BISTATIC RADAR DENIAL BY SPATIAL WAVEFORM DIVERSITY

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INTRODUCTION

Many countries have invested heavily in the development of advanced surveillance systems and technologies. Of increasing concern is the threat that potential adversaries may use bistatic technologies to take advantage of significant investment in advanced sensors by our own countries [1, 2]. With relatively inexpensive receiver systems, an adversary could use our signals as bistatic 'illuminators of opportunity'. A central requirement for non-cooperative bistatic operation is the estimation of a coherent reference signal. This estimate is used to by the bistatic receiver to correlate with the received signals to extract the desired signal. As illustrated in Figure 1, a coherent reference is typically obtained by measuring a direct path signal via the sidelobes of the illuminator [3, 4]. Conventional methods to prevent the interception of the direct path signal include low sidelobe antennas, physical isolation, and the use of spread spectrum waveforms. These methods will become inadequate as surveillance sensors migrate to space.

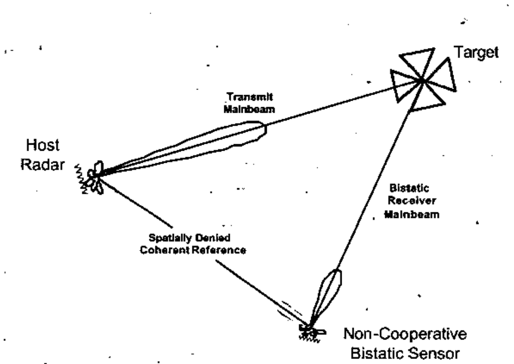

Figure 1. Non-cooperative bistatic receivers require a coherent reference from the host illuminator.

The purpose of this paper is to introduce and evaluate a number of techniques to prevent a radar being used by an adversary as a bistatic illuminator of opportunity. These are all based on the idea of radiating a so-called 'masking signal' (Figure 2) which is arranged to be orthogonal, both in a spatial sense and in a coding sense, to the radar signal, and of a level sufficient to mask the radar signal to an adversary, and hence to deny a reference for bistatic operation.

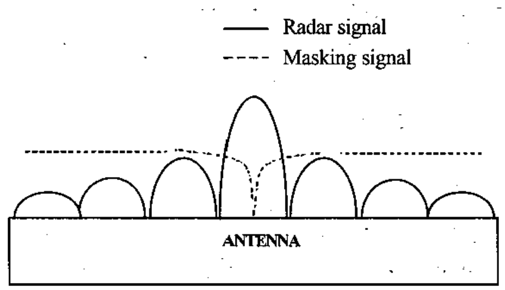

Figure 2. Radar and masking signal radiation patterns.

The problem is therefore one of finding a radar waveform $u_r(t)$ with suitable ambiguity function, and a masking waveform $u_m(t)$ which is orthogonal to the radar waveform over the full range and Doppler domain. The waveforms may be pulsed, quasi-CW or CW. Further, the radar waveform is radiated at a power P_t via a radiation pattern $F_r(\theta)$, and the masking waveform $u_m(t)$ at a power P_m via a radiation pattern $F_m(\theta)$, and we require $F_r(\theta)$ and $F_m(\theta)$ to be spatially orthogonal, over the full bandwidth of the radar.

The overall performance of the scheme is quantified in terms of two parameters: (i) the degree of masking of the radar signal by the masking signal, and (ii) the degree of suppression of echoes (from targets or from clutter) of the masking signal in the channels of the radar receiver.

The next section considers the choice of waveforms. This is followed by a description of two approaches to the choice of radiation patterns, and then the results of some simulations of the overall performance,

WAVEFORM ANALYSIS AND DESIGN

The performance of waveform codes is quantified in terms of their auto-ambiguity [5] and cross-ambiguity [6] functions:

$$|\chi(\tau, f_D)|^2 = \left| \int_{-\infty}^{\infty} u_r(t) u_r^*(t-\tau) \exp(j2\pi f_D t) dt \right|^2 \quad (1)$$

$$|\chi_{r,m}(\tau, f_D)|^2 = \left| \int_{-\infty}^{\infty} u_r(t) u_m^*(t-\tau) \exp(j2\pi f_D t) dt \right|^2 \quad (2)$$

Several different waveform codes have been analysed in this way, including co-channel chirp waveforms of opposite slope [7], pseudo-random binary sequences [8], and Costas codes [9, 10]. For this work the Costas signal is adopted for the host radar waveform because it yields a thumbtack-shaped ambiguity function with a relatively low pedestal. For a fixed number of frequency hops within a radar pulse there are many different hopping patterns that result in essentially the same thumbtack-shaped ambiguity function. Hence, different frequency hopping patterns can be utilized to further complicate the coherent reference estimation task of the non-cooperative radar.

INTERFEROMETER

The first approach to the design of radiation pattern is to use a linear array for the radar, with the masking signal radiated via N additional elements which form an interferometer. The two interferometric elements are driven separately with an independent waveform generation, timing and control circuit. Ideally, the interferometer antenna pattern will overlay the sidelobes of the host radar main antenna pattern with minimal overlay of the radar main beam. This will mask that portion of the host radar signal emitted through the radar sidelobes denying a coherent reference signal to a non-cooperative bistatic receiver.

In Figure 3 we show the azimuthal radiation pattern of the interferometer array factor for N equal to 4 and 5. We notice major lobes at $\alpha = 0°$ and $180°$ as well as major lobes or nulls at $\alpha = \pm 90°$ for N odd or even respectively.

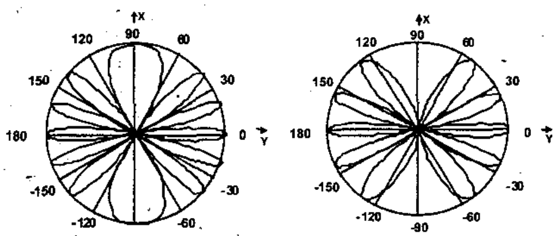

Figure 3. Azimuthal radiation patern for a) N=4 and b) N=5.

In the case where nulls appear at $\alpha = \pm 90°$ we notice something of particular interest: The derivative of the array factor at $\alpha = \pm 90°$ is zero, which means that the nulls at those angles have zero slope. This makes those particular nulls be broader than the rest of the nulls. To avoid self-jamming of the radar waveform, it may be desirable to steer the interferometer pattern such that the main beam of the host radar is centred in this broad null. To achieve this interferometer steering and obtain a broad null at $\alpha = 0$ we place the interferometer on the y-axis keeping the linear array of the main radar along the x-axis. Assuming that the interferometer elements spacing is measured in units of half wavelength $d_{IFM} = k_s(\lambda/2)$ we notice that a broad null exists at $\alpha = 0°$ only for odd k_s. This is shown in Figure 4 for $k_s = 7$.

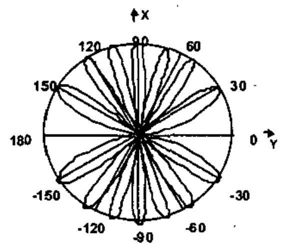

Figure 4. Azimuthal radiation patern for steered interferometer and k_s=7.

In this case broad nulls occur broad side to the main radar antenna in both the horizontal and vertical planes guaranteeing the orthogonality property between radar and masking signal. Since the interferometer excitation is likely to be considerably smaller than the radar excitation, placement of the broad null of the interferometer at the centre of the main beam of the radar is likely to be an effective technique from preventing the interferometer signal from interfering with the desired radar target returns.

Notice that as the number of interferometer elements N increases, both the broad null as well as the spacing between sidelobes widens, thereby decreasing masking coverage in the direction of a potential non-cooperative radar. One possible method to overcome this deficiency is to change the configuration of the interferometer so as to form a triangle with three elements. This pattern is more irregular, but does have increased coverage despite being at a lower amplitude.

BUTLER MATRIX

The second approach uses an N-element linear antenna array. Suppose initially that the array is fed by a Butler Matrix [13] (Figure 5). This generates a set of spatially-orthogonal antenna beams, each of the form

$$|E| = \frac{1}{N} \frac{\sin(N\psi/2)}{\sin(\psi/2)} \quad (3)$$

with

$$\psi = \frac{kd}{\lambda} \sin(\theta - \delta) \quad (4)$$

where d is the element spacing, λ is the wavelength, $k = 2\pi/\lambda$, θ is the azimuth angle and δ is the angle of the maximum of the particular beam. For an N-element array

$$\delta_m = \frac{(2m-1)\pi}{N} \tag{5}$$

so the normalized far-field pattern of the m^{th} beam is

$$E_m = \frac{1}{N}\frac{\sin N\{(kd/2)\sin\theta - [(2m-1)/N](\pi/2)\}}{\sin\{(kd/2)\sin\theta - [(2m-1)/N](\pi/2)\}} \tag{6}$$

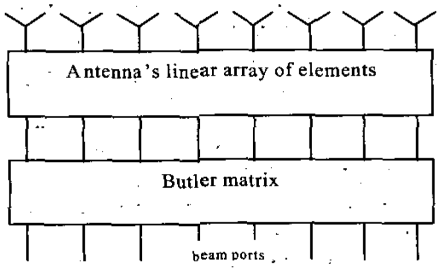

Figure 5. Linear array and Butler matrix

The orthogonality of this set of beams is maintained over a broad bandwidth, dictated by the hardware of the Butler Matrix, but typically an octave or more. In order for this to be so, the beamwidths and directions of the beams must change with frequency. The beams have a first sidelobe level of –13.2 dB, which is rather high for radar purposes; the sidelobe level can be lowered by an amplitude taper across the array in the usual way, but this destroys the orthogonality condition. The set of beams may be steered electronically by a set of phase shifters, either at the antenna elements or at the beam ports.

Suppose that one of the central beams is used for the radar, both for transmitting and receiving. One or more of the remaining beams is used to radiate the masking signal or signals, at an appropriate relative power level. Furthermore, if the radar signal and masking signal(s) were to be generated at the beam ports of the Butler Matrix by direct digital synthesis, which could include the effect of phase shifts to steer the beams electronically, then since the signals radiated from each element are simply weighted combinations of the beam port signals, the element signals may be calculated and generated directly, without any need for the Butler Matrix hardware.

RESULTS

To evaluate the performance of a given system it is necessary to specify a value for the degree of masking. In practice this will vary with direction θ, so may be specified as a peak value or as a mean averaged over the sidelobe region of $F_r(\theta)$. For that reason let us define

$$L(\theta) = \frac{P_r F_r(\theta)}{P_m F_m(\theta)} \tag{7}$$

with P_m and P_r the power levels at which the masking and radar signals are transmitted. This is the radar signal to masking signal ratio for the case of an adversary listening from a particular angle θ. This ratio depends on the geometry of the antennas used and their radiation pattern. The required degree of masking represents a compromise on one hand by the need to disrupt the coherent reference, and on the other hand not to disrupt the operation of the radar. From a knowledge of the effect of ECM, a value of about 13 dB is likely to be adequate. The value of L at $\theta = 0$ (i.e. at the centre of the host radar main lobe) will obviously take negligible values first because of the broad null of the masking signal at this angle and secondly because of the coding of the signals. An adversary could only recover the radar signal if listening from that specific direction.

Once $L(\theta)$ is chosen, the degree of suppression of the masking signal in the radar receiver can be evaluated, both due to the orthogonality of the waveform codes and of the radiation patterns. To do so we introduce the radar signal echo to masking signal echo ratio

$$R(\theta) = L(\theta)\frac{\int_{-\infty}^{\infty} u_r^*(t)u_r(t-\tau)\exp\left(j2\pi f_D t\right)dt}{\int_{-\infty}^{\infty} u_r^*(t)u_m(t-\tau)\left(j2\pi f_D t\right)dt} \tag{8}$$

with $u_r^*(t)u_r(t)$ the response to the radar signal of the filter matched to the radar signal and $u_r^*(t)u_m(t)$ the response to the masking signal of the filter matched to the radar signal. This ratio includes echoes received from the target and clutter. The masking signal levels will be further suppressed because the filter at the receiver is matched to the radar signal.

We see that the above ratio depends on the relative transmitted levels of the radar signal and masking signal, as well as the responses of the radar and masking signals at a given Doppler shift to the filter matched to the radar signal at zero Doppler. All of the above is a function of the position of the target, so it is not controllable by the radar designer. However, the power levels P_m and P_r of the masking and radar signals are controllable and can be set in order to achieve the desired results.

Figure 6 shows an example with a Costas code of length 7, and P_r/P_m = 20 dB. The result is in the form of the cross-ambiguity function (equation 2), taking account of the ratio P_r/P_m and weighted by the further suppression provided by the orthogonality of the radiation patterns. It can be seen that the suppression of the masking signal is of the order of 30 or 40 dB.

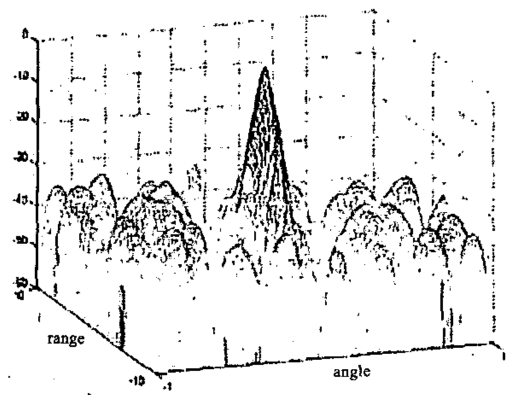

Figure 6. Radar signal echo to masking signal echo as a function of angle and range and P_r/P_m = 20 dB for a Costas code of length 7 (vertical scale in dB).

The higher the power level of the masking signal at the transmitter the more unlikely it is for the radar to be used by an adversary as a bistatic illuminator of opportunity. In addition different waveforms for the masking signal give different results. In Figure 7, for example, we show what happens when keeping the same masking to radar signal power ratio as in Figure 6 but this time applied on a Costas code of length 30.

We notice that in this case the suppression of the masking signal is of the order of 45 or 50dB.

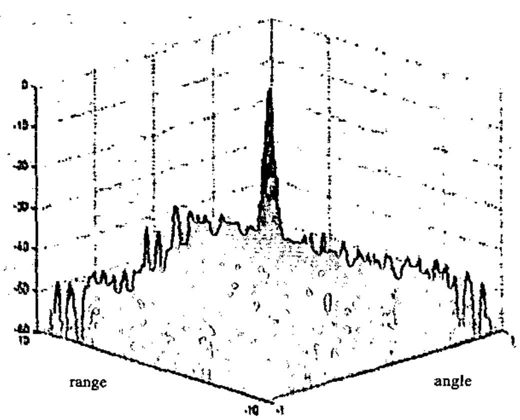

Figure 7. Radar signal echo to masking signal echo as a function of angle and range and P_r/P_m = 20 dB for a Costas code of length 30 (vertical scale in dB).

EMBEDDED COMMUNICATIONS

Another feature of the masking signal is that it may also simultaneously be used for communications. By use of Orthogonal Frequency Division Multiplexing, simple codes embedded in the phase and amplitude of the masking signal can be used to send messages, for example as telemetry to the ground. This is a topic that requires further investigation, especially with regards to security and cryptography.

CONCLUSIONS

This paper has introduced and analysed a set of techniques to prevent a radar being used by a bistatic receiver as a non-cooperative illuminator, by radiating in addition to the radar signal waveform a 'masking signal' waveform. This is designed to be orthogonal to the radar signal waveform, both in the coding domain and the spatial domain. A number of waveform coding techniques have been considered and from those, Costas codes appear to offer best performance and flexibility. Two spatial coding techniques have been devised and analysed; one based on an interferometer, and one based on a Butler matrix. Expressions as a function of the system parameters, have been derived for the degree of suppression of the radar signal by the masking signal, and for the suppression of the masking signal in the host radar echo. Evaluation and plotting of these expressions have demonstrated that it is possible to obtain adequate masking of the radar signal, whilst at the same time achieving suppression of echoes from the masking signal of the order of 30-50 dB. In this respect the performance of the interferometer and Butler matrix schemes are comparable.

The ideas presented are just one example of the idea of 'waveform diversity', in which waveform coding techniques are used with multiple transmit and receive beams, possibly with adaptive processing in spatial and temporal domains. Central to the analysis of such schemes is the cross-ambiguity function, weighted by the appropriate transmit and receive radiation patterns. It should be possible to generalise the formulation in this paper to other types of system, to situations where the transmit and receive antenna patterns are different, and to broadband signals.

ACKNOWLEDGEMENTS

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