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A Technique for Calibrating the Phase Detector of Wideband Radars Using a Phase Modulation and Demodulation Scheme

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A Technique for Calibrating the Phase Detector of Wideband Radars Using a Phase Modulation and Demodulation Scheme

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Abstract

A signal processing method is presented for correcting imbalances in the phase-detection channels of a coherent, wideband radar. Several papers have addressed this problem by the use of the fast Fourier transform (FFT) as a narrowband filter (see F. E. Churchill, G. W. Ogar, and B. J. Thompson, *The Correction of I and Q Errors in a Coherent Processor,* IEEE Trans. Aerosp. Electron. Syst., **AES-17** (January 1981), pp 131–137, and H. Bruce Wallace and Thomas J. Pizzillo, *A Technique for Calibrating the Phase Detector of a Wideband Radar Using an External Target,* Army Research Laboratory, ARL-TR-1521 (March 1998)). The present technique relies upon phase modulation of the transmitted waveform, then demodulation of the phase of the received waveform, and finally the integration and normalization of the waveform. There is one constraint; the number of phase-modulation/demodulation steps is restricted to 4 k, where k is an integer greater than 0. The technique is not dependent upon the target or the phase and gain flatness of the radar waveform. Errors remaining after application of this technique depend on the signal-to-noise ratio and errors in the phase modulator.

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1. Introduction

Inverse synthetic aperture radars (ISARs) transmit a wideband waveform to derive range information. Most systems use a linear- or stepped-frequency modulated waveform, generated by either analog or digital means, that may be processed with a fast Fourier transform (FFT) to create a high-resolution range profile.[1] To be effective, the returned signal that the radar measures must be related to the transmitted signal or to an internal reference signal in a known fashion. While this comparison may be made in a wideband phase-comparison receiver, this report concentrates on the use of a narrowband phase-detector system with stepped frequency. In this class of system, the received signal is down-converted into a narrowband signal and then separated into the received two coherent signal channels that are then mixed with two orthogonal local oscillator (LO) signals. The calibration technique presented here is an improvement of the method in Wallace and Pizzillo[2] and Churchill[3] in that the calibration does not require the FFT and reduces processing time. In addition, it improves on the method in Wallace and Pizzillo[2] in that the dc components are removed as part of the process and it does not generate correction factors; thus it eliminates errors associated with estimates in the corrected data.

This report introduces our basic assumptions and develops a signal model based on them. This technique will then be applied to simulated data and performance efficiency will be considered.

2. Development of the Signal Model

Figure 1 is a block diagram of the pertinent portions of the transmit and receive sections of the radar. A 4-GHz coherent oscillator (COHO) is split before being phase-modulated in the transmitter and used as the LO for the phase detector in the receiver. The resultant in-phase (I) and quadrature-phase (Q) signals define the real and imaginary parts of the received signal. Before the phase detector, this signal is of the form

$$S\left(f,m\right)=Ae^{j\left(\theta(f)+\frac{2\pi m}{M}\right)}\ , \qquad (1)$$

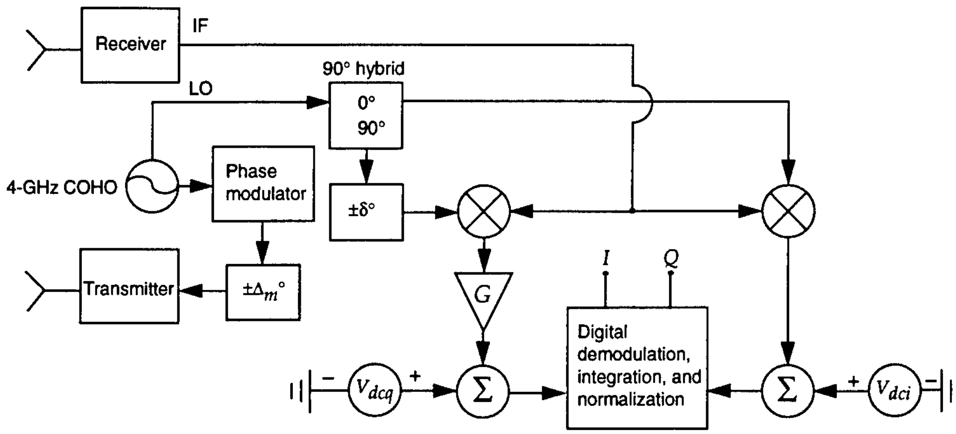
where $f=[f_l \cdots f_n \cdots f_N]$ represents the N frequency steps of a pulse compression system; $\theta(f)$ represents the relative phase that is linearly

[1]*D. L. Mensa, High Resolution Radar Cross-Section Imaging, Artech House, Norwood, MA (1991), chapter 4.*

[2]*H. Bruce Wallace and Thomas J. Pizzillo, A Technique for Calibrating the Phase Detector of a Wideband Radar Using an External Target, Army Research Laboratory, ARL-TR-1521 (March 1998).*

[3]*F. E. Churchill, G. W. Ogar, and B. J. Thompson, The Correction of I and Q Errors in a Coherent Processor, IEEE Trans. Aerosp. Electron. Syst.,* ***AES-17*** *(January 1981), pp 131–137.*

Figure 1. Generalized narrowband phase-detector system.



dependent on frequency; $m = [1 ... m ... M]$ is the number of phase-modulation steps for each frequency step; and A is the amplitude of the received signal that has been scattered by the target. With the exception of noise corruption, this is the ideal form of the signal to be processed by the phase detector. Additionally, if the phase detector were perfect, the measured outputs from each channel for a point target would be represented by two $M \times N$ arrays that are then digitally demodulated, integrated, and normalized to produce two $1 \times N$ row vectors:

$$\vec{I}(f) = [A \cos(\theta(f_1)) \ldots A \cos(\theta(f_n)) \ldots A \cos(\theta(f_N))] \text{ , and}$$

$$\vec{Q}(f) = [A \sin(\phi(f_1)) \ldots A \sin(\phi(f_n)) \ldots A \sin(\phi(f_N))] \text{ .} \qquad (2)$$

In reality, the radar modifies the signal when it is transmitted and received due to imperfections in the system components. Figure 1 shows circuit elements that represent these imperfections: the phase modulator has a fixed, differential phase error, $\pm\Delta_m{}^\circ$, associated with each step, m. The 90° hybrid actually shifts the LO 90° $\pm\delta^\circ$, where δ° is a fixed differential phase error. The mixers have dc offsets represented as a voltage source referenced to ground, and the gain throughout the phase-detector system is different for the I and Q channels represented by G. Because there is no loss in generality, all the error signals due to these imperfections, except dc offset and the phase-modulator error, are represented as occurring in the Q channel.

The measured signal is that which is actually produced by the radar phase detector before the digital processing. It includes the effects of each of the imperfections diagrammed in figure 1 as well as corruptions due to imperfections in the transmitted waveform, the wideband receiver, and any effects due to targets that are not purely pointlike. Because these are introduced before the phase detector, each channel is affected equally in both amplitude and phase. The effect on the n^{th} component of equation (1) due to a measurement made from the combined, imperfect system is

$$\vec{I}_m(f_n) = A\cos\left(\theta(f_n) + \frac{2\pi m}{M} + \Delta_m\right) + V_{dci}\ , \quad \text{and}$$

$$\vec{Q}_m(f_n) = GA\sin\left(\theta(f_n) + \delta + \frac{2\pi m}{M} + \Delta_m\right) + V_{dcq}\ , \tag{3}$$

where $\vec{I}_m(f_n)$ and $\vec{Q}_m(f_n)$ are the measured I and Q signals of the n^{th} frequency step and the m^{th} modulation step, G represents the gain imbalance in the phase-detector channels (assumed to be positive and real), δ represents the phase imbalance introduced by the imperfect 90° hybrid, Δ_m is the error in the m^{th} modulation step, and V_{dci} and V_{dcq} are the dc offsets. If we assume that the target of opportunity from which we would like to measure our calibration is a point target, we need only one complete $M \times N$ measurement to correct for all errors. This assumption is reasonable, provided the target response remains within one range cell for the duration of the measurement. The signal that is to be demodulated, integrated, and normalized is formed with equation (3) as two components of a complex pair:

$$\vec{S}(f_n) = \sum_{m=0}^{M-1} e^{-j\left(\frac{2\pi m}{M}\right)}\left[\vec{I}_m(f_n) + j\vec{Q}_m(f_n)\right]\ . \tag{4}$$

By substituting the Euler form for the trigonometric functions in equation (4), combining V_{dci} and V_{dcq} into a single term V, and dropping the functional dependencies to simplify notation, we have

$$\vec{S} = \frac{A}{2}\sum_{m=0}^{M-1} e^{-j\frac{2\pi m}{M}}\left[e^{j\left(\theta + \frac{2\pi m}{M} + \Delta_m\right)} + e^{-j\left(\theta + \frac{2\pi m}{M} + \Delta_m\right)} + G\left(e^{j\left(\theta + \delta + \frac{2\pi m}{M} + \Delta_m\right)} - e^{-j\left(\theta + \delta + \frac{2\pi m}{M} + \Delta_m\right)}\right) + V\right]. \tag{5}$$

Multiplying through by the demodulation factor, factoring $e^{j\theta}$ from each term, and rearranging we get

$$\vec{S} = \frac{A}{2}e^{j\theta}\sum_{m=0}^{M-1}\left[\left(1 + Ge^{j\delta}\right)e^{j\Delta_m} + e^{-j\left(2\theta + \frac{4\pi m}{M} + \Delta_m\right)} - Ge^{-j\left(2\theta + \delta + \frac{4\pi m}{M} + \Delta_m\right)} + Ve^{-j\left(\theta + \frac{2\pi m}{M}\right)}\right]\ . \tag{6}$$

Next we consider our sum, term by term:

$$\vec{S} = \frac{A}{2}e^{j\theta}\left[\left(1 + Ge^{j\delta}\right)\Delta + \left(1 - Ge^{-j\delta}\right)e^{-j2\theta}\beta + Ve^{-j\theta}\Gamma\right]\ , \tag{7}$$

where $\Delta = \sum_{m=0}^{M-1} e^{j\Delta_m}$, $\beta = \sum_{m=0}^{M-1} e^{-j\left(\frac{4\pi m}{M} + \Delta_m\right)}$, and $\Gamma = \sum_{m=0}^{M-1} e^{-j\frac{2\pi m}{M}}$ are complex constants. If we now constrain $M = 4k$, $k = 1, 2, \ldots$, then $\Gamma = 0$ and equation (7) becomes

$$\vec{S} = \frac{A}{2}e^{j\theta}\left[\left(1 + Ge^{j\delta}\right)\Delta + \left(1 - Ge^{-j\delta}\right)\beta\right]\ . \tag{8}$$

If the phase modulator were perfect and the Δ_m's were 0, then the complex constant Δ would evaluate to the real value M and the complex constant β would evaluate to 0. This would reduce equation (8) to

$$\vec{S} = \frac{M}{2}\left(1 + Ge^{j\delta}\right)Ae^{j\theta} \; . \tag{9}$$

This shows that the correct phase and amplitude of the target may be recovered having only been modified by a complex constant:

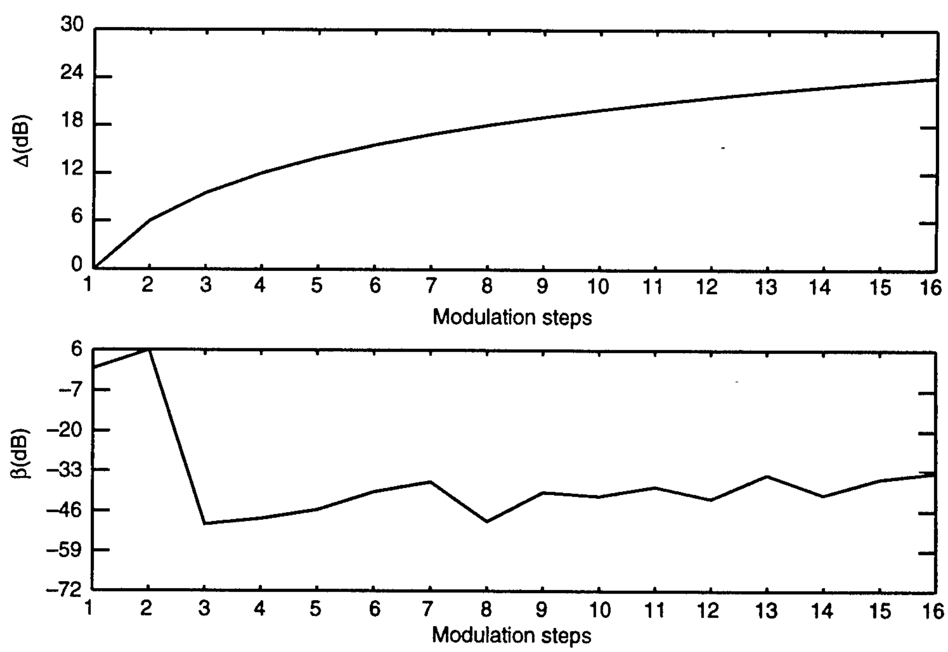
$$C = \frac{M}{2}\left(1 + Ge^{j\delta}\right) \; . \tag{10}$$

Because the calibration reflector measurement is modified by the same coefficient, C is normalized in the same manner as all other range and radar constants and equation (10) reduces to the ideal signal of equation (2):

$$\vec{S}(f) = Ae^{j\theta(f)} = \vec{I}(f) + j\vec{Q}(f) \; . \tag{11}$$

An analysis of each of the three coefficients from equation (8), assuming a uniformly distributed phase-modulator error, indicates that a more relaxed constraint than $M = 4k$ may suffice depending on the sensitivity of the system, namely $M > 2$ as indicated in figure 2. These plots were generated with a Monte Carlo simulation of 50 data sets with the Δ_m's chosen from a uniform distribution U[−3°, 3°]. If the error due to β is intolerable, one may measure the exact phase shift for each step desired and store these values in a lookup table so that the exact value may be used in the demodulation portion of this process. This ensures that β goes to 0 for $M = 4k$.

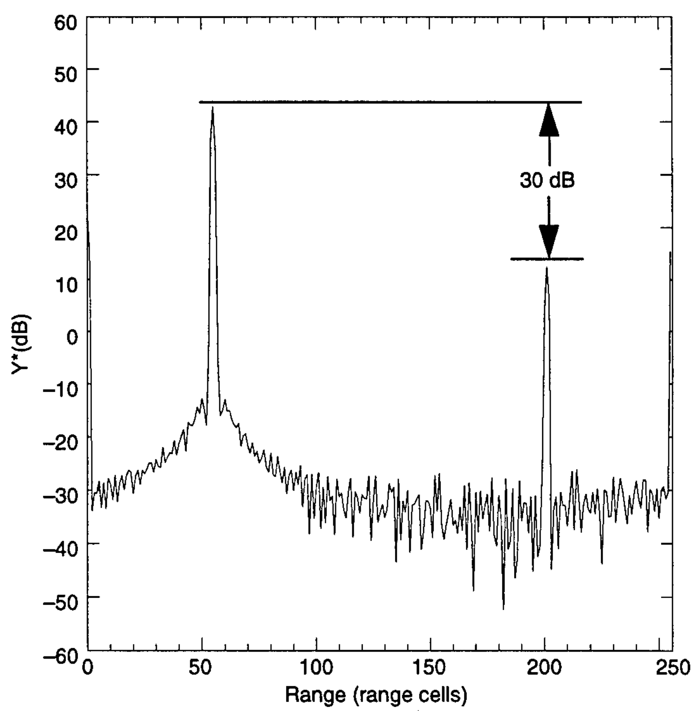
Figure 2. Coefficients of equation (8) with uniformly distributed phase modulator errors of ±3°.



3. Example of Calibration Technique With Simulated Data

First we consider a single step; that is, let $M = 1$ in equation (3). A full discussion of the spectral characteristics of an ideal complex pair as well as the individual effects of dc offset and gain and phase distortions on the spectral components may be found in Scheer and Kurtz.[4] It concludes that a gain imbalance provides amplitude errors at the target response of $(A/2)(1 + G)$ and of $(A/2)(1 - G)$ at the image response. The effect due to nonorthogonality may be expressed as an amplitude error of $(A/2)(1 + e^{j\delta})$ at the target response and $(A/2)(1 - e^{-j\delta})$ at the image response. Extending this argument, it is easy to show that the combined phase and gain distortions provide the target response with an amplitude error of $(A/2)(1 + Ge^{j\delta})$ and the image response with an amplitude error of $(A/2)(1 - Ge^{-j\delta})$. These are two of the terms of equation (7) in addition to the dc term that would be present for the case $M = 1$. Figure 3 shows the effect of a 3-percent gain imbalance, $G = 1.03$, a 3° phase imbalance, $\delta = 3°$, and a dc offset in the I and Q channels of 10 percent. Figure 4 shows the same data as figure 3 for the case $M = 4$ and no phase-modulation errors; that is, the Δ_m's = 0. Both the image response and the dc response have been eliminated and the target response has increased as a result of the $M = 4$ multiplier. Figure 5 shows the same data as

Figure 3. FFT of simulated response to point target with a 3-percent gain imbalance, a 3° phase imbalance, and a 10-percent dc offset.

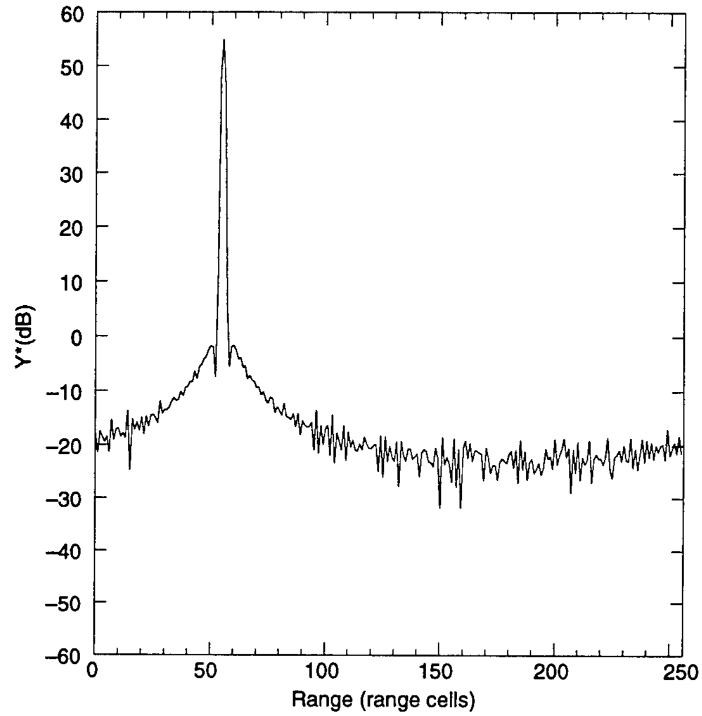


*Y = FFT of $\vec{S}(f_n)$ in equation (4).

[4] *James A. Scheer and James L. Kurtz, Coherent Radar Performance Estimation, Artech House, Norwood, MA (1993), chapter 3.*

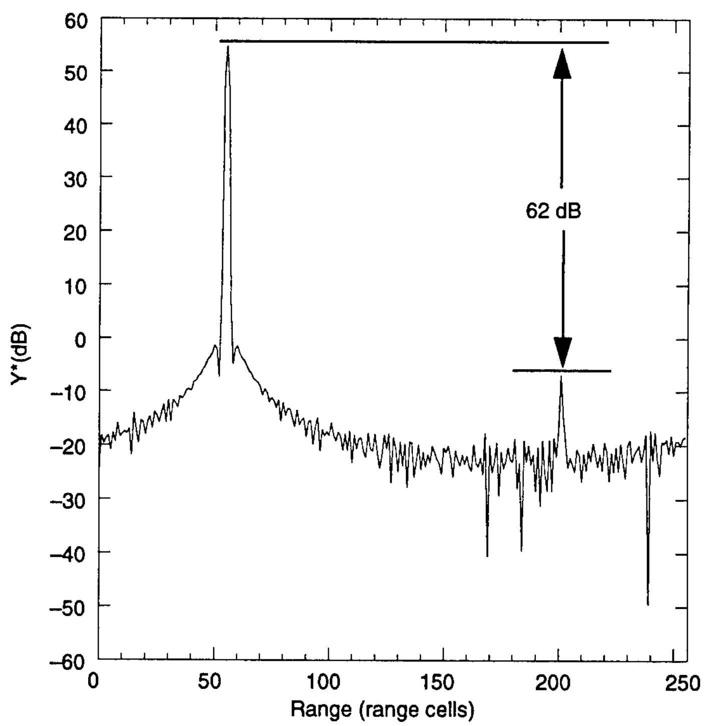
figure 4 but with a uniformly distributed phase-modulation error of ±3°. The effect of the phase-modulation error is an image response 62 dB down from the target response that results from the combined errors of the system, $\beta^*(A/2)(1 - Ge^{-j\delta})$.

Figure 4. Data of figure 3 with $M = 4$ and no phase-modulator errors.



*Y = FFT of $\vec{S}$ in equation (9).

Figure 5. Data of figure 3 with $M = 4$ and uniformly distributed phase-modulator errors of ±3°.



*Y = FFT of $\vec{S}$ in equation (8).

4. Conclusions

A method for correcting the I and Q imbalances of a wideband radar has been presented that requires no internal phase-calibration hardware. The technique relies upon phase modulation of the transmitted signal and then digital demodulation, integration, and normalization of a single data set to eliminate distortions due to gain and phase imbalances as well as dc offsets in the signal channels. Some image response remains after processing if errors in the phase modulator are not accounted for; however, these errors may readily be resolved with the exact modulator values stored in a lookup table.

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