THE DIGITAL IFM RECEIVER REVISITED

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by S. V. Potter

1 Introduction

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Since the outbreak of world War 2 two varieties of radar ESM have developed, namely, elint, wich is concerned with gathering particulars of specific radars and tactical ESM which is concerned with the tactical use of radar . The former aims to gather as much detail as possible with little regard to the time taken, while the latter aims to collect data in a short time so that the information is tactically useful. The measurement of signal source bearing is the primary ESM requirement in a busy, naval, radar environment. Having established the bearing of a contact the next requirement is to establish what it is. Discussions about the best way to do this have ranged ever since radar was invented, indeed the problem is still discussed today. Scan frequency and potentially achieve good resolution, but suffer low probability of intercept, or employ wide open receivers that potentially provide good probability of intercept but poor frequency resolution and limited sensitivity. At the end of World War 2 the British considered it difficult to know what frequency an enemy might use and so opted for the wide open receiver approach while the Americans loved their scanning superhet receivers, and still do.

2 History

The earliest tactical naval ESM receivers usually employed some form of wide band frequency receiver with an operator manually tuning a bearing receiver system to an intercept of interest. With the advent of digital processors and larger and larger digital memories a move to ESM receivers with a digital output became essential. This led to the development of a digital IFM based on a bank of harmonically related, microwave, phase discriminators. A simplified circuit diagram of a phase discriminator is shown in Figure 1. Signals are assumed to arrive one at a time. When a signal arrives the time of arrival is latched, and the frequency measuring process begins. As shown in figure 1 a signal entering the discriminator drives a 50/50 splitter which energizes two transmission lines that have a path length differnce of L. The phase difference between the two emergent signals is say, phi, where phi = 2*pi*f*L. A phase discriminator produces two outputs, one proportional to cos(phi) and the other proportional to sin(phi). One bit A/D converters convert the trigonometric outputs into digital form, see Figure 2. A bank of such discriminators in which the delay line lengths are, say, L, 2L, 4L, 8L.. can be made to give ever increasing resolution (in theory only, in practice noise is a limiting factor).

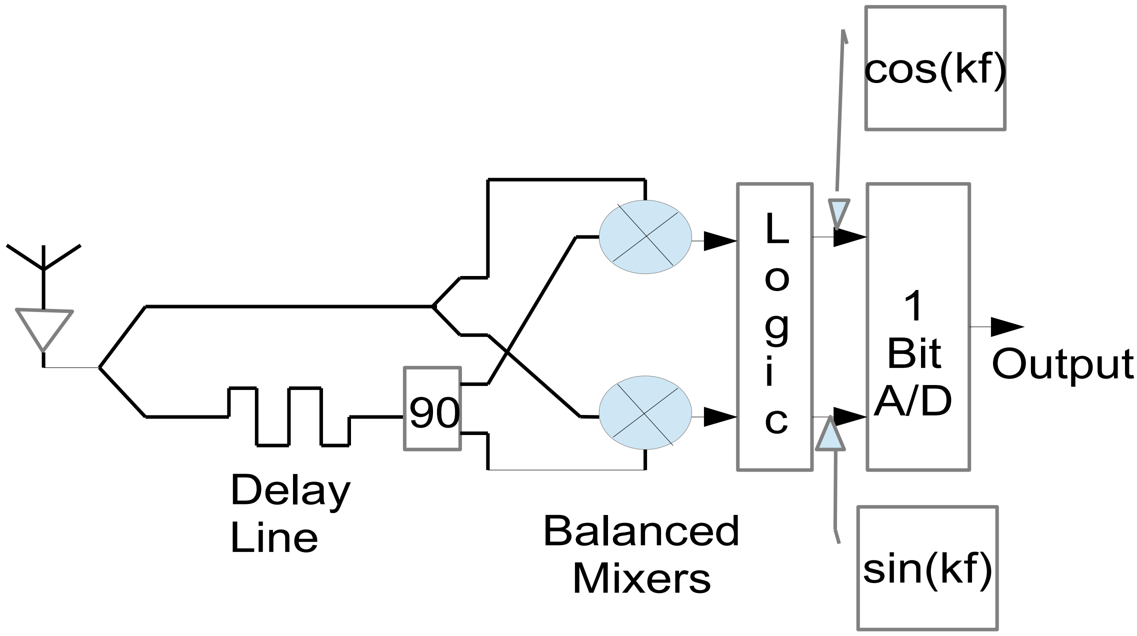

Figure 1 A Single Microwave Discriminator

Figure 2 Output From A Five Discriminator IFM

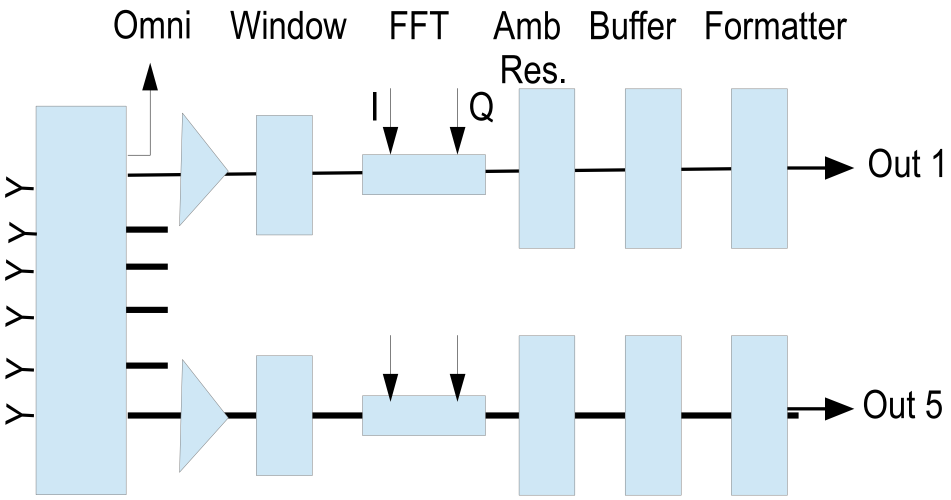

Figure 3 Combined Array and IFM

3 Candidate Techniques

Some techniques that might be used to measure frequency and bearing simultaneously are shown in Table 1.

Technique	Merits	Short Comings
FFT + Array	Heritage, well researched.	Dynamic range
Bragg Cell	Signal sorting	Bulky, dynamic range
Compressive Receiver	Sigal sorting	Input bandwidth
Slow Scan Superheterodyne	Heritage, dynamic range	Low POI

This paper only addresses the FFT + Array option in view of its flexibility and heritage. In a busy environment the traditional IFM can only report the strongest intercept accurately when it is has at least 3dB stronger than the sum of all the other signals and noise combined. Furthermore its performance is degraded considerably when intercepting the now popular, Pulse Doppler signals. A schematic diagram of a

possible FFT + Array system is shown in figure 3.
Each channel is the same, consisting of a twister, an antenna, an RF band-defining filter, an amplifier, a sub-Nyquist sampler, a Kaiser windowing function and a series of ambiguity resolving procedures to produce an amplitude and phase angle for each array channel and the omni channel.

The bins containg the local peaks in the omni-FFT output are located and stored. The data contained in the corresponding bins in the array-FFTs is extracted and stored also. RF frequency is derived from the omni data while the angle of arrival of corresponding intercepts is found from the difference in phase between the Array-FFTs and the omni-FFTs. Taking the phase difference between chanels cancels out systematic channel phase shifts.

4 Module Charateristics

a) Polarization Twister The use of a series of closely spaced, fine wire gratings to produce a fixed amount of twist is established practice.

b) Antenna Array Vivaldi 'horns', or possibly printed dipoles, are contenders as array elements, both hving fairly well established heritage.

c) RF Amplifiers The RF amplifies required for a phased array need to be linear and have a wide dynamic range. A special development would be essential.

d) Samplers would certainly require a special development. While fast sampling diodes are made the drivers are not. Developers in Israel have produced fast sampling pulses based on lasers and light switches; the results are encouraging.

e) Digital Windows Formule Kaiser windows have been published. 100 dB sidelobes can be achieved when the second, or beta factor, is set equal to about 15.

f) FFT Module Texas Instruments offer a complex arithmetic, FFT module that works up to 1.7 gs/s. Something slightly better than this would be required.for the exaple given below

5 Discriminator Tracking

Manufacturing tolerances, comparator dead bands and RFnoise can all cause erroneous discriminator tracking. Consider the case of an input signal close to band centre. As shown in Figure 4 , D1 can distinguish between a mid-band signal and a lower frequency band-edge signal by reference to the sign of both the sine and cosine outputs. To decide whether the signal is above or below the centre frequency reference must be made to D2. If D1 is in error by less than 2*pi/8, ie 45 degrees, then D1 can be corrected. The errors in D2 should be smaller than those in D1 because it has a longer delay line and hence sharper zero crossings. The procedure of correcting Dn by Dn+1 must be carried out for all n.

A designer of a family of discriminators would have to make an apportionment of the discriminator error budget. Clearly, as one moves from a binary to a quarternary to an octal arrangement the error budget would get tighter and tighter; the binary arrrangement allows the largest error margin; the best but the most expensive.

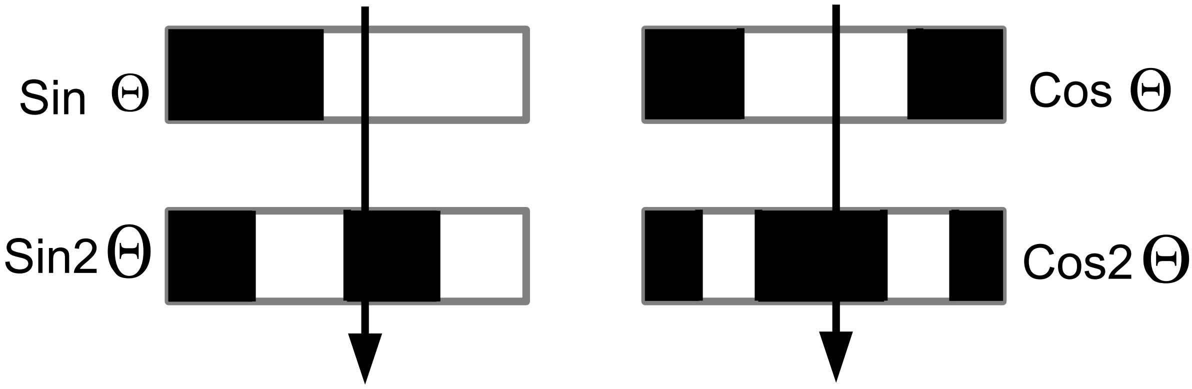

Figure 4 Error Correction Between Discriminators

6 Noise Limitation

In a well designed, multi-discriminator receiver, the RF-input sensitivity is limited by the ability of the receiver to maitain tracking between discriminators.

Following the split in the RF path from the RF amplifier there are two outputs, namely, the delayed and the undelayed, which can be described mathematically as follows:

$V_1 = \exp(j \omega t)$

and $V_2 = \exp[-j(\beta L + \pi/2)]\exp(j \omega t)$ [4.1]

$V_1 + V_2 = 1 + \sin(\beta L) - j\cos(\beta L)$ [4.2]

$V_1 - V_2 = 1 - \sin(\beta L) - j\cos(\beta L)$ [4.3]

$$[4.2] - [4.1] = 2\sin(\beta L) - j2\cos(\beta L) \qquad [4.4]$$

where $\exp(j\omega t)$ is understood
and $\beta = \omega/c$; c velocity of light and L = delay line length.

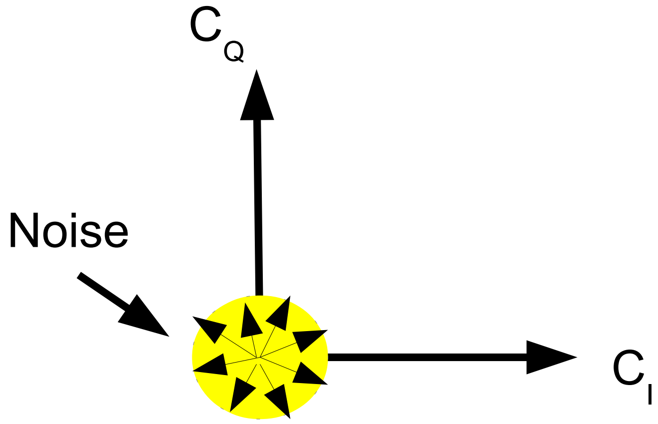

Figure 5 Ricean Model Of band Limited Noise

Adopting the Rice model, the input to the comparitor (A/D convertor) can be described by a rotating vector, as shown in figure 5.

Assume the carrier voltage = C = $E\cos(\omega_0 t)$, where $\omega_0 = 2*pi*f$, and that the accompanying noise is Rayleigh distributed with a magnitude σ, then the in-phase component of the input noise will be σ_{cI}, where

$$\sigma_{cI} = \sigma\cos(\omega_0 t),$$

$$P(\sigma_{cI} > C) = \mathrm{erfc}(2.3)$$ that is C/N > 8.8 dB, whence P < 0.0001

The citerion for discriminator tacking limits the total phase error of each discriminator to plus or minus one half of a discriminator cell width. In the case of D1, this total error is pi/8 (45deg). To maintain P < 0.0001 at 45deg from the ideal cross-over angle between adjacent logic states, C must be increased by 2dB. C must be inreased by a further 3dB because ϕ is derived from two measurements. Thus to limit noise from changing the phase of an in-phase vector by more than 45 then the signal must be 13.8 dB greater than the video band noise.

7 Conclusions

While the proposed linear array covers only 2 GHz bandwidth and about 60 ° in azimuth, many modules would be needed to provide complete coverage of all the popular radar bands. However a lot of the digital hardware could be miniaturized, alieviating the size and mass problems. However the scheme naturally produces frequency and bearing simultaneously.

A lot of thought would have to be given to the problems of assembling the parameters measured in a short time into meaningful groups so that other parameters such as pulse duration and PRF can be derived and a reasonable source identification made.