channels generated in the harmonic generator of Fig. 2b. Consequently, the phase reference which is conserved between signals of

(a)

(b)

(c)

Fig. 2. Spatial and harmonic coherence.

- a) Spatial coherence
- b) Harmonic coherence
- c) Space-frequency transform

the same frequency propagating along separate paths in spatial coherency applications, can also be conserved between signals of separate frequencies propagating along the same path as shown in Fig. 2c. The new reference called the *dispersive-phase reference* is defined as the phase of the signal at the higher frequency minus the phase of the signal at the lower frequency multiplied by the frequency ratio.

Harmonic coherency then is the conservation of a dispersive-phase reference between several, harmonically distributed and differentially delayed signals, propagating along the same spatial channel. The two fundamental properties of maximum dispersive-phase resolution and maximum differential time delay again determine important theoretical constraints.

The maximum dispersive-phase resolution which can be achieved between two harmonically isolated, noise embedded signals can be derived from the following analysis. Using Eqn. (2), the random variables for the time-suppressed phase of each CW signal can be written:

$$\phi_1 = \frac{Y_1}{A_1} \quad (14)$$

$$\phi_2 = \frac{Y_2}{A_2} \quad (15)$$

Assuming that the two CW signals each have amplitude A and that Y_1 and Y_2 are independent, Gaussian and have zero mean and variance σ_ϕ^2 , the probability density functions are:

$$f_{\phi_1}(\phi) = \frac{1}{\sqrt{2\pi} \sigma_{\phi}^2} e^{-\phi^2/2\sigma_{\phi}^2} \quad (16)$$

and

$$f_{\phi_2}(\phi) = \frac{1}{\sqrt{2\pi} \sigma_{\phi}^2} e^{-\phi^2/2\sigma_{\phi}^2} \quad (17)$$

Therefore, the characteristic functions are:

$$\theta_1(\omega) = E\{e^{j\omega\phi_1}\} = e^{-\sigma_{\phi}^2 \omega^2/2} \quad (18)$$

and

$$\theta_2(\omega) = E\{e^{j\omega\phi_2}\} = e^{-\sigma_{\phi}^2 \omega^2/2} \quad (19)$$

The random variable for the dispersive-phase between the two CW signals becomes:

$$\phi = \phi_2 - \phi_3 \quad (20)$$

where

$$\phi_3 = R\phi_1 \quad (21)$$

$$f_{\phi_3}(\phi) = \frac{1}{|R|} f_{\phi_1}\left(\frac{\phi}{R}\right) \quad (22)$$

and R is the frequency ratio of the two signals.

Now, the probability distribution function for ϕ is defined as:

$$\begin{aligned} F_{\phi}(\phi) &= P\{\phi < \phi\} = P\{(\phi_2, \phi_3) \in D_{\phi}\} \\ &= \iint_{D_{\phi}} f_{\phi_2\phi_3}(\phi_2, \phi_3) d\phi_2 d\phi_3 \end{aligned} \quad (23)$$

and therefore, using Eqn. (20), Eqn. (23) yields:

$$F_{\phi}(\phi) = \int_{-\infty}^{\infty} \int_{-\infty}^{\phi+\phi_3} f_{\phi_2\phi_3}(\phi_2, \phi_3) d\phi_2 d\phi_3 \quad (24)$$

Differentiating with respect to ϕ_2 , Eqn. (24) becomes:

$$f_{\phi}(\phi) = \int_{-\infty}^{\infty} f_{\phi_2\phi_3}(\phi-\phi_3, \phi_3) d\phi_3 \quad (25)$$

But ϕ_2 and ϕ_3 are statistically independent since Y_1 and Y_2 are independent. Therefore,

$$f_{\phi}(\phi) = \int_{-\infty}^{\infty} f_{\phi_2}(\phi-\phi_3) f_{\phi_3}(\phi_3) d\phi_3 \quad (26)$$

Using Eqn. (22), Eqn. (26) becomes:

$$f_{\phi}(\phi) = \frac{1}{|R|} \int_{-\infty}^{\infty} f_{\phi_2}(\phi-\phi_3) f_{\phi_1}\left(\frac{\phi_3}{R}\right) d\phi_3 \quad (27)$$

Fourier transforming Eqn. (27) and using Eqns. (18) and (19) gives:

$$\begin{aligned} \Theta(\omega) &= \Theta_1(\omega) \Theta_2(R\omega) \\ &= e^{-(R+1)\sigma_{\phi}^2 \omega^2 / 2} \end{aligned} \quad (28)$$

Consequently;

$$f_{\phi}(\phi) = \frac{1}{\sqrt{4\pi(R+1)\sigma_{\phi}^2}} e^{-\phi^2 / 2(R+1)\sigma_{\phi}^2} \quad (29)$$

and the variance of the dispersive-phase σ^2 equals

$$\sigma^2 = (R+1)\sigma_\phi^2 \quad (30)$$

Substituting the above equation in Eqn. (7) gives:

$$\sigma^2 = \frac{(R+1)N_0 BW}{A^2} \quad (31)$$

and Eqn. (11) becomes:

$$\phi_o = \frac{1.6(R+1)}{\sqrt{A^2/N_0 BW}} \quad (32)$$

This powerful equation allows a desired error in the dispersive-phase of a potential system to be related to a required value for the signal to noise ratio. It acts then as an important theoretical constraint for the resolution limit which can be achieved in all applications of harmonic coherency.

The second fundamental property of harmonic coherency, the maximum differential time delay which can be tolerated between the separate signals (harmonics), is governed by the method chosen for generating the separate signals. In one potential method, an oscillator is phase-locked to the second harmonic of a signal given by Eqn. (12) to obtain:

$$z_2(t) = \text{Re} \left\{ m(t^2) e^{j\Psi(\omega_o, \omega_o', \omega_o'', \dots)} e^{j2\theta} e^{j2\omega_o t} \right\} \quad (33)$$

where Ψ is a time varying probabilistic phase due to the tracking error of the phase-locked loop which results from the AM and FM noise associa-

ted with the two sources. If $z_1(t)$ and $z_2(t)$ are differentially delayed, a deterministic dispersive phase relationship will result only for a delay less than the autocorrelation interval of Ψ ; $R_\Psi(t)$.

A second alternative method generates the second harmonic of the modulated source by means of a non-linear two port⁽²⁵⁾:

$$z_3(t) = \text{Re} \left\{ m(t^2) e^{j2\theta} e^{j2\omega_0 t} \right\} \quad (34)$$

If $z_1(t)$ and $z_3(t)$ are differentially delayed, a deterministic dispersive phase relationship will now result for any delay less than the autocorrelation interval of $m(t^2)$; $R_{m(t^2)}(\tau)$. Since typically $R_{m(t^2)}(\tau) \gg R_\Psi(\tau)$, frequency diverse coherent signals capable of much larger differential time delays can be generated by the non-linear two port rather than by the phase-locked loop.

The above two properties of maximum dispersive-phase resolution and maximum differential time delay provide important constraints on all applications of harmonic coherency. The new field of harmonic interferometry which is among the most important of these applications is outlined in the following section.

C. Harmonic Interferometry

The definition of *dispersive-phase* leads in a natural manner to the concept of harmonic interferometry. By an application of the space-frequency duality relationship, a conventional interferometric

phase measurement system, shown in Fig. 3a, can be transformed into the novel *dispersive-phase* measurement system of Fig. 3b. This measurement system represents a new diagnostic technique which can be used in a similar manner to conventional interferometry but which has several unique characteristics. The most powerful of these characteristics is the independence of the amplitude and phase measurements made by the harmonic interferometer on the distance to the target.

Analogous to the measurement of absolute-phase and absolute-loss of a target made by conventional interferometers, harmonic interferometers measure the dispersive-phase and dispersive-loss of a target. The dispersive-phase is defined again as the phase of the signal at the higher frequency minus the phase of the signal at the lower frequency multiplied by the frequency ratio. The dispersive-loss can be defined as the ratio of the absolute-loss of each signal (frequency). Based on these definitions, dispersive-phase and dispersive-loss remain constant during both guided and free-space propagation.

Since the measurement of dispersive-phase and dispersive-loss are independent of the propagation distance, the harmonic interferometer becomes independent of the requirements of constant and equal path lengths which restrict the use of conventional interferometric systems. Since this technique readily lends itself to totally remote measurements, interferometry can for the first time, leave the laboratory

Fig. 3. Spatial and harmonic interferometric transmission measurement systems.

- a) Spatial interferometer
- b) Harmonic interferometer

environment and perform field measurements under conditions of uncertain and time-varying path lengths.

An important application of harmonic interferometry is the high accuracy radar measurement of the thickness of remote layered media presented in the next chapter.

CLASSIFICATION

CLASSIFICATION

CHAPTER 3MULTIPLE FREQUENCY HOMODYNE RADARA. General Radar Design Considerations

Modulated carrier radar systems, such as FM or impulse radars, possess an inherent range precision capability dependent on the signal to noise ratio at the receiver and on the time spread of the pulse⁽²⁶⁾. The precision capabilities of these radar systems are unsatisfactory in certain applications such as the measurement of the thickness of ice layers or industrial sheets and films, and an alternative method of approach becomes desirable. Such a method using the concepts of coherency presented in Chapter 2 will be developed in this section.

A comparison of the single target range precision analyses governing a modulated carrier radar and a single frequency continuous wave radar based on spatial interferometry is first presented. The capability of this continuous wave radar is then extended to include multiple targets by the simultaneous use of several single-frequency, continuous wave systems. Severe limitations of this arrangement are pointed out and are avoided by the introduction of a continuous wave radar based on the new field of harmonic interferometry. A description of this novel radar, called the *multiple frequency homodyne radar* is then given.

i) Single Target Range Precision

Radar range precision analyses are usually derived in terms of matched filter front-end performances as shown in Fig. 4a⁽²⁶⁾. However, these analyses are restricted to finite energy signals such as modulated carrier (MC) signals. For infinite energy signals such as continuous wave (CW) signals, a matched filter is impossible to build since it is required to store an infinite amount of energy for an infinite duration of time. Consequently, a bandpass filter must be used as shown in Fig. 4b. Since the present analysis will be used to compare the realizable performance of an MC system and a CW system, a comparison of the matched filter response and the bandpass filter response of the two systems respectively, will be made.

The broadband noise at the input of a matched filter of an MC radar system is changed at the output into narrow band noise fluctuating at the same average frequency as the filter response. Superposition of this noise onto the matched filter response gives:

$$z(t) = \text{Re} \left\{ [v(t) + x(t) + jy(t)] e^{j\omega_0 t} \right\} \quad (35)$$

where $v(t)$ is the complex envelope of the filter response and $x(t)$ and $y(t)$ are the in-phase and phase quadrature components of the noise. To simplify the calculation, the filter response can be represented mathematically by a pure sinusoid of the same amplitude and half-power width as the central peak as shown in Fig. 5. The noise will now

(a)

(b)

Fig. 4. Radar receivers for single target range precision analyses.

- a) Modulated carrier radar receiver
- b) Continuous wave radar receiver

Fig. 5. Modulated carrier radar response with sinusoidal equivalent.

manifest itself predominantly as a phase shift of this sinusoid and the analysis of Chapter 2A can be used to obtain a phase error, ϕ_0 , which is exceeded only 1% of the time. Letting A be the amplitude of the central peak of the signal, N_0 the one-sided power spectral density of the noise, and BW the effective bandwidth of the matched filter, Eqn. (11) yields:

$$\phi_0 = \frac{1.6}{\sqrt{A^2/N_0 BW}} \quad (36)$$

If the matching cosine has a frequency f , the relation between phase and delay is

$$\tau = \frac{1}{f} \frac{\phi}{2\pi} \quad (37)$$

Since

$$f = \frac{1}{4T} \quad (38)$$

where T is the half-power width of the signal, Eqn. (37) becomes:

$$\tau = \frac{2}{\pi} T \phi \quad (39)$$

With this, Eqn. (36) gives:

$$\tau_0 = \frac{T}{\sqrt{A^2/N_0 BW}} \quad (40)$$

As a result, the required signal to noise ratio necessary to achieve a desired error in delay for an MC waveform with a specified half-power width can be found.

In certain applications, the distance to a single, isolated target may be known to within four range bins* but must be located more accurately. In this case, a CW spatial interferometric system matched to the pulse envelope of an MC system will provide a large theoretical improvement in precision.

Since the amplitude and period of the CW signal is matched to the pulse of the MC signal, the ratio of the errors in delay given by Eqn. (40) can be written:

$$\frac{\tau_{O \text{ CW}}}{\tau_{O \text{ MC}}} = \sqrt{\frac{BW_{\text{CW}}}{BW_{\text{MC}}}} \quad (41)$$

This powerful relationship compares for the first time the resolution achievable for an MC and a CW radar and is dependent only on the ratio of the bandwidths of the two front-end filters. For an MC system, the bandwidth of the matched filter is approximately equal to the inverse of the half-power width of the signal pulse:

$$BW_{\text{MC}} = 1/T \quad (42)$$

For a CW system, bandpass filters with .04% to 4% relative bandwidths are easily constructed. Using Eqn. (38); this gives:

$$\frac{.0001}{T} < BW_{\text{CW}} < \frac{.01}{T} \quad (43)$$

* One range bin is defined as that range interval which corresponds to the half-power width of the central lobe of the radar response.

Substituting Eqn. (42) and (43), Eqn. (41) yields:

$$.01 < \frac{\tau_o \text{ CW}}{\tau_o \text{ MC}} < .1 \quad (44)$$

This equation demonstrates that one or two orders of magnitude improvement in range precision can be achieved for the CW system as compared to the MC system. Alternatively, one or two orders of magnitude reduction in centre frequency becomes possible for the CW system while maintaining the high resolution capability of the MC system.

ii) Multiple Target Range Considerations

In a multiple target environment, the envelope returns of an MC system are uncorrelated and consequently the various distances to the targets are dependent only on the wave velocity. For a CW system based on spatial coherency, the returns are correlated⁽²⁷⁾ and the situation increases in complexity. If the number of independent target returns is matched with a corresponding number of CW spatial interferometric systems, a system of equations can be set up for the various distances. In general, the system of equations is dependent on the velocity of propagation, return cross-sections, and attenuation constants. For targets such as ice layers or industrial films and sheets, which are composed of a known number of layers, each having a known dielectric constant and loss tangent, the system of equations becomes solvable for the various thicknesses.

This becomes evident from a consideration of the return signals from two dielectric interfaces illuminated by two harmonically related CW spatial interferometric radars as shown in Fig. 6a. Assuming for the moment that no multiple reflections occur, the absolute-phases at the receiver for both the fundamental and second harmonic CW radars are shown in Figs. 6b and c respectively, and can be written:

$$Ph_1 = 2\beta\ell_1 + \omega_0 t + \Lambda_1(\epsilon_2, \sigma_2, \ell_2) \quad (45)$$

$$Ph_2 = 4\beta\ell_1 + 2\omega_0 t + \Lambda_2(\epsilon_2, \sigma_2, \ell_2) \quad (46)$$

where β = propagation constant

ω_0 = angular frequency in rad/sec

t = time in sec.

If each of these phase responses are measured and the dispersive-phase is calculated (i.e. twice the phase of the fundamental CW radar is subtracted from the phase of the second harmonic CW radar), the dispersive-phase is found to be completely independent of time and distance to the target, and is only a function of the dielectric constant ϵ_2 , the conductivity σ_2 and the thickness ℓ_2 of the dielectric layer:

$$Ph_2 - 2Ph_1 = \Lambda_2(\epsilon_2, \sigma_2, \ell_2) - 2\Lambda_1(\epsilon_2, \sigma_2, \ell_2) \quad (47)$$

Since ϵ_2 and σ_2 are known apriori in applications such as the measure-

Fig. 6. CW radar responses of a single layer target.
 a) Reflections from the target
 b) Fundamental harmonic CW radar response
 c) Second harmonic CW radar response

ment of ice layers and industrial film and sheet thicknesses, a function $Ph_2 - 2Ph_1$ involving only ℓ_2 is obtained.

Similarly, if both of the absolute losses are measured and the dispersive-loss is calculated (i.e. their ratio is taken), the dispersive-loss is found to be independent of time and distance to the target, and is a function only of ϵ_2 , σ_2 and ℓ_2 . Again since ϵ_2 and σ_2 are known apriori, a function involving only ℓ_2 is left.

These two functions for the dispersive-phase and dispersive-loss are each single valued over an interval of ℓ_2 corresponding to $\lambda/4$ of the difference signal of the two harmonics. Since these functions will also resolve each other, this interval can be extended to an unambiguous thickness corresponding to $\lambda/2$.

The rigorous mathematical description governing the relationship between the spatial interferometric measurements of Ph_1 and Ph_2 and the physical characteristics of the dielectric layers is based on the physical interaction of the electromagnetic waves. Assuming two uniform plane waves incident on the target, the Fresnel reflection coefficient for each frequency is:

$$\frac{E_r}{E_i} = \text{Re} \left\{ \frac{\left(\frac{1-\epsilon_2}{1+\epsilon_2} \right) \left(\frac{\epsilon_2+\epsilon_3}{\epsilon_2-\epsilon_3} \right) + e^{-2\gamma_2 \ell_2}}{\left(\frac{1+\epsilon_2}{1-\epsilon_2} \right) \left(\frac{\epsilon_2+\epsilon_3}{\epsilon_2-\epsilon_3} \right) + e^{-2\gamma_2 \ell_2}} \right\} \quad (48)$$

where

E_r = reflected electric field intensity in region 1 at $z=0$

E_i = incident electric field intensity in region 1 at $z=0$

$$\xi_j = \eta_1 / \eta_j = \sqrt{\frac{\epsilon_{rj}}{\epsilon_{r1}}} ; j=1,2 \quad (49)$$

$$\gamma_2 = \frac{\sigma_2 \eta_1}{2\xi_2} + j \frac{\omega \xi_2}{v} \quad (50)$$

η_j = characteristic impedance of medium j

ϵ_{rj} = relative dielectric constant of medium j

σ_j = electrical conductivity of medium j in mhos/m

ℓ_j = physical thickness of medium j in m

ω = angular frequency in rad/sec

v = free space velocity of light in m/sec

Letting:

$$A = \frac{1+\xi_2}{1-\xi_2} \quad (51)$$

$$B = \frac{\xi_2 + \xi_3}{\xi_2 - \xi_3} \quad (52)$$

$$C = \frac{\sigma_2 \eta_1}{2\xi_2}$$

and transforming E_i and E_r to a reference plane at $z=-\ell_1$ (the plane of the radar sensor), the received signal becomes:

$$\begin{aligned}
E_r = \operatorname{Re} \left\{ E_{t1} \left[\frac{B/A+e^{-2(C+j\frac{\omega_o}{v}\xi_2)\ell_2}}{AB+e^{-2(C+j\frac{\omega_o}{v}\xi_2)\ell_2}} \right] e^{j\left[\theta - \frac{2\omega_o}{v}\ell_1\right]} e^{j\omega_o t} \right. \\
\left. + E_{t2} \left[\frac{B/A+e^{-4(C+j\frac{\omega_o}{v}\xi_2)\ell_2}}{AB+e^{-4(C+j\frac{\omega_o}{v}\xi_2)\ell_2}} \right] e^{j\left[2\theta - \frac{4\omega_o}{v}\ell_1\right]} e^{j2\omega_o t} \right\} \quad (54)
\end{aligned}$$

Adding narrow band noise, normalizing independently the complex envelopes of each harmonic and considering the near field phase factors, Eqn. (54) yields:

$$\begin{aligned}
E_r = \operatorname{Re} \left\{ \left[1+x_1(t)+jy_1(t) \right] e^{j\left[\Lambda_1 + \theta - \frac{2\omega_o}{v}\ell_1 + 2\tan^{-1}\left(\frac{\omega_o\ell_1}{v} - \frac{v}{\omega_o\ell_1}\right)\right]} e^{j\omega_o t} \right. \\
\left. + \left[1+x_2(t)+jy_2(t) \right] e^{j\left[\Lambda_2 + 2\theta - \frac{4\omega_o}{v}\ell_1 + 2\tan^{-1}\left(\frac{2\omega_o\ell_1}{v} - \frac{v}{2\omega_o\ell_1}\right)\right]} e^{j2\omega_o t} \right\} \quad (55)
\end{aligned}$$

where

$$\begin{aligned}
\Lambda_1 = \tan^{-1} \left\{ \frac{e^{-2C\ell_2} \sin\left(-\frac{2\omega_o}{v}\xi_2\ell_2\right)}{B/A+e^{-2C\ell_2} \cos\left(-\frac{2\omega_o}{v}\xi_2\ell_2\right)} \right\} \\
- \tan^{-1} \left\{ \frac{e^{-2C\ell_2} \sin\left(-\frac{2\omega_o}{v}\xi_2\ell_2\right)}{AB+e^{-2C\ell_2} \cos\left(-\frac{2\omega_o}{v}\xi_2\ell_2\right)} \right\} \quad (56)
\end{aligned}$$

and

$$\Lambda_2 = \tan^{-1} \left\{ \frac{e^{-4Cl_2} \sin\left(-\frac{4\omega_0}{v} \xi_2 l_2\right)}{B/A + e^{-4Cl_2} \cos\left(-\frac{4\omega_0}{v} \xi_2 l_2\right)} \right\}$$

$$= \tan^{-1} \left\{ \frac{e^{-4Cl_2} \sin\left(-\frac{4\omega_0}{v} \xi_2 l_2\right)}{AB + e^{-4Cl_2} \cos\left(-\frac{4\omega_0}{v} \xi_2 l_2\right)} \right\} \quad (57)$$

Assuming the noise power is much smaller than the signal and letting

$$2 \tan^{-1} \left(\frac{\omega_0 l_1}{v} - \frac{v}{\omega_0 l_1} \right) \approx \pi - \frac{2v\omega_0 l_1}{\omega_0^2 l_1^2 - v^2}; \quad l_1 > \lambda_0/2$$

$$= \pi - 2 \sum_{n=0}^{\infty} \left(\frac{v}{\omega_0 l_1} \right)^{2n+1}; \quad l_1 > \lambda_0/2$$

$$= \pi - \left(\frac{2v}{\omega_0 l_1} \right); \quad l_1 > \lambda_0/2 \quad (58)$$

and

$$2 \tan^{-1} \left(\frac{2\omega_0 l_1}{v} - \frac{v}{2\omega_0 l_1} \right) \approx \pi - \left(\frac{v}{\omega_0 l_1} \right); \quad l_1 > \lambda_0/4 \quad (59)$$

Eqn. (55) becomes:

$$E_r = \text{Re} \left\{ e^{j \left[y_1(t) + \Lambda_1 + \theta - 2 \frac{\omega_o \ell_1}{v} + \pi - \frac{2v}{\omega_o \ell_1} \right]} e^{j \omega_o t} + e^{j \left[y_2(t) + \Lambda_2 + 2\theta - \frac{4\omega_o \ell_1}{v} + \pi - \frac{v}{\omega_o \ell_1} \right]} e^{j 2\omega_o t} \right\}; \ell_1 > \lambda_o/2 \quad (60)$$

Upon reception, the phase of the fundamental signal is measured and averaged by the CW radar and doubled to give:

$$2\text{Ph}_1 = 2 \left[\Lambda_1 + \theta - \frac{2\omega_o \ell_1}{v} + \pi - \frac{2v}{\omega_o \ell_1} \right] + 2\omega_o t; \ell_1 > \lambda_o/2 \quad (61)$$

This can then be subtracted from the measured and averaged phase of the second harmonic signal yielding the dispersive-phase:

$$\text{Ph}_2 - 2\text{Ph}_1 = \Lambda_2 - 2\Lambda_1 - \pi + \frac{3v}{\omega_o \ell_1}; \ell_1 > \lambda_o/2 \quad (62)$$

This equation relates the measured phases of the two reflected signals to the physical characteristics of the target contained in functions Λ_1 and Λ_2 . For measurement applications where the distance to the target is greater than λ_o , the phase term arising from the near field consideration, $3v/\omega_o \ell_1$ rapidly approaches zero. The calculated dispersive-phase is then left completely independent of the distance to the target. This allows an arrangement of two spatial interferometric radar systems to perform useful diagnostic field measurements under conditions of uncertain path length.

For specified values of A, B and C, $Ph_2 - 2Ph_1$ is single valued over an interval of approximately $\epsilon_2 \ell_2 = \pi/4$. This ambiguity interval can be extended to $\epsilon_2 \ell_2 = \pi/2$ with the additional information obtained from the dispersive-loss of the layers (i.e. the ratio of the absolute losses at each frequency). Employing the radar range equation, this becomes:

$$\frac{P_{R1}/P_{T1}}{P_{R2}/P_{T2}} = \left(\frac{G_1}{G_2} \right) \left(\frac{\frac{c}{2\ell_1 \omega_0}}{\frac{c}{4\ell_1 \omega_0}} \right) \left| \frac{\Lambda_1}{\Lambda_2} \right|^2 \quad (63)$$

$$= 4 \left(\frac{G_1}{G_2} \right) \left| \frac{\Lambda_1}{\Lambda_2} \right|^2 \quad (64)$$

$$= 4 \left(\frac{G_1}{G_2} \right) \frac{\operatorname{Re} \left\{ \frac{B/A+e^{-2 \left[C+j \frac{\omega_0}{v} \epsilon_2 \right] \ell_2}}{AB+e^{-2 \left[C+j \frac{\omega_0}{v} \epsilon_2 \right] \ell_2}} \right\}}{\operatorname{Re} \left\{ \frac{B/A+e^{-4 \left[C+j \frac{\omega_0}{v} \epsilon_2 \right] \ell_2}}{AB+e^{-4 \left[C+j \frac{\omega_0}{v} \epsilon_2 \right] \ell_2}} \right\}}^2 \quad (65)$$

where

P_{T1} is the transmitted power of the fundamental harmonic

P_{T2} is the transmitted power of the second harmonic

P_{R1} is the received power of the fundamental harmonic

P_{R2} is the received power of the second harmonic.

G_1 is the antenna gain of the fundamental harmonic receiver

G_2 is the antenna gain of the second harmonic receiver

Plots of $Ph_2 - 2Ph_1$ versus ℓ_2 and of $P_{R1}/P_{T1}/P_{R2}/P_{T2}$ versus ℓ_2 for various values of dielectric constant ϵ_{r2} and for conductivity $\sigma_2=0$ and frequency $\omega_0/2\pi=58\text{MHz}$ are given in Fig. 7. As can be seen, two possible values for ℓ_2 less than 150cm occur for any measured value of $Ph_2 - 2Ph_1$ and can be resolved by a comparison of the corresponding value of $P_{R1}/P_{T1}/P_{R2}/P_{T2}$ with the theoretical threshold of 4. For example, from Fig. 7a, a calculated value of -140° for the dispersive-phase of a target will give two possible thickness values of 32cm and 66cm. The dispersive-losses associated with each of these thickness values, however, lie on opposite sides of the threshold shown in Fig. 7b. Consequently, the additional calculated value for the dispersive-loss of the target will resolve which of the two thickness values is correct.

The data presented in Fig. 7 represents the situation encountered in the measurement of ice thickness and is used by the VHF radar system described in Section 3.Bi.

iii) Advantages and Limitations of CW Radar

As has been shown in the previous sections, a CW radar system used to measure the thicknesses of layered targets provides an improvement in range precision when compared to conventional MC radars due to the narrow system bandwidths and hence low noise powers achievable. In addition, the CW system provides a simplified and inexpensive design resulting from greatly relaxed component-bandwidth requirements.

Fig. 7. Dispersive-phase and dispersive-loss for a single-layer target ($\sigma_2=0$, $\omega_0/2\pi=58\text{MHz}$).
 (a) Dispersive-phase
 (b) Dispersive-loss

Synchronous detection capabilities can also be easily implemented in CW radars whereas sampling operations are necessary for MC systems.

The requirement that a target consists of a known number of layers, each of a known maximum thickness, dielectric constant and loss acts as a fairly severe restriction on possible applications of the CW spatial interferometric radars. However, in diagnostic applications where these requirements can be met, such as the measurement of ice layers or industrial film and sheet thicknesses, the CW radar becomes attractive compared with MC systems.

The synchronous detection process used in CW radars is based on a comparison of the phases of the transmitted and received signals, and consequently, is an application of conventional spatial interferometry. This process is commonly used to determine the distance to a single, stationary target in surveying applications⁽²⁸⁾. However, three serious limitations of this comparison technique exist in situations where the distance to the target can be large and varying as in remote sensing applications.

The first limitation arises in situations where the distance to the target is larger than one-half the product of the speed of light and the signal autocorrelation interval, $\ell > vR_s(\tau)/2$. For such large distances, the transmitted and received signals will lose their coherency and synchronous detection becomes impossible. This situation, although infrequently encountered, may arise in environmental monitoring from the

upper atmosphere or space.

A second and more important limitation also arises from the large difference in path lengths between the transmitted reference signal and the received signal. The phase detection networks form high Q resonant circuits and consequently any local oscillator frequency drift will completely obscure initial phase alignments. The high frequency stability required to avoid such misalignments is very difficult to achieve in environmental and industrial testing due to the extreme temperature differentials and long periods of operation which can be encountered.

The third and most severe limitation arises from any motion between the instrument and the target during the finite time averaging interval used in the synchronous detection process. Relative motion which is impossible to avoid during continuous profiling experiments will completely disrupt the detected phases.

These three limitations of the multiple-frequency CW radars can all be avoided by use of the concepts of harmonic coherency and harmonic interferometry discussed in Chapter 2. CW radar systems which use these principles are called *multiple-frequency homodyne radars*.

iv) Multiple-Frequency Homodyne Radar

The severe limitations of CW radars discussed in the previous section can be avoided by employing a synchronous detection process based on harmonic interferometry. By employing the harmonic interferometer shown in Fig. 3b, the dispersive-phase can be measured

directly rather than calculated from the absolute-phases of the separate signals (frequencies) as must be done using CW spatial interferometric radar techniques. As a result, the three limitations will not arise.

The first of these limitations, the maximum distance to the target, is no longer limited by the signal autocorrelation interval since a comparison of just the received signals and not the transmitted signals is made using the harmonic interferometric technique.

In addition, the phase comparison circuits will no longer form high Q resonators since each received signal traverses the same path length.

Also, the measured dispersive-phase can be averaged over the finite time integration interval regardless of relative motion.

A schematic diagram representing a bi-frequency homodyne radar capable of resolving two dielectric interfaces is shown in Fig. 8. This system is a reflection mode configuration of the harmonic interferometer of Fig. 3b. In the transmitter, the second harmonic of the fundamental is generated in a non-linear X_2 frequency multiplier. The two frequencies are then directed towards the 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.

Fig. 8. Bi-frequency homodyne radar.

The construction and testing of two experimental bi-frequency coherent radars, one a VHF radar for the measurement of ice layer thicknesses, the other a C-band radar for the measurement of industrial film and sheet thicknesses, is described in the remainder of this thesis.

B. Homodyne Radar Design and Experimental Results

i) VHF Bi-Frequency Homodyne Radar

A view and system block diagram of the 58MHz and 116MHz VHF radar constructed for measuring the dispersive-phase and dispersive-loss introduced by reflections from an ice sheet are shown in Figs. 9a and 9b. In addition to the functional elements previously presented in Fig. 8, this system includes amplifiers, attenuators and filters, and two resonant phase cancellation networks required to isolate the transmitter and receiver and to reduce the space requirements of the radiating elements.

In the block diagram of Fig. 9b, the 58MHz oscillator is fed to the Harmonic Generator circuit shown in detail in Fig. 9c. Two harmonically related outputs at 58MHz and 116MHz respectively drive the 58MHz and 116MHz Transmitter Power Detector circuits shown in Fig. 9d. The detected voltages, V_1 and V_2 , provide monitoring of the radiated power levels at each frequency. The output from each Transmitter Power Detector is then fed to a Quadrature Hybrid, shown in Fig. 9b, where

Fig. 9. VHF bi-frequency homodyne radar.
(a) Radar instrument tray.

CAUTION: HIGH VOLTAGE

CAUTION: HIGH VOLTAGE

(b)

Fig. 9. VHF bi-frequency homodyne radar.
b) Schematic diagram

(c)

(d)

Fig. 9. VHF bi-frequency homodyne radar.

c) Harmonic generator

d) Transmitter power detector 58 MHz (116 MHz)

(e)

Fig. 9. VHF bi-frequency homodyne radar.
e) Receiver power detector

Fig. 9. VHF bi-frequency homodyne radar.
f) Detected voltage function circuit

Fig. 9. VHF bi-frequency homodyne radar.
 g) Receiver frequency doubler
 h) Phase detector

it is split between a control path and a radiating element.

Return signals from the target are received with the same antennas used by the transmitter and are fed through the quadrature hybrids and narrow band, noise suppressing bandpass filters to the Receiver Power Detector circuits shown in Fig. 9e. Antenna reflections which are also fed to the Receiver Power Detectors are suppressed to typically -50dB by adjustments of the variable phase shifters and variable attenuators (situated in the Receiver Power Detector circuits) in the control paths. Resonant phase cancellation is then achieved by combining the control signals with the antenna reflections in the 3dB power combiners of the Receiver Power Detectors.

The detected voltages in the Receiver Power Detector circuits, V_3 and V_4 , provide monitoring of the received power levels from the target at each frequency. They are fed together with the transmitter detected voltages, V_1 and V_2 , to the Detected Voltage Function Circuit of Fig. 9f which determines the dispersive-loss of the target.

The output from the 58MHz Receiver Power Detector is fed to the Receiver Frequency Doubler shown in Fig. 9g. The 116MHz output of this circuit is then fed along with the output from the 116MHz Receiver Power Detector to the 116MHz Phase Detector shown in Fig. 9h. The detected voltages, V_5 and V_6 of this circuit, determine the relative phase between the two 116MHz input signals and consequently the dispersive-phase of the target.

The power supply circuit for the VHF homodyne radar and the variable current sources used to drive the variable attenuators in the resonant phase cancellation networks are shown in Fig. 9i.

The calibration data for the VHF homodyne radar constructed for ground based ice thickness field trials is given in Figs. 10 to 13. The measured transmitter power levels for the 58MHz channel and the 116MHz channel as a function of the local oscillator drive level and as functions of the transmitter detector voltages, V_1 and V_2 , are given in Fig. 10a,b,c. The measured receiver power levels for each channel, as functions of the receiver detector voltages, V_3 and V_4 , are given in Figs. 11 and 12, respectively with inserts of the power levels driving the inputs to the phase detector network. The 116MHz phase detector characteristic as a function of these input driving power levels is shown in Fig. 13. Acceptable operation is found for driving power levels in the range of .2mW to .5mW with a maximum phase error of ± 7 electrical degrees.

Optimum driving powers at the phase detector are obtained by adding known values of attenuation in each channel between the transmitter and receiver until detector voltages representing a desirable phase characteristic are obtained at the receivers (inserts of Figs. 11 and 12). The attenuation settings are then added to the receiver power levels obtained in Figs. 11 and 12 to obtain the actual received power. For example, a driving power level of .25mW in the phase detector will provide an optimum characteristic in Fig. 13. Consequently, attenuation should be added between the transmitters and receivers until detected

CALIBRATION DATA

CALIBRATION DATA

Fig. 10. Transmitter power characteristics.
a) Power level characteristic

Fig. 10. Transmitter power characteristics.

b) 58MHz power detector characteristic

Fig. 10. Transmitter power characteristics.
c) 116MHz power detector characteristic

Fig. 13. 116MHz phase detector characteristic.

voltages of -104mV and -172mV are obtained for V_3 and V_4 respectively in Figs. 11 and 12.

The theoretical data presented earlier in Fig. 7 represent the situation encountered by the 58MHz and 116MHz bi-frequency homodyne radar used to measure the thickness of ice. For fresh water ice, the dielectric constant is equal to $3.18 \pm 2\%$ ⁽¹⁾ and as a result the data for $\epsilon_r = 3.2$ can be used in Fig. 7 with negligible error. Also, perturbations introduced in the data due to non-zero conductivity of fresh water ice are small; less than 1% in Fig. 7a and less than 2% in Fig. 7b and consequently, the effect of any uncertainty can be ignored. The phase detector characteristics of Fig. 13 are used to precisely measure the dispersive-phase $Ph_2 - 2Ph_1$ of the ice layer from which the two possible thickness values can be determined from Fig. 7a. The 58MHz and 116MHz transmitter and receiver power detector curves of Figs. 10, 11 and 12 are then used to approximately determine the dispersive-loss, $P_{R1}/P_{T1}/P_{R2}/P_{T2}$ of the ice layer, in order to resolve which of these thickness values represents the actual ice depth.

Field tests of the VHF radar have been made on ice formed on the Ottawa and Rideau Rivers during the 1974-75 and 1975-76 winter seasons. By use of the resonant phase cancellation networks, which allow a monostatic, single antenna arrangement for each frequency, a reduction in the large space requirements for the radiating elements was achieved. A further reduction in size was obtained by mounting both antennas, each a high gain, five element YAGI, orthogonally along one mast.

The phase and amplitude adjustments in the control paths necessary for this monostatic antenna arrangement require a calibration procedure in the absence of a return from the target. For the ground based measurements, this was achieved by rotating the antennas to direct the main lobes above the horizon and the deep nulls to the target area. For airborne applications, this calibration could be performed at high altitudes where the returns from the target are negligible.

Due to the broad antenna beamwidths of 45 to 55 degrees, non-uniform plane waves exist at the ice surface and introduce perturbations in the dispersive-phase and dispersive-loss data for the ice layer. For ground based measurements, these perturbations will result in a maximum thickness offset of 5% which could be accounted for using an averaging technique. For airborne applications, this effect of the beamwidth becomes negligible since the reflections from the ice surfaces are non-specular in the VHF spectrum. However, an averaging of ice thickness over a large target area is unavoidable. This averaging is advantageous for volumetric estimation studies of an ice sheet but makes identification of localized faults or thin sections difficult.

A second calibration procedure is initially required to account for the total insertion-phase and insertion-loss of the radar system due to the time delays and losses introduced by the various components. This calibration is accomplished by comparing the measured values for

the dispersive-phase and dispersive-loss of the radar obtained with a flat metal sheet as the target, and the theoretical values for a layer of zero thickness. An offset in phase is then obtained which is subtracted from all future measurements before the theoretical data of Fig. 7a is used. Also, a shift in the threshold level for the dispersive-loss from the theoretical value of 4 is obtained in Fig. 7b.

During the ground-based field tests on the Ottawa and Rideau Rivers, each radar measurement was repeated between two and four times in order to ensure proper operation of the RF systems (the transmitter and receiver systems), the control systems (the resonator circuits) and the detection systems. The data obtained during these trials are presented in Table I. In the first column, the mean value for the dispersive-phase (accounting for the calibration offset) as well as the spread for each measurement is shown. The two possible thickness values for the ice, as determined by Fig. 7a with $\epsilon_r = 3.2$, are given in the second column. The approximate value of the dispersive-loss is given in the third column and is used to resolve between the two possible thicknesses by comparing its value with the experimentally obtained threshold of 3.8. The selected value for the ice thickness is given in column four. Ground truth thickness obtained by drilling up to eight cores over a twelve foot by twelve foot target area beneath the antenna is presented in column five. A comparison between the radar measurements and the ground truth data is presented in Fig. 14.

Table I. Fresh water ice thickness data obtained with a bi-frequency homodyne radar ($\omega_0/2\pi=58\text{MHz}$).

$\phi_2 - 2\phi_1$	POSSIBLE CALCULATED VALUES FOR ℓ_2	$\frac{P_{R1}/P_{T1}}{P_{R2}/P_{T2}}$ (EXPERIMENTAL THRESHOLD =3.8)	DEDUCED VALUE FOR ℓ_2	GROUND TRUTH VALUE FOR ℓ_2	DATE
-124°	55cm 74cm	data invalid due to satur- ation of the 176MHz channel assumed value less than thresh- old	74cm	69±5cm	19 Mar 75 (measured using an 88MHz system)
-181±1°	40±.3cm 57±.3cm	2.6	57±.3cm	53±2cm	19 Feb 76
-178±1°	40±.3cm 58±.3cm	3.0	58±.3cm	53±2cm	19 Feb 76
-175±3°	39±.7cm 58±.7cm	2.4	58±.7cm	53±2cm	19 Feb 76
-172±1°	38±.3cm 59±.3cm	2.2	59±.3cm	57±4cm	23 Feb 76
-114±3°	28±1.cm 69±.5cm	2.0	69±.5cm	62±5cm	26 Feb 76
-112±3°	27±1.cm 69±.5cm	2.0	69±.5cm	62±5cm	26 Feb 76
-119±3°	29±1.cm 68±.5cm	2.0	68±.5cm	64±5cm	26 Feb 76
-123±3°	30±1.cm 67±.5cm	2.2	67±.5cm	64±5cm	26 Feb 76

Fig. 14. Fresh water ice thickness data obtained with a bi-frequency homodyne radar ($\omega_0/2\pi=58\text{MHz}$).

Excellent agreement is found over the range of ice thicknesses that were located and measured near Ottawa.

In order to test the operation of the radar for both larger and smaller values of thickness, coaxial models of the ice sheets were constructed. A resistive termination of 10 ohms in a 50 ohm coaxial system was used to simulate the reflections from the ice-water interface at the bottom of an ice sheet. A lossless reactive discontinuity which consisted of a shunt, short circuited 50 ohm transmission line, 57cm long was used to provide reflections of the same amplitude as the air-ice interface at the top of an ice sheet, at both 58MHz and 116MHz. The difference in phase of the reflections of the reactive discontinuity and the air-ice interface was accounted for in the mathematical description of the reflections. Various thicknesses of ice sheets were then simulated by changing the length of 50 ohm coaxial cable between the discontinuity and the termination. A comparison between the radar measurements on these coaxial models and the length of cable between the discontinuity and the termination is presented in Fig. 14 again with excellent agreement.

ii. C-Band Bi-Frequency Homodyne Radar

A view and block diagram of a 4GHz and 8GHz bi-frequency homodyne radar capable of measuring the dispersive-phase and dispersive-loss introduced by reflections from industrial films and sheets is shown in Fig. 15. In a similar manner to the VHF system, amplifiers, attenuators and filters are necessary to maintain proper signal control.

Due to the higher frequencies and shorter wavelengths used, a bistatic antenna arrangement employing two waveguide horns for each frequency is used. Consequently, the resonant phase cancellation circuits are not required.

The calibration data for the laboratory based, C-band homodyne radar is given in Figs. 16 to 19. The measured transmitter power levels for the 4GHz channel and the 8GHz channel as a function of the local oscillator drive level is given in Fig. 16. The measured receiver power levels for each frequency channel as functions of the receiver detector voltages, V_7 and V_8 , are shown in Figs. 17 and 18 respectively, with inserts of the power levels driving the inputs to the phase detector network. The phase detector characteristic is shown in Fig. 19.

Fig. 15. C-band bi-frequency homodyne radar.
a) Radar prototype

(b)

Fig. 15. C-band bi-frequency homodyne radar.
b) Schematic diagram

Fig. 16. Transmitter power level characteristic.

24.0 RECEIVED POWER (μ W-dBm)

66.

6.0 7.8 $\uparrow$ DRIVING POWER AT PHASE DETECTOR
(mW-dBm)

5.0 7.0

25.2

4.0 6.0

3.0 4.8

27.0 2.0 3.0

1.0 0.0

0.0 0.0

30.0

Fig. 17. 4GHz receiver power detector characteristic.

V_7 , DETECTED VOLTAGE (mV)

10

20

30

40

50

60

70

27.0 RECEIVED POWER
($\mu\text{W-dBm}$)

67.

60 17.8 DRIVING POWER AT PHASE DETECTOR (mW-dBm)

28.0 50 17.0

40 16.0

29.2 30 14.8

20 13.0

31.0 10 10.0

0

34.0

Fig. 18. 8GHz receiver power detector characteristic.

V_8 DETECTED VOLTAGE (mV)

10

20

30

40

50

60

70

0

Fig. 19. 8GHz power detector characteristic.

The optimum phase detector driving power of +4dBm is obtained by inserting known values of attenuation after the local oscillator (to control the 4GHz receiver driving power) and after the .8GHz receiver (to adjust its driving power to match the 4GHz level). Again, the detected voltages V_9 and V_{10} of the phase detector characteristics and V_7 and V_8 of the power detector characteristics are used to determine the dispersive-phase and dispersive-loss of the target layer.

Experimental testing of the C-band radar has been undertaken using a combination of plexiglass ($\epsilon_r=2.6$ at 4GHz) and a metal sheet as the target. The data obtained from these tests is given in Table II in the same format as Table I and a comparison between the radar and the physical thickness measurements of the plexiglass is presented in Fig. 20. Excellent agreement is also found in all cases with this system, with a maximum error of 1.5mm.

Table II. Plexiglass thickness data obtained with a bi-frequency homodyne radar ($\omega_0/2\pi=4\text{GHz}$).

Ph_2-2Ph_1	POSSIBLE CALCULATED VALUES FOR ℓ_2	$\frac{P_{R1}/P_{T1}}{P_{R2}/P_{T2}}$ EXPERIMENTAL THRESHOLD = 4.3)	DEDUCED VALUE FOR ℓ_2	MEASURED VALUE FOR ℓ_2	ERROR IN WAVE- LENGTHS
-91.4°	1.8mm 11.6mm	4.3	1.8mm	1.6mm	.0003 λ
-126.0°	5.0mm 10.5mm	5.8	5.0mm	5.0mm	0
-174.6°	7.5mm 7.5mm	-	7.5mm	6.0mm	.02 λ
-165.4°	6.7mm 9.0mm	4.3	9.0mm	10.0mm	.013 λ
- 83.4°	12.0mm 20.25mm	3.7	12.0mm	13.0mm	.013 λ
- 76.2°	12.1mm 19.5mm	9.8	19.5mm	20.0mm	.007 λ

PLEXIGLASS THICKNESS (mm)

PLEXIGLASS THICKNESS OBTAINED BY RADAR (mm)

Fig. 20. Plexiglass thickness data obtained with a bi-frequency homodyne radar ($\omega_0/2\pi=4\text{GHz}$).

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CHAPTER 4
CONCLUSIONS

A. Summary

The design and construction of a novel radar system capable of ultra-high resolution thickness measurements has been described in this thesis. This system is an application of a new field of harmonic interferometry, based on the novel concept of harmonic coherency, which is derived through a space-frequency transformation applied to spatial coherency.

To present this conceptual development, spatial coherency and two of its primary properties were first discussed. The space-frequency transformation was then introduced and harmonic coherency defined. A relationship governing the maximum dispersive-phase error between two frequency diverse signals in the presence of noise was derived and the optimum method for generating these signals was established. The concept of harmonic coherency then led in a natural manner to the field of harmonic interferometry which makes remote interferometric measurements possible for the first time.

A discussion of the resolution capabilities of conventional modulated carrier radars and continuous wave radars based on spatial interferometric techniques was then presented. It was found that a continuous wave radar is capable of at least an order of magnitude increase in resolution compared to a modulated carrier radar system. Severe limita-

tions imposed on the continuous wave radars led to an implementation of a system based on harmonic interferometric techniques. Two prototypes of this type of radar, called the multiple-frequency homodyne radar, were constructed, calibrated and tested.

Experimental data obtained by the VHF bi-frequency homodyne radar for measuring the thickness of ice on the Ottawa and Rideau Rivers has been presented with excellent agreement to ground truth measurements. Also, data obtained by the C-band radar for measuring the thickness of plexiglass sheets (with application to industrial production line monitoring) have been presented. Again, excellent agreement between the radar measurements and the physical thicknesses was found.

B. Future Work

Continued work in many areas of multiple-frequency homodyne radar design is required on both technological and analytical levels, and concerning potential applications.

Various subsystems in the two designs presented in this thesis require improvement before advancing from the prototype stage to production models. Antenna studies should be undertaken aimed at reducing the size of the radiating element of the VHF radar. Also, automatic control circuits should be implemented in the resonant phase cancellation networks used to reduce the antenna reflections of this radar. In addition, non-dispersive automatic level control circuits should be added before the VHF radar is considered for airborne applications.

Although the resonant phase cancellation networks are necessary for ground based measurements of the VHF radar, a pulsed technique may be required to isolate the transmitter and receiver for airborne measurements. The power levels of the return signals for this measurement are too low for proper operation of the resonant phase cancellation networks but are sufficiently delayed in time to allow pulsing operation. An investigation of this method for isolating the transmitter and receiver should be undertaken.

Even though the homodyne radars allow an order of magnitude reduction in frequency, the cost of amplifiers in the microwave spectrum is still high. An introduction of heterodyne frequency translation techniques in systems such as the C-band radar would avoid the use of these high frequency amplifiers, thereby significantly reducing the cost of the system.

On the conceptual level, an extension of the work presented in this thesis to include the capabilities and limitations for measuring the thicknesses of more highly structured media should be pursued. Both the mathematical modeling of the target and the construction of an experimental system promise to be interesting and rewarding challenges.

Finally, concerning potential applications of the multiple-frequency homodyne radars, many studies ranging from the monitoring of ocean oil spill thicknesses for environmental protection and clean-up to the characterizing of aircraft and missile wakes for identification purposes could be undertaken.

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