Television-based bistatic radar

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Indexing terms: *Radar and radionavigation, Television*

Abstract: The paper discusses the use of 'illuminators of opportunity' for bistatic radar systems. Experiments in the London area using the Crystal Palace transmitters are reported, including the use of TV pictures designed to make the transmission more closely resemble a pulsed radar signal. It is shown that the separation of targets from the direct signal and clutter requires extensive signal processing under all but the most favourable conditions.

1 Introduction

There has been growing interest over the past 20 years in the application of bistatic and multistatic techniques for both long- and short-range surveillance radar, and a number of experimental systems have been built and evaluated [1–3]. The majority of these have employed as illuminators either (i) existing radar transmitters or (ii) dedicated transmitters whose characteristics are optimised for the bistatic system. Another approach, however, employs as the illuminator any convenient radar-like transmission, or 'illuminator of opportunity'. This has the merit of being a completely covert radar system, which does not advertise the presence of receiver or transmitter stations. This paper considers a system of this kind, using a UHF television transmitter as the illuminator. The television signal is repetitive and essentially pulsed in nature, and indeed the 'ghosting' effect sometimes seen on a domestic television set is a simple example of radar operation.

The advantages and disadvantages of bistatic radar over the monostatic case have been discussed by a number of authors [4, 5] and elsewhere in this special issue. These are summarised as follows:

(*a*) The receiving system is completely passive, and hence undetectable, and is immune to the effects of deliberate directional interference. Because it is passive, it may also be simple and cheap.

(*b*) The dynamic range of signals to be handled is reduced, because of the defined minimum range.

(*c*) It is necessary to provide synchronisation between transmitter and receiver in respect of (i) instant of transmission of pulse, (ii) transmit antenna azimuth (in the case of a scanning transmit antenna), and (iii) transmit signal phase (if coherent moving target indication (MTI) is to be employed). This may take the form of a land-line link, although with a co-operative transmitter it is possible to realise a totally independent bistatic system by means of 'flywheel' clocks at the receiver which are resynchronised each time the transmitter beam sweeps past [6]. Additionally, a coherent reference for MTI cancellation may be obtained from close-in stable clutter echoes [7].

(*d*) There is a co-ordinate distortion effect; targets on the transmitter-receiver baseline have zero bistatic range. Elsewhere, contours of constant bistatic range are ellipses with the transmitter and receiver sites as the two foci. The distortion can be corrected with a knowledge of the bistatic geometry, and a system providing the correction in real time, using a fast microprocessor-based system to produce a corrected plan-position indicator (PPI) display centred on the transmitter site, has been described [2].

(*e*) Similarly, contours of zero Doppler are the same ellipses of constant bistatic range. Contours of maximum Doppler are hyperbolas crossing the ellipses orthogonally. Moving targets cannot present zero Doppler to two receiving sites simultaneously.

(*f*) Milne [4] has discussed several configurations of transmitters and receivers. Azimuthal discrimination (i.e. directional antennas) at both transmitter and receiver is very desirable, since otherwise the radar system is vulnerable to responses from sidelobes. A directional receive antenna must scan at a non-uniform rate, which practically rules out mechanical scanning. The form of scanning required follows the position of the RF pulse through space, and is known as 'pulse chasing' [8]. Thus, either a small number of electronically agile beams or a static set of contiguous beams will be required. Such a receiving system is unlikely to be either simple or cheap, but because the receiver is passive it is much less vulnerable to attack.

(*g*) The target bistatic cross-section σ_b is not the same as the monostatic cross-section, although the range of values of σ_b for a particular target will be comparable with the range of values of monostatic cross-section [9]. Thus a particular target is unlikely to present a low cross-section to more than one transmitter-target-receiver geometry. Additionally, at large angles target glint is drastically reduced, eliminating this contribution to radar inaccuracies [3].

(*h*) Receiving or transmitting stations can be used interferometrically to obtain high azimuthal discrimination; sources of interference can be located by triangulation, and passive detection (correlation) techniques can be employed to locate noise-like interference sources.

(*i*) The constraints on radar waveform design can be different to those for the monostatic case. As a simple example, a higher-than-normal pulse repetition frequency (PRF) may be used, and the range ambiguities resolved by triangulation from several receiver sites or by using staggered pulse repetition intervals (PRIs).

(*j*) The transmitter may be located remotely. An example of this is the US Sanctuary programme [3]. This may be particularly desirable for short-range surveillance applications. In addition, multiple 'winking' transmitters may be employed, as a counter to homing missiles.

Another possibility is to use as the bistatic transmitter one or more illuminators of opportunity — i.e. transmissions present for other purposes and with favourable location, power level and modulation. To do this, it is necessary to know a number of parameters of the source, e.g. location, motion (if any), antenna beam direction (if the transmitter is directional, and especially if it scans in azimuth and/or elevation) and modulation. As examples, if the modulation

Paper 4917F (E15, E14), first received 28th February and in revised form 25th July 1986

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is predetermined and regular, as would generally be the case with an existing radar transmitter, then it is just necessary to know the instant of transmission of each pulse [6]. On the other hand, if the modulation is random, such as with a communications or broadcast transmission, the modulation function must be found explicitly.

The questions that this paper seeks to answer are (i) what characteristics are desirable in illuminators of opportunity?; and (ii) what forms of signal processing are necessary to take advantage of them? We discuss first of all the properties of the television modulation scheme, and the application of these to a radar system. The approach taken in this work has been to consider simplified test-card-like modulation as the basis for experiments, and then to extend the results to *ad hoc* TV programme material, and to more general transmissions. A brief description is given of an experimental system using the television transmitter located at Crystal Palace in south London, and of results achieved, and these are discussed in the light of the questions posed above.

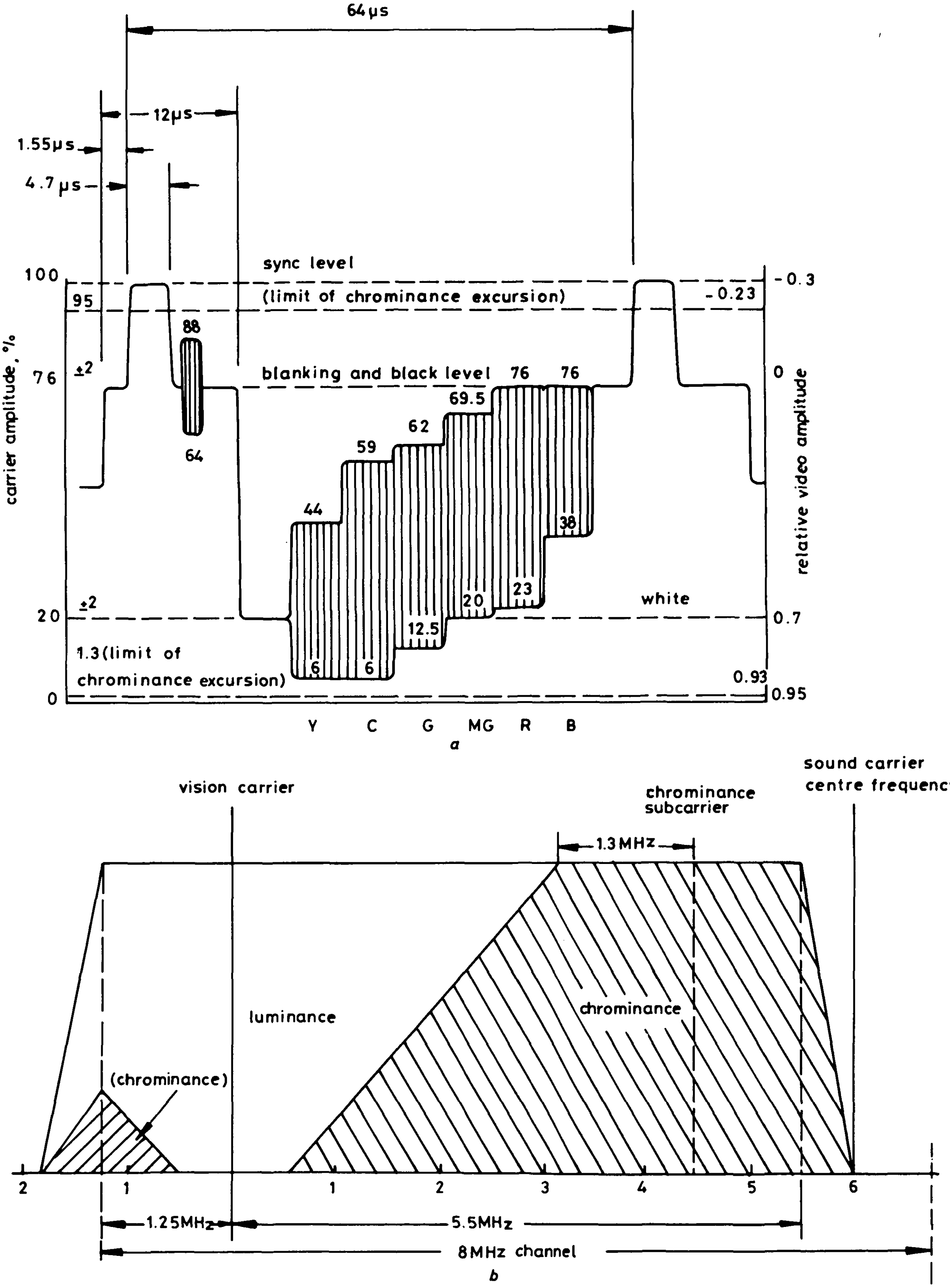


Fig. 1 *Idealised carrier amplitude/time variation and frequency bands*

a Waveform showing variation of carrier amplitude with time for a line of 100.0.100.25 colour bars (95%)

b Frequency bands occupied by colour picture components and sound signal from an ideal transmitter

2 Television waveforms and their use as radar signals

2.1 The TV waveform

Details of the 625-line PAL TV transmissions used in the UK can be found in a joint BBC and IBA specification [10]. Fig. 1*a* shows the idealised carrier amplitude for a picture of line colour test bars. The transmitted signal is essentially filtered AM and the waveform shown is that recovered in an ideal receiver with the correct frequency response.

An important feature of the transmission is that it has asymmetric sidebands. The terms 'single-sideband (SSB) full carrier' or 'vestigial sideband' are sometimes used to describe it, but incorrectly so. It is worthwhile examining this in more detail since the idealised traces published in the specification will not be those received on systems with different passbands.

Whatever method a TV transmitter actually uses to construct its signal, the result should be the same as using the video signal to amplitude-modulate the vision carrier fully and passing the modulated signal through a filter; the result is shown in Fig. 1*b*.

From the standpoint of conventional modulation techniques, the difficulty with the TV vision signal is that it changes from AM to SSB as the video frequency rises. Below 1.25 MHz the signal is pure AM; from 1.75 MHz to 5.5 MHz it is pure upper sideband. In particular, this means that while the lower video frequencies can modulate the carrier fully, the higher ones can reach only 50%. Normal detection would, therefore, result in reduced high-frequency gain. In a TV receiver this is overcome by applying a filter response that reduces the levels of the carrier and the low-frequency video information.

The exact spectral distribution of the vision signal depends on the picture content. From Fig. 1 it can be seen that the higher video frequencies relate to chrominance information, and that for a black-and-white picture the energy will be concentrated closer to the carrier.

It is therefore fair to assume that the bulk of the vision signal energy falls within the ± 1.25 MHz band that is transmitted as AM. Exceptions might be pictures with saturated colours or with fine luminance detail all over large areas, such as a page of text.

The sound carrier, with FM audio information, is 6 MHz above the vision carrier; it can be considered either as a separate carrier accurately spaced or as an upper sideband modulation. Its power level is approximately one-fifth of the peak vision carrier. The sound carrier, however, (officially) has a modulation limit of 15 kHz, and so it is not suitable for pulsed or frequency-shift keyed (FSK) radar purposes as poor range resolution would be obtained, about 20 km.

2.2 Received video signals

The dominant feature of a TV signal is the line sync pedestal and pulse. In the UK PAL system a higher carrier level represents a blacker part of the picture. The line blanking pedestal is at the black level of 76% of peak, but the sync pulse itself extends blacker-than-black to the 100% level (Fig. 1*a*). The pure white level is at 20%; during the picture information part of the line the luminance signal must be within this 20% to 76% band. The combined luminance and chrominance signal can, however, exceed these limits and reach 1.3% or 95%.

The chrominance signal is centred on the colour subcarrier of 4.43 MHz and will be lost in systems of reduced bandwidth. For filtered video, then, or for black-and-white transmissions, the picture video level is constrained to the 20% to 76% band. This modulation format does not allow the transmitted power to be reduced to zero at any time and, therefore, accurately mimicking a conventional pulsed radar signal is not feasible.

A conventional radar signal is useful for ranging purposes because of the form of its autocorrelation function; this is a measure of how easily and accurately timing information can be derived from a signal and of how well a mixture of variously delayed signals can be separated. It is given by

$$A_v(\tau) = \int v(t) v^*(\tau - t)\, dt \tag{1}$$

The autocorrelation function of an ideal radar signal is zero everywhere except for well defined tall and narrow peaks at sufficient spacing to avoid any ambiguity. Since a TV signal never goes to zero, it will not prove so good.

If no control over the picture is assumed and normal *ad hoc* programme material or a test card is radiated, the system would then have to work with a typical waveform such as that in Fig. 1*a*. The time autocorrelation function of this would depend on the picture content, but it can be expected to show a series of small peaks at the line repetition rate of 16 kHz.

On the other hand, if control is possible then waveforms can be chosen, within the transmitter system limits, that provide better autocorrelation functions. One of the simplest and easiest to generate is the 'sync-plus-white' signal which has the high-level sync pulses separated by periods of 20% (white) level. The autocorrelation function of this can readily be calculated but with a max/min ratio of about 2.8 it is not especially useful. Furthermore, it will provide only poor range resolution and has severe ambiguity in multiples of 9600 m.

It may be possible to construct coded waveforms with finer resolution and without ambiguity. Initially, however, we will be considering short-range targets for which range ambiguity is not a problem, and we will be more concerned with detection and MTI than with accurate ranging.

The sync-plus-white waveform is useful because its pulsed nature allows some simple time-domain processing. This signal with superimposed echoes would look something akin to a radar A-trace, and it should be possible to pick out targets by eye. There is, however, one fundamental difference. The echoed sync pedestal may combine constructively or destructively with the residual 20% carrier level of the direct signal. Thus, on the A-trace an isolated target can appear as a positive- or negative-going pulse, or, if the phases are in quadrature, not at all.

This effect has one advantage; it would not be necessary to supply a coherent reference for the implementation of an MTI scheme. A simple MTI system would consist of simple differentiation, either AC coupling or taking the difference between successive lines. This could give quite useful results with sync-plus-white signals and may even be profitable with *ad hoc* pictures. This of course depends on the line-to-line consistency of the picture, and in any case might only indicate the presence of a target.

2.3 Target/clutter ratios

The received signal at any location is the RF sum of the direct signal and of multipath propagation. Most of the multipath energy will arrive by reflection from fixed objects, such as buildings, but some may come from moving aircraft targets.

The bistatic radar equation gives the power received from a target:

$$P_r = \frac{P_t G_t G_r \lambda^2 \sigma_b}{(4\pi)^3 (r_1 r_2)^2} \tag{2}$$

where P_t is the transmitted power, P_r is the received power, G_t is the transmit antenna gain, G_r is the receive antenna gain, σ_b is the target bistatic cross-section, r_1 is the transmitter-to-target range, r_2 is the target-to-receiver range and propagation losses have been taken as negligible.

This can be compared with the power received directly at range r from the transmitter:

$$P'_r = \frac{P_t G'_t G'_r \lambda^2}{(4\pi)^2 r^2} \tag{3}$$

The ratio of target to direct signal is therefore

$$\frac{P_r}{P'_r} = \frac{1}{4\pi} \frac{\sigma_b r^2}{(r_1 r_2)^2} \frac{G_t}{G'_t} \frac{G_r}{G'_r} \tag{4}$$

The last two terms represent the fact that the antenna gains may be different in the direct line-of-sight and target directions. If isotropic antennas, or at least ones with uniform azimuthal patterns are assumed, then

$$\text{power ratio} = \frac{1}{4\pi} \frac{\sigma_b r^2}{(r_1 r_2)^2} \tag{5}$$

The distance from Crystal Palace to University College London (UCL) is 11.8 km. Inserting this and choosing $r_1 r_2 = 400\ \text{km}^2$, which represents an aircraft between UCL and Heathrow Airport, and $\sigma_b = 20\ \text{m}^2$, we find the power ratio is approximately -88 dB.

Clearly this is unusable as, unless the signals were separated in time, the target would be buried under the distortion products caused by the direct signal in the receiver circuits, quite apart from the signal processing problem.

The answer, therefore, is to rely on antenna patterns to give some improvement. Naïvely, one could consider steering a null onto the transmitter and reversing the tables completely. The required depth of the null, however, is so great that only an adaptive closed-loop technique would be useful [11]. The problem would then arise of determining exactly what constitutes the unwanted signal that must be eliminated. Pointing a reference antenna beam at the transmitter is not a perfect solution, as this would pick up narrow-angle multipath as well and possibly even the target signal via a sidelobe. Even if a direct reference via a land line were available, the problem of distortion in the line circuits or in the transmitter after the pickoff point would remain.

In the limit, though, the best achievable target/clutter ratio is set not by adaptive antenna performance but by ground clutter.

If the receiver used an antenna with a beam pointed at the target and very low sidelobes everywhere else, the signal received would be the echo from the target plus the echoes from any clutter in the area commonly covered by both receive and transmit antennas. The criterion is then the relative echoing areas and properties of clutter and target.

The power returned from illuminated ground clutter is equal to that from a target of $\sigma^\circ A$, where σ° is the relative reflectivity and A is the plan area illuminated. For low incidence angles, Skolnik [12] gives values of σ° of -20 dB for a city and -30 dB for cultivated terrain.

Because the ground clutter is not confined to a specific range, it is not possible to ascribe to it an effective area. The target/clutter power ratio is given by

$$\frac{\text{target power}}{\text{clutter power}} = \frac{\dfrac{G_r \sigma_b}{(r_{1t} r_{2t})^2}}{\displaystyle\iint \frac{G_r \sigma^\circ}{(r_1 r_2)^2}\, dA} \tag{6}$$

where r_{1t}, r_{2t} are the target-to-transmitter and target-to-receiver distances.

For conventional radar, the integration is over a range/beamwidth cell; for this case the limits are set by the receiver antenna pattern. In neither case can the integration be carried to $r_2 = 0$, so assumptions about minimum range have to be made.

Taking no clutter returns within 1 km of the receiver, for our geometry we have an approximate numerical solution of

$$\frac{\text{target power}}{\text{clutter power}} \simeq \frac{2.4 \times 10^7}{\sigma^\circ \theta} \frac{\sigma_b}{(r_{1t} r_{2t})^2} \tag{7}$$

where θ is the antenna beamwidth; with $\sigma_b = 20\ \text{m}^2$, $r_1 r_2 = 400\ \text{km}^2$, $\theta = 0.1$ rad and $\sigma^\circ = -20$ dB, then the target/clutter power ratio $= 3 \times 10^{-6}$ or -55 dB.

This is with a 6 m-wide antenna mounted in a very clear site. It might be possible to rely on some improvement in this figure if a suitable elevation pattern is used. This, however, is only likely for short-range targets where the problem is less severe anyway.

These indicate target/clutter ratios at the receiver detector of some -50 dB or more. With a TV illuminator, unlike normal radars, the target return is compared with a fixed clutter level rather than one that decreases with range. For a high probability of detection of distant targets, though, the signal processing following the detector must use the features of the waveforms to realise more than 50 dB processing gain.

A further problem with such a system is multipath illumination of the target. Delays with respect to the transmitter-target-receiver path of up to a few microseconds might be expected from terrain or objects close to and beneath the transmitter, and since these receive the benefit of the full transmit antenna illumination they could cause significant blurring of the target echo. A set of tapped delay-line equalisers, one for each azimuth direction, might be used to combat this problem.

2.4 MTI possibilities

With a sync-plus-white signal, the 20% residual carrier level will act as the reference for a simple two-pulse canceller. This is true whenever there are repetitive features of the transmitted waveform; this is easily achieved by pictures without horizontal features.

If adjacent lines are compared, the high line repetition frequency of 16 kHz results in an MTI filter response that peaks at quite high Doppler frequencies. A better response may result if the MTI delay is chosen as, say, 8 or 16 lines.

A further complication is that in a bistatic system the Doppler frequency produced by a moving target depends not only on the target speed and direction but also on the position relative to the two sites. Bistatic Doppler frequencies will always be lower than those found in monostatic systems in similar circumstances.

Fig. 2 shows the Doppler frequencies to be expected for certain conditions. This assumes targets travelling perpendicular to the baseline between Crystal Palace and UCL.

Also shown are the filter responses for 8- and 16-line delay two-pulse cancellers.

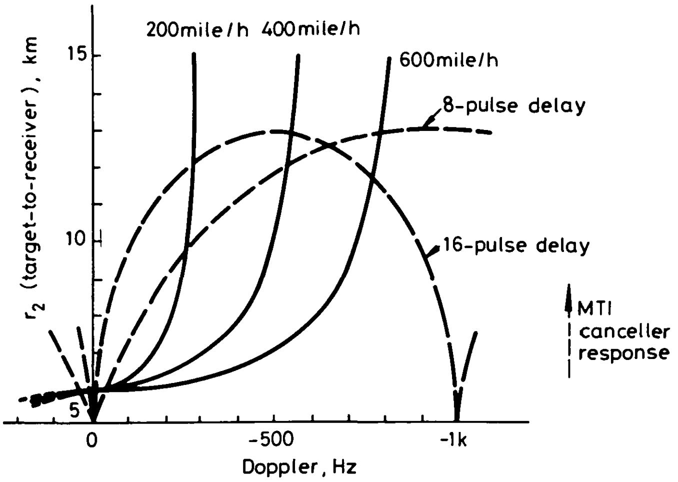


Fig. 2 *Variation of target Doppler as a function of range and velocity*
Also shown are MTI canceller responses

The processing gain (or, more specifically, improvement factor or subclutter visibility) of an MTI canceller depends very much on the relative stabilities of the transmitter and the reference oscillator (see Reference 12, Chap. 17). These must be accurate to within one cycle over the MTI delay period. From this point of view a TV bistatic system is excellent, and the zero velocity null can be expected to be very deep. The line frequency and colour subcarrier are normally controlled to within 0.25 parts per million [10]. What will limit the processing gain are more prosaic and unpredictable factors such as circuit noise level and linearity; the processing gain cannot exceed the dynamic range of the system.

2.5 The radar system

The Crystal Palace station transmits four TV channels, as shown in Table 1. The effective radiated power (ERP) is 1 MW; the radiation is horizontally polarised with uniform azimuthal distribution.

Table 1 : Crystal Palace transmission

	Channel number	Vision carrier, MHz
ITV	23	487.25
BBC 1	26	511.25
Channel 4	30	543.25
BBC 2	33	567.25

Vertical-plane radiation patterns of the Crystal Palace transmitter show that the main beam is angled downwards by about 1°, with the below-horizontal pattern following an approximately cosec^2 function. The reason for this pointing is to avoid wasting power above the horizontal, which is unfortunately just where we want to illuminate aircraft targets. Above the horizontal, in all azimuthal directions, there is a null at about 2.5° above the horizontal, about 15 dB deep, with a lobe at about 5° about -6 dB. Aircraft targets therefore will not receive the benefit of full illumination by the transmitter.

The ERP for airborne targets is therefore reduced to at best 250 kW in total. Furthermore, since it is generally only practical to work with one TV channel at a time, the effective ERP is reduced further to 62.5 kW.

The requirement for detection against thermal noise can be written as

$$P_r \geqslant R_t kTB \tag{8}$$

where K is Boltzmann's constant, T is the system noise temperature, B is the bandwidth and R_t is the threshold ratio dictated by false alarm rates etc.

Combining this with the bistatic radar eqn. 2 yields the condition

$$(r_1 r_2)^2 \leqslant \frac{1}{(4\pi)^3} \frac{P_t G_t G_r}{B} \frac{\lambda^2 \sigma_b}{R_t kT} \tag{9}$$

Although a TV channel is 8 MHz wide, most of the energy is concentrated into a smaller bandwidth centred on the vision carrier. As shown above, 2.5 MHz is a reasonable value to take for B.

The detection threshold ratio might be 10 dB and a large aircraft might have a bistatic radar cross-section of 20 m^2. Inserting these figures into eqn. 9, with the wavelength for the BBC2 signal and choosing 15 dB receive antenna gain, we find that

$$r_1 r_2 \leqslant 2.1 \times 10^8 \text{ m}^2$$

The local geometry of Crystal Palace, UCL and Heathrow Airport is shown in Fig. 3. Superimposed on this is the curve given by $r_1 r_2 = 2 \times 10^8$ m^2. Such a locus is an oval of Cassini.

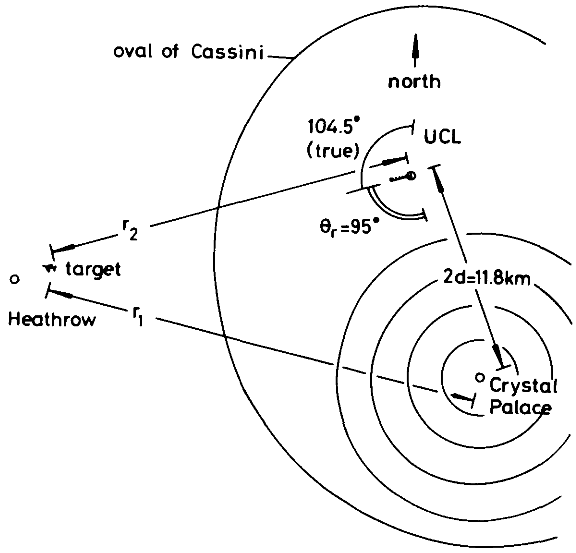


Fig. 3 *Horizontal-plane geometry of Crystal Palace, UCL and Heathrow*
National Grid references:
Crystal Palace 340 712
UCL 298 822
Heathrow 075 758
Indicated oval of Cassini is the locus of $r_1 r_2 = 2 \times 10^8$ m^2

A large receiving aperture would be required for increased range since, although the transmitted power is of the same order as some radar systems, it emanates from a low-gain aerial. By comparison, normal monostatic radar systems have large-aperture (and hence high-gain) antennas that are used on both transmit and receive.

2.6 Pulsed sideband modulation

To assist with the signal processing, the use of more complicated special transmissions was investigated. Clearly these had to be in the normal TV format, and so must be realisable either from video signals or from specially prepared pictures. What sort of picture, then, might result in a transmission with radar-like features?

It was noted earlier that for video frequencies between 1.75 and 3.5 MHz the TV transmitter (and receiver) are straightforward SSB systems. Signals in this band thus have a direct equivalent in the RF spectrum, and can be separated from the rest of the transmission by simple frequency filtering at RF, IF or baseband (video). If a pulsed signal is introduced here then something akin to a normal radar waveform can be achieved.

One way of doing this is with a picture containing narrow vertical stripes. This will cause the transmitted signal to contain high-energy bursts of a sideband repeated at the line rate. It would be a simple matter to filter this pulse train from the recovered TV signal and examine it in the time domain for target echoes.

It was discovered that this is essentially similar to one of the patterns produced by some standard video test equipment. The pattern in question, called 'multiburst', is a black-and-white video signal in which each line consists of a white, a black and a grey level followed by full black to white tone bursts at frequencies of 0.5, 1, 2, 3, 4 and 5 MHz. Each burst is about 6 μs long, which is potentially a more useful pulse length than the 12 μs sync pedestal. Arrangements were made to incorporate this into a test transmission schedule.

3 Description of experimental system

The experimental system used to investigate some of the aspects of TV bistatic radar comprised two parallel receiver channels. The idea was that one channel could be used to receive a direct or reference signal with minimal ghosting while the other would be the one studied for radar echoes.

The receivers were built around standard commercial tuner and IF units. This was the simplest and cheapest way of assembling a working system, but did mean that the performance, in terms of noise figure and intermodulation products, might be inferior to a 'custom' system. It also meant using the prescribed passband shape rather than one better suited to AM reception around the vision carrier, although it was possible to alleviate this somewhat by overriding the automatic frequency control (AFC) and retuning it slightly.

The video output from these strips was not of the full 5.5 MHz bandwidth; in particular, the colour subcarrier had been filtered out. For our purposes, though, this is more of an advantage than disadvantage.

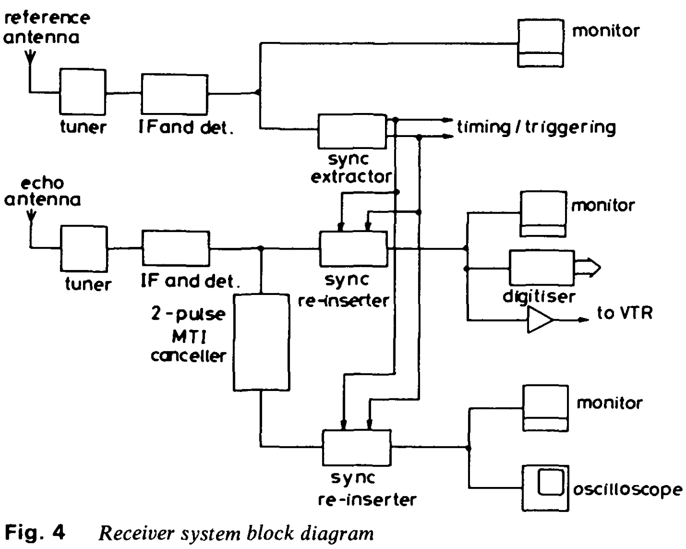


Fig. 4 *Receiver system block diagram*

Various other modules were built to process the video signals; in this manner, it was possible to rearrange the parts of the system according to what video handling was required. Fig. 4 shows the receiver system block diagram.

3.1 Antennas

A 10-element Yagi, mounted on the roof of the UCL building and directed at the Crystal Palace TV transmitter, was used to provide the reference signal. Observed on a monitor, this signal had no significant ghosting and an oscilloscope examination of the video showed clean distinct sync pulses; this yielded -18 dBm vision carrier level at the receiver input.

A vertical array of four 17-element Yagis was used for most of the echo reception. This was mounted on the north side of the building to provide some shielding from the direct signal. This array was spaced to direct its first null at 15° elevation and tilted upwards by this angle to reduce ground clutter problems. The array could be manually set to azimuthal directions around that of Heathrow. The gain over isotropic of this system was calculated as 17 dB. The total level of signal (i.e. direct + clutter) received from it is -32 to -35 dBm (vision carrier).

At a later stage a single Yagi on a rotator was used for some experiments. On the same mast was fixed a video camera feeding a monitor in the laboratory. This enabled manual tracking of promising targets. The camera was also useful for checking the presence or absence of aircraft.

3.2 Video processing

The received echo signal has such a distorted sync pulse that it cannot be properly displayed on a monitor. The solution to this was to splice in the sync pulses from the direct signal. The line repetition interval is long compared to the bistatic delay, so this technique does not disturb any information that might be present.

Several methods of digitally processing the video signals were available. The first used an 8-bit digitiser and store constructed at the Royal Signals & Radar Establishment (RSRE). This could catch a block of data in real time and transfer it to a computer for off-line processing. Digitisation to N bits sets a limit to the dynamic range of 20 $\log_{10} 2^N$ dB; thus for eight bits the limit is 48 dB.

This system could only capture data in small windows, with a significant recovery time before the next set of samples. Recording fleeting events was therefore not easy, and so it was augmented by a hardware digital MTI canceller that operated continuously but could not record data. This employed 8-bit digitisation of the raw 'echo' video at a 1 μs rate; the output could be selected as the difference between the current video line and that 1, 8 or 16 lines previously. This system, therefore, could give results in real time.

4 Experimental results

4.1 Sync-plus-white experiments

The first tests with this system were done using the sync-plus-white picture signal described earlier. This was radiated by the BBC at various times outside programme hours on Channels 26 (BBC 1) and 33 (BBC 2).

Fig. 5 shows a typical single-line A-trace video from these tests. Static echoes of the sync pulse (or sync blanking pedestal) are clearly visible. It was found that the patterns of these echoes would change markedly as different azimuthal directions were tried. Sometimes, the dominant features of the received signals could be correlated with large buildings in the general area. These are, of course, the result of many overlapping target returns, which accounts for the complicated nature of the A-trace.

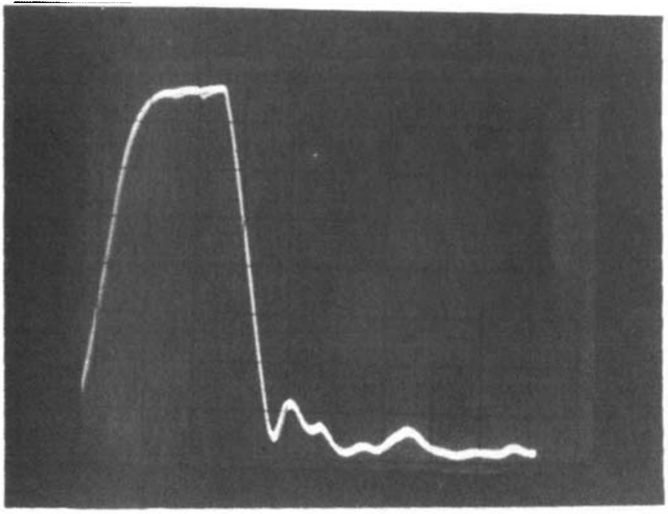


Fig. 5 *Sync pulse from echo antenna*
Horizontal scale: 5 μs/div; vertical scale: linear

The digitiser and store was used to capture data at instances when aircraft were visible. These were processed later to give output in a variety of forms. One particular result is illustrated: Fig. 6*a* shows an A-scan display of signal level against range, and the characteristic 'butterfly' of coherently detected Doppler-shifted returns is evident about one-third of the way across the plot, superimposed on top of the clutter and direct signal. Implementing a two-pulse MTI canceller on this data gave the plot shown in Fig. 6*b*.

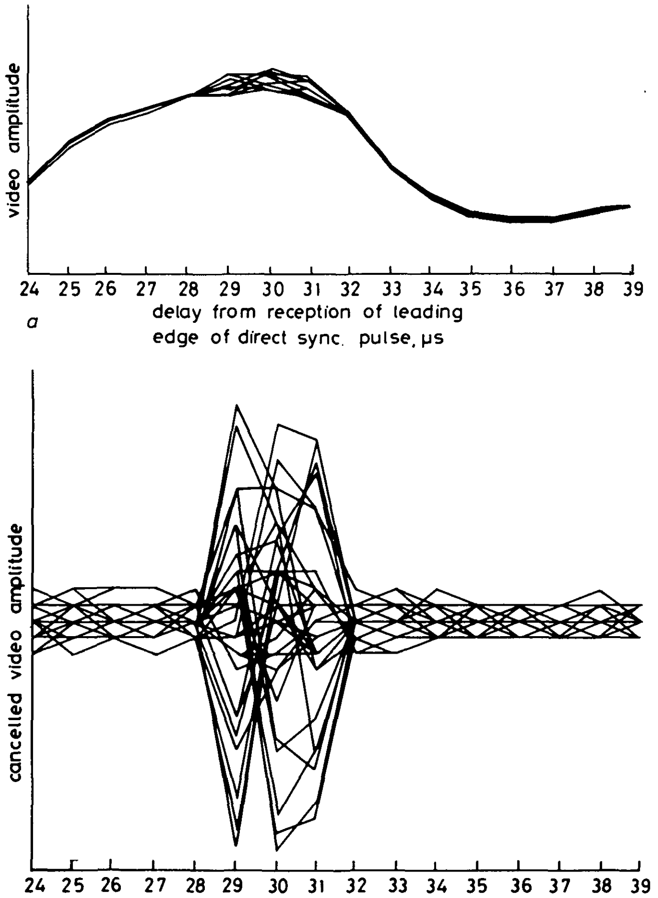


Fig. 6 *A-scan display of signal against range*
a Superimposed on the direct signal and clutter is the 'butterfly' of a moving target
b Same data as (*a*), but processed with a two-pulse MTI canceller, implemented in software. The moving target is considerably enhanced

The signal/thermal noise ratio was estimated from this plot as follows. The maximum difference between successive pulses would be twice the amplitude of one of them (assuming 180° phase change from one pulse to the next). The ratio of the maximum difference to the residual noise level (measured from the plot in other range bins) is some 16 dB; thus the single-pulse signal/noise ratio is approximately 13 dB.

Inserting this value of signal/noise ratio, plus the target range estimated from the delay, into the bistatic radar equation yields a value of approximately 34 m^2 for the target cross-section. The error in this value is estimated to be ± 3 dB, being the accumulation of errors in all the other parameters in the equation. This represents a factor of two in the target cross-section.

The output of the hardware MTI canceller was also observed. This was arranged to give an analogue output representing the difference between lines of video. A means of displaying this cancelled video on a monitor was devised. Effectively this meant just splicing in the sync signals from the direct signal, although care must be taken not to destroy any low Doppler frequencies by level clamping or AC coupling.

When a moving target is present, a vertical band of patterning should be seen. The horizontal position of the band will depend on the bistatic range of the target, and the vertical modulation rate will depend on the Doppler shift of the return. The intensity of the pattern is of course set by the video gain, but more important is the signal/noise ratio, which is set by considerations such as target size and position and antenna gain. The 'noise' floor in this case is defined by the digitisation process, which limits the dynamic range at the canceller input to 48 dB.

During the observation of this display, the experimenters can claim to have seen faint images of this sort at the times that aircraft were present. These pictures are somewhat fleeting, and it proved impossible to photograph them.

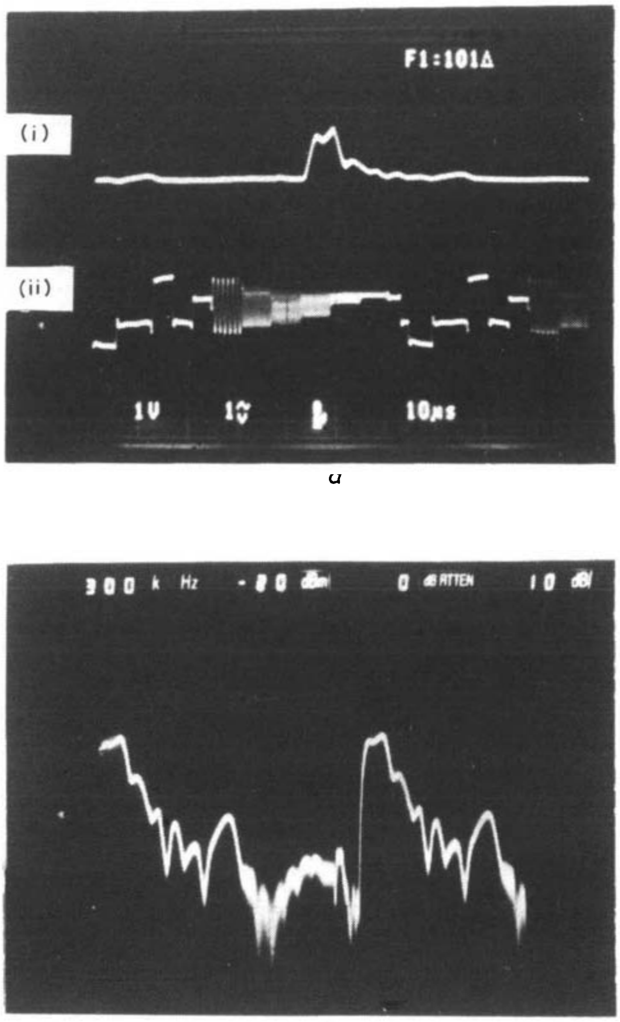
To confirm the success of this system, it is felt that further processing of the cancelled signal would be necessary, preferably with automatic target detection. Without this, digital recording of the cancelled data suffers the same 'hit and miss' problem as the software canceller. An attempt was made to record signals on a video tape recorder (VTR) for later analysis on an image processing system. This was unsuccessful, however, because of the inadequate signal/noise ratio of the recording process.

4.2 Pulsed sideband experiments

Channel 4 kindly made several transmissions of the multiburst signal described in Section 2.6. Frequency filtering of the received signal was performed by tapping off the signal just before the IF filtering and routing it to a spectrum analyser. The centre frequency and bandwidth of the analyser could be adjusted to separate out an individual burst from the multiburst. The detected result could then be displayed on the analyser screen, or the vertical output used to feed an oscilloscope or, with syncs reinserted, a monitor.

Fig. 7*a* is an A-trace from the spectrum analyser output when it has been tuned to centre on the 3 MHz burst with a bandwidth of 300 kHz. Also shown is the video trace from the reference direct receiver; the tailing off of burst amplitude is due to the limited frequency response of this receiver. Echoes from static objects can be seen trailing after the pulse; these are even more apparent on a logarithmic amplitude display, as can be achieved on the spectrum analyser screen (Fig. 7*b*).

These traces do show slight low-frequency variations. The only analysis methods available, however, were the spectrum analyser display itself or assembling the analyser vertical output into an intensity-modulated monitor display, as with the hardware MTI canceller. It has not been possible to detect any Doppler frequencies consistent with aircraft targets. Two factors could have a bearing on this. First, the energy level of an individual multiburst pulse is less than that of the sync pedestal used for previous tests. The difference is more than the 3 dB allowance for single sideband modulation, as the multiburst video excursion provides less than 100% modulation. The full allowance is 7 to 7.5 dB depending on what assumptions are made about the harmonic content of the multiburst signal. This is the theoretical minimum difference between the average power levels in the sync pedestal and a multiburst pulse. In practice the difference may be more, and it may be that in this experiment the moving target returns are below the noise threshold of the system.



a

b

Fig. 7 *Multiburst signal on different displays*
Horizontal scale: 10 μs/div
a Linear display
(i) filtered echo signal
(ii) direct signal
b Logarithmic display

Nevertheless, the A-traces have shown clearer pictures of large static targets than the sync-plus-white transmissions, demonstrating that this technique is useful as a means of introducing radar-like features into a TV transmission.

5 Discussion and conclusions

5.1 Discussion

The experimental results proved negative for the most part, and it is necessary to account for this. Part of the explanation is due to the difficulties in capturing adequately long data records, but mostly it is because the dynamic range of the MTI cancellation system used (48 dB) is inadequate to cope with the high clutter/signal ratios of the quasi-CW radar system, which may be as much as 40 dB higher than this value. The positive results obtained occurred on the few occasions when the target aspect gave a high value of bistatic cross-section, the receive antenna direction coincided with that of the target and the data were satisfactorily captured.

It might be suggested that the effects of passing aircraft on television reception are actually quite common; any TV watcher will claim to have seen the very effects we have searched for in his own home in the form of 'aircraft flutter'. Aircraft flutter, though, is not quite the same thing as Doppler-shifted target returns. It involves a low-frequency beating between the direct signal and the one reflected off the aircraft. It appears as a simple fluctuation in received signal strength. For this to happen a number of criteria must be simultaneously met: (i) the delay of the reflected signal must be small compared with the line length and picture features, so that similar signals are beating together; (ii) the rate of change of the delay must be low, so that the effect is not masked by the persistence of screen phosphor and viewer's vision; (iii) the reflected signal strength must be a significant fraction of the direct; and (iv) the whole effect must last long enough to be noticed.

All these conditions are met when an aircraft crosses the bistatic baseline close to either the transmitter or receiver. Exactly on the baseline, the bistatic range is zero and the Doppler shift is zero whatever the direction and speed of motion. At such grazing incidence angles (forward scatter), the bistatic cross-section also assumes rather higher values than usual [9]. By going close to one end of the baseline the $r_1 r_2$ product can be made low, giving strong reflected signals, and if the track crosses the line obliquely then even high-speed targets pass slowly through the zero Doppler null. Aircraft flutter is, therefore, a short-range phenomenon and somewhat different from the effects sought here.

There is also the possibility that flutter or fluctuation can occur on TV pictures without aircraft being involved. For a reflecting object well away from the baseline, the movement required to give a 180° phase shift in the received echo is only a quarter-wavelength, about 13 cm. To give 1 Hz Doppler frequency a speed of 0.25 m/s is all that is required. These figures are not inconsistent with the movements expected of high buildings, and even of the Crystal Palace tower itself, when the wind blows. Such low rate variations have been observed on echo signal A-traces in the absence of visible aircraft, and this may be the explanation.

5.2 Conclusions

The bistatic radar function is dictated by the nature of the illuminating signal, in terms of frequency, modulation bandwidth, transmit power and antenna directionality. The simplest case is when a pulsed radar transmitter is available as the illuminator. Here the receiver signal processing will be similar to that of the illuminating radar, and the expected system performance can be readily calculated from standard expressions.

The situation when the illuminator is not radar-like is more difficult. Of key importance is the autocorrelation function (and hence ambiguity function) of the transmitted waveform. This will determine:

(*a*) The achievable range resolution, given approximately by c/B, where B is the modulation bandwidth.

(*b*) The spacing of range ambiguities; in a practical system, range ambiguities may be tolerable if they can be resolved by triangulation from multiple receiver sites.

(*c*) The range sidelobe level; if this is high, targets may be obscured and the detection performance of the radar will be poor.

(*d*) Similarly, the achievable Doppler resolution, Doppler ambiguity spacing and Doppler sidelobe level.

We can draw the following conclusions about desirable properties of illuminators of opportunity:

(i) The transmit power should be commensurate with the coverage required. In the case of a complex modulation function such as that of a television, the calculation must be made on the basis of the power of that part of the modulation spectrum used for radar purposes.

(ii) The radiation pattern of the illuminator should be either omnidirectional (floodlight coverage) or a single pencil beam.

(iii) The modulation bandwidth of the source should be commensurate with the required range resolution. The ambiguity function of the signal should approximate the 'thumbtack' ideal (i.e. high resolution without ambiguity or sidelobes in both range and Doppler co-ordinates).

In most of these respects the television waveform is not ideal. The autocorrelation function for a TV waveform shows broad peaks at 64 μs intervals corresponding to the line sync pulses; if the sync pulses are gated out, the autocorrelation function peaks will be sharper (depending on the picture content), but will still recur at 64 μs intervals. In the case of any quasi-CW illuminator, the range sidelobe level will be high, and the radar best suited for Doppler rather than range measurements. The television transmit power is high and the azimuth coverage omnidirectional, but the elevation plane coverage is deliberately restricted.

The signal processing in the receiver is also dictated by the nature of the illuminating waveform. It is necessary either to know explicitly the form of this signal, or to receive separately an undistorted direct version of it. In the case of an unpredictable modulation scheme, such as *ad hoc* TV picture material or other broadcast signals, direct signal reception is mandatory.

In the case of quasi-CW waveforms, such as television transmissions, performance is severely limited by the unfavourable target/clutter ratio given by eqn. 6. This ratio is improved if a higher receive antenna gain is employed, both because the target signal level is higher and because the clutter contribution is reduced by the narrower antenna beamwidth.

We may make the following observations about the necessary signal processing:

(*a*) The dynamic range must be adequate to cope with the full range of direct and target echo signals.

(*b*) Real-time crosscorrelation with adequate dynamic range represents a substantial processing burden.

(*c*) As with bistatic radar using a dedicated transmitter, a static set of contiguous receive beams or a small number of electronically agile beams will be necessary if wide sector coverage is desired.

In spite of the problems encountered, bistatic radar based on illuminators of opportunity has substantial attractions. While television transmissions are in several ways not ideal for this purpose, and require substantial processing to extract target echoes, a system of adequate dynamic range using real-time crosscorrelation would represent an intriguing prospect.

6 Acknowledgments

The authors would like to thank the BBC and IBA for their co-operation in providing out-of-hours test transmission, in particular Mr. Eric Picketts and Dr. Paul Ratliffe of the BBC, and Mr. Mike Sage of the IBA. Thanks are also due to Mr. Tom Eustace and Mr. Mike Holloway for their assistance during the experimental programme, and to Mr. Ira Clarke, Professor D.E.N. Davies and Dr. Ken Milne for valuable discussions.

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