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Project AY4490- A study into techniques for improving radar spectrum utilisation

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Executive summary

This report has been prepared for Ofcom (formally Radiocommunications Agency) under study contract AY4490 by QinetiQ with contributions from its partners University College London, University of Bath and CCLRC Rutherford Appleton Laboratory through the 2003-04 Spectrum Efficiency Scheme (SES) initiative.

Project scope

The primary requirement is to investigate methods that reduce interference between radars and other services and considers the practical and theoretical possibilities of sharing radar spectrum with other radio services.

Civil radar in the frequency range 1-16GHz has been considered. In this frequency range radar tend to be pulsed systems which can interfere with communication systems, especially digital systems, which are susceptible to interference from pulsed radar and from interference from radar assigned to operate in adjacent bands. For these reasons pulsed radar systems generally have exclusive allocations, whereas conventional communications services share a frequency band, since in-band interference can often be avoided.

By applying interference reduction techniques to radar systems there may be the opportunity to release spectra or allow radar and communication systems to co-exist. Ofcom has recognised the potential of releasing current radar spectrum allocations to other communication services in the frequency bands 1-16GHz, after receiving the findings of the Ofcom sponsored radar study, AY4501[1].

The frequency bands in the range 1-16GHz are currently used for a variety of radar applications that include air traffic control, marine navigation, coastal surveillance and weather observation. These frequency bands have desirable propagation properties and could be used to support other communication services, such as cellular telephony, messaging and wireless local area networks. Making more efficient use of the radio spectrum is a key requirement for Ofcom and technologies that can promote this philosophy are currently being supported through the SES initiative.

Three techniques that have the potential to improve radar spectrum utilisation have been investigated in this study:

1. Rapidly tunable filters.
2. Ultra narrow band radar.
3. Waveform design.

In order to assess the amount of spectrum that could be released, each technique was applied to three exemplar radar bands within the 1-16GHz frequency range. A simplified cost benefit analysis (CBA) was produced to provide a comparison of the level of benefits between each technology.

Key project conclusions

1. It has been shown that each technology investigated has the potential to reduce the current occupied bandwidth of the radar systems in all of the exemplar bands analysed.

[1] Ofcom radar study project AY4051: "Investigation into the characteristics, operation and protection requirements of civil aeronautical and maritime radar systems".

2. There is a potential benefit to be gained by releasing spectra to other services. This amount of benefit is determined by:
   (i) The choice and application of technology that is implemented to improve spectrum utilisation.
   (ii) The service that would take advantage of any spectral release.
3. In line with current thinking, some of the existing regulations have been shown to be flawed. In addition, the regulations were found not to significantly prejudice waveform with theoretically noisy spectra, and in some cases perform the opposite. This is summarised by:
   - The practice of adjusting waveform characteristics to give increased allowed bandwidths, which is a common practice for radar manufactures to achieve type approval.
   - Allowing spurious signals levels to be relative to peak envelope power rather than absolute levels.
4. The filter and waveform solutions could be implemented as either a retrofit to existing radar systems or incorporated in a new radar design. The cost benefit analysis has shown that the cost to introduce a new design is an order of magnitude greater than retrofitting to existing radar.
5. Retrofitting to existing radar is considered a near term solution and should allow implementation within 5 years. New radar systems, of which the ultra narrow band solution is one, are considered medium term solutions and are not likely to be implemented for 10 to 15 years.
6. The waveform solution is the best near term solution offering an improvement in spectrum utilisation across all frequencies. Filtering techniques are limited as it becomes more difficult as frequency increases since the power handling capabilities of waveguide filters becomes an issue. Generating appropriate high powered waveforms from power devices at acceptable cost may also be difficult. Hence a combination of filters and waveforms, optimised for a particular frequency and application, is probably the best solution for exiting radar systems.
7. Considering the low-cost nature of marine radar for smaller pleasure craft, the X-band magnetron tube transmitter is likely to remain the chosen output device for the near future. The cost of solid-state output amplifiers is likely to remain higher than magnetrons in the near term. For modest output powers, filtering offers an effective solution to emission control in this band.
8. Improvements to magnetron modulator design will ensure that the magnetron is kept in the correct mode throughout its life and would significantly improve the spurious emissions by using, currently available, constant-current modulation and/or a correct modulation techniques.

Recommendations for future work

The cost benefit analysis has concluded that there is benefit to be gained by implementing any one of the technologies in the exemplar bands. The cost benefit analysis carried out was not a rigorous exercise, but merely an indicator to compare different technology options.

Recommendation 1: It is recommended that a more rigorous cost benefit analysis be carried out to fully investigate the true cost and benefits of the proposals.

Recommendation 2: In parallel with recommendation 1, it is recommended that a detailed analysis of all radar bands within the 1-16GHz frequency range be carried out to determine which are the most appropriate bands for the implementation of these techniques, given the proposed timescales for rollout of the spectrum trading initiative.

Although military systems have not been considered in any great detail, it is clear that they are embedded within the whole of the 1-16GHz frequency range. Military systems, for issues of National security, could not be treated in the same manner as civilian radar systems.

Recommendation 3: In order to understand the true benefit and practicalities of the proposals set out in this report, it is recommended that further dialogue with MoD Defence Spectrum Management be conducted. It is recommended that a dialogue with MoD be established to investigate the possibility of military systems adopting some of the proposed techniques in order to assist the process of improving spectrum utilisation.

Recommendation 4: It is recommended that the waveform and filter investigations be taken forward to demonstration on candidate radar systems to prove the principals outlined in this report and to show the user community the benefit of introducing such technology.

Recommendation 5: The ultra narrow band solution has the greatest potential to improve spectrum utilisation in the long term. However, it is an immature technology therefore further investigation is recommended to help understand the advantages and constrains of this solution.

Recommendation 6: In order to improve the spurious emissions of X-band marine radar it is recommended that further investigation be carried out to demonstrated the potential improvement of implementing an optimum modulator and filter design.

List of contents

Glossary

ASDE	Airport Surveillance and Detection Equipment
ATC	Air Traffic Control
BPSK	Binary Phase Code Shift Keying
BWE	Bandwidth Extrapolation
CAA	Civil Aviation Authority
CBA	Cost Benefit Analysis
CCLRC	Council for the Central Laboratory for the Research Councils
CFA	Cross Field Amplifier
CW	Continuous Wave
DOA	Direction Of Arrival
DSP	Digital Signal Processor
ETSI	European Telecommunications Standards Institute
FBAR	Firm Bulk Acoustic resonator
FD	Frequency Domain
FDC	Frequency Domain Concatenation
FET	Field Effect Transmitter
FSK	Frequency Shift keying
FSS	Frequency Selective Surface
FWA	Fixed Wireless access
GUI	Graphical User Interface
HRR	High Range Resolution
ICAO	International Civil Aviation Organisation
IEC	International Electrotechnical Commission
IMO	International Maritime Organisation
INR	Interference to Noise ratio
ITU	International Telecommunications Union
LFM	Linear Frequency Modulation
LP	Linear Prediction
LPI	Low Probability of Intercept
LPF	Low Pass Filter
MCA	Maritime & Coastguard Agency
MEMS	Micro Electro Mechanical Systems
MF	Matched Filter
MIMO	Multiple input Multiple Output
MoD	Ministry of Defence

MLS	Maximum length sequence
MSK	Minimum Shift keying
NATS	National Air Traffic Services
NM	Nautical Miles
NLFM	Non Linear Frequency Modulation
OOB	Out Of Band
PC	Pulse Compression
PRI	Pulse Repetition Interval
PRF	Pulse Repetition Frequency
PSR	Primary Surveillance Radar
PSF	Point Speed Function
PSR	Peak Sidelobe Ratio
RAL	Rutherford Appleton Laboratory
RCS	Radar Cross section
SARTS	Search & Rescue Transponders
SC	Stepped Chirp
SES	Spectrum Efficiency Scheme
SF	Stepped Frequency
SFPC	Stepped Frequency Polyphase Code
SNR	Signal to Noise Ratio
SP	Stretch Processing
SSR	Secondary Surveillance Radar
SSSC	Single Sample Stepped Chirp
TB	Time Bandwidth
TD	Time Domain
TFA	Time Frequency Analysis
TQ	Taylor Quadriphase
TWT	Travelling Wave Tube
UCL	University College London
UWB	Ultra Wide Band
VTS	Vessel Traffic Surveillance
WLAN	Wireless Local Area Network
WP	Work Package
WT	Wireless Telegraphy

1 Introduction

This report has been produced under Ofcom (formally Radiocommunications Agency) study contract AY4490 by QinetiQ with contributions from its partners; the University College London (UCL), University of Bath and CCLRC Rutherford Appleton Laboratory (RAL) under the Spectrum Efficiency Scheme (SES) initiative for 2003-04.

The Spectrum Efficiency Scheme (SES) involves research in support of Ofcom's duty to ensure the optimal use of the electromagnetic spectrum and further Ofcom's aim to remain at the forefront of technological understanding. It also aims to foster the early sharing of information between members of the radio community.

The 2003/04 SES was the inaugural year of the scheme and a range of technologies were selected for research to support the initiative, these include:

- Frequency-selective surfaces (FSS)
- Wireless local area networks (WLAN)
- Fixed wireless access (FWA)
- Point-to-point links
- Radar
- Antennas
- Multiple-input multiple-output (MIMO) antenna

Each project was carried out in collaboration with industry, academia and government agencies ensuring technology transfer and dissemination of information to all sectors of the radio community.

To provide Ofcom assurance of the value of the research there is encouragement to demonstrate the principles being studied and to provide and an estimate of the likely cost and benefit that could be released by adopting the proposals.

1.1 Background

Frequency bands are allocated for terrestrial radionavigation in all parts of the radio spectrum from 9kHz, up to 265GHz. The radionavigation service is a subdivision of the radiodetermination service. Radionavigation is defined as [1]:

"The determination of the position, velocity and /or characteristics of an object, or the obtaining of information relating to these parameters, by means of propagation properties of radio waves."

Radionavigation services are divided into two categories: marine and aeronautical. As both services have safety implications they are given a high degree of protection from interference. For this reason allocated bands are primary and many are world-wide and exclusive.

The systems for these services fall into two frequency categories:

- Below about 500MHz where transmissions are from one or several fixed stations transmitting continuous wave.
- Above about 500MHz where a typical system is pulsed radar.

The occupied bandwidth of a pulsed radar is wide, typically tens of Megahertz. Most radars, using no frequency determining device other than a cavity magnetron, have operating frequencies that, compared with other kinds of radio systems, are ill-defined and of poor stability.

Pulsed radars sometimes interfere with communication systems, especially digital systems, which are susceptible to interference from pulsed radar and from interference from radar assigned to operate in an adjacent bands. For these reasons pulsed radar systems generally have exclusive allocations, whereas conventional communications devices share a frequency band, since in-band interference can often be avoided.

By applying interference reduction techniques to radar systems there may be the opportunity to release spectra or allow radar and communication systems to co-exist. Ofcom has recognised the potential of releasing current radar spectrum allocations to other communication services in the frequency bands from 1-16GHz, after receiving the findings of the Ofcom sponsored radar study, AY4051 [2].

The frequency bands in the range 1-16GHz are currently used for a variety of radar applications that include air traffic control (ATC), marine navigation and weather observation. These frequency bands have desirable propagation properties and could be used to support other communication services, such as cellular telephony, messaging and wireless local area networks. The sum total of all current radar allocations within this frequency range amounts to more than 1GHz of bandwidth.

The findings of AY4051 [2] have encouraged Ofcom to make further investigations into the possibility of civil radar operators sharing their spectrum allocations with other communication services.

If this principal can be realised and the civil radar services can maintain performance to the required specification without “unwanted” interference from other services operating within the same band, spectrum allocations for “new” services can be created.

Once this is successfully achieved it will assist Ofcom to meet its mission statement, which is to be a World leader in spectrum efficiency management.

1.2 Requirement

This report is the output to contract AY4490: “A study into techniques for improving radar spectrum utilisation”, issued by Ofcom in June 2003.

The primary requirement is to investigate methods that reduce interference between radars and other services and considers the possibility of sharing spectrum without affecting a radars operational need. The work conducted complements the studies carried out in report AY4051 [2].

A number of study topics were originally suggested and are covered by the following three requirements:

1. Techniques which reduce interference from radars to other systems.
2. Techniques which reduce the interference from other systems to radars.
3. Techniques which simultaneously reduce interference to/from radar from/to other radio systems.

A response to the requirement is discussed in §2.1.

The output covers UK civil radar in the frequency band 1-16GHz and has taken into account military and future allocations, where appropriate.

Consideration has been given to the R&TTE (Radio & Telecommunications Terminal Equipment) Directive, which was introduced in 2002. The directive has been introduced to provide greater flexibility and opportunity at the price of conformity to products. Manufacturers will no longer be able to rely on the type approval scheme but they must carry the consequence of full product liability. The Directive aims to provide the European Radio and TTE industry with a more deregulated environment than at present. The involvement of third parties in conformity assessment is not necessary in most cases. The person who places equipment on the market will, in general, be regarded as taking full responsibility for its conformity to essential requirements, and for properly informing users of its intended use. Only in the case of radio equipment for which harmonised standards are not available, or are not used, is it mandatory to consult a notified body.

For certain applications there are specific minimum performance requirements which must be met. In these cases the study has taken into consideration the requirements set out by ITU-R, IMO, ICAO, EUROCONTROL, IEC, ETSI, CAA/NATS and users of radars, proposed changes in radars that may affect search and rescue transponders (SARTS), radar equipment manufacturers and other bodies concerned with the regulation of radar systems.

1.3 Project structure

The project team was a consortium of specialists drawn from the UK radar industry and academia, who can assist Ofcom to formulate a strategy for the improving spectrum utilisation in the frequency band 1-16GHz. The structure of the project organisation is given in Figure 1-1.

The members of the consortium are QinetiQ, University College London (UCL), University of Bath and CCLRC Rutherford-Appleton Laboratories (RAL). Each member of the consortium provided contributions to at least one of the principal tasks in the work programme. Each member of the consortium has representation on the “Project Steering Group” which was established to provide guidance and recommendations to both the project team and Ofcom.

Although not part of the consortium, the radar industry had an important role to play by providing advice on the practicality and cost implication of the technical proposals. Their views have been taken into consideration during the project through consultation and review meetings.

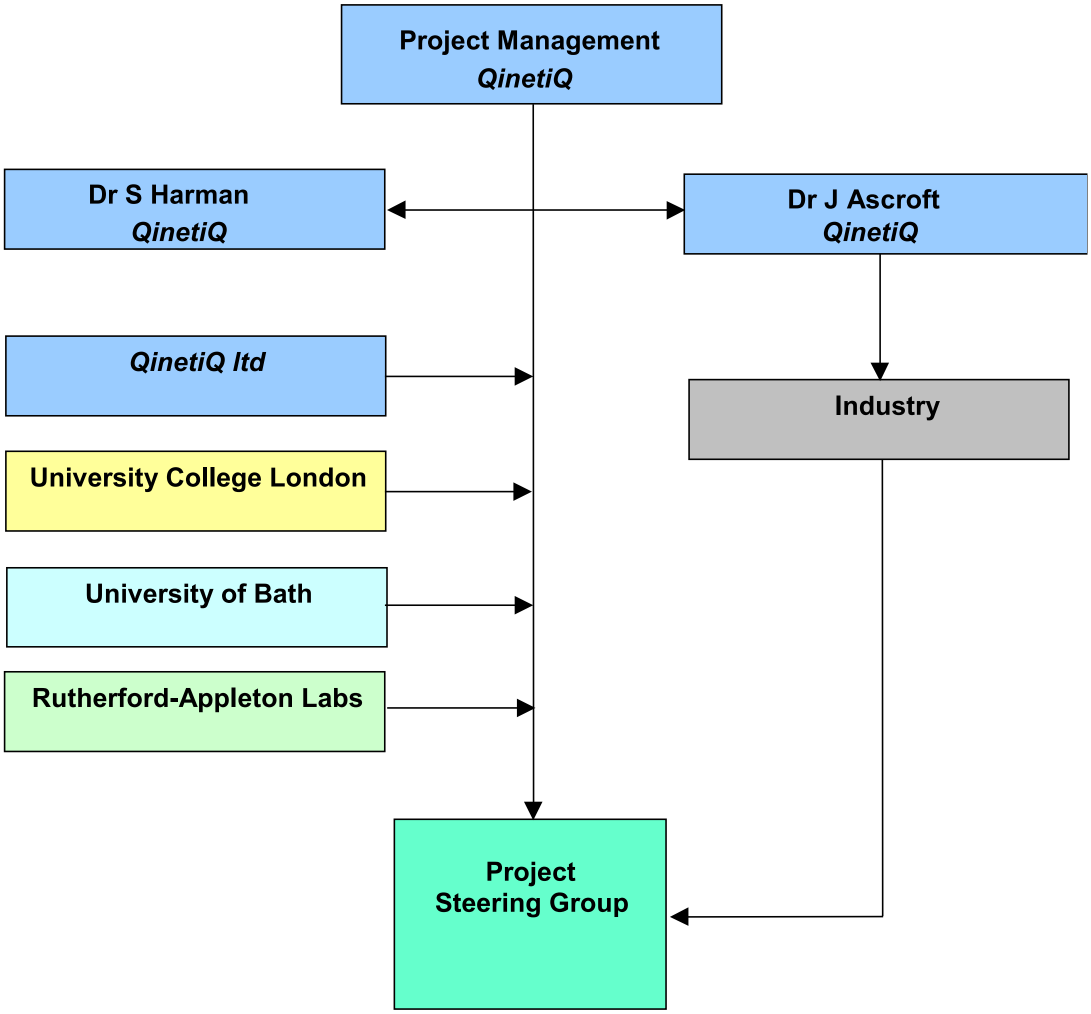


Figure 1-1 Project organisation

1.4 Aims and objectives

The aim of the study was to investigate the practical and theoretical possibilities of sharing radar spectrum with other radio services, together with methods for reduction in the use of bandwidth used by UK civil radar in the frequency band 1-16GHz.

The output demonstrated the ability to free up spectra using the appropriate technology and identify the associated costs and benefits. In collaboration with the radar industry the project identified the practicality issues of introducing such technologies and therefore helped to form a strategy for taking any credible options forward.

1.5 Document structure

The main body of the report contains the conclusions and recommendations from the technical work packages as well as a cost benefit analysis (CBA) and comments from the radar industry. The detailed technical work packages are contained within the Appendices of the report. Below is a breakdown of the document structure:

§2 – The approach to the project including an investigation into the UK civil radar spectrum between 1-16GHz and the technologies investigated. This section also presents the key outputs from each of the individual technical work package.

§3 – The cost benefit analysis (CBA).

§4 & 5 – The key conclusions and recommendations together with comments from the radar industry.

Appendix A – The technical output from WP2: Rapidly tuneable filters.

Appendix B – A report on MEMS technology for filter applications in support of WP2.

Appendix C – The technical output from WP4: Ultra narrow band radar.

Appendix D – The technical output from WP6: Waveform design.

Appendix E – A report on the demonstrations in support of WP6.

Appendix F – A list of radar industry who participated in the project.

Appendix G – A literature search of filter technology.

2 Spectrum efficiency techniques investigated

2.1 Introduction

The technical approach taken by the project is highlighted in Figure 2-1 where the aim of the study is to investigate the practical and theoretical possibilities of sharing radar spectrum with other radio services, together with methods for reduction in the use of bandwidth used by civil radar.

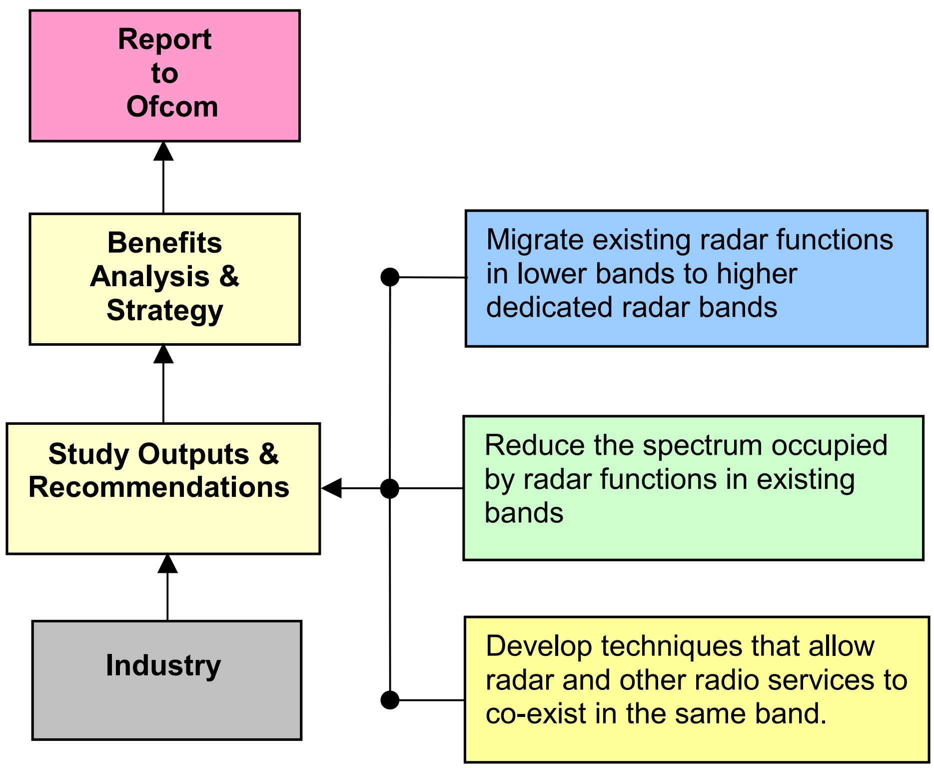


Figure 2-1 Technical approach to AY4490

In response to the original topics suggested in the ITT i.e.:

1. Techniques which reduce interference from radars to other systems.
2. Techniques which reduce the interference from other systems to radars.
3. Techniques which simultaneously reduce interference to/from radar from/to other radio systems.

Three principal approaches were identified for making available more “spectrum” for other radio services by reducing the spectrum required by civil radar:

1. Migrate existing radar functions in the lower frequency bands to the higher dedicated radar frequency bands.
2. Reduce the spectrum occupied by radar functions in their existing bands.

3. Develop techniques that allow radar and other radio services to co-exist in the same band.

From these subject areas several technical work packages were identified to support the project.

The key technical measure of success for any of the approaches is the prevention or mitigation of interference between radars operating within the same band, by radars on other systems or by other systems on radars. In addition any new techniques must not compromise the required radar performance and must be capable of cost effective implementation.

The approach to the project was to identify, develop, and where appropriate demonstrate, techniques for bandwidth reduction and interference suppression, assess the practicality (technically and commercially) for implementation and finally suggest a strategy to Ofcom for their implementation. The practicality issues were addressed through the involvement of the radar industry. The purpose of involving the radar industry was to ensure that early visibility of the proposals is more likely to provide a smoother path to implementation. The involvement of the radar industry was through a review of the technical proposals and participation in a series of workshops.

Ofcom selected three technical work packages (WP) for investigation in this contract. In addition to the technical work package a cost benefit analysis and strategy was required. The responsibilities for delivery of each work packages were:

- WP2 – Rapidly tunable filters: University of Bath.
- WP4 – Ultra narrow band radar: University College London.
- WP6 – Waveform design: QinetiQ.
- WP10 – Cost benefit analysis: QinetiQ.

Rutherford Appleton Laboratory (RAL) was responsible for delivering the output of a series of demonstrations to help support the techniques developed in WP6.

Below is a brief description of each technical work package:

2.1.1 WP2 – Rapidly tunable filters

Many simple low cost RF sources, such as magnetrons, generate frequency components (sidebands and spurious emissions) which cover a wide frequency band than required for a range resolution. A recently published report for Ofcom AY4051 [2] has considered the use of high-Q filters for the control of the transmitted radar spectrum.

In line with the recommendations of AY4051 [2], this study has considered further the technology options to provide high-Q, rapidly re-tunable filters for use in tube based frequency hopping radars, and also high-Q filters for use in active array radars. This task has also considered the use of frequency selective antenna structures for bandwidth filtering.

2.1.2 WP4 – Ultra narrow band radar

The ultimate in bandwidth reduction would be to move to essentially CW operation, which means in practice a bandwidth of perhaps a few Hz. However, such CW radars suffer from range ambiguity and leakage of the transmitter signal directly into the receiver, which can limit the sensitivity. These shortcomings can be overcome by the use of a multistatic system, i.e. multiple receiver sites. Multiple receiver sites allow target location to be determined via a combination of interferometry and triangulation.

Current installations could also provide much of the necessary hardware and infrastructure, for example the dish antenna and turning gear. Narrow band radars are inherently low cost and exhibit superior noise performance as a result of operating over a narrow bandwidth.

The task has determined the suitability of narrow band distributed radar as a replacement for current monostatic radars, particularly in S-band and C-band. Performance has been benchmarked against that of existing radar systems, including those used for air traffic control, weather observation, coastal surveillance and marine navigation. As it is possible that more than one frequency of operation may be required to meet the demands of some applications, this work has included an in depth analysis of the fundamental aspects of the concept, the engineering implications and an initial cost estimate.

2.1.3 WP6 – Waveform design

The detection range of radar is dependent on the total energy contained in the transmitted waveform, irrespective of the structure of the waveform and assuming a matched filter receiver. Simple pulsed radars use a short pulse, of width commensurate with the required range resolution, typically 100ns, and a peak pulse power high enough to achieve the required range.

Alternatively the same transmit energy, and hence range, can be achieved by using a longer, lower power waveform. The range resolution can be maintained by coding the long waveform at a rate commensurate with the required range resolution. The most common coding schemes use frequency or phase change, but more complex schemes utilising combinations of frequency, phase and amplitude modulation have been devised.

A receiver matched to the transmit waveform provides coherent integration of the energy in the waveform into a range interval determined by the code "chip" length. Codes not matched to the receiver will not achieve coherent integration. In addition to the potential for rejection of interfering signals, low peak power coded waveforms are likely to produce less interference in other systems. The development of software controlled waveform generation and receivers is a key enabling technology.

This task has exploited advances in military systems for low probability of intercept (LPI) and communications systems e.g. CDMA. Key issues in the design of the waveforms include range resolution, unambiguous range and range sidelobes, occupied bandwidth, code length and sensitivity to nature of the interfering signals. Consideration has also been given to adaptive control of the radar transmit power as a means of reducing interference to other users. This task has also assessed the potential of coded, spread-spectrum waveforms for interference suppression.

2.2 Investigation into UK civil radar in the frequency range 1-16GHz

In order to undertake an analysis of the technologies being investigated it was necessary to develop an understanding of both the UK radio frequency allocation in the frequency range 1-16GHz and the types of civil radar that use particular parts of that spectrum.

2.2.1 Understanding the UK radio frequency allocation

The UK radio frequency is a shared resource between the civil and the military communities. Although this project is an investigation into UK civil radar, it would reduce the value of this exercise if military allocation frequencies were completely ignored. It is the aim of this project to analyse the appropriate frequency bands and identify, where possible, military users. Having identified military users then a strategy can be formed to either work around existing military frequencies or to engage with the MoD in an attempt to release some of the spectrum they occupy in the relevant bands.

The civil UK radio frequency allocations for 2002 [3] are readily available from the Ofcom website and detail the following information:

- The primary services within a band.
- The secondary services within a band.
- ITU-R footnotes.
- Specific comments relating to allocations in each band.

In order to visualise the civil UK radio frequency spectrum in the range 1-16GHz, a diagrammatic representation of the allocations has been produced. Figure 2-2 is taken from data in the UK frequency allocation table 2002 [3] for the 1-16GHz only. At the time the diagram was produced it was assumed that all allocations were correct. The diagram does not indicate any military allocation but does distinguish between “Primary” and “Secondary” allocations, where “Secondary” is denoted by a “s” in the band. However, there are some special cases where specific users in “Secondary” allocation bands have the authorisation to operate on a basis of equality with other service in that band i.e. “Primary” status. The diagram only represents allocations defined in 2002 and there are a number of allocations which have been, or are in the process of being, defined. One such allocation is Galileo [4], which will be allocated within the 1215-1350MHz band. Where possible these anticipated allocations have been considered in the overall analysis.

Figure 2-2 shows all radio allocations. In order to identify areas within the spectrum where radar is allocated a further diagram has been produced. Figure 2-3 indicates where, within the 1-16GHz band, there are radar allocations. It is clear that radar is spread across all the 1-16GHz frequency range, although in almost all bands there are frequency bounds within which systems must operate as there are allocations to other service providers be it primary or secondary allocations – these bands are not exclusive to radar.

It is important to make the distinction between frequency assignment and frequency allocation. In some cases, even though agreement has been reached at International level for the use of these frequencies for radar, they have not been assigned by Ofcom for radar in a geographical area. In these cases, the secondary allocations may be the primarily use for these frequencies.

It is from these identified radar bands that examples were selected. This allowed the principle of improving spectrum utilisation using the technologies identified in §2.1 to be demonstrated.

Fig. 2-2

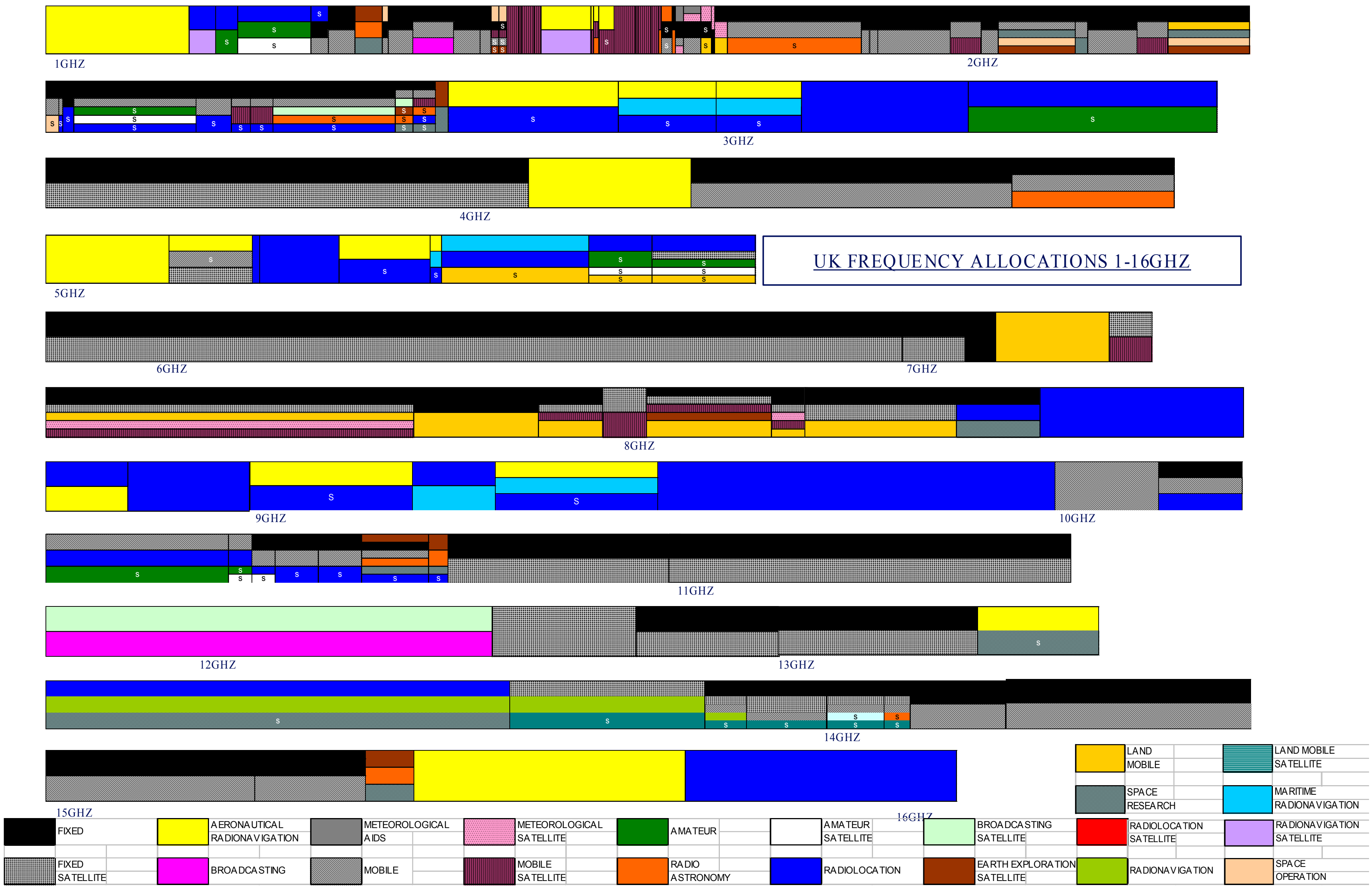
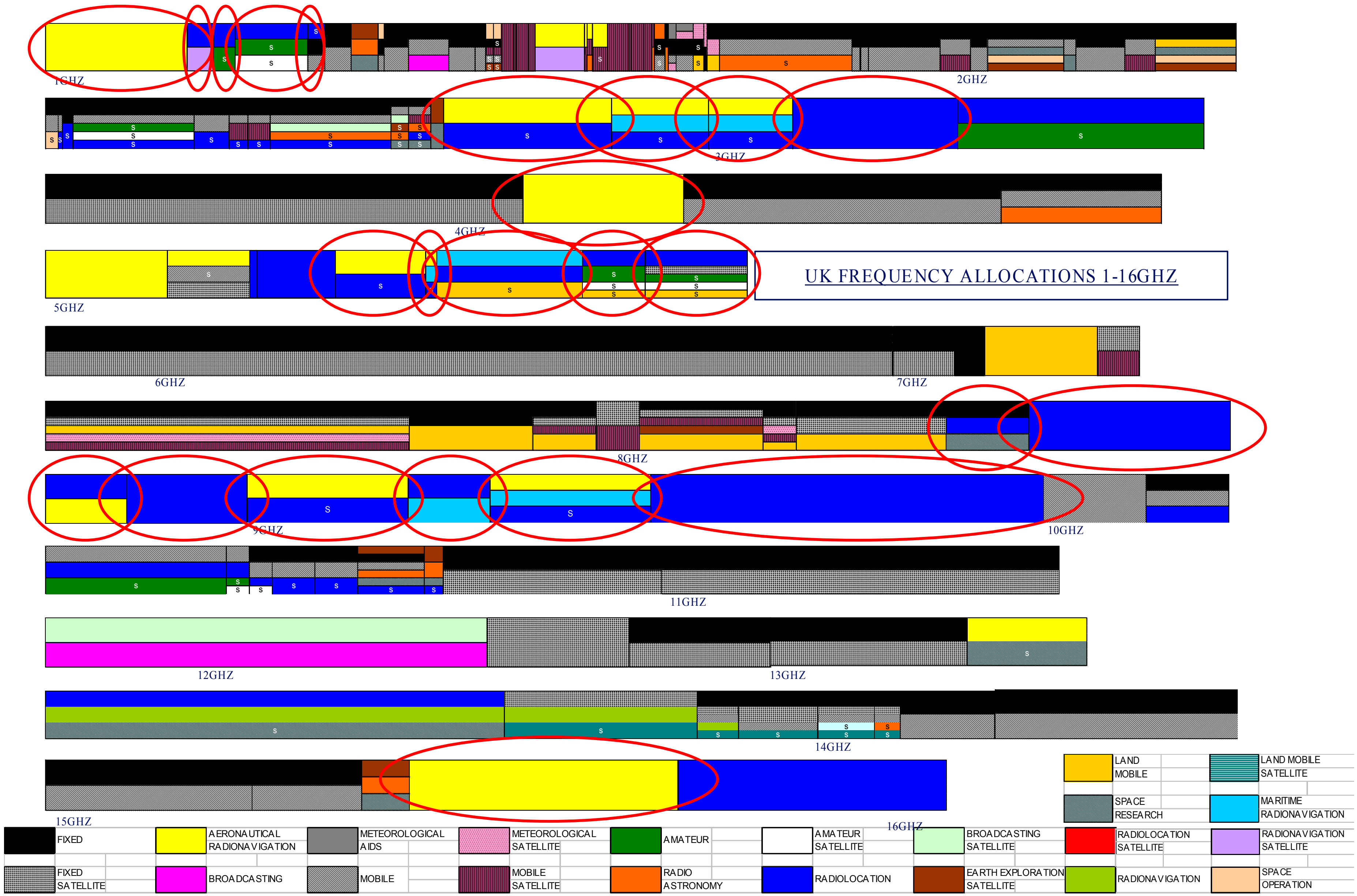

UK FREQUENCY ALLOCATIONS 1-16GHZ
1GHZ
2GHZ
3GHZ
4GHZ
5GHZ
6GHZ
7GHZ
8GHZ
9GHZ
10GHZ
11GHZ
12GHZ
13GHZ
14GHZ
15GHZ
16GHZ
LAND MOBILE
LAND MOBILE SATELLITE
SPACE RESEARCH
MARITIME RADIONAVIGATION
FIXED
AERONAUTICAL RADIONAVIGATION
METEOROLOGICAL AIDS
METEOROLOGICAL SATELLITE
AMATEUR
AMATEUR SATELLITE
BROADCASTING SATELLITE
RADIOLOCATION SATELLITE
RADIONAVIGATION SATELLITE
FIXED SATELLITE
BROADCASTING
MOBILE
MOBILE SATELLITE
RADIO ASTRONOMY
RADIOLOCATION
EARTH EXPLORATION SATELLITE
RADIONAVIGATION
SPACE OPERATION

Fig 2-3


1GHZ
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UK FREQUENCY ALLOCATIONS 1-16GHZ
6GHZ
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11GHZ
12GHZ
13GHZ
14GHZ
15GHZ
16GHZ
FIXED
FIXED SATELLITE
AERONAUTICAL RADIONAVIGATION
BROADCASTING
METEOROLOGICAL AIDS
MOBILE
METEOROLOGICAL SATELLITE
MOBILE SATELLITE
AMATEUR
RADIO ASTRONOMY
AMATEUR SATELLITE
RADIOLOCATION
BROADCASTING SATELLITE
EARTH EXPLORATION SATELLITE
LAND MOBILE
SPACE RESEARCH
RADIOLOCATION SATELLITE
RADIONAVIGATION
LAND MOBILE SATELLITE
MARITIME RADIONAVIGATION
RADIONAVIGATION SATELLITE
SPACE OPERATION

2.2.2 Radars in the 1-16GHz frequency band

To determine whether two radar systems interfere with each other then information such as:

- Radar transmit power
- Bandwidth
- Antenna gain
- Sensitivity
- Geographic location
- Antenna separation
- Terrain
- Propagation data

is required. For a network of radar, such as ATC, then information relating to all systems and locations is also required.

To provide the project with the relevant information a database of radar types and radar parameters has been established which cover:

- Air traffic control – approach radar
- Air traffic control – en-route radar
- Air traffic control – airport ground movement radar
- Secondary surveillance radar
- Weather radar
- Coastal radar
- Maritime radar
- Altimeters radar
- General surveillance and tracking radar

The database was complied from information gathered through discussions with the radar community and from published data. The number of fixed and mobile radars operating in the UK was taken from AY4051 [2] and is detailed in Tables 2-1 and 2-2:

Radar Type	**Band**	**Number**
SSR: secondary surveillance	L	31
ATC: en-route	L	11
ATC: approach	S	36 + 84(1)
ATC: airfield control	X	9
ASDE: airport surface detection	X	5
ASDE: airport surface detection	Ku	3
Weather	C	13
VTS: vessel traffic system	S	2
VTS: vessel traffic system	X	11
VTS: vessel traffic system	Unknown	13

Table 2-1 Fixed civil radars operating in the UK

(1) S-band ATC radar in the UK is divided into 36 civilian and 84 MoD.

Radar Type	Band	Number
Airborne weather	X	700 at any time
Altimeters	C	700 per day
Marine - Pleasure	X	50-60,000
SOLAS vessels	X & S	1200 per day
Marine - Fishing	X	6625

Table 2-2 Summary of mobile radars operating in the UK

2.3 Summary of the outputs of the technical work packages

A detailed report of the work undertaken within each technical work package is contained within the Appendices.

2.3.1 WP2 – Rapidly tunable filters

The objective of this work package was an evaluation of rapidly tunable filters in tube-based frequency hopping and array radars to improve spectrum utilisation, which will release spectrum that can be allocated to other users.

This study was broken down into a number of activities:

- Review of technology options for tube-based radars.
- Implementation of high-Q filters into civil radar systems and a discussion of the key requirements and technology options for rapidly tunable filters with particular emphasis on ubiquitous vacuum electron tube-based radar systems.
- Review of technology options for array radars, which has centred on Micro-Electro-Mechanical Systems (MEMS) technologies. This is one of the most viable technologies for future array radars in which the prospect of further integration of the T-R modules shows considerable promise.

A search of published papers and patents to establish the state of current filter research, and discussions with different manufacturing sectors on the ability to produce filters, were carried out. The search considered current and future technology of high-Q, rapidly re-tunable filters for use in tube-based and active array radars.

The high peak-power of the typical radar systems almost certainly dictates the use of waveguide based filters. This is almost certainly true if the power is generated from a single source such as a magnetron as opposed to a distributed power source such as an array of solid-state power amplifiers. Waveguide filters remain the best option for high peak and average power handling combined with low insertion loss. A review of technology options for tube-based radar systems has been carried out i.e.:

- High power fixed waveguide filters.
- Technologies for high power frequency agile filters.
- Mechanically and electronically tunable filters.
- Magnetically tunable filters.
- Frequency selective surfaces (FSS).

RF filters at the transmitter output can be very useful in suppressing harmonic emissions. Filters can also be used to suppress out-of-band (OOB) and non-harmonic spurious emissions that prefer closer to the fundamental emission than the second harmonic. However, their utility in controlling relatively close-on portions of the emission spectrum is limited. This is partly due to the required small fractional bandwidth (ratio of filter bandwidth to the centre frequency) which places demands on Q-factor. This is also partly due to their additional cost, weight, size and the fact that many radars are tunable and/or use multiple waveforms, some of which have much wider necessary bandwidths than others have. There are key problems with the implementation of high-power RF filters that can be reconfigured to accommodate changes in a carrier frequency or a pulse waveform, especially when it is considered that such changes are required to occur within a few milliseconds.

Transmitter architecture is also an important determinant of the achievable degree of spectrum control. Where multiple power amplifiers are used emission spectrum falloff rate and level are influenced by whether the outputs of those power amplifiers are combined within the waveguide or only in space after being radiated. In-waveguide combining effectively creates a severe impedance mismatch for the mutually incoherent components of the output waveforms from the power amplifiers, which can dramatically lower the radiated noise power relative to the sum of the available noise powers of the amplifiers. Conversely, array radars fed by multiple amplifiers each of which is radiated reduces combination losses but it allows all of the noise power of the amplifiers to be radiated. Current opportunities for RF filtering are also limited in such arrays. This is partly because a separate filter would be needed for each amplifier, where the number of amplifiers can be hundreds or thousands. This situation could change with advances in MEMS technologies.

The radar output device has a major effect on the requirement for filtering non-harmonic spurious emissions. However, the selection of the radar output device cannot be made entirely on spurious emission characteristics. Because of the inherently high levels of harmonic spurious emissions from all types of output devices, the suppression of these harmonic components is generally performed where practical. Such filters are typically based on so-called “waffle iron” designs. To mitigate the non-harmonic spurious emissions from radar bands to other nearby bands band-pass filters would be required.

The use of such filters will involve a trade-off in the radar system performance; such harmonic and non-harmonic filters will have a small insertion loss. Depending on the size (order) of the filter and the operating frequency typical values are 0.1 to 1.0dB for each. Hence if two filters are required (one for harmonic suppression and another for non-harmonic in-band suppression) the insertion loss will double. These losses although small may have a significant effect on the performance of a radar: a 0.2dB loss is equivalent to a loss of 47kW of peak power from a nominal 1MW radar. This assumes of course, that the transmitter power is itself stable to better than 0.1dB. Alternatively a loss of 0.4dB corresponds to a 2.3% reduction of detection range, which might be significant in some applications. If the filters are connected to the output device via an isolator (formed from a “three-port” circulator and a termination) further losses will be introduced and the performance further degraded.

Power handling, size, and weight of the filter are factors that need to be considered in the feasibility of using an output filter on the radar, particularly in mobile radars. Size and weight can be overriding considerations in the case of mobile, active-array radars. Filtering bands close to the radar operating band requires steep selectivity skirts and hence high energy storage, which raises the risk of breakdown (or lowers the power-handling capacity) and can also introduce phase distortion in the pass-band, which is another major consideration for active-array radars. The higher the radar power, the more attenuation is needed to suppress spurious outputs to a given level, so the more sections of the filters will be needed, and hence the higher their insertion loss, size, and weight will tend to be.

The use of narrow bandwidth, high-Q band-pass filters will have a significant effect on waveform coding. These filters tend not to be linear phase. Clearly this will modify the waveform. This is correctable in the same way that communications systems employ channel equalisers. Although it is a further complication to the overall system design, it may be justified.

The low-cost of the magnetron devices compared to linear beam devices means that it is likely to remain the prevalent electron tube output device. The classic problems of magnetron operation still exist. However, they are now better understood, specified and controlled. The key problems pertaining to the frequency stability of magnetrons are moding, thermal drift, frequency pushing, frequency pulling, noise and time-jitter.

A number of solutions exist for cleaning-up the spectra of magnetrons, including adding additional filtering. A large step to cleaning up magnetron spectra, particularly as the device ages, is correct modulation. Here, by correct modulation we mean that the magnetron is kept in the pi-mode throughout its lifetime. One way to achieve correct modulation is to ensure constant current drive. Low-pass filters can adequately suppress harmonic contents. Once these problems have been addressed, should further suppression of spurious output be required, in-band filters can be added. However, the connection of a relatively small fractional bandwidth, high-Q filter to a magnetron device is not without problems. In order not to contribute to the problem of frequency pulling in magnetrons, it is necessary to use an isolator between the output device and the filter. Additionally due to the reflective nature of the filters, this function is best performed by a three-port circulator and a termination, as OOB energy would otherwise be reflected back into the magnetron.

The conclusions and recommendations from this study are summarised as follows:

- For crossed-field devices little can be done to change the modulator system, in terms of changing pulse rise- and fall-times of the pulses with a view to controlling output sidebands. However it appears that there are gains to be had from necessitating improved modulator design.
- High-power, high-Q RF filtering remains the best option for improving spectrum utilisation, once issues of correct modulation have been addressed. That is of course provided the issues of insertion loss etc. do not significantly degrade the radar function.
- Solid state devices with waveform coding perhaps offer the best opportunity for achieving the ultimate in spectral efficiency. However, there are distinct cost disadvantages to this approach.
- A large range of options are available for improving the spectrum efficiency of linear beam devices, in particular through the use of appropriate waveform coding. Again, filtering remains a key option with similar caveats to those for crossed-field devices applying.

2.3.2 WP2 – MEMS technology for tuneable filters in radar applications

Although a sub-section of WP2, the subject of MEMS technology for active array radar is far enough detached from the issue of cleaning up the spectra of tube-based devices, it was dealt with as a separate topic.

In recent years significant progress has been made in the use of MEMS technology to realise a range of RF components. These include tuneable capacitors, high Q micro-inductors, micro-mechanical filters, RF switches, resonators and cavity structures.

Whilst many of these devices are aimed primarily at the communications market and some of them are likely to find an application in radar. A specific example of this is the development of MEMS RF switches in active array antennas.

This is a review of the state-of-the-art RF MEMS components against key performance criteria identified at being of interest to the array radar community for rapidly tuneable filters and to provide some insight on the potential of these devices to impact this area. The issues that have been examined are:

1. Switching/ tuning time
2. Filter Q/bandwidth
3. Insertion loss
4. Reliability/ power handling
5. Size/ cost

Significant progress has been made on the demonstration of MEMS based RF switches in recent years. There are two generic types of MEMS RF switch architecture that are commonly presented in the literature - Ohmic and capacitive.

The most simple of these device structures is the Ohmic switch. This structure uses mechanical action to open or close electrical dc contacts. Provided contact resistances are kept below approximately 2Ω the switch insertion loss will be less than 0.2dB.

The capacitive style of switch uses mechanical action to change the capacitance through the switch between the signal line and ground. In the “up” position the loading of the capacitor on the signal is minimal and the signal is passed unattenuated. In the “down” position the signal line is effectively shorted to ground for high frequencies thus reflecting the signal back into its input port.

The key point to note for the capacitive switch, unlike Ohmic switches, is that the window of low insertion loss and high isolation is relatively narrow and consequently structures have to be individually designed to operate at a specific frequency. It can readily be shown that that this class of device is most suitable for higher frequency signals, greater than a few tens of GHz, due to the otherwise unwieldy size of the MEMS switch element.

The main down side with this technology is the relatively long switching speed associated with mechanical movement of the switch, typically >2μs compared to 10-100ns. At a most simple level the mechanical switching speed of a MEMS switch is limited by the mechanical resonant frequency.

There has been significant interest in using micromachining techniques to implement high performance RF passives, such as tunable capacitors and high-Q micro-inductors for silicon based RF systems. For frequencies in excess of 3-4GHz or for applications requiring higher Q components than can be readily realised using passives, filters based on thin film bulk acoustic resonators (FBAR) or cavity resonators have become attractive.

It is generally accepted that the components within a given bandpass filter should have Q’s that are 20-25 times the Q of the filter. This means that the use of filters based on the use of MEMS capacitors and micro-inductors are currently relatively limiting due to the low Q of inductors.

Tuneable filters based on MEMS capacitors with off-chip inductors have been reported with centre frequencies up to 2GHz and RF circuit functions such as phase shifters based on semi-lumped circuit implementations using MEMS capacitors lower-loss microstrip transmission lines to replace lumped inductors have been realised. At higher frequencies MEMS switches have been used to tune the bandwidth and centre frequency of bandpass filters.

Clearly as well as designing individual filters with a tuneable frequency response MEMS switches could be used to direct signals through one of a bank of predefined fixed response filters.

The conclusions and recommendations from this study are summarised as follows:

Improvements in the performance of a range of RF components using MEMS/ micromachining technologies appear to have a significant future in helping to realise RF functions for future RF systems especially where reduction in size is important.

The evidence in work published in the open literature shows that these technologies show a lot of promise for a number of applications. Based on evidence available to date, tuneable filters based on MEMS technology that cover the frequency range of interest will be available to the array radar community. Although the technology is maturing with components for the telecommunications market becoming available, there are a number of questions that still need to be addressed before a definitive answer can be given as to their applicability to the problem of tuneable filters for array radar systems.

Two of the key areas of concern for this application are switching speed and the ability of this technology to handle the high power signals seen during the transmit operation.

The tuning time for MEMS based filters systems is unlikely to be much faster than 1-2ms due to the relatively slow mechanical response of these structures compared to semi-conductor based RF switches.

It is generally accepted that any filtering in RF circuits occurs after the power amplifier stage. Consequently the filter stage, including any MEMS components, sees the full transmit power. One route that has been proposed to ameliorate this is by moving the filter to a point preceding the amplifier. Clearly this means that the linearity of the amplifier stage has to be very good but if a system architecture is used with one amplifier per antenna channel then the maximum power that each amplifier is required to handle can be significantly reduced. To enable a full understanding of the benefits of such an approach it is recommended that a system level study is initiated to look at possible novel architecture changes that could be made to exploit the maximum potential benefits of these devices.

Finally, although the developments in RF MEMS technologies have been rapid during the past 7-8 years, the technology is still at a relatively early point in its life cycle. Consequently, it is currently impossible to predict how these technologies will develop in the way that one does for silicon IC technology using, for example, using Moore's law. This problem is exacerbated by the divergent needs of the different potential customer bases which will require components to be optimised for their own market needs – for example high power handling for radar applications and low voltage, low power for mobile communications and that is without considering frequency.

2.3.3 WP4 – Ultra narrow band radar

This work package has investigated and evaluated an ultra-narrow band distributed radar system as a replacement technology for current systems operating in the L, S, C and X-bands of the electromagnetic spectrum. The performance was benchmarked against that of existing systems. In this way an examination of whether or not ultra narrow band distributed radar can achieve the same performance as existing systems whilst utilising significant spectral savings and thus offer a credible and attractive future option.

The bandwidth occupancy of current Civil Aviation, Weather, Marine Navigation, Coastal Surveillance radar systems is, in part, determined by historical factors. These types of systems employ a high power transmitter, duplex switching and short pulses to achieve the required levels of isolation, sensitivity and location accuracy from a single radar site. It is these characteristics that give rise to the relatively large bandwidth occupancy of current systems, since bandwidth is approximately inversely proportional to pulse length. For example, a 100ns pulse width equates to a bandwidth of 10MHz and results in a range resolution of 15m. The frequency spectrum allocations for radiolocation in L, S, C and X-bands occupy more than 1GHz of these bands.

In principle, as a potentially very attractive alternative, monochromatic or very narrow band radar waveforms could be employed occupying a bandwidth of just a few Hz. This implies continuous wave (CW) operation, which in turn implies range ambiguity. That is to say an infinitely long continuous tone has no range resolving capability. In addition the detection range may be limited due to ‘spill-over’ or ‘leakage’ of the transmitter signal directly into the receiver thus competing with target returns. This will limit ultimate sensitivity and hence the maximum detection range.

However, these restrictions can, in principle, be overcome if multiple transmitter and receiver sites are used. Three or four sites may be required in order to replace current systems. Separation of the transmitter and receivers reduces the transmitter ‘leakage’ problem and appropriately high powers consistent with the required detection ranges may be able to be employed. Multiple distributed receivers also allow targets to be located via a combination interferometry and triangulation rather than relying on a single radar beam. Accuracy will depend on the location, the number of the radar receivers and signal to noise ratio (SNR). The location accuracy can be further improved by exploiting Doppler (which is easily determinable in a CW radar) and angle at each receiver. Indeed the combination of these parameters can be used to determine location uniquely. This may be necessary for Marine Navigation Radar systems where there are limitations on the possible locations of the transmitter and the receiver. However, initial calculations suggest that narrow band CW radars could, potentially, be an alternative to current systems.

Current installations could also provide much of the necessary hardware and infrastructure to support a distributed narrow band system. For example the parabolic dish antenna, turning gear and housings could all be re-used thus affording a considerable financial saving. Furthermore, narrow band radar systems are inherently low cost with suitable transmitters being readily available. They also exhibit superior noise performance as a result of only operating over a narrow bandwidth. The completed system is envisaged as utilising largely ‘off the shelf’ components.

The metric of any proposed system will be that it is able to replicate the performance of currently used systems.

A theoretical analysis of narrow band radar for representative ATC, weather, marine and coastal radar systems have been carried out. Two cases were considered:

1. The first case carried out some simple analysis of performance trade-offs using pulsed and CW radar equations.
2. The second case investigated the geometrical considerations in order to make Doppler and SNR comparisons between monostatic and bistatic pulsed and CW systems.

The simple analysis has shown that the SNR and Doppler performance can be maintained to a reasonable level using the ultra narrow band approach. The weather radar shows some rather larger drop in the SNR particularly for the monostatic case modelled. However, these results need to be treated with caution since the extended scatterer case influences the definition of CW bandwidth and the applicability of this methodology in this case is still being investigated.

In many radar applications target location accuracy, in some form of co-ordinate space, is of vital importance where safety is paramount. It is critical that the narrowband CW radar system is able to offer the same level of performance as current pulsed systems.

The location accuracy of current systems may be expressed in terms of range resolution (usually a few tens of metres) and beamwidth (dependent on the size of the antenna used). Ultra narrow band distributed radar has no range resolving ability. Hence, to make up for this, multiple receivers are used to calculate the location of targets. Simple triangulation can then be invoked to find target range from each receiver and thus position. Assume an ATC radar with transmit and receive beamwidths of five degrees and further assume monopulse or another form of within beam processing. At a range of 100km a target may be located to an accuracy of the order of 50m in X-Y co-ordinate space. This may be further improved by combining triangulation with Doppler history data. These combinations will be unique for each target enabling them to be separated spatially without ambiguity. However, a direct comparison with the pulsed radar is not straightforward due to the fundamentally different techniques employed.

In addition, this is a technique that is yet to be fully developed and further work is required to be able to confidently apply this approach and make comparisons with the pulsed case. Studies will continue to investigate achievable location accuracy, developing the triangulation method, applying it to the three candidate systems and taking scanning into account.

In a narrow band CW radar, there is the problem of isolation between the transmitter and receiver. A single antenna can be used for both transmission and reception with the separation between the two signals being achieved by the Doppler shift. However, in practice, it is not possible to eliminate fully the transmitter signal leaking into the receiver. One solution is to use two antennas that will achieve the largest isolation.

In the proposed narrowband system, there is likely to be a number of bistatic system pairs in a multistatic arrangement. This implies separated transmitter and receiver antennas. It is very important to determine how much isolation is required between each transmitter/receiver pair to give the required sensitivity. Having done that, the next step was to determine how much isolation can be achieved.

For typical ATC radar the receiver sensitivity was set by the noise power in the receiver and is calculated at –188dB for a CW equivalent system. This figure is based on a 18Hz CW effective bandwidth and assumes a noise figure of 2dB. Thus the signal leaking directly from the transmitter to the receiver should sit below this level if genuine targets are to be detected. This will lead to slightly pessimistic results, as the true compression gain that may be accrued is not factored into these calculations. The amount of leakage will also set the dynamic range i.e. if the transmitter leakage is above the noise floor of the receiver, increasing the transmit power will only increase the leaked signal and no benefit will accrue.

It is assumed that power from the transmitter can only reach the receiver via transmit and receive antenna sidelobes. The magnitude of leakage will be fundamentally determined by the transmitter power, the sidelobe levels and the physical separation of the transmitter and receiver. However, this can be further reduced by placing nulls in the antenna beam pattern, dynamic cancellation of any breakthrough by Fourier processing.

Isolation is achievable through a number of different factors. For example, consider a case with the following parameters to illustrate the requirements to maintain leakage below the noise floor of the receiver. The transmitter power is taken to be 100W and it may be assumed that transmit and receive antenna sidelobes have an effective gain of -10dB. If the transmitter and receiver are separated by 1km then this provides of the order -70dBW of the received signal strength. Thus, a further 120dBs of cancellations are necessary. This might be achieved using combinations of techniques. For example creating nulls using auxiliary antennas on both transmit and receive may provide another 60dBs of attenuation. Dynamic cancellation of the breakthrough signal can yield up to a further 50dB of attenuation. Lastly Fourier processing could yield a further 30dBs of isolation (except in the DC term in the case of Fourier processing). Thus the leakage signal should appear at a level of some 20dBs below the receiver noise floor.

However, it remains to be seen in practise as to the degree of direct signal attenuation that can be achieved and at what cost. Further research is required to determine exact levels of isolation achievable and these will subsequently fix the range of design freedoms. This has been examined in the project although a full analysis lies out side of the scope of this contract.

The conclusions and recommendations from this study are summarised as follows:

- Ultra narrow band distributed CW radar can achieve similar sensitivities to a pulsed counterpart.
- The location accuracy of ultra narrow band distributed CW radar is superior in angle and inferior in range. Further work is required to fully evaluate performance and assess the impact of the difference in the form of measurement of location accuracy.
- The potential saving in spectrum is of the order of 90% of that occupied by current pulsed radar systems.
- Although current studies have focused on fixed radar installations, ultra narrow band radar techniques may be applied to marine navigation as a replacement for the current magnetron based systems. However, there are two principle aspects of performance that must be addressed. The first is the restricted sighting of the multiple receivers that will determine the accuracy to which targets can be located. The second problem is that of sea clutter. Little data exists for narrow band CW radar and hence it is not possible to predict with confidence the influence this may have. One area of particular concern is that of close in clutter returns ambiguously competing with target returns which consequently reduce overall system sensitivity. Distinguishing targets from sea clutter on the basis of velocity will generally be aided in the CW case but the achievable performance remains to be clarified. It should be noted that there are FMCW systems currently in use in a maritime environment therefore offering some evidence that sea clutter problems are surmountable.

This is very much a preliminary feasibility study and much work remains to be done to fully validate this approach. In particular the study identified the following as requiring further attention. Most of these items have been the subject of some initial investigations during the project but all require further detailed analysis:

- Extending and validating the isolation requirements to evaluate the design restrictions and cost issues.
- Validating the weather radar data for extended volume scatterers.
- Investigate the effect of moving from monostatic to bistatic to multistatic geometry.

- Investigate target location methods including modulation schemes to increase location accuracy without increasing bandwidth excessively.
- Investigate practical implementation issues such as location/site availability.
- Perform a cost benefit analysis including trade off between financial benefits for better spectrum utilisation and capital cost of existing equipment modifications.

2.3.4 WP6 – Waveform design

The objectives of this work package was to demonstrate techniques that allow the mitigation of interference between radars and other service users within a given band by the use of waveform coding and hence allow more users in the band.

This work package presents the finding of literature search, study and simulations in addition to demonstrating of some of the techniques discussed.

Currently most civil radar systems use low cost magnetrons that transmit short uncontrolled pulses to achieve adequate resolution and long range performance. The transmissions from such radar systems have broad operating bandwidths and suffer from spectral splatter, or spurious emissions. Therefore the available spectrum is not being efficiently used by these systems. In addition, the number of applications that require use of the electromagnetic spectrum is increasing, which places pressure on existing allocated spectral bands and their boundaries. With recent advances in technology and demand for bandwidth these pressures have transferred from the below 1GHz bands to all bands between 1-16GHz. This sets the scene for the focus of this study.

The nominal range resolution for a bandwidth *B*:

$$\Delta r = \frac{c}{2B}$$

Equation 2-1

Where *B* is typically the half power bandwidth and termed as a resolution bandwidth.

Basic radars generate bandwidth by transmitting narrow pulses: the nominal resolution bandwidth being the reciprocal of the pulse-width. This includes most of the civil radar systems under consideration. CW radars have very narrow bandwidths and detect moving targets by the Doppler shifts in the received signal but no range information is obtained. More sophisticated radar systems employ pulse-compression (PC), where the transmitted waveform is a long modulated pulse that either sweeps the resolution bandwidth e.g. linear frequency modulation (LFM) or have a wide instantaneous bandwidth throughout the pulse e.g. binary coded waveforms. Such systems traditionally require sophisticated, bespoke waveform generators and receivers. Therefore, currently only the large high cost assets and military systems use such waveform. However, advances in waveform generation technology and digital signal processing allow PC technology to be accessible to lower cost radar systems.

This work package has examined such sophisticated waveform techniques with an aim to satisfying reduced spectrum and cross system interference objectives, whilst presenting performance issues resulting from use of such waveforms.

Interference between systems:

This study shows that for a typical system to work within a nominal 3km of a radar or communication system with a similar bandwidth. The interference rejection needs to be in excess of 100dB. Whilst the antenna sidelobes may be expected to reject 30-40dB of the energy, this leaves a challenging 60dB or more to be dealt with by a waveform design and digital signal processing (DSP).

Regulations:

In line with current thinking, some of the current regulations have been shown to be flawed. In addition, the regulations were found not to significantly prejudice waveform with theoretically noisy spectra, and in some cases perform the opposite. This is summarised by:

- The practice of adjusting waveform characteristics to give increased allowed bandwidths, which is a common practice for radar manufactures to achieve type approval.
- Allowing spurious signals levels to be relative to peak envelope power rather than absolute levels.

Simple spectral noise reduction:

A matched transmitter technique based on closing the loopholes in the regulations was presented which was shown to be effective in cleaning the spectral noise produced by current systems. These techniques constrict the transmitted waveform to the minimum required to maintain it resolution, typically its 3dB bandwidth. This concept could form the basis of future regulations. In addition, peak power reductions due to pulse compression and coherent processing were shown to be effective in reducing noise over the entire spectrum.

Reductions of allowed bandwidths of factors between 10 and 20 were produced over the current radar systems, if these techniques are implemented.

Orthogonal waveforms:

The practicality of rejecting interference by employing orthogonal PC waveforms was investigated. These would normally give a rejection of the order of $1/TB$ and $1/\sqrt{TB}$ in power, where TB is the Time-Bandwidth (TB) product. Achieving more would require complementary spectra and that might lead to a major degradation in sidelobe and Doppler performance.

LFM and digitally derived equivalents were found not to offer sufficient orthogonal variants for multiple systems operating in a close proximity. Polyphase coded and Costas coded waveforms offer the best candidates for operation of a large number of radar systems, in a shared spectrum. However large TB products are required to achieve adequate range sidelobe levels and an unfeasibly large TB product would be required to meet the interference rejection targets.

When analogue orthogonal waveforms are used in conjunction with the technique of operating on different centre frequencies the use of orthogonal waveforms give no significant advantages. Whereas combining the techniques of frequency separation with coded orthogonal waveform can enable interference performance as good as analogue waveforms with more compact spectrum.

Magnetron radars:

Modern coaxial magnetrons, magnetron modulation and injection locking have been discussed to offer greater frequency stability and less noise than current magnetrons. This may enable tighter packing of such radar systems into a given band, reduced harmonic emissions and lower power levels.

Pulse compression waveform with narrow spectrum:

The radar's spectral footprint can be reduced by careful selection of a waveform or by modifying its existing waveform. The ideal case would be were all the transmitted energy lies within the resolution (half power) bandwidth, which is essential for maintaining the radar range resolution.

Taylor quadrIphase:

Taylor quadriphase (TQ) is a variant of the four-phase coded waveform that can reduce the wide spectral splatter normally associated with coded waveforms whilst maintaining the performance of its conventional counterpart. Its spectrum falls off at a much steeper 12dB/octave, compared to 6dB/octave for conventional waveforms, reaching –40dB in a 1/13th of the bandwidth. In addition, as there is much less energy in the spectral sidelobes, its spectrum can be truncated at the first nulls without significantly degrading the performance.

Frequency modulation:

Given a reasonably high time-bandwidth product, the near rectangular spectrum of LFM waveforms are naturally more spectrally efficient than coded waveforms. This can be improved furthermore, and a number of techniques for reducing its spectral impact were investigated.

Non-linear frequency modulation (NLFM) can significantly narrow the spectral width, but only about -30dB below peak value. Below this level, the NLFM spectral skirt fall-off is set by the sharp edges of the transmitted pulse. It is an attractive option because it removes from the transmitted spectrum the part that would have been removed in the receiver by windowing. Furthermore, NLFM improves receiver SNR performance as no power is lost by windowing and the non-linear slope does not significantly affect its Doppler tolerant performance.

A somewhat drastic method of generating LFM waveforms with much reduced out-of-band emission was also examined. It has a considerable potential for spectral cleanliness, but as the familiar LFM spectral shape as now been transposed to the pulse envelope, it has the disadvantage that it requires a degree of linearity in the radar's power amplifiers. It is an effective technique, however when the maximum bandwidth is required within a constrained spectral environment.

Frequency separation:

Frequency separation of waveforms to reduce interference between standard analogue and digital waveforms has been examined. Predictably it was found that waveforms with compact spectra can operate with closer frequency separations for the given interference targets. For example, LFM waveform operating on frequencies separated by only five times their range resolution bandwidth can reject each other by greater than 60dB.

However, orthogonal digital waveforms were found to be superior in some aspects when frequency shifted, especially if combined with a waveform filtering.

Spectral nulls:

One solution to interference rejection is to generate nulls in the waveform's spectrum that coincide with the interference signals. Interference by the radar on the signal source would also be reduced along with the sensitivity of the matched filter (MF) to those signals. Two methods of generating such nulls were discussed.

Such methods could be used to form sets of waveforms with non-overlapping spectra, giving a truly orthogonal waveform, this possibility has been left for further work.

Interference filtering/rejection:

The ideal solution to interference is to remove it prior to the matched filter (MF) and a method for so doing was shown to be effective. Rejection methods have potential where a degree of co-operation between the radar systems is present. However, simple interference rejection methods, currently used in incoherent radar, were thought to be less effective for pulse to pulse coherent modern radar because of the need for a staggered pulse repetition interval (PRI).

This area of interference filtering/rejection is outside the remit of this report. However, it has been covered briefly because when used in conjunction with radar waveform techniques presented, it makes the discussed techniques and waveforms viable. For example, interference rejection techniques require that the interfering signal has high signal to noise (>30dB), and the interference may not be completely removed, whereas the waveform techniques reduce interfering signals at any signal to noise values, often by smaller amounts than the interference rejection. Therefore, the dual use of interference filtering/rejection and the waveform techniques can satisfy interference rejection targets.

This is considered as an area in which a further study is required.

Stepped chirp and FSK waveform:

The stepped chirp waveform splits the full bandwidth LFM chirp into a sequence of overlapping, narrow-band sub-chirps, which may overlap in frequency. As the receiver is only tuned to the narrow-band sub-chirp it offers high rejection to other radars transmitting at other sub-bands.

The sub-chirps need not be transmitted in order (i.e. Costas waveform) but this adds to the complexity of processing. However, in this case, each sub-chirp could be transmitted opportunistically when its sub-band is free.

Bandwidth extrapolation/interpolation for gap filling:

Bandwidth extrapolation (BWE) is a well-established technique that applies linear prediction principals to increasing the radar's range (and angular) resolution. A model is formed from the known spectrum, which is then used to extrapolate the spectrum by a factor of two or more to give a commensurate increase in range resolution. The resolution enhancements of a factor of two can be made without prejudicing the original data.

This technique has been applied to the allied problem of interpolating the spectrum in order to fill gaps in the spectrum. Gaps could be due to null forming in the transmitted waveform, or due to interference blanking in the frequency coverage. The technique has been found to be successful but requires good SNR to be effective.

Application of waveform techniques to bands of interest:

Application of a selection of the waveform concepts to three exemplar bands of interest exposed the need for assessment of the waveform spectrums through demonstrations.

However, results show that considerable improvements can be made in all the bands of interest, particularly in the cases, where filtered waveforms and co-operative step chirp FM waveforms are used. Reductions of the selected radar bands are calculated to less than half of the current radar allocations.

To summarise, a number of waveforms, schemes and techniques that have been explored, which have the potential to reduce the spectral occupation of radar systems. In addition, the problem of interference between systems has been addressed. Whilst no single panacea was found that fully met the requirements for band sharing in all circumstances, a combination of techniques could be effective.

Such combinations will form complex systems that require design, simulation, demonstration and evaluation. As such this is a topic that will need further work.

3 Cost benefit analysis

3.1 Cost benefit analysis methodology

In order to establish an economic value of the technical proposals a cost benefit analysis (CBA) is presented. The CBA carried out is not a rigorous analysis, but simplified to gain and understanding of the differences in costs and benefits of the various technology options. A number of assumptions have been made to support the CBA and are outlined in §3.2.

The CBA has been developed by taking advice from the project team, Ofcom, and the civil radar user and manufacturing community. A list of contributors from the radar industry is detailed in Appendix F.

A high level methodology was developed from which the CBA was carried out. It was decided that there were two parts to the analysis; each of which would have its own level of accuracy:

1. The determination of cost/MHz of released bandwidth for the technology options investigated.
2. The determination of a range of benefits brought about by the release of any new spectra.

The calculation of benefits depends upon the assumption of the type of system that is likely to take up any released spectra:

The high level methodology is to:

- Identify exemplar frequency bands across the 1-16GHz frequency range.
- Neglect secondary users within each of the selected bands to simplify the CBA.
- Identify the techniques that have the potential to improve spectrum utilisation.
- Apply the techniques to the exemplar bands to assess an amount of spectral release.
- Identify the costs associated with implementing the techniques.
- Calculate the cost/MHz.
- Identify the benefits associated with a release of spectrum.
- Calculate the relevant benefits.

In identifying suitable bands to carry out the investigation, a number of factors were considered:

- Try to establish a single primary band to reduce the amount of complexity.
- Marine radars:
  - There are between 50,000 and 100,000 systems to consider.
  - They share bands with other systems.
  - They are mobile with International implications.
- ATC & Met radars sit within shared bands but they also have specific non-shared primary bands just for that application.
- Lower frequency bands are more likely to generate most interest in terms of spectrum trading and generation of revenue.

Having taken all of these factors into consideration three frequency bands across the 1-16GHz spectrum were selected. It was decided to select bands at different points within the spectrum to determine whether benefits were frequency dependent. Having chosen the bands an initial strategy for freeing up as much spectra as possible was proposed:

1. Band 1: 1.215-1.350GHz see Figure 3-3: The approach is to apply the relevant techniques to civil airport radar systems in the band and then re-tune to the 1.215-1.260GHz band thereby releasing spectra in the 1.260-1.350GHz band.
2. Band 2: 2.700-3.100GHz see Figure 3-4: The approach is to apply the relevant techniques to civil airport radar systems in the band and then re-tune to the 2.900-3.100GHz band thereby releasing spectra in the 2.700-2900GHz band.
3. Band 3: 15.40-15.70GHz see Figure 3-5: The approach is to apply the relevant techniques to airport ground movement radar systems in the band and then re-tune to one end of the band thereby releasing the maximum single amount of defined bandwidth.

Having defined the relevant bands then the next step is to decide which technology or combinations of technology could be used to improve the spectral utilisation of the relevant bands. The techniques are:

- Filters (F) – which is classed as a near term solution.
- Waveforms (W) – which is classed as a near term solution.
- Filters + Waveforms (F+W) – which is classed as a near term solution.
- Ultra narrow band (U) – which is classed as medium term solution.

The definition of near term being implementation within the next five years and medium term within the next five to fifteen years.

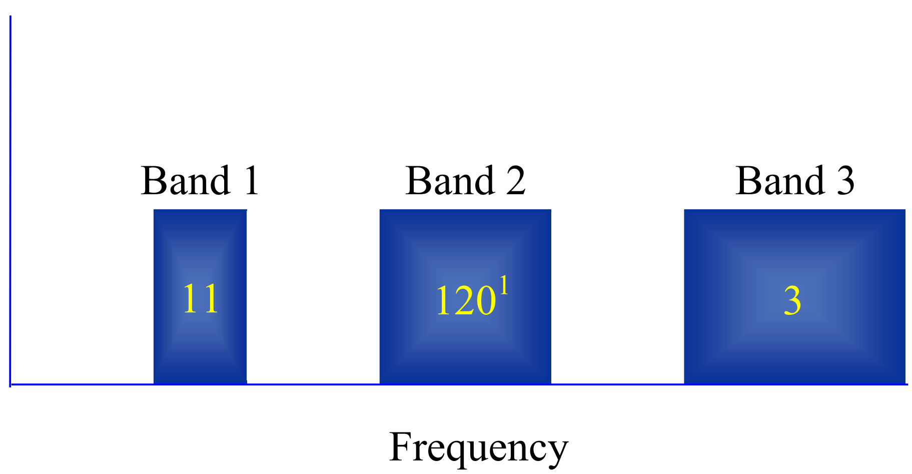


Figure 3-1 Indication of the number of primary radar systems in each band

[1] Includes 84 military radar

Therefore, the frequency release for each of the bands is:

- Band 1 [(F),(W),(F+W),(U)] MHz
- Band 2 [(F),(W),(F+W),(U)] MHz
- Band 3 [(F),(W),(F+W),(U)] MHz

The cost of implementing the technology is:

- Band 1 £ [(F),(W),(F+W),(U)]
- Band 2 £ [(F),(W),(F+W),(U)]
- Band 3 £ [(F),(W),(F+W),(U)]

Costs are determined from a combination of the following:

- Demonstrating the technique.
- Procurement of the new technology.
- Implementation/qualification of the new technology.
- Systems down time.
- Impact of any potential loss in performance.
- Any change in running/maintenance costs.
- Write off costs of existing radars that are replaced.
- Cost of introducing new regulations.
- The number of systems in which the technology has to be applied.

Having determined the costs then the cost per MHz can be calculated for each band and each technology:

- Band 1 £/MHz [(F),(W),(F+W),(U)]
- Band 2 £/MHz [(F),(W),(F+W),(U)]
- Band 3 £/MHz [(F),(W),(F+W),(U)]

This is achieved through a combination of applying the spectrum efficiency techniques studied and then re-tuning each radar in the defined band to maximise the releasable bandwidth defined as “New Band”.

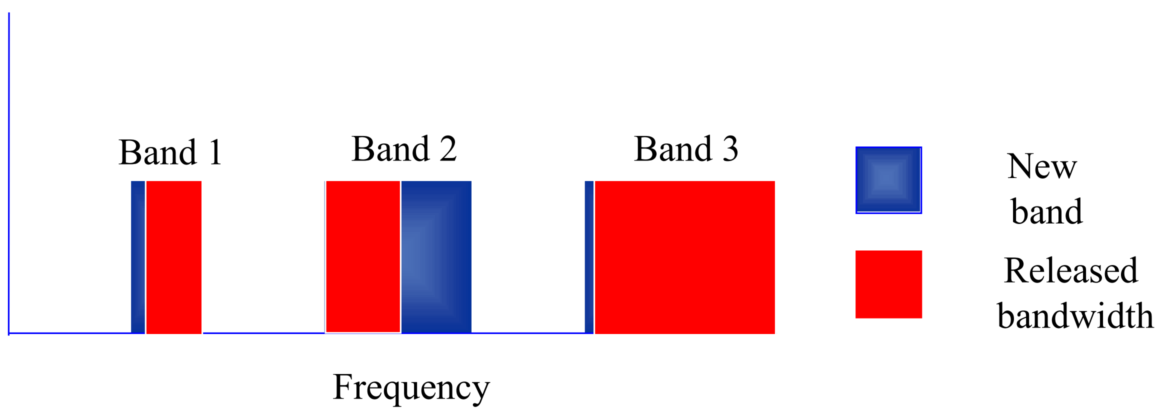
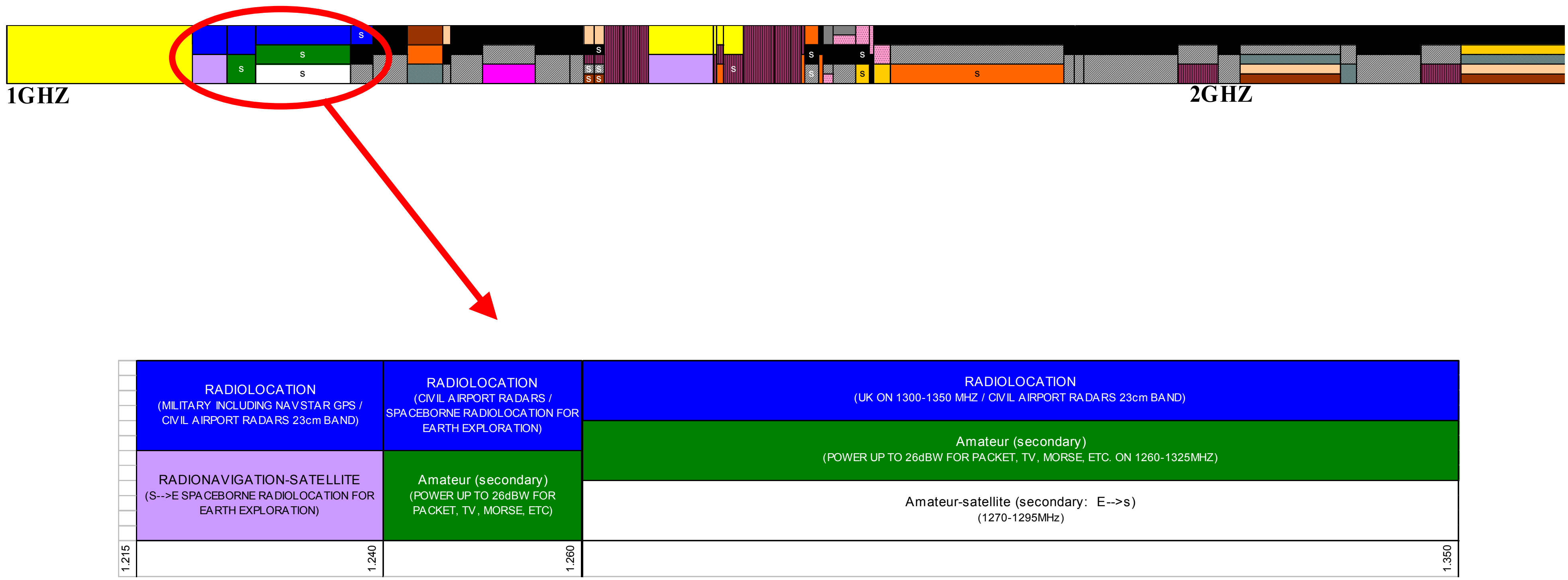


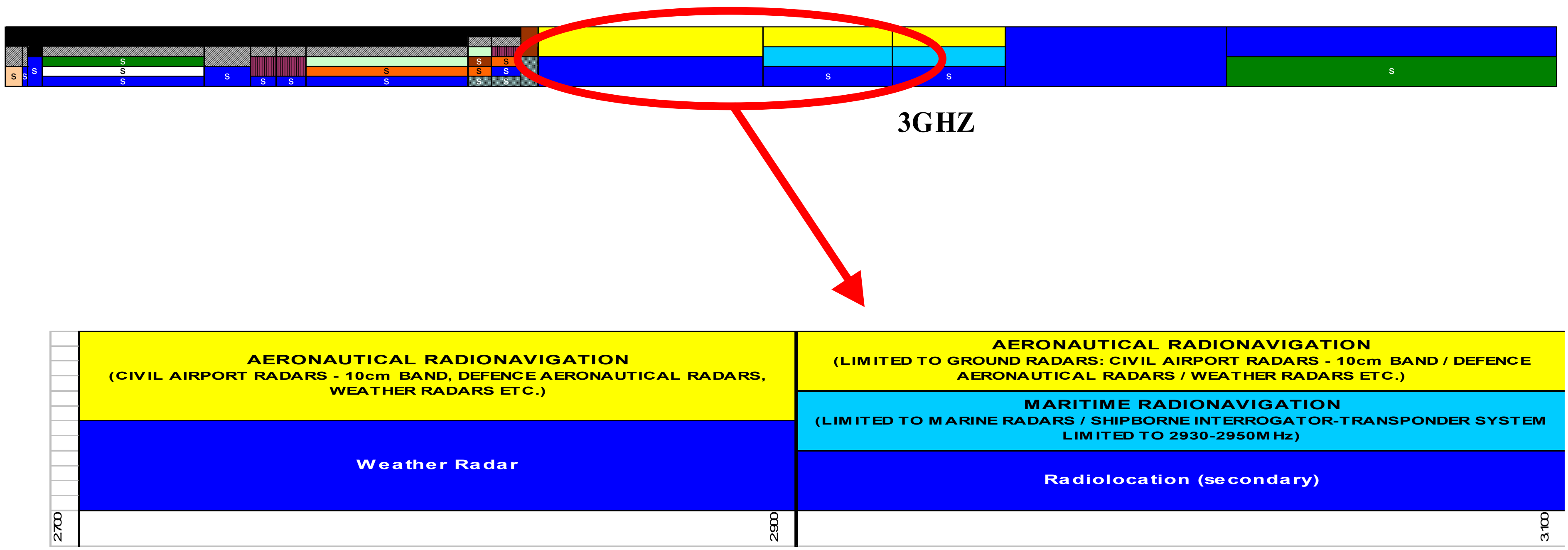
Figure 3-2 Illustration of released bandwidth

Figure 3-3 Band 1



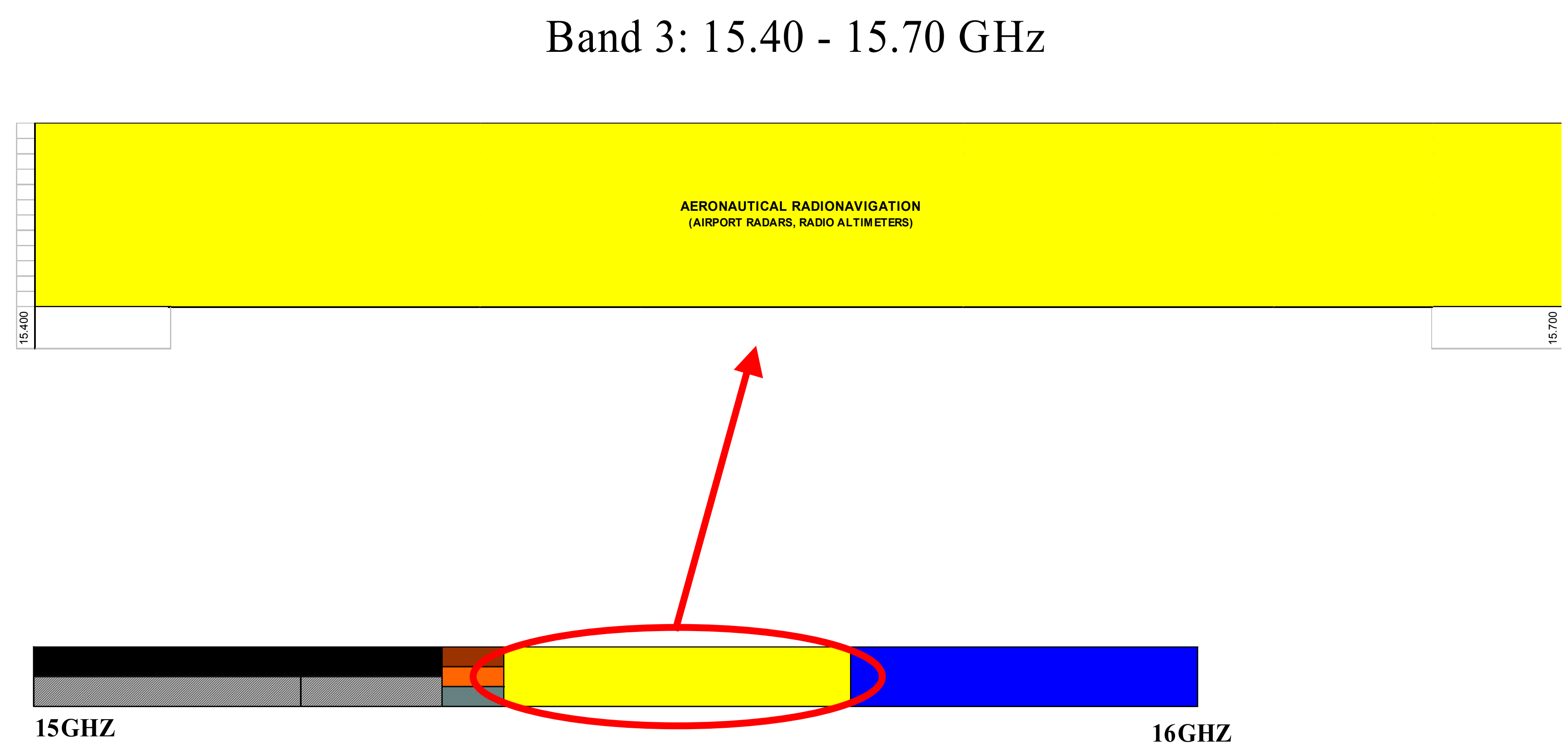
Band 1: 1.215 - 1.350 GHz

Figure 3-4 Band 2



Band 2: 2.700 - 3.100 GHz

Figure 3-5 Band 3



The second phase of the CBA was to establish a range of likely benefits. These fall into the following categories:

- UK industry:
  - Providers of technology to upgrade existing systems to release spectra.
  - Providers of new technology that will benefit from any released spectra.
  - Additional services associated with an introduction of new technology.
- Radar user community:
  - Increasing the robustness of radar systems.
  - Compliance with ITU-R regulations.
  - Allowing spectrum trading.
- Ofcom:
  - To manage the spectrum efficiently and effectively.
- UK consumer/citizen:
  - Individuals that would benefit from any new technology introduced.
  - Firms that would benefit by enhancing their businesses from any new technology introduced.

3.2 Assumptions

Below is a list of assumptions made during the CBA:

- Neglect secondary allocations in bands of interest.
- The correct number of systems in each band has been accurately captured.
- MoD systems will also adopt the techniques for the purpose of the analysis. This will help determine a maximum amount of available spectrum releasable and hence the maximum benefits.
- There are no new allocations designated for the bands of interest other than those already identified.
- Technology to take up released spectra was based on the frequency bands being used in the analysis. It was agreed that the most likely services that would wish to take advantage of any spectral release in the defined bands would be:
  - Band1: 1.260-1.350GHz; Cellular, radio LAN
  - Band 2: 2.700-2.900GHz ; Cellular, radio LAN
  - Band 3: 15.40-15.70GHz; Fixed wireless, line-of-sight communications
- For ultra narrow band radar it is estimated that the potential saving in spectrum is of the order of 90% of that occupied by current pulsed radar systems. If the technique were applied to all radar within the band then it reasonable to assume that at least 90% of that band should be released by using this technology.
- The required separation bandwidth for ATC radar (B_s) is 60MHz.

- From all the waveform techniques investigated, the best value performance for each band was selected.
- Cost associated with demonstrating the technique is a one-off cost.
- Cost associated with the procurement of technology is estimated on a for sale basis. However, commercial practices may dictate a licensing agreement or a profit share from revenue generated through spectrum trading.
- Cost of re-qualification is based on a cost to the airport operator and a cost to the airline industry for a system down time of 10 working days.
- Only one service will take up the allocation that is made available from the introduction of new technology.

3.3 Defined costs

In order to determine a cost/MHz of released bandwidth then a closer analysis of the three selected bands, in terms of existing and future allocation, is required.

Proposed future frequency allocations which affect the bands of interest are:

- Earth exploration satellite: 1.215-1.24GHz
- Space research: 1.215-12.4GHz
- Earth exploration satellite: 1.24-1.30GHz
- Radionavigation satellite: 1.24-1.30GHz
- Space research: 1.24-1.30GHz

These allocations are a result of the development of Galileo as well as the modernisation plans for current GNSS (GPS, GLONASS) [4]. Because of these findings the strategy for freeing up spectra in the 1.125-1350GHz bands will have to be modified. The 1.215-1240GHz band is already a shared primary band with radio satellite. Therefore, with the addition of the above services it is unlikely that there will be the available spectrum to relocate any additional radionavigation services from the 1.260-1.350GHz bands into this band. It has therefore been decided to omit the 1.125-1.240GHz band from the original assumption and concentrate on 1.240-1.350GHz.

Considering the 2.700-3.100GHz bands, the 2.700-2.900GHz band has a secondary allocation that includes weather radar. The Table of current allocations and users ITU footnote 5.423 [3] states that ground based radar used for meteorological purposes are authorised to operate on a basis of equality with stations of the aeronautical radionavigation service. Therefore any meteorological radar in this band must be treated as primary allocations and considered the same as ATC radar. Information has suggested that UK weather radar, operated by the Met Office, operate in other bands, although there may be other independent and mobile weather radar facilities within the UK that are currently operated in the 2.700-2.900GHz band. The proposal is to relocate 2.700-2.900GHz primary radar systems (ATC) to 2.900-3.100GHz; the analysis has taken into account marine radar that has primary allocation in 2.900-3.100GHz band.

Considering the 15.40-15.70GHz band, there is only a small number of airport ground movement radar operating in this frequency band. There is also a desire to move out of this band to a lower frequency because of component costs. The benefit from this proposal is that this would release the whole of this band for an alternative system.

From the outputs of the technology studies an estimation of the amount of spectrum that could be released for each of the frequency bands is detailed in Table 3-1.

	Band 1	Band 2	Band 3
Frequency range (GHz)	1.240 – 1.350	2.700 – 3.100	15.40 – 15.70
Current bandwidth (MHz)	110	400	300
Filter technology (F)	Based upon the use of fixed frequency inductive-iris band-pass filters reducing bandwidth occupation to necessary bandwidth (as defined by ITU-R SM.1541-1, Annex 8) for the number of systems for a given emission type.		
Waveform technology (W)	Based on the results of applying different waveform techniques to the number of systems in each band to ensure a minimum interference between radar systems.		
Filter + Waveform (F+W)	Based on a combination of the analysis of the filter and waveform technology.		
Ultra narrow band technology (U)	Based on the conclusion that this technology should yield a potential saving in spectrum of the order of 90% of that occupied by current pulsed radar systems.		
Required bandwidth(1) - F(MHz)	40(2)	295(2)	300(3)
Required bandwidth – W(MHz)	70	200	100
Required bandwidth – {F+W}(MHz)	40	200	100
Required bandwidth - U(MHz)	10	40	30
Released bandwidth – F(MHz)	**70**	**105**	**0**
Released bandwidth – W(MHz)	**40**	**200**	**200**
Released bandwidth – {F+W}(MHz)	**70**	**200**	**200**
Released bandwidth – U(MHz)	**100**	**200**	**270**

Table 3-1 An estimate of releasable bandwidth within the context of all assumptions

(1) Required bandwidth is the total frequency required to accommodate all radar systems in that band.

(2) Assuming a 100μs non-linear FM chip pulse.

(3) Filter design would require a bandwidth equal to the current band 3 bandwidth.

The costs associated with implementing each technology are shown in Table 3-2. During the discussions with the radar industry it was suggested that in some cases it may be easier, from a technology point of view, to procure a new radar system with the proposed technology incorporated rather than try to retrofit the proposals to existing systems.

Therefore two cost options have been provided:

Option 1 – The cost to retrofit the proposed technology to existing radar systems.

Option 2 – The cost of a new system with the proposed technology incorporated.

Costs	Band 1 (£k)	Band 2 (£k)	Band 3 (£k)
Filters Option 1:			
- Demonstrating the technique[2]	50	50	[1]
- Procurement of the new technology[3]	10	10	
- Implementation/qualification of the new technology per day[4]	100	100	
- Systems down time	10 days	10 days	
- Impact of any potential loss in performance[5]	Nil	Nil	
- Any increase in running/maintenance costs[6]	Nil	Nil	
- Cost of introducing new regulations[10]	–	–	
Filters Option 2:			
- Demonstrating the technique	50	50	[1]
- Development, procurement and qualification of a new radar system	10,000– 50,000	10,000–50,000	
- Any increase in running/maintenance costs	Nil	Nil	
- Write-off costs per radar system[11]	5,000	5,000	
- Cost of introducing new regulations[10]	-	-	
Waveforms Option 1:			
- Demonstrating the technique	50	50	50
- Procurement of the new technology	50	50	50
- Implementation/qualification of the new technology per day	100	100	100
- Systems down time	10 days	10 days	2 day
- Impact of any potential loss in performance	Nil	Nil	Nil
- Any increase in running/maintenance costs	Nil	Nil	Nil
- Cost of introducing new regulations[10]	–	–	–
Waveforms Option 2:			
- Demonstrating the technique	50	50	50
- Development, procurement and qualification of a new radar system	10,000–50,000	10,000–50,000	2,000–4,000
- Any increase in running/maintenance costs	Nil	Nil	Nil
- Write-off costs per radar system	5,000	5,000	1,000
- Cost of introducing new regulations[10]	-	-	-

Filters + Waveforms Option 1[9]:			
- Demonstrating the technique	75	75	50
- Procurement of the new technology	60	60	50
- Implementation/qualification of the new technology per day	100	100	100
- Systems down time	10 days	10 days	2 day
- Impact of any potential loss in performance	Nil	Nil	Nil
- Any increase in running/maintenance costs	Nil	Nil	Nil
- Cost of introducing new regulations[10]	–	–	–
Filters + Waveforms Option 2[9]:			
- Demonstrating the technique	75	75	50
- Development, procurement and qualification of a new radar system	10,000–50,000	10,000–50,000	2,000–4,000
- Any increase in running/maintenance costs	Nil	Nil	Nil
- Write-off costs per radar system	5,000	5,000	1,000
- Cost of introducing new regulations[10]	-	-	-
Ultra narrow band:			
- Demonstrating the technique	1,000	1,000	[8]
- Development, procurement and qualification of a new radar system[7]	20,000–250,000	20,000–250,000	
- Any increase in running/maintenance costs	Nil	Nil	
- Write-off costs per radar system	5,000	5,000	
- Cost of introducing new regulations[10]	-	-	
Number of systems per band	11	36+84	3

Table 3-2 An estimate of costs within the context of all assumptions

(1) The application of filters is not practical in this band as it would require a 300MHz bandwidth and efficiencies would be poor as highlighted by the radar industry (§4.5.5).

(2) Cost associated with demonstrating the techniques is a one-off cost.

(3) Cost associated with the procurement of technology is estimated on a for sale basis. However commercial practices may dictate a licensing agreement or a profit share from revenue generated through spectrum trading.

(4) Cost of re-qualification is based on a cost to the airport operator and a cost to the airline industry for a system down time of 10 working days and a cost per day is used.

(5) The introduction of a filter into a radar system will cause a <0.5dB loss at the centre frequency, which is not considered to be a significant loss in performance.

(6) Many of the proposed methods for increasing spectrum efficiency have either never been tried operationally before or are only used in high-cost and specified military equipment. Currently, it is therefore impossible to determine whether system robustness of civilian radar systems would be improved or reduced by incorporating these technologies. Further work to determine this will be required, possibly requiring operational data over several years, and hence have not been included in these calculations

(7) Since ultra narrow band is both a new technique and is likely to employ active array technology, the cost of the radar system could vary from between 2-times and 5-times that quoted for a standard replacement system. These systems are relatively unknown outside specialized military systems and more investigation is required to establish a more accurate figure for the civilian market.

(8) To replace 3 airport surface movement radar, based on the assumptions for the development, procurement and design of this technology, investing in ultra narrow band radar for this band is not considered practical.

(9) For band 3 same as for waveform because filters are not applicable.

(10) The cost of introducing new regulations is difficult to quantify. Ofcom have advised that this cost would be negligible in terms of the overall cost therefore should not affect the analysis.

(11) Write-off costs for existing radar is based on an assumption of an original cost of £10M per radar systems (£2M for ground movement radar) with a life expectancy of 25 years and that on average the remaining lifetime for all radar is half the total life expectancy. Therefore a 50% write-off for all radar systems. Further studies would be required to determine a more accurate estimate of radar write-off as in some instances independent airports update less frequent and changes in ITU-R recommendations can introduce a step change in procurement policy.

It is clear from the cost break down that for retrofitting any technology in bands 1 & 2, the overwhelming portion of cost is due to the re-qualification of the equipment. However, the cost of retrofitting is an order of magnitude less than the cost of re-designing a new radar to incorporated these techniques.

From the above a summary of the total costs for the number of systems in each band are given in Table 3-3:

Total costs	**Band 1 (£M)**	**Band 2(1) (£M)**	**Band 3 (£M)**
Filters Option 1	11.1	36.4	-
Filters Option 2	166 – 606	542 – 1982	-
Waveforms Option 1	11.6	37.9	0.8
Waveforms Option 2	166 – 606	542 – 1982	9.2 – 15.2
Filters + Waveforms Option 1	11.7	38.2	0.8(2)
Filters + Waveforms Option 2	166 – 606	543 – 1983	9.2 – 15.2(2)
Ultra narrow band	286 – 2816	936 – 9216	-

Table 3-3 Summary of total costs

(1) Based on the cost of modifying 36 civil radar systems only.

(2) Same as for waveform because filters are not applicable for band 3.

Cost/MHz	Band 1 (£M)	Band 2 (£M)	Band 3 (£M)
Filters Option 1	0.16	0.35	-
Filters Option 2	2.37 – 8.66	5.16 – 18.88	-
Waveforms Option 1	0.29	0.19	0.004
Waveforms Option 2	4.15 – 15.15	2.71 – 9.91	0.05 – 0.08
Filters + Waveforms Option 1	0.17	0.19	0.004(1)
Filters + Waveforms Option 2	2.37 – 8.66	2.72 – 9.92	0.05 – 0.08(1)
Ultra narrow band	2.86 – 28.2	4.68 – 46.08	-

Table 3-4 Cost/MHz

(1) Same as for waveform options because filters are not applicable for band 3.

3.4 Defined benefits

The assumptions in §3.2 state that the most likely services to take advantage of any spectral release in the defined bands would be:

- Band1: 1.260-1.350GHz; Cellular, radio LAN
- Band 2: 2.700-2.900GHz ; Cellular, radio LAN
- Band 3: 15.40-15.70GHz; Fixed wireless, line-of-sight communications

In order to determine whether the release bandwidth, identified in Table 3-1, could support the above technologies, then knowledge of current allocation is required. From the table of current allocations and users [3] there are allocations already assigned for:

- Mobile communications: UMTS frequency bands are 1920-1980MHz (up link) and 2110-2170MHz (down link); hence bandwidths of 60MHz.
- High performance radio local area networks that have an allocation to operate in the 5250-5300MHz band; hence a bandwidth of 50MHz.
- Fixed point-to-point links where there are several bands in the frequency range 7-16GHz where a typical allocation is between 30MHz and 150MHz

From this information it is reasonable to assume that these services could be supported by the spectra released in Table 3-1, and on this assumption a range of benefits can be estimated.

3.4.1 UK industry

The benefits to UK industry are through a combination of:

1. The suppliers of new technologies to the current radar community which will allow them to release a portion of the radio spectrum.
2. The radio manufacturing industry that will make use of the radio spectrum which is made available.
3. The radio service industry that will provide a range of services to any new radio service which is established, this may include operators through to equipment retailers.

3.4.2 The radar user community

The benefits to the current radar user community are:

1. The implementation of these technologies will improve system robustness and sensitivity by reducing the level of interference between radar systems.
2. Implementation of these technologies will ensure conformance to new ITU-R standards for interference and spurious emissions, due for release in 2008.
3. Adopting these technologies will release bandwidth that the user community will be able to trade, through the spectrum trading initiative [5], thereby generating income.

The spectrum trading initiative will allow holders of Wireless Telegraphy (WT) Act licences to buy and sell all or part of their rights to use the spectrum. Trading may be conducted on the basis of an outright transfer, a lease or hire. The current assumption is to commence trading in 2004 by rolling out the scheme band by band over a four-year period. It has been recognised that there are some special cases where safety of life is paramount. Therefore, trading for these services will take longer to establish. It is the aim of Ofcom to allow everyone the opportunity to trade, although there will be no obligation for a licensee to trade.

Marine and aviation services are special cases where the service they provide is directly linked to safety of life. Not only are there safety implications but also International harmonisation to consider. The proposal is to have agreement for introducing trading for ground based radionavigation aids by 2007-2009.

It is understood that trading should only proceed with the agreement of the relevant regulators and International harmonisation practices may have to be amended, which can take up to 7 years.

For other non-safety services, the scheme will be less complicated. A trader will issue the rights to use a licence with an undertaking that the new user ensures protection from interference as required by the owner. Non compliance means that licences can be revoked without compensation.

In an attempt to harmonise the radio spectrum, Ofcom have established links with MoD, who have allocations throughout the spectrum, to discuss issues such as improving spectrum utilisation. This is carried out through the Cabinet Office Spectrum Strategy Committee. It is Ofcom's proposal that the MoD should be able to realise the benefit of trading parts of the spectrum that it controls. For aviation radar this is essential because both civil and MoD radar services currently co-operate to provide the UK air picture.

In order to make an assessment of the value of spectrum trading to the radar community, examples of the likely revenue to be generated for a transfer of licences for a different use has been made. The spectrum trading consultation document [5] has set out the potential annual benefit of spectrum trading by licence class.

The assumptions made in [5] are:

- For a simple transfer of ownership with no change of use, this results in a gain of economic value of 25% of the value of the licence to the previous owner. This is taken to be the minimum likely gain.
- On average 8% of licences are to be traded each year.

When a trade involves a change in use of the licence, then it is much more difficult to estimate the financial benefit, as it will be determined by the buyers and sellers valuation of the spectrum in its new use.

Table 3-5 presents an estimate of the benefit of spectrum trading for each of the bands of interest based on the service that is likely to take up any new allocation.

Spectrum Trading	Licence Product	Licences	New (1) Use	Current (1) Value (£k)	Trade(2), (3) Value (£k)	Benefit (4) (£k)
Band 1	ATC: En-route	11	Cellular: GSM	84,500(1)	18,590	18,590
Band 2	ATC: Approach	36	Cellular: GSM	84,500(1)	60,840	60,840
Band 3	ATC: ASDE	3	Fixed: 13-20GHz	16.007(1)	0.96	0.96

Table 3-5 An estimate of the benefit of spectrum trading on a per annum basis

(1) Data taken from Spectrum Trading Consultation [5].

(2) Assume that the value of each traded licence is equal to the current value of the licence for which is being traded.

(3) Based on 25% increase on the current licence value and an 8% trade of licences each year.

(4) There is a question on whether cellular mobile could use the spectrum released. Firstly, current 3G licence holders might have a legitimate expectation that no new competing cellular operators would be licenced during their licence period. Secondly, there is other internationally agreed expansion spectrum for 3G, and manufacturers will make equipment to work in these expansion bands and are unlikely to do so for other bands. Therefore the value of other spectrum for cellular may be much smaller than stated in these tables. There may even be no demand for radar spectrum from cellular operators. This area needs further study as it impacts on many aspects of efficient spectrum utilisation in the UK.

3.4.3 Ofcom

It is Ofcom's duty to ensure the optimal use of the electromagnetic spectrum. By improving spectrum utilisation through the implementation of technology, Ofcom is benefiting by achieving its mission statement as well as providing the opportunity for spectrum trading.

3.4.4 UK consumer/citizen

The final beneficiary from the introduction of a new service is the UK consumer. Without a requirement a new service would not attract investment, therefore a service where there is a requirement will have a benefit.

With the introduction of a new service there would be two separate beneficiaries; the individual who would use the service and companies who would use the service to further their business.

By referring to recent CBA publications relating to digital switchover [6] and the economic impact of radio, 2002 update [7], it is possible to get an agreed value of the likely benefit for the introduction of the proposed services in this report. It is assumed that only one service will take up the allocation that is made available from the introduction of new technology, however in practice, there could well be more.

Table 3-6 presents a summary of all benefits for each of the bands of interest.

Benefit	Band 1 (£M)	Band 2 (£M)	Band 3 (£M)
UK industry:			
Existing systems: Option 1(1)	1.1	3.8	0.08
Existing systems: Option2(1)	11 - 275	36 - 900	0.6 – 1.2
Introduction of new systems(2)	676 p.a.	676 p.a.	18(3) p.a.
Radar user community(4)	18.6 p.a.	60.8 p.a.	0.001 p.a.
Ofcom(5)	-	-	-
UK consumer/citizen(6)	11,966 p.a.	11,966 p.a.	313 p.a.

Table 3-6 Summary of benefits

(1) Assume a producer surplus equal to 10% of the cost to implement option proposals.

(2) Taken from The Economic Impact of Radio, 2002 Update [7], which is a producer surplus to those who sell the services or goods to other companies within the radio industry.

(3) No data available so assume the same ratio of producer/consumer surplus as for band 1 and 2.

(4) Data taken from Table 3-5.

(5) There is no direct financial benefit to Ofcom for improving spectrum utilisation. All Ofcom financial benefits are accredited to UK consumer/citizen.

(6) Taken from "The Economic Impact of Radio, 2002 Update" [7], which is a consumer surplus based on a combination of both private and business benefits.

Note: The producer surplus figures used to calculate the benefit of the introduction of new systems and the consumer surplus figures used to calculate the benefit to UK consumer/citizen are average values, and these will overstate significantly the value of additional services - the marginal value. This is because the marginal value of any service typically declines as more people use the service, because those who value it most are the early users and those who value it least typically are the last to subscribe. Both the value to consumers, and to producers from offering additional services to what we have already will be lower than the current average. More recent figures on the marginal value of the spectrum can be found in a recent report [8]. Further investigation, outside the scope of this report, will be required to ascertain the effect of this.

Spectrum trading for ATC radar is unlikely to be agreed until the end of the decade at the earliest. In order to get an estimate of how the costs and benefits are going to occur over a period of time a model has been developed to calculate the NPV for future costs and benefits for the years 2008 to 2015:

Band 1	2008 (£M)	2009 (£M)	2010 (£M)	2011 (£M)	2012 (£M)	2013 (£M)	2014 (£M)	2015 (£M)
Filters Opt 1	(2.9)	(2.8)	(2.7)	9866	9521	9188	8866	8556
Filters Opt 2: Min(1)	(16.5)	(15.9)	(15.4)	9863	9517	9184	8863	8553
Filters Opt 2: Max(1)	(19.1)	(18.4)	(17.8)	9849	9504	9172	8851	8541
Waveform Opt 1	(3.0)	(2.9)	(2.8)	9866	9521	9188	8866	8556
Waveform Opt 2: Min	(16.5)	(15.9)	(15.4)	9863	9517	9184	8863	8553
Waveform Opt 2: Max	(19.1)	(18.4)	(17.8)	9849	9504	9172	8851	8541
Filters + Waveforms Opt 1	(3.0)	(2.9)	(2.8)	9866	9521	9188	8866	8556
Filters + Waveforms Opt 2: Min	(16.5)	(15.9)	(15.4)	9863	9517	9184	8863	8553
Filters + Waveforms Opt 2: Max	(19.1)	(18.4)	(17.8)	9849	9504	9172	8851	8541
Ultra narrow band: Min	(17.2)	(16.6)	(16.0)	9859	9514	9181	8860	8550
Ultra narrow band: Max	(31.7)	(30.6)	(29.6)	9780	9438	9108	8789	8481
Do nothing	0	0	0	0	0	0	0	0

Table 3-7a NPV (2004) for all options in band 1

(1) Min and Max defines the spread of cost related to the introduction of a new technology.

Band 2	2008 (£M)	2009 (£M)	2010 (£M)	2011 (£M)	2012 (£M)	2013 (£M)	2014 (£M)	2015 (£M)
Filters Opt 1	(9.4)	(9.1)	(8.8)	9899	9553	9218	8896	8584
Filters Opt 2: Min	(54.1)	(52.2)	(50.4)	9888	8542	9208	8885	8574
Filters Opt 2: Max	(41.6)	(40.2)	(38.8)	9862	9516	9183	8862	8552
Waveform Opt 1	(10.9)	(10.5)	(10.2)	9899	9553	9218	8896	8584
Waveform Opt 2: Min	(54.1)	(52.2)	(50.4)	9888	8542	9208	8885	8574
Waveform Opt 2: Max	(41.6)	(40.2)	(38.8)	9862	9516	9183	8862	8552
Filters + Waveforms Opt 1	(9.9)	(9.6)	(9.2)	9899	9553	9218	8896	8584
Filters + Waveforms Opt 2: Min	(54.1)	(52.2)	(50.4)	9888	8542	9208	8885	8574
Filters + Waveforms Opt 2: Max	(41.6)	(40.2)	(38.8)	9862	9516	9183	8862	8552
Ultra narrow band: Min	(56.2)	(54.2)	(52.3)	9877	9531	9197	8875	8565
Ultra narrow band: Max	(104)	(100)	(96.9)	9618	9282	8957	8643	8341
Do nothing	0	0	0	0	0	0	0	0

Table 3-7b NPV (2004) for all options in band 2

Band 3	2008 (£M)	2009 (£M)	2010 (£M)	2011 (£M)	2012 (£M)	2013 (£M)	2014 (£M)	2015 (£M)
Filters Opt 1	-	-	-	-	-	-	-	-
Filters Opt 2: Min	-	-	-	-	-	-	-	-
Filters Opt 2: Max	-	-	-	-	-	-	-	-
Waveform Opt 1	(0.21)	(0.20)	(0.19)	258	249	240	232	224
Waveform Opt 2: Min	(0.91)	(0.88)	(0.85)	258	249	240	232	224
Waveform Opt 2: Max	(0.95)	(0.91)	(0.88)	257	248	240	231	223
Filters + Waveforms Opt 1	(0.21)	(0.20)	(0.19)	258	249	240	232	224
Filters + Waveforms Opt 2: Min	(0.91)	(0.88)	(0.85)	258	249	240	232	224
Filters + Waveforms Opt 2: Max	(0.95)	(0.91)	(0.88)	257	248	240	231	223
Ultra narrow band: Min	-	-	-	-	-	-	-	-
Ultra narrow band: Max	-	-	-	-	-	-	-	-
Do nothing	0	0	0	0	0	0	0	0

Table 3-7c NPV (2004) for all options in band 3

Assumptions:

- A discount rate of 3.5% p.a. following Treasury guidance in the “Green Book” has been used.
- The cost of upgrading existing radar systems will be spread over a 3-year period from 2008.
- The cost of introducing new radar systems is spread over a 25-year depreciation period from 2008, which equates to the expected life of a new radar system.
- Existing radar replaced by new systems will be written-off over a 3-year period from 2008 equating to the period in which new systems are introduced.
- Producer surplus from upgrading existing radar systems will be spread over a 3-year period from 2008.
- Spectrum trading will not be able to commence until all systems have been fully implemented from 2011.
- Filters and ultra narrow band radar are considered impractical for band 3.

4 Conclusions

4.1 Cost benefit analysis

It has been shown that each technology investigated has the potential to reduce the current occupied bandwidth of the radar systems in all of the exemplar bands analysed.

It has also been shown that there is a potential benefit to be gained by releasing spectra to other services. The amount of benefit will be determined by:

1. The choice and application of technology that will be implemented to improve spectrum utilisation.
2. The service that will take advantage of any spectral release.

The benefit to the user community would depend on the value at which it could trade licences that are freed up by implementing these technologies.

The NPV for all technologies and for all bands is dominated by:

1. The cost of radar re-qualification for retrofitting the technology: Option 1.
2. The cost of development and re-qualification for a newly design radar: Option 2.
3. The benefit to the UK consumer/citizen.

It is clear from the cost break down that, for retrofitting any technology in bands 1 & 2, the overwhelming portion of cost is due to the re-qualification of the equipment. However, the cost of retrofitting is an order of magnitude less than the cost of re-designing a new radar to incorporated the techniques.

From the frequency bands selected, according to the radar industry, there is less pressure to release spectrum in band 3: 15.40-15.70GHz. There are only a small number of radar systems operating in this band and there are proposals to move to a lower frequency because of component costs. Since the development cost for an ultra narrow band radar solution is high and it is technically difficult to use filter technology, then as there are only a small number of systems to consider, the only practical solution for band 3 is the application of waveforms.

In order to get the best value from spectrum trading a defined bandwidth to trade across the UK would be necessary. This would require all radar systems within the relevant band to comply, including any military applications.

Although military radar systems have not been considered in any great detail, it is clear that they are embedded within the whole of the 1-16GHz frequency range. Military radar systems, for issues of National security, could not be treated in the same manner as civilian radar systems. Therefore in order to understand the true benefit and practicalities, further dialogue with MoD Defence Spectrum Management should be conducted.

Without MoD co-operation it is going to be difficult to release defined useable bands within the 1-16GHz frequency range to allow spectrum trading.

The timescale for introduction of the proposed technologies are:

- Filters (F) – which is classed as a near term solution.
- Waveforms (W) – which is classed as a near term solution.
- Filters + Waveforms (F+W) – which is classed as a near term solution.
- Ultra narrow band (U) – which is classed as medium term solution.

The definition of near term being implementation within the next five years and medium term within the next five to fifteen years. The timescale for introducing filters and waveforms is relatively short term. Considering ATC radar, changes would have to be made serially and could take anything up to 2-3 years due to the re-qualification process.

4.2 WP2: Filter technology

The application of fixed frequency waveguide filters has the potential to reduce in-band and out of band spurious emissions from L-band, S-band and C-band radars, with an acceptable insertion loss (typically 0.2-0.5dB). At these frequencies the likely power handling without recourse to expensive pressurisation or other flashover protection is acceptable. However, high power filtering above X-band is problematic. The increase in frequency increases the Q factor required to achieve the filtering function. This necessitates the use of more filter sections to ensure the power handling requirements can be met.

For reasons of cost compared to solid-state transmitters, the magnetron tube transmitter is likely to remain the output device of choice for X-band marine systems for the near future. For modest output powers, filtering also offers an effective solution to emission control in this band. This is tempered by the physical constraints of the enclosure size typically used by marine radar. However, improvements to magnetron modulator design to ensure that the magnetron is kept in the pi-mode throughout its life would significantly improve the spurious emissions.

4.3 WP4: Ultra narrow band radar

The key conclusions from this study are:

- Ultra narrow band radar has the potential to be a replacement technology for a number of existing civil radar applications
- The potential spectrum saving is very large with at least 90% of existing pulsed radar spectrum being released.
- Although current studies have focussed on fixed radar installations, ultra narrow band radar techniques may be applied to marine navigation as a replacement for the current magnetron based systems. However, there are two principle aspects of performance that must be addressed. The first is the restricted sighting of the multiple receivers that will determine the accuracy to which targets can be located. The second problem is that of sea clutter. It should be noted that there are FMCW systems currently in use in a maritime environment, therefore offering some evidence that sea clutter problems are surmountable.

Considerably more research is needed on practical implementation of this technology before its viability can be established.

4.4 WP6: Waveform technology

Methods for reducing radar's spectral occupancy whilst retaining its performance have been investigated. Methods for mitigating the effects of interference from other radars and communication systems are also considered.

The methods investigated form a portfolio of different techniques for spectral efficiency which cover:

- Simple modifications to current waveform.
- Spectrally compact waveforms and waveform that can operate with minimal frequency separation without interfering.
- Waveform families with reduced interference/cross talk.

- Radar concept changes, perhaps using adaptive waveform that change according to the varying spectral conditions, whilst using processing to recover losses in a range resolution.

Application of a selection of the waveform concepts to three exemplar bands of interest exposed a need for an assessment of the waveform spectrums through a demonstration. However, results show that considerable improvements can be made in all the bands of interest, particularly in the cases where filtered orthogonal waveforms and co-operative step chirp FM waveforms are used. Reductions of the selected radar bands are calculated to be less than half of the current radar allocations.

A number of waveforms, schemes and techniques have been explored, which have the potential to reduce the spectral occupation of radars. Whilst no single panacea was found that fully met the requirements for band sharing in all circumstances, a combination of techniques could be effective. Such combinations will form complex radar systems that require design, simulation, demonstration and evaluation. As such this will be a topic for further work.

The problem of interference between systems has been addressed and a simple method was derived for calculating worst-case interference rejection requirements.

In line with current thinking, some of the current regulations have been shown to be flawed. In addition, the regulations were found not to significantly prejudice waveform with theoretically noisy spectra, and in some cases perform the opposite. This is summarised by:

- The practice of adjusting waveform characteristics to give increased allowed bandwidths, which is a common practice for radar manufactures to achieve type approval.
- Allowing spurious signals levels to be relative to peak envelope power rather than absolute levels.

A scheme is presented to show benefits of co-operation, based on step chirp waveforms. The waveforms allow a small numbers of radars to co-operate with each other by scheduling use of different sub-bands. It is shown that all the radars systems can use all of the same bandwidth without interfering with each other.

4.5 Radar industry comments

Below is a selection of comments that came out of two industry workshops. The purpose of the workshops was to get the views from both the user community and radar manufacturers on the practicalities and costs associated with implementing the techniques investigated. Comments are presented on a per band basis for each technology in turn.

4.5.1 Comments on the selection of band 1 and the application of filters:

- Filters at these frequencies are physically large devices, have an insertion loss, can cause oscillator pulling and would require isolation from the source.
  - The introduction of isolators and circulators could generate unwanted spurious as well introducing further losses.
  - Any loss >0.25dB could be an issue and would have to be agreed between user and manufacturer before any such implementation could take place.
- Two approaches could be taken:
  - Retrofitting filters to existing systems.

  - Re-designing a radar system to include a suitable filter design.
- The estimated cost to re-design, manufacture and qualify a new radar system would be in the range £10M - £50M.
- Retrofitting would only be worthwhile in radar systems that are less than 10 years old.
- Re-tuning the radar within the band should not be a major problem.
- There may be technical difficulties because of ATC multi-frequency operations. This would require a tunable or switchable system and may require a re-design of the receiver.
- Typical filter presented for this band was a 3MHz bandwidth filter with a loss of 0.5dB at centre frequency rolling of to 1dB across the band. Length of filter would be approx. 800mm and would have to be manufactured from invar and require cooling.
- In practice, the introduction of a filter should not be a major problem and have no effect on the Doppler. However, the requirement for high speed switching in ATC may introduce unwanted losses and questions reliability.
- The major cost driver associated with a retrofit in ATC radar would be re-qualification.
- For ATC there could be no rapid introduction of this technology. Any modification to the ATC radar network would have to be introduced site by site, with each site requiring its own re-qualification and the cost associated with it.

4.5.2 Comments on the selection of band 1 and the application of ultra narrow band radar:

- Technology is immature and requires further research.
- A demonstration of the principle is required.
- The technique may be operationally vulnerable to jamming.
- Three fundamental problems were highlighted:
  - At the limits of range detection and target size there was a question on whether the required range accuracy was achievable.
  - The system location accuracy relies on one target in one cell. A question was raised as to what happens when more than one target is in a cell. Investigating the issue of target separation should be a high priority.
  - CW is susceptible to noise sidebands from its own transmitter, clutter etc. and will reduce the sensitivity of the radar system.
- The location of the receivers needs investigating.
- How does synchronisation between transmitter and receivers translate into a range and separation accuracy. What happens when one of the receivers is slightly out of sync with the rest.
- There would be a cost factor of 2x to 5x that defined for a new radar design as proposed in the filter case because the development costs, given a design based on active array technology.
- Adopting a CW system would totally use up the bandwidth in which it operates and not allows any other secondary users.
- There were questions raised on range resolution for distributed and isotropic scatters.

4.5.3 Comments on the selection of band 1 and the application of waveforms:

- It would be easier to implement the proposed waveform techniques in solid state systems.
- Legacy equipment may be a problem, specifically magnetrons.
- ATC radar use TWTs that may require modification to allow some of these techniques to be implemented.
- Orthogonal codes spread the energy rather than cancel – is this acceptable or does it cause de-sensitisation.
- Implementation into ATC systems is the same as for filters.

4.5.4 Comments on the selection of band 2 for all technologies:

- There are many different types of military systems in this band.
- Filters are not as practical in this band as there would be power handling & heating problems. A typical filter would have a 20MHz bandwidth.
- ATC radar in this band is for aircraft approach, which requires higher resolution and may therefore introduce higher re-qualification costs.
- Met office weather radar does not require the level of re-qualification set out by ATC, therefore the main costs for weather radar are the component and installation costs.
- Met office does not have weather radar in this band but there may be other weather radars in the UK that occupy this band.
- All comments on ultra narrow band radar in the previous band apply to this band.

4.5.5 Comments on the selection of band 3 for all technologies:

- The validity in studying this band was questioned as it was felt that this band was not an issue.
- There are discussions on whether to move airport ground movement radar out of this band.
- Application of filters may not practical in this band, as filter bandwidth increases with frequency and efficiencies become poor.
- It is an expensive band to work in terms of component costs.
- All comments on ultra narrow band radar in the previous bands apply to this band.

4.6 Summary

In order to summarise the above conclusions; Table 4-1 provides a statement relating to the suitability of each technology to improve spectrum utilisation in the context of the specific radar application in the 1-16GHz frequency range.

Technology	Radar Application					Comments
	ATC	Weather	Coastal	Marine	ASDE	
Filter	✓	✓	✓	✓	✗	For improving spectrum utilisation: best suited for frequencies at the lower end of the 1-16GHz band. Would improve the spurious emissions of current L, S, C and X-band magnetrons. Relatively low cost, low risk and near term solution.
Waveform	✓	✓	✓	✓	✓	For improving spectrum utilisation: suitable for all frequencies but may demand modifications to some transmitter types. Relatively low cost, low risk and near term solution.
UNB	✓	✓	✓	✓	✗	Best solution for improving spectrum utilisation. Technology is immature. Marine radar may be difficult because of the restricted sighting of the multiple receivers and the impact of sea clutter. Relatively high risk, high cost and medium term solution.

Table 4-1 Summary of conclusions

✓ - Signifies the technology being suitable for the specific radar application.

✓ - Signifies the technology having potential for the specific radar application but requires further investigation.

✗ - Signifies the technology being unsuitable for the specific radar application.

5 Recommendations

5.1 Cost benefit analysis

The conclusions from the CBA have shown that there is a benefit to be gained by implementing any of the technologies to the radar systems in the exemplar bands. The CBA carried out was not a rigorous exercise, but merely an indicator to compare different technology options. It is therefore recommended that a more rigorous CBA be carried out to fully investigate the true cost and benefits of the proposals.

A band that does not have such an emphasis on safety of life would, in hindsight, have been worthwhile investigating. It is recommended that a detailed analysis of all radar bands within the 1-16GHz frequency range be carried out to determine which are the most appropriate bands for the implementation of these techniques, given the proposed timescales for rollout of spectrum trading.

It is recommended that a dialogue with MoD be established to investigate the possibilities of military systems adopting some of the proposed techniques in order to assist the process of improving spectrum utilisation.

It is recommended that, where appropriate, the technologies investigated be taken forward to demonstration on candidate systems to prove to the user community the benefit of introducing such a technology.

5.2 WP2: Filter technology

It is recommended that a demonstration be performed to test the effectiveness of filtering. Ideally this should be performed on one of the candidate radars, or in the first instance on an experimental system such as the RAL S-band radar, or other transmitter test-bed. The use of harmonic filters for the suppression of out-of-band emissions is to be fully recommended.

In order to improve the spurious emissions of magnetron radars it is recommended that further investigation be carried out to demonstrate the potential improvements with improved modulator design and addition of filters.

5.3 WP4: Ultra narrow band radar

The key recommendation from this study is that further research is needed to determine the potential performance of this technology.

This would need to include:

- Further theoretical analysis to address the following key points:
  - Target location accuracy obtainable and potential for improvement.
  - Tracking of multiple targets within a receiver beamwidth.
  - Location, synchronisation and beamshape of transmitter and receiver to ensure adequate coverage of the surveillance area.
  - Potential/cost of agile beam receivers to perform the above tasks.
  - Effect of variable multistatic geometry on above.
  - Potential interference effects on other users.
- If ultra narrow band radar techniques are to be applied to the marine navigation environment as a replacement for the current magnetron based systems, there are two principle aspects of performance that must be addressed:

- The first is the restricted sighting of the multiple receivers that will determine the accuracy to which targets can be located.
- The second problem is that of sea clutter and system sensitivity.

- Initial experimental investigation of some of the above ideas are recommended. This would include an examination of practical implementation issues such as site availability and terrain obscuration. An option would be to demonstrate the principal on a short-range system before transfer to a suitable field system.

5.4 WP6: Waveform technology

The key recommendations from this study are:

5.4.1 Optimisation and demonstration of identified waveforms and techniques

This study covered theoretical aspects of waveform design and the practicalities of implementing the waveform in civil radar systems need further examination. In addition, spectral spread and hence rejection of waveforms with different frequency shifts and bandwidths is likely to be equally dependent of the radar system as the waveform. Therefore, measurement of implemented waveforms on representative systems is essential for an assessment of spectral spread and interference.

In addition, because of the broad nature of this study it is recommended that further in-depth study and optimisation of the most promising techniques identified in this report and an investigation into the effects of combining techniques be carried out.

5.4.2 Interference rejection techniques

This study briefly reviewed the benefits of a small selection of interference rejection techniques, and found them useful in support of efficient waveform design. Therefore it is urged that a further broad investigation into Interference rejection techniques be made, to fully assess the assumptions made in this report and to investigate further benefits.

5.4.3 Co-operative/sympathetic radar techniques

A co-operative radar technique has been explained which shows the benefits of co-operation. In addition, several techniques have been introduced that can be used to adaptively use available bandwidth (or segments of bandwidth) and hence reduce interference to and from radar systems. The subjects of co-operation, dynamic bandwidth allocation and adaptation is thus a subject for further investigation and study which could enhance constrained band usage by multiple users.

6 References / Bibliography

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A WP2: Rapidly tunable filters

A.1 Introduction

Following the recommendation of Ofcom contract AY4051, the objective of this work package is an evaluation of rapidly tunable filters in tube-based frequency-hopping and array radars to preserve spectrum and provide “free space” that can be allocated to other users.

This work package is broken down into a number of activities:

- WP210: Review of technology options for tube-based radars.
- WP220: Implementation of high-Q filters into civil radar systems.
- WP230: Review of technology options for array radars.

The remainder of this appendix will discuss the key requirements and technology options (WP220) for rapidly tunable filters with particular emphasis on ubiquitous vacuum electron tube-based radar systems (WP220).

As part of this work package a detailed literature review of rapidly tunable filters has been undertaken. A summary of the findings of the literature search is given in §A2. Work package WP230 has been centred on considerations of Micro-Electro-Mechanical Systems (MEMS) technologies. This is one of the most viable technologies for future array radars in which the prospect of further integration of the Transmit-Receive modules, (e.g. integration of attenuators and phase shifters filters and RF amplifiers) shows considerable promise. Appendix B discusses MEMS technologies for use in tunable radar filters.

A.2 Summary from literature search

A.2.1 Introduction

This section summarises a search of published papers and patents to establish the state of current filter research. Discussions with different manufacturing sectors on the ability to produce filters were also carried out. The search considered current and future technology of high-Q, rapidly re-tunable filters for use in tube-based frequency hopping radars and high-Q filters for use in active array radars.

A.2.2 Literature and patent search

A search of technical papers and publications revealed many papers on fixed frequency and tunable filter technology for both tube-based devices and array radars. A patent search has provided fourteen references that are relevant to this study. This information has been used to support this work package and a list is presented in Appendix G.

A.2.3 Contact with manufacturers

The design departments of two UK filter manufactures were contacted. Below are specific comments that came out of the discussions:

Neither company provides any re-tunable filters and does not know of any other supplier. The general view is that the microwave filter market is currently driven by the satellite communications industry. Most filter requirements are determined at the design stage and purchased at the same time as the system. It was stated that it is rare for filters to be fitted retrospectively to radars. Filters used for a specific radar application must be designed to meet individual requirements and safety factors. There are many trade-offs between characteristics such as Q, power handling, physical size and insertion loss which need to be taken into consideration.

Both manufacturers are specialists in the manufacture of microwave filters, diplexers and specialize in waveguide, interdigital combline, cavity, lumped element and microstrip devices. They do not have any re-tunable/active filter capabilities. Passive filters can be manually re-tuned within the range of +/- 5%. This involved adjusting a series of tuning screws. Filters can be designed against a specific requirement but at lower end of the 1-16GHz frequency range the physical size of waveguide filters will be very large, heavy and expensive. Suspended stripline filters have many advantages; they have a coaxial input and are compact, however, stripline filters can only handle powers of a few Watts so they are ideal for use on receivers. Switched banks of these filters are available.

Many radar in the UK already have filters fitted. If there is a requirement for more stringent control of emissions, then these filters will have to be replaced with filters with a higher Q. Radars that use two magnetrons will already have filters incorporated into the diplexer. Where higher Q filters are required, it will be necessary to replace the diplexer in these cases.

For radar systems that switch between a number of fixed frequencies, a recommendation is for a bank of filters that can be switched using waveguide switches. An alternative to this is a broad band filter that covers the entire frequency range. This option would be appropriate where very rapid switching would be required e.g. a frequency hopping radar that changes frequency between pulses. However there are limitations and disadvantages with these broadband filters.

There are several concerns about using high-Q filters. The first is that there is the potential to mistune the filter with the result that the signal is reflected back to the transmitter causing damage. There appears to be little knowledge of the practical use of re-tunable filters; switching between a bank of filters or using one broadband filter with a bandwidth that covers all of the radar's frequency range are more practical options.

One of the characteristics of high-Q filters is that, due to the large number of poles in the filter, they take a long time to settle. This suggests that they may not be suitable in situations where fast re-tuning is required, as the first pulse will suffer degradation. Another concern is the phase response at the edges of the filter. This can cause group delay, which may affect the performance of the radar.

Fitting a filter on the transmitter may introduce reflections, which will be damaging to the radar. It may be necessary to use a circulator with the third port terminated. These extra components will add additional cost. The introduction of filters will also produce heat so heat dissipation must be considered.

Active array radars contain many low-power transmitting elements. In order to apply filtering techniques to this type of radar, a large number of amplitude and phase matched filters are required. This has cost, size and weight implications, and also introduces the problem of heat dissipation. The structure of the radar is such that the amplifiers and filters are co-located with the transmitting elements. As these are located high above the ground, the additional weight may cause problems if the radar becomes unbalanced. It is also very importance that any filters used are small as there is not usually very much space available in the antenna.

The filters are likely to introduce extra delay. The array must be re-calibrated to compensate for this in order to maintain the beamforming capabilities. This technique can also correct for any amplitude and phase mismatching in the filters.

Other considerations for phased array radars are that banks of switched filters are physically too big to fit into the existing structure, so re-tunable filters are required. Ideally the same filter will be used for transmit and receive instead of two separate filters, this will conserve space and weight. However, the two signals will have to be separated either with a circulator or another technique such as polarisation.

For future active phased array radar, MEMS technology may provide the most appropriate filtering solutions. MEMS filtering is discussed in Appendix B.

Both manufacturers, on request, would be able to provide a the most appropriate filter design for a defined application.

A.3 Requirements for rapidly tunable filters

A review of the database generated as part of WP6 (Waveform design), revealed that there are currently no civilian frequency-hopping radar systems. However, other spectrum efficiency techniques such as those under consideration in WP6 may lead to a requirement for rapidly tunable filters. In this context and in the context of frequency hopping the term “rapidly tunable” may considered to imply re-tuning on a pulse-to-pulse basis. Examination of the database reveals that the pulse repetition interval (PRI) is typically 400-2000pps; that is the filter re-tuning every 0.5-2.5ms.

From the complete list of radar systems WP6 has also derived a summary specification of some typical systems. An extract of this information is given in Table A3-1.

Function	**Commercial Marine S-band**	**Commercial Marine X -band**	**Leisure Marine X-band**	**ATC 1 L-band**	**ATC 2 S-band**	**Weather C-band**
Frequency	3050 MHz	9410 MHz	9410 MHz	1325 MHz	2800 MHz	5600 MHz
Output device	Magnetron	Magnetron	Magnetron	TWT	TWT	Magnetron
TX power	25kW	25kW	4kW	160kW	60kW	500kW
Modulation	Pulse	Pulse	Pulse	NLFM chirp	NLFM chirp	Pulse
PRI (pps)	1400/700	1400/700	2000/500	100/400	1000/400	1200/300
Frequency agile	No	No	No	No	No	No

Table A3-1 Typical system parameters for a variety of civil radar functions

A.4 WP210: Review of technology options for tube-based radar systems

The high peak-power of the typical systems shown in Table A3-1 almost certainly dictates the use of waveguide based filters. This is almost certainly true if the power is generated from a single source such as a magnetron as opposed to a distributed power source such as an array of solid-state power amplifiers. Waveguide filters remain the best option for high peak and average power handling capabilities combined with a low insertion loss.

A.4.1 High power fixed waveguide filters

A waveguide is essentially a hollow pipe without a centre conductor and has a finite cut-off frequency. Waveguide band-pass filters have zero transmission at cut-off, which accounts for their asymmetric frequency response (see example in Figure A4-1). The design of waveguide filters is well documented; see [1].

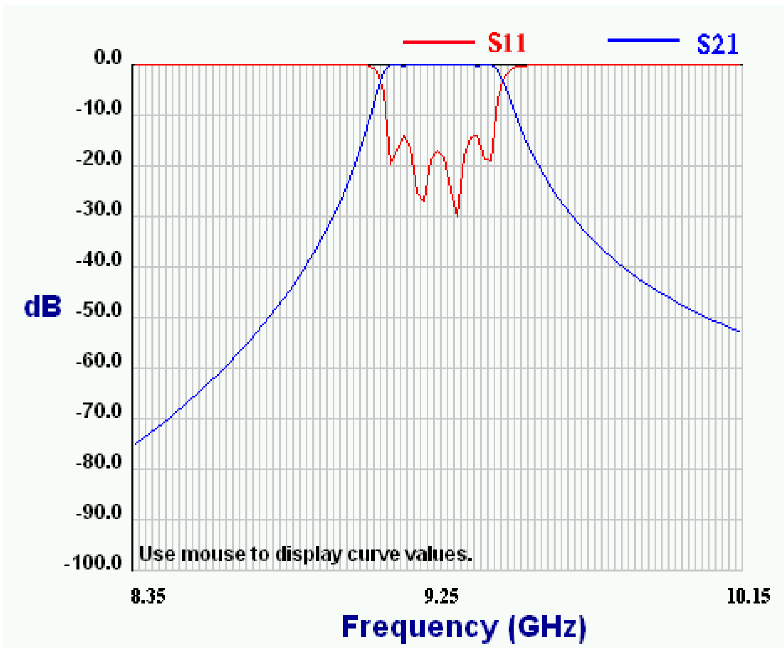

Figure A4-1: 6-pole iris band-pass filter, 9.1-9.4GHz

Some radar output devices produce significant harmonic contents (see §A4.5). These can be suppressed with so-called "waffle-iron" low-pass filters [1] which provide a wide pass-band but are capable of providing high stop-band attenuation for several harmonics. In common with almost all waveguide filters they are reflective, requiring additional components. These issues are considered in §A5.

There are a number of UK manufacturers of fixed waveguide band-pass and waffle-iron harmonic filters (e.g. Flann Microwave and BSC Filters). However, there are no known manufacturers of frequency agile filters. BSC Filters can produce iris coupled resonant cavity filters. These filters can be produced with a frequency range 1.5-75GHz and with a fractional bandwidth of between 1-10%, i.e. a minimum pass-band of ~100MHz at 9410MHz. Q-factors of 1000-20000 are quoted as being achievable. BSC Filters produce waffle-iron filters that are quoted as being capable of 100kW power handling, and stop-band out to four-times the cut-off frequency.

A.4.2 Technologies for high power frequency agile filters

This section considers the following candidate technologies for rapidly tunable filters:

- Mechanically tunable filters (via physical movement of an iris or tuning post).

- Magnetically tunable filters (via application of a magnetic field to a ferromagnetic material).
- Electronically tunable filters (via application of voltage or current).
- Frequency selective surfaces (incorporated into the antenna and/or radome).

Mechanically and electronically tunable filters

For pulse-to-pulse frequency agility, mechanically tunable filters can almost certainly be dismissed. Most forms of electronically tunable filters can also be dismissed as they do not have the power handling capability required for most radar applications. Yttrium Iron Garnet (YIG) based band-pass filters have a Q-factor of between 2000 and 5000 but they are only capable of handling powers of around 10W [2].

Magnetically tunable filters

This leaves magnetically tunable filters as the most promising type of filter technology for rapidly tunable filters. In E-plane pure metal insert filters, the insert bridges the broad walls of a rectangular waveguide. These filters offer low loss and are suitable for high average and peak power applications, making them ideal for radar applications. The inserts can be accurately manufactured using photolithographic techniques or by precision milling. One of the most successful types of magnetically tunable filter implementations are based on ferrite slab E-plane finline filters. In finline filters the metal fins are printed on each side of a dielectric substrate that bridges the broad walls of a rectangular waveguide. The ladder shaped inserts can again be manufactured using photolithographic techniques or by precision milling. If the dielectric constant of the layer is unity, the finline filter degenerates to so-called double metal insert waveguide filter. If the dielectric substrate is replaced with a ferromagnetic material (i.e. a ferrite material) the characteristics of the device can be altered by the application of a magnetic field. The agility of such devices is largely limited by the ability to rapidly collapse the applied magnetic field. Uher et al (1988) [3] present theory and measurements of a magnetically tunable 8-pole band-pass filter with a mid-band frequency of 14-16GHz. This device is tunable across the band by application of a magnetic field in the range $0-2.2\times10^{5}$ A m^{-1}. Uher and Hoefer (1991) [4] present a detailed review of the characteristics of tunable microwave and millimetre-wave band-pass filters. Table A4-1 below (adapted form the Uher and Hoefer (1991) [4]) compares mechanically tunable and magnetically tunable E-plane filters.

Performance parameter	Mechanically tunable waveguide filter	Magnetically tunable E-plane filter
Bandwidth [%]	0.3-3	1-10
Insertion loss [dB]	0.5-2.5	0.7-2.5
Selectivity [dB $octave^{-1}$]	12-24	12-24
Rejection att. [dB]	>50	>50
Max average power [W]	100-500	50-200
Tuning range [% of WG-band]	5-20	60-70
Tuning speed [GHz ms^{-1}]	Very low	$0.5-10^{3}$
Bandwidth variation [%]	10-20	5-10
Tuning linearity [MHz]	+/- 15	+/-15

Tuning stability [MHz]	20	25
Millimetre-wave capability	No	Yes

Table A4-1 Comparison of mechanically and magnetically tunable filters

It can be seen that the magnetically tunable filter compares well to the more conventional mechanically tunable filter. The key issue with the magnetically tunable filter is the size and power dissipation of the field coils required to generate the large magnetic fields. Despite this drawback, such a technology is capable of meeting likely requirements of tunable, but nominally fixed or pulse-to-pulse agile frequency-hopping radars.

The choice of radar output device used determines not only its spectral emission characteristics, but also the type of filter technology that can be employed. For example, high Q filters have, by definition the ability to store large amounts of energy; this can lead to electrical breakdown problems and excess ringing. Similarly, increasing the rise and fall times of the pulse modulator will act to reduce the bandwidth of the radiated signal. However, this can lead to improper operation with some radar output devices.

<u>Frequency selective surfaces</u>

Frequency selective surfaces (FSS) based on metallic screens have been widely used for many years with dual reflector antennas as a form of multiplexer [5]. Early work in this area dates back to the mid-1940s, and has recently been the subject of considerable work in the open literature, particularly concerning the interaction of antennas with FSS radomes [5]. A FSS can be formed from repeated periodic arrangement of simple metallic structures such as square patches [6]. In the pass-band such structures exhibit a low reflection coefficient; outside of the pass-band they have a high reflection coefficient. In the case of monostatic radar, care would have to be given to ensure adequate protection of the receiver during transmit as all out of band energy would be reflected. The behaviour of FSS structures is strongly dependent on the incidence angle. Whilst this would not be an issue for fitting over a slotted waveguide array antenna, of the sort typically used in marine radar, it could be an issue for incorporation into reflector antennas. Much of the early analysis on FSS was based on the study of infinite planar surfaces. More recently, attention has turned to conformal and multi-layered FSS (to improve selectivity). A related form of FSS is the perforated metal plate antenna whereby holes in a metal plate are dielectrically loaded. These have similar properties to the metal on dielectric FSS.

A more esoteric form of FSS is the plasma reflector [7]. In this form of FSS, a plasma is induced by ionising a gas enclosed in a vacuum such as an enclosed evacuated tube. When the plasma is induced it becomes a frequency selective conducting surface. By controlling the plasma properties it is possible to tailor the response of the FSS. A key advantage is that the plasma can be turned off by controlling the source that ionises the gas to modulate the radar pulse. In the off-state, the plasma reflector effectively is absent. This is an interesting technology. However, it is currently too immature to be considered as a viable alternative to metallic FSS for most applications.

FSS have some desirable properties; however, it could be argued that, in the case of radar transmitters at least, the problem is best tackled by improvements in modulation and a waveguide filtering. FSS conformal radomes are likely to be more expensive to produce than waveguide filters.

A.4.3 A review of ITU M.1314 and SM.329-9 regulations

This section reviews the recommendations of the ITU-R presented in M.1314 (Reduction of spurious emissions in radar systems operating in the 3-5GHz bands) and SM.329-9 (Spurious emissions). The present version of M.1314 was ratified in 1997. The ITU-R has approved a number of radio regulations purporting the emissions of radar systems (radio determination systems). The Ofcom report on contract AY4051 by AMS cites a number of these with regard to the measurement of radar emissions, however M.1314 is not reviewed. The view of the ITU-R presented in M.1314 for the prospect of rapidly tunable filters is quite a pessimistic one. Despite this, its overall conclusions are encouraging.

The ITU-R M.1314 makes the following recommendations:

- That the information on radar transmitter design factors affecting unwanted emission characteristics of radars be used to reduce unwanted emissions.
- That where practical, linear beam or solid-state radar output devices should be used in radars to reduce the non-harmonic radar spurious emission levels
- That, when necessary and where possible, radar output filters should be used to reduce non-harmonic and harmonic spurious emissions.

ITU-R SM.329-9 presents limits for different radar systems; class A and class B. The different classes apply depending on a radar function. Unlike the out-of-band limits, a level relative to the peak-envelope-power in a reference bandwidth specifies these limits rather than being specified by a mask. The reference bandwidth used by ITU-R M.1177 is specified at 1MHz. This means that the actual radar PEP must be converted to effective PEP in 1MHz to determine compliance. An ambiguity here highlighted by AY4051 concerns the measurement of PEP; on one hand it is considered to be the PEP EIRP (i.e. at the antenna) and on the other PEP (i.e. at the transmitter output).

A.4.4 Impacts of radar design on spurious emissions

The function of a radar system largely determines the design of the radar and the selection of the output device. Radar functions vary widely and generally require unique performance specifications. Some radar functions involve design factors that are not under the control of the designer, for example the radar cross-section of the target to be located or characterised. These factors directly impact the design factors such as the required transmitter power, transmitter waveform, transmitter output device, antenna gain, receiver sensitivity, range and azimuth resolution and Doppler range. The judicious trade-off between these factors to improve the emission spectrum is key to enhancing the compatibility between radar systems and other services.

Of these, the selection of the transmitter output device arguably plays the largest role as this affects not only the transmitter design, but also the receiver and antenna systems. Additionally, the design of multi-function radar can further complicate the selection of the output device. For example, there may be diverse requirements of duty cycle and average power level values.

Other major considerations in the selection of an output device may include energy efficiency (conversion of DC energy to RF energy), instantaneous bandwidth (intrinsically available bandwidth without adjustments), pulse to pulse coherency, mass, size, mechanical ruggedness, device lifetime and cost.

Table A5-1 shows the output device performance for major design factors need to be considered in the design of radar systems. As seen in Table A5-1, there is a wide variation in the output device characteristics for the major design factors of peak power, instantaneous bandwidth and energy efficiency. It should be noted that the above design factors must be given a primary consideration in the selection of the radar output device to ensure that the radar function can be achieved. In general, radar output device spurious emission characteristics are only fully considered once the other radar function objectives are satisfied. The spurious emission characteristics should of course be considered as important as the radar function.

A.4.5 A brief review of the characteristics of radar output devices

This section briefly reviews the operating characteristics and spurious emissions of the radar output devices most commonly used below 16GHz.

Cross -field devices

Magnetrons require carefully controlled voltages if they are to start properly in the correct mode. If the cathode voltage rise rate is too high the magnetron may fail to start, if the rate of rise is too low this may excite a lower current order mode. The preferred mode of operation in a magnetron, since it is the one that yields the most power, is the so-called π-mode in which the E-field in adjacent cavities of the anode has a phase difference of π radians. The constraints imposed by the conditions for correct oscillation place a limit on the amount of pulse shaping (pulse modulator rise and fall times) that can be applied to control the emission spectrum.

Co-axial magnetrons as their name suggests employ a co-axial resonant cavity around the anode. This high Q cavity couples the energy from the anode structure to the output port. Most of the energy produced by a co-axial magnetron is stored in the outer cavity this acts to reduce the stress on the anode and cathode structures which is why it generally has a longer expected lifetime. The large Q factor of co-axial magnetrons renders them unsuitable for generating short pulses, since the cavity will continue to “ring” for a period after the removal of the cathode voltage.

Crossed field amplifiers (CFAs) are small and relatively high efficiency devices that share similar properties to magnetrons. CFAs have relatively low gain and are, typically, driven by a TWT or used in cascade. Since they are power amplifiers (c.f. a power oscillator such as a magnetron) the RF drive is present prior to the application of power by the modulator. This tends to ensure that the device starts reliably and quickly in the correct mode (in this case the π-mode is undesired as it increased the tendency to oscillate) and is more tolerant of changes in the cathode voltage rise rate. Note: using the π-mode results in a slight reduction in energy efficiency (see Table A5-1).

Linear beam devices

Linear beam devices include klystrons, travelling wave tubes and hybrid devices such as twystrons. Klystrons provide high power capability and high gain. The instantaneous bandwidth of Klystron devices is typically between 1-12% although clustered cavity designs are capable of 20%. Travelling wave tubes employ slow-wave structures provided either by a helix or coupled cavities.

TWTs generally produce much lower peak powers, but have larger instantaneous bandwidths and are capable of relatively long pulse lengths. Careful modulator design is required for TWT amplifiers as an oscillation can occur on the rising and falling edge of the RF pulses. Most linear beam devices produce significant harmonic content.

Solid-stated devices

Solid-state devices can be used in radar output systems in two distinct ways. Firstly as a replacement for a high power tube in which the outputs from multiple solid-state amplifier modules are combined in a single output to feed a conventional antenna, and secondly in which each amplifier modules feeds an antenna element as part of an array.

The current generation of solid-state power amplifier devices are not capable of generating the high peak powers of tube devices. The requirement to achieve equivalent energy at the target as a tube transmitter necessitates the transmission of longer pulses, and generally pulse compression in order to meet the range resolution requirements.

The use of long pulses increases the blind range of the radar, as the receiver must be blanked during the pulse transmission. The use of class-C stages (i.e. the output stages are run in saturation) prohibits the use of amplitude shaping to control the output spectrum.

A.5 WP220: The use of radar output filters

RF filters at the transmitter output can be very useful in suppressing harmonic emissions. Filters can also be used to suppress out-of-band (OOB) and non-harmonic spurious emissions that are closer to the fundamental emission than the second harmonic. However, their utility in controlling relatively close-on portions of the emission spectrum is limited. This is partly due to their additional cost, weight, size and the fact that many radars are tunable and/or use multiple waveforms and, some of them have much wider necessary bandwidths than others have. There are key problems with the implementation of high-power RF filters that can be reconfigured to accommodate changes in carrier frequency or pulse waveform, especially when it is considered that such changes are required to occur within a few milliseconds.

Transmitter architecture is also an important determinant of the achievable degree of spectrum control. Where multiple power amplifiers are used, emission spectrum falloff rate and level are influenced by whether the outputs of those power amplifiers are combined within the waveguide or only in space after being radiated. In-waveguide combining effectively creates a severe impedance mismatch for the mutually incoherent components of the output waveforms from the power amplifiers, which can dramatically lower the radiated noise power relative to the sum of the available noise powers of the amplifiers. Conversely, array radars fed by multiple amplifiers each of which is radiated reduces combination losses but it allows all of the noise power of the amplifiers to be radiated. Current opportunities for RF filtering are also limited in such arrays. This is partly because a separate filter would be needed for each amplifier, where the number of amplifiers can be hundreds or thousands. This situation could change with advances in MEMS technologies.

Table A5-2 shows that the radar output device has a major effect on the requirement for filtering non-harmonic spurious emissions. However, as mentioned earlier the selection of the radar output device cannot be made entirely on spurious emission characteristics. Because of the inherently high levels of harmonic spurious emissions from all types of output devices, the suppression of these harmonic components is generally performed where practical. Such filters are typically based on so-called “waffle iron” designs. To mitigate the non-harmonic spurious emissions from radar bands to other nearby bands band-pass filters would be required.

The use of such filters will involve a trade-off in the radar system performance; such harmonic and non-harmonic filters will have a small insertion loss. Depending on the size (order) of the filter and the operating frequency typical loss values will be between 0.1 to 1.0dB for each. Hence, if two filters are required (one for harmonic suppression and another for non-harmonic in-band suppression) then the insertion loss will be doubled. These losses although small, may have a significant effect on the performance of a radar: a 0.2dB loss is equivalent to a loss of 47kW of peak power from a nominal 1MW radar. Alternatively a loss of 0.4dB corresponds to a 2.3% reduction of detection range, which might be significant in some radar applications. The importance of this loss assumes that the output power variability is smaller than the insertion loss. If this is not the case, the effect of the additional filter insertion loss is negligible. If the filters are connected to the output device via an isolator (formed from a “three-port” circulator and a termination) further losses will be introduced and the performance further degraded.

Power handling, size, and weight of the filter are factors to be considered in the feasibility of using an output filter on the radar, particularly in mobile radars. Size and weight can be overriding considerations in the case of mobile, active-array radars. Filtering bands close to the radar operating band requires a steep selectivity skirts and hence high energy storage, which raises the risk of breakdown (or lowers the power-handling capacity). This can also introduce phase distortion in the pass-band, which is another major consideration for active-array radars. The higher the radar power, the more attenuation is needed to suppress spurious outputs to a given level, so the more sections the filters will need, and hence the higher their insertion loss, size, and weight will tend to be.

The use of narrow bandwidth, high-Q band-pass filters will have a significant effect on waveform coding. Narrow bandwidth filters tend not to be linear phase. Clearly this will modify the waveform. This is correctable [8] in the same way that communications systems employ channel equalisers. Although it is a further complication to the overall radar system design, but it can be justified.

A.5.1 Magnetron spectra

The low-cost of the magnetron devices compared to linear beam devices (estimated by E2V to be an order of magnitude less for a typical 30kW magnetron device) means that it is likely to remain the prevalent electron tube output device. The classic problems of magnetron operation still exist. However, they are now better understood, specified and controlled. The key problems pertaining to the frequency stability of magnetrons are as follows [9]; moding, thermal drift, frequency pushing, frequency pulling, noise and time-jitter. A brief discussion of these follows:

Moding

If other possible operating-mode conditions exist too close to the normal-mode current level, stable operation is difficult to achieve. Reduction of the harmonic outputs of magnetrons can be controlled by improvements to the modulator. This is particularly true as the cathode ages, or by mode suppression techniques such as internal strapping (conducting bridges in the resonator system). The basic function of the straps is to establish a wide separation in wavelength between the π-mode and all other modes [10].

Output device	Peak output power range (kW)	Energy efficiency (%)	Instantaneous 1dB bandwidth (% of carrier frequency)	Pulse-to-pulse coherency	Weight (kg)	Size	Mechanical ruggedness	Relative life expectancy(1)	Relative cost(2)
Crossed-field:									
Crossed field amplifier	60-5,000	40-65	5-12	Yes	25-65			1.0	
Cavity magnetron (free)	20-1,000	35-75	(3)	No	1-25	Small	Good	1.0	Low
Cavity magnetron (locked)	20-1,000	35-75	(3)	Yes	1-25			1.0	
Coaxial magnetron	10-3,000	35-50	(3)	No	2-55			5.4	
Linear beam:									
Coupled cavity TWT	25-200	2-40	10-15	Yes	10-135			7.4	High
Klystron	20-10,000	30-50	1-12	Yes	25-270	Large	Good	13.5	Medium
Twystron	2,000-5,000	30-40	1-12	Yes	55-65			10.4	High
Solid state transistors									
(parallel class-C modules)									
Si BJT	10-90	20-30(5)	10-30	Yes	0.5-2.5				
GaAs FET (4)	0.5-5	10-25(5)	10-30	Yes	(per module)	Small	Excellent	15	High

(1) Life expectancy is normalised relative to a conventional magnetron.

(2) Depends on production volume.

(3) Although magnetrons do not have an instantaneous bandwidth, tuning frequency ranges up to 10% of the operating frequency can be achieved.

(4) Silicon (Si) BJT modules are generally used below 3.5GHz, and Gallium Arsenide (GaAS) modules above.

(5) Depends on the number of modules combined to provide the desired peak power.

Table A5-1 Radar output device performance for major design factors considered for radar systems operating in the 3-5GHz bands

Output device	Spurious emission level			
	Non-harmonic (dBc) in 1MHz	Harmonic (dBc)		
		2nd	3rd	4th
Crossed field:				
CFA	-35 to –50	-25	-30	-45
Cavity magnetron (free running - unlocked)	-65 to –80	-40	-20	-45
Cavity magnetron (primed - locked)	-75 to –90	-40	-20	-45
Coaxial magnetron	-60 to -75	-40	-20	-45
Linear beam:				
Coupled cavity TWT	-105 to –115	-20	-25	-35
Klystron	-110 to –120	-20	-25	-35
Twystron	-105 to –115	-20	-25	-35
Solid state transistors (parallel class-C modules):				
Si BJT	-100 to –110	-45	-55	-65
GaAs FET	-100 to –110	-35	-45	-55

Table A5-2 Typical spurious emission characteristics for radar output

Note:

1. Harmonic spurious emission levels listed are nominal values. The range of harmonic spurious emissions is typically +5 to –10dB of the nominal values.
2. Harmonic emission levels can be reduced to below –100 dBc with a low-pass filter.
3. Magnetron devices have a number of modes other than the preferred π-mode of operation. The output power in some of these modes may only be 40dB below the main carrier. These modes are intermittent and of short duration occurring during the start-up of the oscillations.
4. Non-harmonic emission levels in crossed field devices can be reduced to below –100dBc with waveguide bandpass filters. These filters require very low insertion loss (a few tenths of a dB).
5. Linear beam output devices may have non-harmonic spurious emissions close to the carrier typically –80 to –90dBc depending on the selectivity of the cavity arrangement.
6. Magnetron information is based on older magnetron design data.

Thermal drift

The magnetron frequency varies with ambient temperature according to the temperature coefficient of the anode structure. There may be significant variations on start-up, but typically stabilise after around 10-30 minutes.

Frequency pushing

The frequency of operation depends on the electron density which in turn is a function of the anode current. This implies that the driving current pulse from the modulator needs to be flat so as not to cause FM as well as AM.

Frequency pulling

This results as a trade-off between good coupling (power output) and load sensitivity. The incorporation of a ferrite isolator can substantially reduce this problem.

Noise

Once the main pulse has been produced, excessive inverse voltage or even small forward post-pulse voltages may be applied to the magnetron; this produces noise. Noise can also be generated by the modulator falling edge on turn-off. The modulator supplies energy and the magnetron itself have significant energy storage, which increases the turn-off time. One solution, covered by patent (US4424545, 1984), involves additional switching within the modulator to rapidly dump unwanted pulse energy into a capacitor.

Time jitter

This is due to the random variations in delay from the application of the modulator pulse, to the set-up of fields and generation of the RF pulse.

A.5.2 Correct modulation

A key requirement to maintain spectral purity throughout the life of the magnetron is to ensure that it is subject to a correct modulation. In this context, the term correct can be taken to imply constant current pulses. The correct modulation reduces the probability of undesirable modes and hence undesirable spurious emissions. The provision of correct modulation implies driving the magnetron with a constant current (E2V, 2003); constant current implies constant electron density. A result of this is that as the magnetron ages and the cathode and anode coatings deteriorate, the modulator voltage is required to increase. The frequency of operation however does change with time due to the deposition of the cathode coating on the anode structure which changes the resonant frequency.

Unless the magnetic assembly is struck or subjected to other magnetic disturbances it should maintain its field constant over the life of the device (fundamentally determined by the cathode lifetime) (Collins, 1948). The reported deterioration in quality of output spectra of magnetrons is almost entirely due to poor modulation (as the anode and cathode deteriorate) rather than changes to the magnetic circuit.

A.5.3 Cleaning up magnetron spectra

A number of solutions exist for cleaning-up the spectra of magnetrons, including adding additional filtering. A large step to cleaning up magnetron spectra, particularly as the device ages, is the correction modulation. Low-pass filters can adequately suppress harmonic contents. Once these problems have been addressed, should further suppression of spurious output be required, then in-band filters can be added. However, the connection of a relatively small fractional bandwidth, high-Q filter to a magnetron device is not without problems. In order not to contribute to the problem of frequency pulling in magnetrons, it is necessary to use an isolator between the output device and the filter. Additionally due to the reflective nature of the filters, this function is best performed by a three-port circulator and a termination (see Figure A5-1), as out-of-band energy would otherwise be reflected back into the magnetron, or directed back towards the antenna.

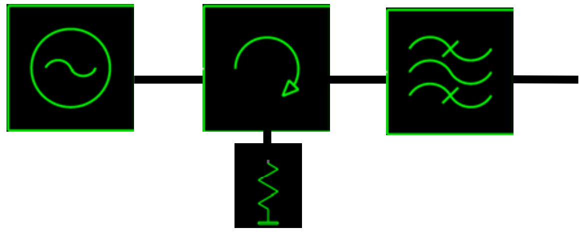

Figure A5-1: Additional circulator and termination arrangement required for additional filtering

A.5.4 Filter design considerations

In principle at least it is possible to design a filter that is as narrow as the "necessary bandwidth" defined by the ITU-R (SM.1138-0, 1995, SM.853-1, 1997, SM328-10, 1999 and SM1541-1, 2002. The filter design is, as discussed previously, is complicated by the management of a number of issues. The major issues are:

- Ohmic losses – The insertion loss of the filter needs to be minimised so as not to substantially degrade the radar function. This may lead to a requirement to silver plate the interior surfaces of the waveguide filter.
- Power handling – This is related to the energy storage, or Q factor of the filter. Reducing the Q factor of an individual resonator implies the use of a larger number of filter sections, which implies higher insertion loss. Given the relatively low average powers it is likely that forced air cooling of waveguide filters will be satisfactory.
- Temperature coefficient – The use of high Q sections implies close tolerance cavity dimensions. It is likely that such filters will have to be fabricated from a low thermal expansivity material such as invar.
- Size – To achieve the peak power requirements, typically of the order of 100-200kW, it is necessary to restrict the Q factor of the waveguide (unless other measures e.g. waveguide pressurisation are used). The use of lower Q factor sections as mentioned previously necessitates the use of a larger number of filter sections. In the case of the L-band and S-band ATC radars this is not seen as problematic. Fixed iris filters of the type considered here are of the order of 1-2m long, which can be accommodated within the standard waveguide runs from the transmitter.

As an example, BSC Filters estimates that a typical filter for L-band with a 3MHz bandwidth, 120kW peak power handing and 0.5-1.0dB insertion loss across the pass-band, fabricated from invar will be approximately 800mm long. The design of filters will need to be considered on a case-by-case basis, as the issues are hard to generalise. A further complication occurs with ATC radars employing frequency diversity, however these are tractable by incorporating appropriate filters and duplexers. In almost all cases, the capital costs of designing and manufacturing an individual filter is likely to be of the order of £5k to £30k including appropriate circulators. A larger issue is the other costs involved associated with re-qualification in the case of safety-of-life services such as ATC.

The application of filters to reduce spectral occupancy down to the necessary bandwidth for a given emission type represents an upper bound to the spectral efficiency achievable. Assuming that this is achievable typical reduction in bandwidth requirements are:

- L-band NLFM ATC radar, reduction of bandwidth by a factor of 4.

- S-band pulsed ATC radar, reduction of bandwidth by a factor of 20.
- Ku-band pulsed ASDE radar, reduction of bandwidth by a factor of 10. Note that the addition of filters at Ku-band is complicated by the power handling and short pulse (~40ns) requirements. Filters are unlikely to be a practical solution for Ku-band ASDE radars.

A.6 Impacts of filters on waveform design

The choice of pulse waveform type and the way in which the waveform is shaped can have an important influence on spectrum control and hence compatibility and inter-operability. This is recognised and is considered in WP6. Most radars, especially those using a single power oscillator (e.g. a magnetron based marine radar) or power amplifier, are subject to constraints of heat dissipation and energy efficiency. Such systems ostensibly use pulses that have a constant envelope (a summary of issues arising in magnetron systems has been presented in §A4.5).

Radar waveforms can be categorised at the first level into either unmodulated pulse waveforms or intra-pulse modulated waveforms. Intra-pulse modulation is usually employed to achieve some form of pulse compression, or some form of band sharing schemes (as is the subject of WP6). However, intra-pulse coding can also be used as a means to drive a frequency-steered array antenna.

Intra-pulse modulation can be further categorised into the following subcategories:

- Continuous FM or “chirp” pulses.
- Stepped frequency pulses (e.g. for use in frequency steered arrays).
- Digitally coded pulses (e.g. Barker sequence).

From the pure viewpoint of emission control, the guiding principle is to ensure smoothness of the derivatives of the waveform, since it is this that governs the ultimate spectrum roll-off. The various waveforms therefore differ by their differences in the amplitude, phase and frequency within the pulse.

All pulses have finite rise and fall times. One might be tempted to employ smooth and gradual rises and falls from the pulse modulator in order to reduce the spectral occupancy. However, as discussed in §A4.5, crossed-field devices have constraints on the rise and fall speeds if spurious modes of oscillation are to be avoided. Even when amplifiers or oscillators other than crossed field devices are used there may be problems. In order to limit power dissipation in solid-state amplifiers for example, the output transistors usually operate in saturation (class-C) even though spurious oscillations are no longer a concern (for a properly designed amplifier).

Continuous FM techniques by definition are phase continuous and so offer good spectral efficiency, these are subject to detailed consideration in WP6. Stepped frequency pulses, generally have poorer emission spectra than continuous FM techniques as a consequence of discontinuities in phase. It is feasible to remove these transitions in such a way that removes the discontinuities in phase. However, discontinuities are likely to remain in the first and higher derivatives, which do not occur in continuous FM waveforms.

Digitally coded pulses usually have discontinuities in phase (in common with BPSK and QPSK communications waveforms) that result in relatively poor emission spectra. In principle this can be improved via baseband (modulating) waveform or by applying filtering at IF or indeed RF. When baseband filtering is performed, the chip-to-chip transitions are made gradual rather than abrupt, but when using coding schemes that involve 180 degree phase shifts (i.e. traversing through the origin on a I-Q phase plot), this can lead to the signal envelope briefly being zero. Again, this may cause problems with crossed-field power amplifiers, but can also lead to addition phase modulation due to the inherent AM-to-PM conversion that tends to occur in most power amplifiers. The rapid phase transitions lead to unwanted sidebands and a spectral skirt that rolls-off at only 20dB/decade. Assuming that the baseband signal can be modified so as to provide a continuous phase signal (akin to Minimum Shift Keying used in communications systems). This would result in the phase being continuous, but its derivative would still contain some discontinuities. This results in the ultimate roll-off being 40dB/decade. In many communications systems (e.g. GSM) it is common to perform both baseband filtering and MSK. This is, theoretically at least, possible with radar waveforms too. This would result in the ultimate roll-off being better than 40dB/decade.

It is possible to continue to improve the spectral occupancy almost arbitrarily. However, the ultimate quest for improved spectrum efficiency cannot be pursued without consequences in the time-domain (i.e. range resolution and Doppler). As an extreme example, a linear-FM rectangular pulse waveform with an infinite time-bandwidth product (infinite pulse compression ratio) would have a perfectly rectangular spectrum (aside from the rise and fall ramps). The response of a matched filter would have a sinc (sin(x)/x) shape with range-time sidelobes only 13dB down on the peak. This would be insufficient for many applications (e.g. weather radar). The matched-filter response is not just the Fourier transform of the emission spectrum. For chirped systems a high spectrum roll-off is usually accompanied by high range sidelobes, in a similar manner to low range sidelobes result in high sidebands in the frequency domain. Hence, as with all systems, there will be an inevitable trade-off between spectrum efficiency and radar function.

A.7 Conclusions and recommendations of WP210 and WP220

The findings of this study are summarised as:

- For crossed-field devices little can be done to change the modulator system, in terms of changing pulse rise- and fall-times of the pulses with a view to controlling output sidebands. However, it appears that there are gains to be had from necessitating a change to improved constant-current modulator design.
- High-power, high-Q RF filtering remains the best option for improving spectrum efficiency, once issues of correct modulation have been addressed. That is of course provided the issues of insertion loss etc do not significant degrade the radar function.
- Solid state devices with waveform coding perhaps offer the best opportunity for achieving the ultimate in spectral efficiency. However, there are distinct cost disadvantages to this approach.
- A large range of options are available for improving the spectrum efficiency of linear beam devices, in particular through the use of appropriate waveform coding. Again, filtering remains a key option with similar caveats to those for crossed-field devices applying.

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B WP230: MEMS technology for tuneable filters

B.1 Introduction

In recent years significant progress has been made in the use of Micro-Electro-Mechanical Systems (MEMS) technology to realise a range of RF components. These include tuneable capacitors, high Q micro-inductors, micro-mechanical filters, RF switches and resonators based on bulk acoustic wave (BAW) and cavity structures.

Whilst many of these devices are aimed primarily at the communications market some are likely to find uses in radar applications. A specific example of this is the use of MEMS RF switches in phases array antennas, which are being developed on a number of, largely DARPA funded, programmes in the USA.

The purpose of this report is to review the state-of-the-art of RF MEMS components against key performance criteria identified at being of interest to the array radar community for rapidly tuneable filters, and to provide some insight on the potential of these devices to impact this area. These key criteria include:

1. Operating frequency – spectrum of interest covers 1-16GHz
2. Switching/ tuning time
3. Filter Q/bandwidth
4. Insertion loss
5. Reliability/ power handling
6. Size/ cost

B.2 RF MEMS component review

B.2.1 MEMS RF switches

Significant progress has been made on the demonstration of MEMS based RF switches in recent years. Excellent reviews of much of this work are available in the open literature [1,2]. Although much of this work resides within vertically integrated systems companies in the USA, and is therefore not likely to be sold as a commodity, recent press articles indicate that a number of companies are close to market with products reflecting the growing maturity of this technology [3,4].

There are two generic types of MEMS RF switch architecture that are commonly presented in the literature - Ohmic and capacitive. (These generic classes can be further divided into what are called "broadside" and "inline" switches depending on the orientation of the RF ports and the actuation electrode [2]). The basic operation of each of these generic devices is shown schematically in Figure B2-1.

The most simple of these device structures is the Ohmic switch [5,8-12]. (This class of device is also often referred to as a "resistive" or "series" switch.) This structure uses mechanical action to open or close electrical dc contacts as shown in Figure B2-1a. Provided contact resistances are kept below approximately 2Ω the switch insertion loss will be less than 0.2dB.

In the open (off position) the isolation is dictated by the effective capacitance between the two parts of the signal line forming the input and output ports. This is governed by the capacitance through the substrate in the absence of the switch and the capacitance from each end of the transmission line to the suspended armature of the switch.

The key advantage with this device structure is it's broadband nature in that provided care is taken to minimise these capacitances switches with high isolations of the order of 30dB for frequencies up to 20GHz can be readily realised.

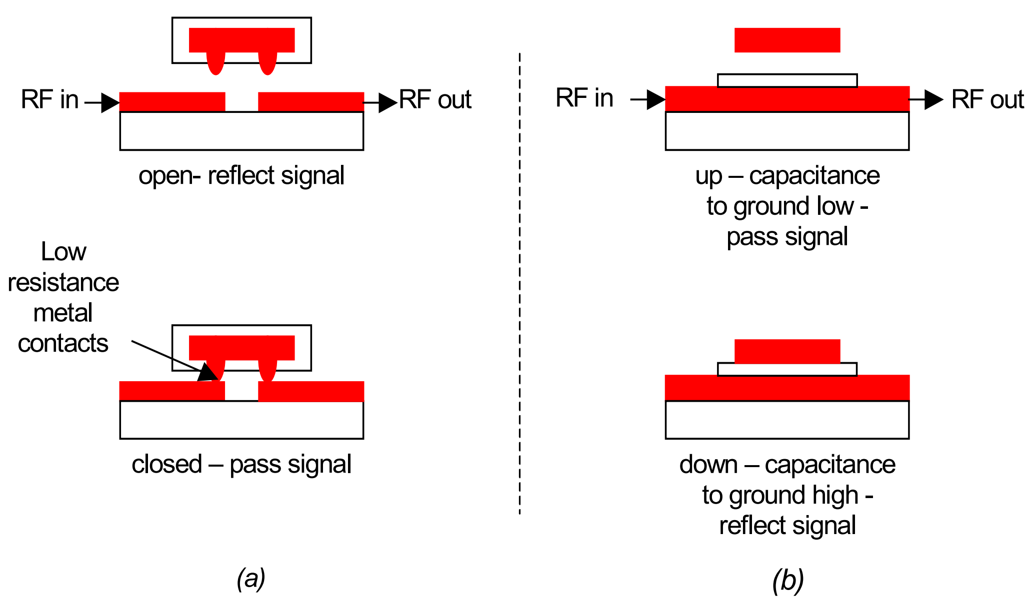

Figure B2-1 – Schematic showing basic operation of (a) Ohmic and (b) capacitive MEMS switches.

The second style of switch (Figure B2-1b) uses mechanical action to change the capacitance through the switch between the signal line and ground [13-15]. In the "up" position the loading of the capacitor on the signal is minimal and the signal is passed unattenuated. In the "down" position the signal line is effectively shorted to ground for high frequencies thus reflecting the signal back into the input port.

The ability of the switch to pass signals in the up position and reflect them in the down position is largely defined by the ratio of the two capacitors $\Gamma\left(=\frac{C_d}{C_u}\right)$ where C_d and C_u are the capacitance values in the down and up positions respectively. For good switch characteristics this need to be as large as possible and hence surface roughness within the switch [13,16] and the use of high K dielectrics [17] are key areas of research for these class of devices.

The key point to note for this class of device is that, unlike Ohmic switches, the window of low insertion loss and high isolation is relatively narrow and consequently structures have to be individually designed to operate at a specific frequency. It can readily be shown that that this class of device is most suitable for higher frequency signals, greater than a few tens of GHz, due to the otherwise unwieldy size of the MEMS switch element.

Table B2-1 presents a comparison of the key parameters of interest for MEMS RF switches along with other competing technologies. It is clear that MEMS devices offer significant advantages in terms of insertion loss, isolation and linearity over competing device technologies such as GaAs FETs and pin diodes. It is also likely, depending on the level of integration and related packaging costs, that MEMS devices fabricated on silicon substrates will have a lower cost than that GaAs based semiconductor components.

The main down side with this technology is the relatively long switching speed associated with mechanical movement of the switch, typically >2μs compared to 10-100ns for creating carriers in a semiconductor-based device. At a most simple level the mechanical switching speed of a MEMS switch is limited by the mechanical resonant frequency which, to a first order, is given by:

$$f_{res} = \frac{1}{2\pi}\sqrt{\frac{k_{eff}}{m_{eff}}}$$

Equation B2-1

Characteristic	**MEMS switch**	**GaAs FET**	**Pin diode**	**Mechanical relay**
Insertion loss	~0.1dB at 4GHz	~2dB at 6GHz	0.5-1.0dB at 4GHz	~0.1dB
Isolation	40dB at 4GHz	20dB at 5GHz	>60dB	-
Switching speed	>2μs	~10ns	~100ns	-
Harmonic production	Negligible	High	High	Negligible
Power consumption	Very low	Low	Low	High
Drive voltage	5-50V	3V	3V	12V
Size, large dimension	~100μm	10-100μm	40μm	10mm

Table B2-1 – RF switch technology comparison

Where k_{eff} and m_{eff} are respectively the effective spring constant and mass of the system. Clearly, fast switching requires high values of k_{eff} and small m_{eff}. However, increasing the effective mechanical spring constant, clearly has the detrimental effect of increasing the force required to actuate the switch, which in turn leads to higher actuation voltages.

It should be noted that most published results are based on electrostatically actuated devices due to their relative ease of fabrication and low drive current. These generally have relatively high drive voltages, typically 40-60V, in order to maintain satisfactory switching speeds. In order to try and improve on this performance other actuation techniques are being considered for this application including piezoelectric actuation.

Other areas where significant work is being undertaken is on the power handling capabilities and long term reliability of MEMS based devices. Although there is relatively little published data on this topic, it is clear that significant effort is being placed to address these topics and a number of groups have published data on switches successfully demonstrating 10^7-10^9 cycles [5-7]. It should be noted that reliability data has to be carefully interpreted, as numbers of cycles for hot switching are often significantly less than that for a cold switch. For example Milhailovich et al have reported best device results of hot-switched small signal (1mA) lifetimes of tens of millions of cycles and cold switched lifetimes an order of magnitude higher.

B.2.2 MEMS RF passives

There has been significant interest in using micromachining techniques to implemented high performance RF passives for silicon based RF systems. Specific examples include:

Tuneable capacitors - A range of techniques have been developed to realise voltage tuneable capacitors based either interdigitated structures fabricated using anisotropic deep etching of silicon [18-19] or plate structures using surface micromachining [20-22]. With Q's in excess of 100 being realised for frequencies up to a few GHz the performance of these components are beginning to look promising for application in a range of RF functions including VCOs and multi-band filters. As in the case of MEMS switches the tuning time for these devices is going to depend on the mechanical resonant frequency of the structure. This is typically between 1 and a few 10's of kHz for these devices.

High Q micro-inductors – It is generally accepted that the quality factor Q for spiral inductors realised on silicon wafers is limited to values of the order of 10 for component values of the order of 1-10nH. A range of micromachining techniques have been used to improve the performance of integrated inductors. Generally, these either locally removing parts of the substrate from beneath the planar inductor coil or suspending them above the substrate [24-28]. These techniques act to both increase the self-resonance frequency via a reduction in parasitic capacitance within the coils of the inductor and increase its Q through reduced losses due to eddy current within the substrate. Recently there has also been interest in building out-of-plane inductors [29]. Figure B2-2 shows the general trend in Q_{max} for these inductor technologies as a

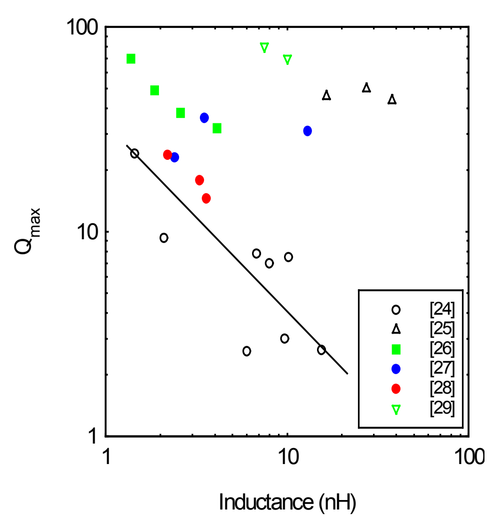

Figure B2-2 Comparison of inductor Q_{max} for conventional design techniques [24] and MEMS techniques [25-29].

function of inductance compared with those using more conventional design and fabrication techniques. It is clear that significant progress is being made in this area with Q's beginning to approach 100 making them more interesting to the RF filter design community.

Currently, there is very little data available on the power handling and reliability of MEMS based RF passives. Consequently it is difficult to advice on the likely impact of these components in array radar applications.

B.2.3 MEMS based FBAR and active resonators

For frequencies in excess of 3-4GHz or for applications requiring higher Q components than can be readily realised using passives, FBAR or cavity resonators become attractive.

In the first of these device types micromachining technologies are used to improve the performance of bulk acoustic devices by reducing losses due to acoustic emission into the substrate. This is done by removing either part of the silicon substrate or a sacrificial layer from beneath the device to leave a quasi free standing structure with two near ideal reflecting boundaries [30]. These devices are commonly using in mobile phone applications and although there have been concerns expressed about the mechanical reliability of these devices under high shock levels companies such are Agilent are shipping components such as duplexers and filters based on this technology.

Whilst most of the work reported to date has concentrated on the 1-2GHz frequency range a number of groups are interested in extending the operating frequency of these devices up to 5-10GHz [31,32] which represents a significant part of the frequency spectrum of interest to the array radar community.

Although the power handling capabilities of these devices meet the requirements of the mobile phone industry no data has been found to date to indicate their suitability for handling the power levels likely to be required for radar applications.

Micromachining technology is also being used to realise a range of cavity resonators operating in the 5.7-35GHz range [33-36]. Together with the technologies discussed previously, would cover the complete frequency spectrum of interest. This is a relatively new area of work and reported Q's generally fall well below predicted values and there is no data reported on power handling to enable a full prediction of the likely impact of this technology.

B.3 Tunable filters using MEMS components

It is generally accepted that the components within a given bandpass filter should have Q's that are 20-25 times the Q ($=f_0/\Delta f$) of the filter. This means that the use of filters based on the use of MEMS capacitors and micro-inductors are currently relatively limiting due to the low Q of inductors.

Tuneable filters based on MEMS capacitors with off-chip inductors have been reported with centre frequencies up to 2GHz [37] and RF circuit functions such as phase shifters based on semi-lumped circuit implementations using MEMS capacitors lower-loss microstrip transmission lines to replace lumped inductors have been realised [38]. At higher frequencies, the use of MEMS switches has been used to tune the bandwidth and centre frequency of bandpass filters [39-42].

Clearly, as well as designing individual filters with a tuneable frequency response MEMS switches could be used to direct signals through one of a bank of predefined fixed response filters – see Figure B3-1.

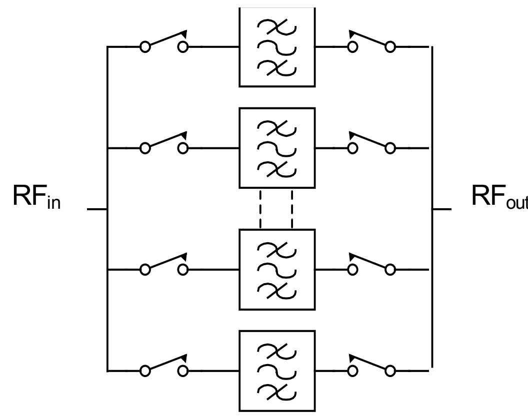

Figure B3-1 – switched filter bank using MEMS switches to provide tuning function.

B.4 Conclusions and recommendations of WP230

Improvements in the performance of a range of RF components using MEMS/ micromachining technologies appear to have a significant future in helping to realise RF functions for future RF systems especially where a reduction in size is important.

The evidence in work published in the open literature shows that these technologies show a lot of promise for a number of applications. Based on evidence available to date, tuneable filters based on MEMS technology that cover the frequency range of interest will be available to the array radar community. Although the technology is maturing with components for the telecommunications market becoming available there are a number of questions that still need to be addressed before a definitive answer can be given as to their applicability to the problem of tuneable filters for applications in array radar systems.

Two of the key areas of concern for this application are switching speed and the ability of this technology to handle the high power signals seen during the transmit operation.

The tuning time for MEMS based filters systems is unlikely to be much faster than 1-2ms due to the relatively slow mechanical response of these structures compare to, for example, semi-conductor based RF switches.

It is generally accepted that any filtering in RF circuits occurs after the power amplifier stage. Consequently the filter stage, including any MEMS components, sees the full transmit power. One route that has been proposed to ameliorate this by moving the filter to a point just preceding the amplifier. Clearly this means that the linearity of the amplifier stage has to be very good but if a system architecture is used with one amplifier per antenna channel then the maximum power that each amplifier is required to handle can be significantly reduced. To enable a full understanding of the benefits of such an approach it is recommended that a system level study is initiated to look at possible novel architecture changes that could be made to exploit the maximum potential benefits of these devices.

Finally, although the developments in RF MEMS technologies have been rapid during the past 7-8 years the technology is still at a relatively early point in its life cycle. Consequently it is currently impossible to predict how these technologies will develop in the way that one does for silicon IC technology using, for example, Moore's law. This problem is exacerbated by the divergent needs of the different potential customer base that will require components to be optimised for their own market needs. For example, high power handling for radar applications and low voltage and low power for mobile communications, and that is before any frequency consideration.

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C WP4: Ultra narrow band radar

C.1 Introduction

The frequency spectrum is a very finite resource. With the continued growth in traditional applications and emergence of new technologies, there is a need to maximise the efficiency with which the spectrum is used so that more effective management strategies may be adopted. For example, there is already considerable congestion in the L and S-bands and this is likely to grow worse. Hence a technique that enables L and S-band radar systems to operate over significantly reduced bandwidths is potentially very attractive as it can free up spectrum for communications, GSM, the new 3G services and others.

Radar systems such as ATC primary surveillance radars operate at such frequencies and hence there is a need to come up with alternatives that use less bandwidth while at the same time achieving the same performance. In addition, the problem of spurious and unwanted radiation from the high power transmitters usually used in these radar systems, can also be addressed. One approach is to use ultra narrow band continuous wave (CW) waveforms rather than the inherently wider bandwidths associated with the pulsed waveforms used in current systems.

This work package investigates and evaluates an ultra-narrow band distributed radar system as a replacement technology for current systems operating in the L, S and C-bands of the Electro-magnetic spectrum. This technique is equally applicable in X-band although pressure for bandwidth occupancy may not be so great at these frequencies. The performance is benchmarked against that of existing systems. In this way, an examination of whether or not ultra narrow band distributed radar can achieve the same performance as existing systems whilst utilising significant spectral savings, and thus offer a credible and attractive future option.

The bandwidth occupancy of current civil aviation, weather, marine navigation, coastal surveillance radar systems etc is in part determined by historical factors. These types of systems employ a high power transmitter, duplex switching and short pulses to achieve the required levels of isolation, sensitivity and location accuracy from a single radar site. It is these characteristics that give rise to the relatively large bandwidth occupancy of current systems. For example a 100ns pulse width equates to a bandwidth of 10MHz and results in a range resolution of 15m. The frequency spectrum allocations for radiolocation in L, S and C-bands are shown in §C7 and occupy more than 1GHz of these bands.

In principle, as a potentially very attractive alternative, monochromatic or very narrow band radar waveforms could be employed occupying a bandwidth of just a few Hz. This implies continuous wave operation, which in turn implies range ambiguity. That is to say an infinitely long continuous tone has no range resolving capability. In addition, the detection range may be limited due to ‘spill-over’ or ‘leakage’ of the transmitter signal directly into the receiver thus competing with target returns. This will limit ultimate sensitivity and hence the maximum detection range.

However, these restrictions can, in principle, be overcome if multiple transmitter and receiver sites are used. Three or four sites may be required in order to replace current systems. Separation of the transmitter and receivers reduces the transmitter ‘leakage’ problem and appropriately high powers consistent with the required detection ranges may be able to be employed. Multiple distributed receivers also allow targets to be located via a combination interferometry and triangulation rather than relying on a single radar beam. Accuracy will depend on the location and number of the radar receivers and will be dependent on signal to noise ratio (SNR). The location accuracy can be further improved by exploiting Doppler (which is easily determinable in a CW radar) and angle at each receiver. Indeed the combination of these parameters can be used to determine location uniquely [1]. This may be necessary for Marine Navigation Radar systems where there are limitations on the possible locations of the transmitter and the receiver. However, initial calculations suggest that narrow band CW radars could, potentially, be an alternative to current systems.

Current installations could also provide much of the necessary hardware and infrastructure to support a distributed narrow band system. For example, the parabolic dish antenna, turning gear and housings could all be re-used, thus affording a considerable financial saving. Furthermore, narrow band radar systems are inherently low cost with suitable transmitters being readily available. They also exhibit superior noise performance as a result of only operating over a narrow bandwidth. The completed system is envisaged as utilising largely 'off the shelf' components. The metric of any proposed system will be that it is able to replicate the performance of currently used systems.

This report presents initial calculations of the detection and location performance of a small number of "straw man" ultra narrow band radar systems, thereby providing an investigation into the potential of this approach as a means of improving spectral efficiency.

This report deals with the initial evaluation of narrow band systems for a range of radar systems applications. ATC, weather detection, marine and coastal surveillance radar have been chosen as being representative of applications and the likely range of performance to be encountered. A brief summary of each of these applications is followed by a theoretical analysis and evaluation of the narrowband radar system as an alternative to the pulsed ATC radar.

In addition some specific issues concerning the deployment of narrowband systems such as geometry, sight of receivers, isolation and hardware are considered. Given the timescale and effort levels available, this investigation is by no means exhaustive. The report concludes with some preliminary views on the feasibility of this technique as an alternative to current systems and some suggestions for further work to clarify these issues.

C.2 Review of current radar systems applications

C.2.1 ATC radar

Figure C2-1 ASR 23 SS - En-route Primary Surveillance Radar with co-mounted Monopulse Secondary

Figure C2-2 ASR 10 SS - Terminal Approach Primary Surveillance Radar with co-mounted Monopulse

Current ATC primary surveillance radars (PRS) generally operate in the L and S-bands (23 and 10cm). They provide en-route and terminal approach radar coverage and are usually co-mounted with secondary surveillance radars (SSRs). SSR is a data link based system operating at around 1GHz that is used for co-operative targets. This is not specifically considered in this study. The systems used in the UK are based on Raytheon's ASR-23-SS and ASR-10-SS are shown in Figures C2-1 and C2-2. A summary of selected UK and USA primary radar specifications relevant to this study are reproduced in the Table C2-1 below. Most of this data is extracted from [3-5]. Although some of these systems are in the process of being upgraded or replaced, the general specifications of future systems are similar.

RADAR	**23 cm** **en-route**	**10 cm** **Terminal Approach**	**FAA** **ASR-9**
Frequency	1250-1350MHz	2.7-2.9GHz	2.7 - 2.9GHz
Transmitter power (kW)	150	10 - 40	1.1MW
Range (nm)	80 - 210	60 - 100	60
Range resolution	250m	230m	1000m (weather)
Pulse length (μs)	3/34/66	1/75 or 1/100	1.04/100

Pulse Repetition Frequency (Hz)	300/800	870/1176	900 / 1200 alternating
Clutter rejection	20dB(MTI) 35dB(MTD) weather Rejection	37dB weather clutter rejection	49dB ground clutter rejection
Azimuth Beamwidth	5 deg	1.4 deg	5 deg
Elevation Beamwidth	Cosec*2 or pencil beam	Cosec*2 Pattern	Cosec*2 Pattern
Antenna rotation rate	15 rpm	15 rpm	12.5 rpm
Update rate (aircraft)	4 - 8 sec	4 sec	5 sec

Table C2-1 UK and FAA Primary Radar Specifications

<u>Operational Requirements:</u>

Various operational specifications are laid down by aviation regulatory authorities. For this study, the main ones of interest have been extracted [6] in order to be able to judge whether these specifications could be met using a narrowband system. These are for the traditional en-route ATC primary surveillance radars that will be used for benchmarking. A summary of these specifications is reproduced in Table C2-2 below.

Parameter	**Value**
Range (average)	300 km
Overall probability of target position detection	> 90%
Average number of false target reports per antenna scan	< 20
Positional accuracy: Systematic errors slant range bias azimuth bias slant range gain error time stamp error Random errors slant range azimuth	 <100 m <0.1° <1m/NM <100 ms <120m <0.15°
Resolution: slant range azimuth	2× nominal (compressed) pulse width or

	3× nominal 3dB beamwidth
Aircraft operating characteristics: ground speed	 40-800 kt
Update rate	< 8 seconds

Table C2-2 ATC Radar Operational Requirements

C.2.2 Weather surveillance radar

There are a large number of weather radars in operation. Many of these are custom built for specific area monitoring or research. An example of a widely deployed system has been selected as an example for this study. The WSR-88D is an advanced doppler weather surveillance radar manufactured by Lockheed Martin. The specifications for this system are shown in Table C2-3 below.

Radar Range : - Reflectivity - Velocity	 460 km (248 NM) 230 km (124 NM)
Antenna: - Size - Beamwidth - Gain - Side Lobes (with radome) - Azimuth Rate (max)	S-band, centre-feed, parabolic dish 9m 0.95 ° 45.5 dB -27 dB 6 RPM
Transmitter: - Frequency Range - Peak Power - Average Power - Pulse Width -PRF	 2.7-3GHz 750 kW 1.56 kW 1.6, 4.5-5 μs 318-452 Hz (long) 318-1304 Hz (short)
Receiver bandwidth	0.795MHz

Table C2-3 WSR-88D Performance Specifications

C.2.3 Marine radar

With respect to commercial ship borne radar, the performance specifications as set out by the Department of Transport, UK are summarised below. These have been extracted from [1]. This type of radar can be installed on board vessels or at the coast.

1. The radar equipment when mounted at a height of 15 metres above sea level and in normal conditions should give a clear indication of:
   a) Coastlines:
      1. Ground rises to 60m at 20NM
      2. Ground rises to 6m at 7NM

b) Surface Objects:
   1. A ship of 5000 tons gross, whatever the aspect as 7NM.
   2. A small vessel of 10m in length at 3NM
   3. An object such as a navigational buoy having an effective echoing area of approximately 10 square metres at 2NM.

2. The minimum range specified is 50m from the radar and the range resolution of two objects 0.5 to 1.5 NM from the radar.

3. The antenna to be used is a slotted waveguide. The length depends on the maximum range scale as follows:

Maximum Range Scale	Less than 48 NM	48-96 NM
Minimum Aerial Length	0.85m	1.2m

4. The horizontal and vertical beamwidths for such aerials at the half-power points are as follows:

Aerial Length	0.85m	1.2m
Horizontal beamwidth	<2.7 degrees	<2 degrees
Vertical beamwidth	24-30 degrees	

5. The antenna sidelobes should be attenuated by more than 23dB with respect to main beam within $\pm 10°$ of it and by more than 30dB outside that.
6. The antenna gain should be better than 25dBi.
7. The radar equipment will be operating in the X-band.
8. The peak power depends on the maximum range as follows:

Maximum Range Scale	Less than 48 NM	48-96 NM
Minimum Power (Nominal)	≤ 5 kW	≤ 10kW

9. The pulse length and PRF used depend on the range as follows:

	Pulse Length (μs)	PRF (Hz)
Short Range	0.05-0.15	800-3200
Long Range	0.35-1	500-1000

C.2.4 Coastal surveillance radar

There are a number of coastal surveillance radars in operation in the UK. These generally operate in X-band. Pulse widths of less than 20ns are typical leading to large bandwidth for these systems. Some example characteristics of a typical antenna are shown in Table C2-4 below (Easat EA3462 X-band shaped reflector).

General and Mechanical: - Type - Projected Aperture Size - Total weight (incl. turning gear, oil and motor)	 Shaped Reflector 5.5×0.7 m

- Height incl. Pedestal from fixed flange - Max Swept diameter - Rotation rate	1200 kg 1.9 m nominal 6.2 m 10-30 RPM
Electrical Specification: - Beam Characteristics - Operating frequency - Gain - Elevation 3dB beamwidth - Azimuth beamwidth - Azimuth sidelobe level (worst case)	 Inverse Cosec^2 elevation pattern X-band, 9.1-9.5GHz 40 dBi 4 ° nominal <0.45° Within 15°: -28 dB, within ±90° -40 dB, backlobes -38 dB

Table C2-4 Coastal Radar Performance Specifications

C.3 Theoretical analysis of narrowband radar capabilities

In this section, the performance of a narrowband radar equivalent to the systems described in §C2 is computed. This can be divided into two sections. The first step is to perform some simple analysis of performance trade-offs using pulsed and CW radar equations. Next is to look at the geometrical considerations in order to make Doppler and SNR comparisons between monostatic and bistatic pulsed and CW systems. The basis of these assessments is briefly described along with a summary of the results. The calculations are described in more detail in §C8.

C.3.1 Comparative analysis of power and bandwidth

It is assumed that for narrow band radars to have the same sensitivity as their pulsed counterparts, they should exhibit the same average (i.e. CW) power.

The average power P_{av} of a pulsed transmitter is given:

$$P_{av} = P_t \tau PRF = \frac{P_t \tau}{PRI}$$

Equation C3-1

Hence, the radar equation can be written as:

$$SNR_1 = \frac{P_{av} G A_e \sigma .n}{(4\pi)^2 k T_0 B_n \tau F_n r^4 L.PRF}$$

Equation C3-2

In terms of energy E_p, the pulsed radar equation becomes:

$$SNR_1 = \frac{E_p G A_e \sigma}{(4\pi)^2 k T_0 B_n \tau F_n r^4 L}$$

Equation C3-3

Since:

$$E_p = P_{av}.t_d = \frac{P_{av}.n}{PRF}$$

Equation C3-4

The radar equation for CW radar is given by:

$$SNR = \frac{P_t G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} F_n r^4 L}$$

Equation C3-5

B_{CW} is the Doppler filter or speed gate bandwidth

Since the duty cycle of a continuous transmitter is 1, the average power is equal to the peak power:

$$P_{av} = P_t$$

Equation C3-6

Therefore:

$$SNR = \frac{P_{av} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} F_n r^4 L}$$

Equation C3-7

In terms of energy E_{CW}, this becomes:

$$SNR = \frac{E_{CW} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} t_d F_n r^4 L}$$

Equation C3-8

Since:

$$E_{CW} = P_{av}.t_d$$

Equation C3-9

To achieve the same *SNR* in both pulsed and CW systems, the radar equations expressed in terms of *SNR* are made equal as follows:

$$SNR_{pulsed} = SNR_{CW}$$

Equation C3-10

$$\frac{E_p G A_e \sigma}{(4\pi)^2 k T_0 B_{nP} \tau F_n r^4 L} = \frac{E_{CW} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} t_{dCW} F_n r^4 L}$$

Equation C3-11

Assuming all the parameters are constant except for receiver bandwidth and transmitter power, this reduces to:

$$\frac{E_p}{B_{nP} \tau} = \frac{E_{CW}}{B_{CW} t_{dCW}}$$

Equation C3-12

Using the above analysis comparative data has been produced for pulsed and CW systems in range of applications. A summary of these results is reproduced below. Details of how these figures were arrived at can be found in §C8.

	Power (W)		**Bandwidth (Hz)**	
Application	**Pulse**	**Equivalent CW**	**Pulse**	**CW**
En-route ATC	1.5×10^5	360	3.3×10^5	18
Weather	7.5×10^5	$1,35 \times 10^3$	2.2×10^5	36
Marine	1.0×10^4	3.8	2.0×10^6	77

Table C3-1 Summary of Comparative Requirements

From the above results in can be seen that a significant improvement in bandwidth is suggested while maintaining the same performance parameters in terms of power on target. In the case of the ATC radar, the Doppler shift of the received echo for moving targets and fluctuations due to target acceleration and antenna scan speed will widen the spectrum further. Calculations (see §C8) show that the bandwidth would be increased to around 4 kHz maximum taking into account all likely aircraft speeds.

Note: The weather radar calculation has been made for a point target rather than the real case of an extended scatterer. A more realistic weather radar situation is modelled in the next section.

C.3.2 Geometrical consideration for narrowband Doppler and SNR

To extend the basic analysis MATLAB simulation software has been developed to allow investigation of the effect of different CW system geometries in terms of received SNR and Doppler. It is able to produce analytical and graphical representations of SNR and Doppler for a range of monostatic and bistatic configurations. This section gives a brief outline of this software and presents some screenshot examples of its use. A more detailed description of the software development is given is §C9.

This software enables a very large range of scenarios to be modelled and is still under development. In this report a range of initial results enabling comparison of monostatic and bistatic pulsed and CW systems for a range of applications has been produced.

Figure C3-1 shows an example of the Graphical User Interface (GUI) illustrating the ease with which parameters may be changed and investigated.

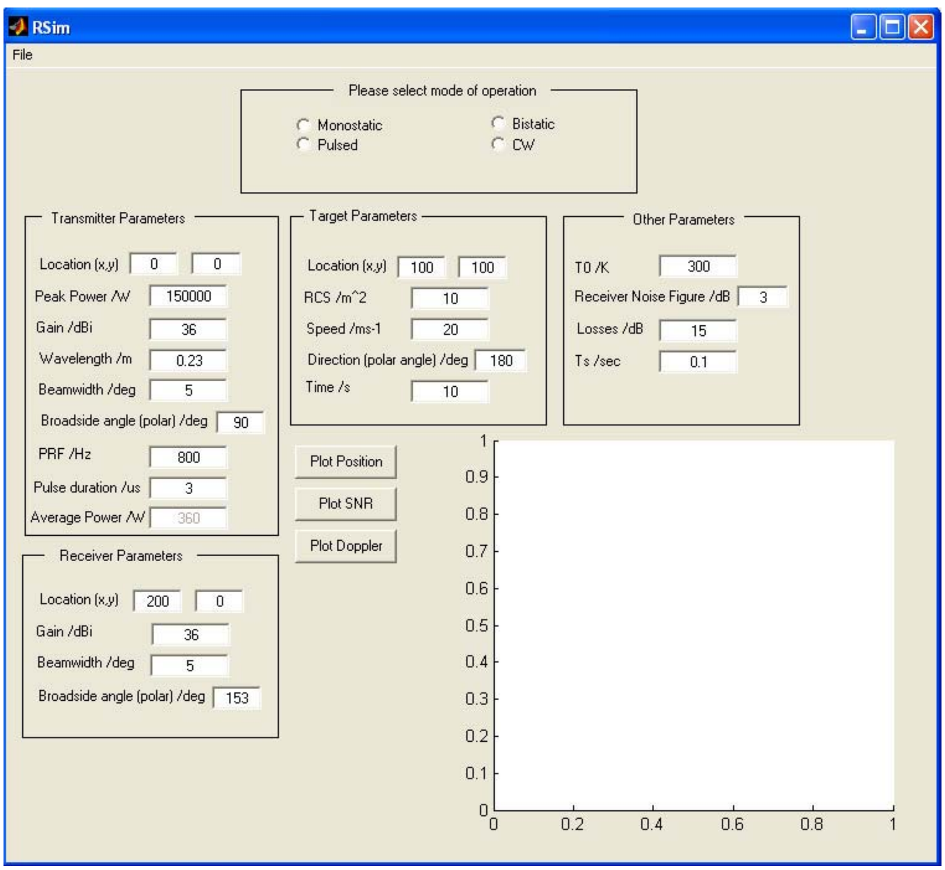


Figure C3-1 Graphical user interface of the simulation software RSim

Requirements Capture:

The initial stage in the design and development of the software tool was to establish a list of the requirements and the expected output of the software. These requirements can be divided up as follows:

Input:

Transmitter Parameters	Location in Cartesian coordinates
	Peak Power (W)
	Antenna Gain (dBi)
	Wavelength (m)
	Beamwidth (deg)
	Broadside angle (polar) (deg)
	PRF (Hz)
	Pulse duration (μs)
Receiver Parameters	Location in Cartesian coordinates
	Antenna Gain (dBi)
	Beamwidth (deg)
	Broadside angle (polar) (deg)
	Receiver noise figure (dB)
	Receiver temperature (K)
Target Parameters	Location in Cartesian coordinates
	RCS (m^2)
	Speed (ms^{-1})
	Direction (polar angle) (deg)
	Time of flight (s)
Other Parameters	System losses (dB)
	Sampling interval (s)
Mode of Operation	Monostatic
	Bistatic
	Pulsed
	Continuous Wave

Output:

Position Plot
Signal to noise power ratio plot
Doppler frequency shift plot

Obviously a large number of scenarios can be modelled. Selections of realistic scenarios taking into account the radar system parameters and likely target characteristics for the ATC and marine radar application have been investigated. The basic parameters are outlined for each case below followed by a summary of the results.

Notes:

1. The SNR values quoted below are rounded to nearest dB as the actual value varies slightly across the beamwidth as the target passes through the beam.

2. The Doppler shift in the ATC case is symmetrical as the target moves through the beam in the monostatic case while a linear variation across the beam is seen in the narrowband distributed case.

3. The receiver beam has been positioned in the narrowband distributed cases to approximately intersect the whole of the target path as it passes through the transmit beam (i.e. transmit and receive beams are synchronised). The transmit and receive beams are assumed static in each of the measured situations

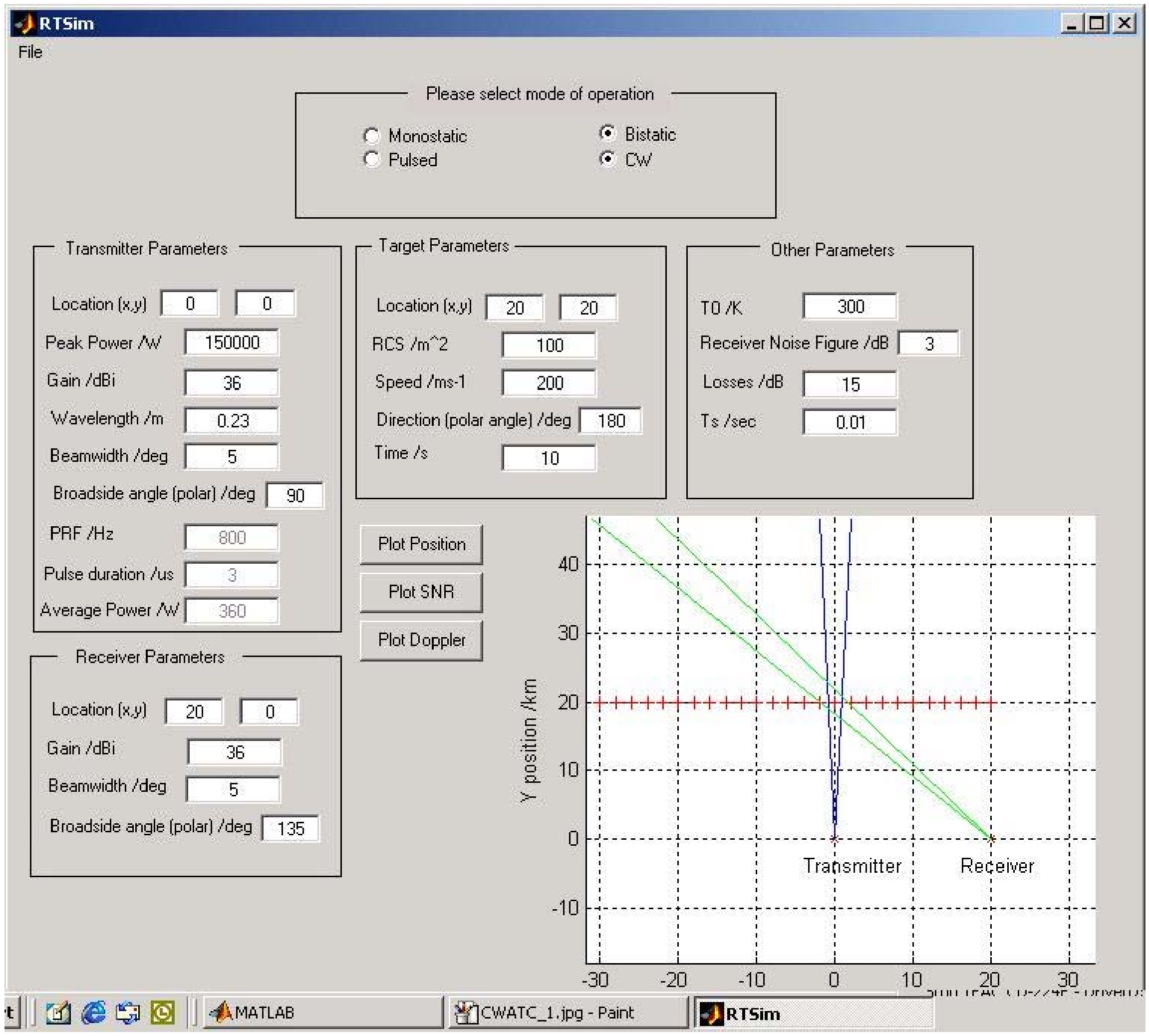
4. The effective bandwidth of the CW system is calculated from the dwell time of the transmit beam on the target (see §C8)

5. All pulses that hit the target are integrated which may not be realistic in many situations but gives a reasonable approximation. In the case of slow moving targets this may give a rather larger than normal integration gain.

<ins>ATC radar examples:</ins>

Radar System:	L-band, 150kW-peak power, primary system
Target RCS:	100m^2 (large passenger aircraft)
Target Speed:	200 m/s
Target Direction	Parallel to transmitter - receiver baseline
Receiver:	36dB, 5 degree beamwidth
Bistatic Geometry:	20km baseline.

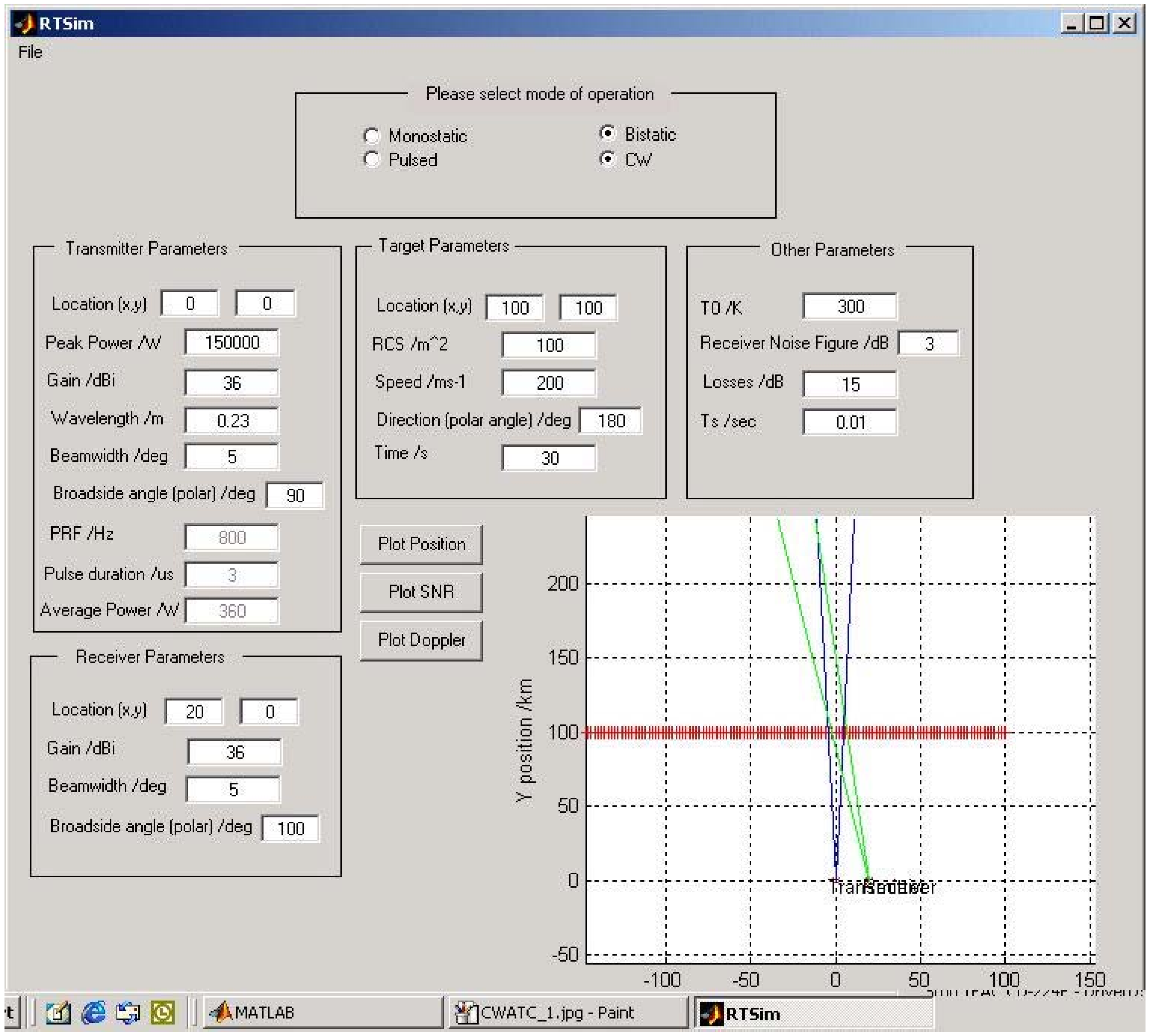
Figures C3-2 and C3-3 shows GUI's and example outputs for the above parameter set.



	Monostatic Pulsed	Monostatic CW	Bistatic Pulsed	Bistatic CW
SNR (dB)	55	81	52	78
Doppler (Hz)	± 60	± 60	-550 to -660	-550 to -660

Figure C3-2 Screenshot of ATC bistatic radar GUI showing system geometry and output results.

(Minimum Range = 20 km)



	Monostatic Pulsed	**Monostatic CW**	**Bistatic Pulsed**	**Bistatic CW**
SNR (dB)	34	60	33	59
Doppler (Hz)	± 80	± 80	-100 to -200	-100 to -200

Figure C3-3 Screenshot of ATC bistatic radar GUI showing system geometry and output results.

(Minimum Range = 100 km)

<u>Marine Radar Example:</u>

Radar System: X band, 10kW peak power system

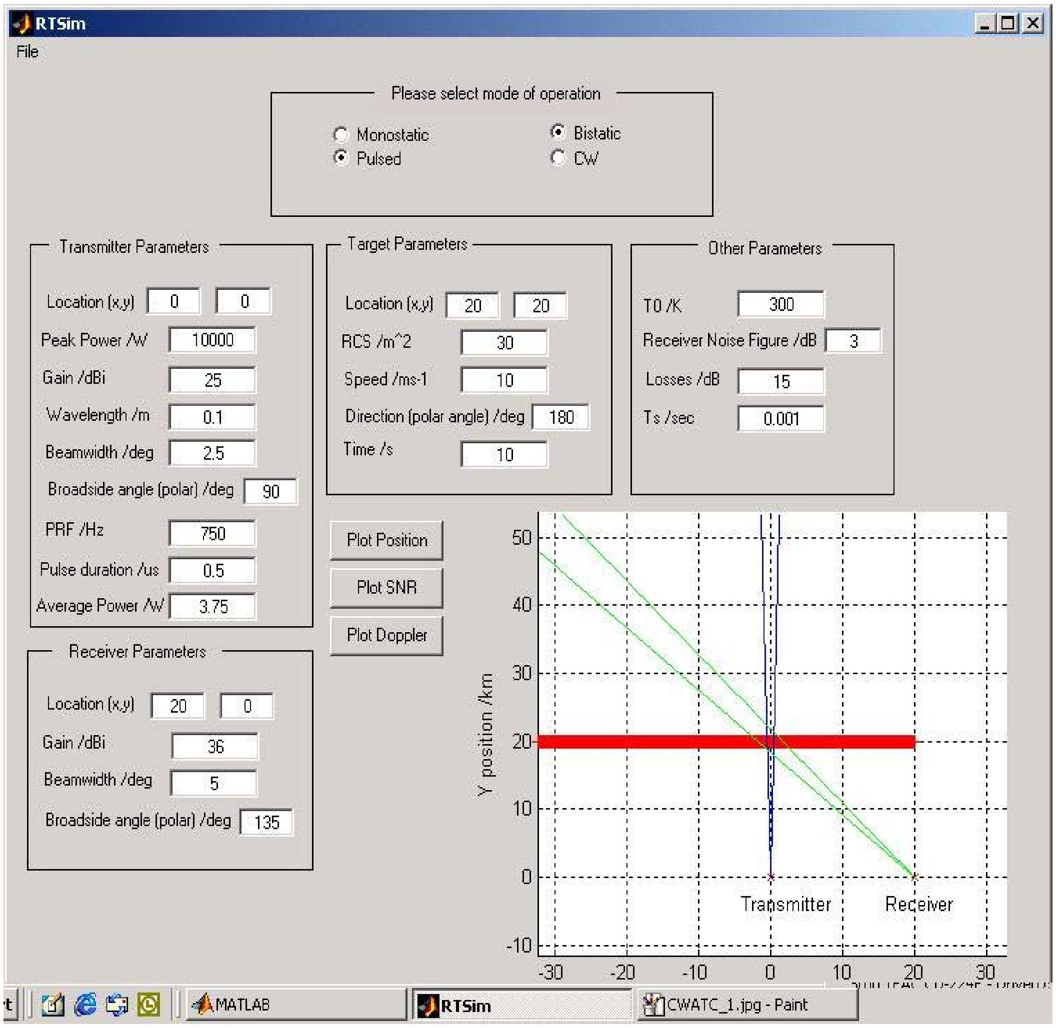
Target RCS: $30m^2$ (small boat)

Target Speed: 10 m/s

Target Direction: Parallel to transmitter - receiver baseline

Receiver: 36dB, 5 deg. beamwidth

Bistatic Geometry: 20km baseline.



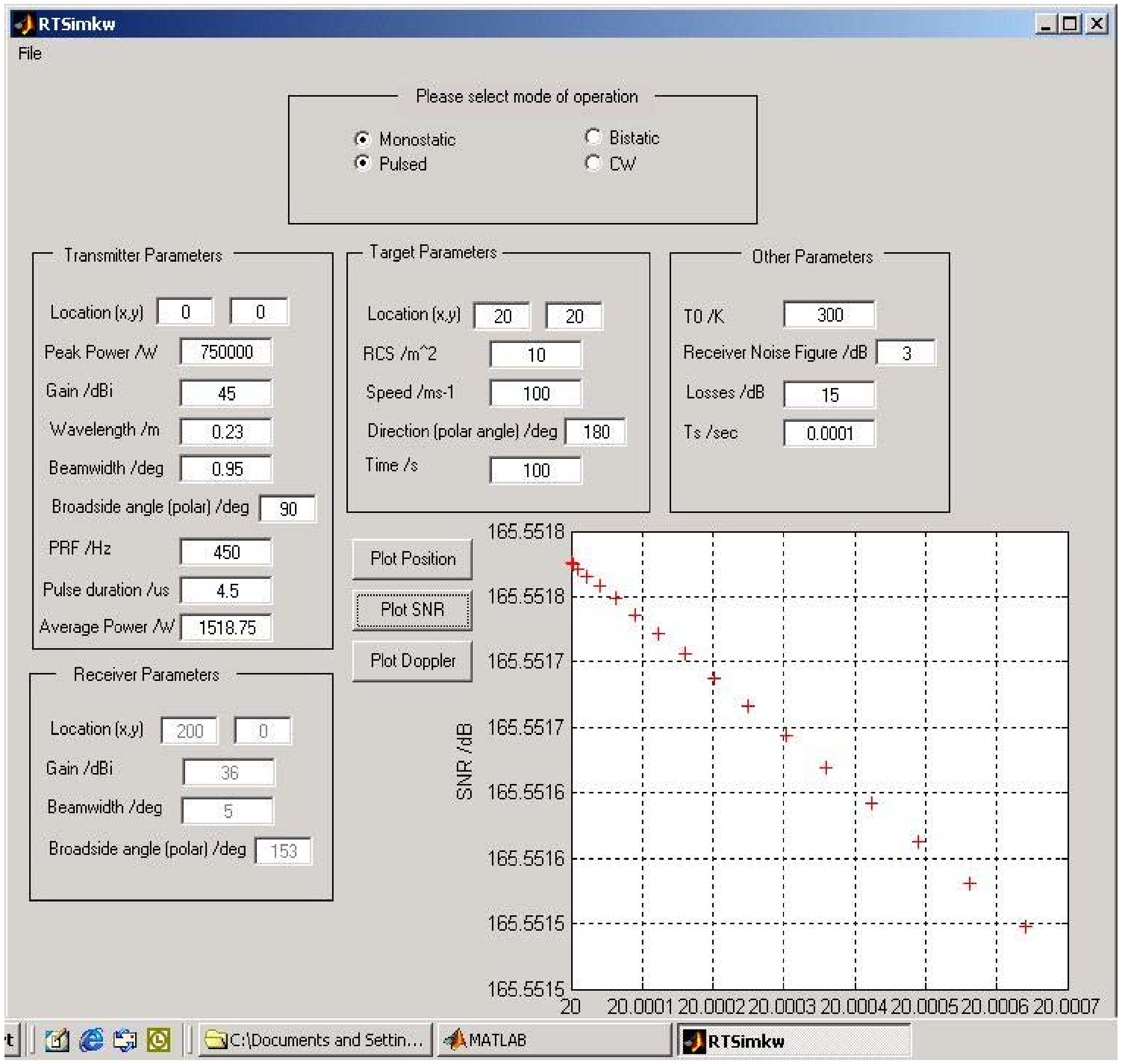
	Monostatic Pulsed	Monostatic CW	Bistatic Pulsed	Bistatic CW
SNR (dB)	36	70	44	78
Doppler (Hz)	± 5	± 5	-68 to -74	-68 to -74

Figure C3-4 Screenshot of Marine bistatic radar GUI showing system geometry and output results.

(Minimum Range = 20 km)

<u>Weather radar example:</u>

Radar System:	WSR-88D S-band Doppler weather radar
Target RCS:	Extended scatterer (Beam filling storm rain rate 10 mm/s)
Target Speed:	N/A
Target Direction:	Parallel to transmitter - receiver baseline
Receiver:	36dB, 0.95 deg. beamwidth
Bistatic Geometry:	Not yet analysed



	Monostatic Pulsed	Monostatic CW
SNR (dB)	165	137

Figure C3-5 Screenshot of monostatic weather radar GUI showing SNR plot and output results.

(Minimum Range = 20 km)

From the example results in Figures C3-2 to C3-5, it can be seen that the SNR and Doppler performance can be maintained to a reasonable level using CW approach. The weather radar shows some rather larger drop in the SNR particularly for the monostatic case modelled. However, these results need to be treated with caution since the extended scatterer case influences the definition of CW bandwidth and the applicability of this methodology in this case is still being investigated.

Additional investigations are continuing into a range of system geometries and target parameters and will be included in the final report.

C.3.3 Location accuracy

In many radar applications, there is a requirement to accurately locate target detections in some form of co-ordinate space. This is of vital importance where safety is paramount. Therefore it is critical that the narrowband CW radar system is able to offer the same level of performance as current pulsed systems.

The location accuracy of current systems may be expressed in terms of range resolution (usually a few tens of metres) and beamwidth (dependent on the size of the antenna used). Ultra narrow band distributed radar has no range resolving ability. Hence, to make up for this, multiple receivers are used to calculate the location of targets. Simple triangulation can then be invoked to find target range from each receiver and thus position. Assuming ATC radar with transmit and receive beamwidths of five degrees and then further assume monopulse or another form of within beam processing. At a range of 100 km a target may be located to an accuracy of the order of 50m in X-Y co-ordinate space. This may be further improved by combining triangulation with Doppler history data. These combinations will be unique for each target enabling them to be separated spatially without ambiguity. However, a direct comparison with the pulsed radar is not straightforward due to the fundamentally different techniques employed.

In addition, this is a technique that is yet to be fully developed and further work is required to be able to confidently apply this approach and make comparisons with the pulsed radar case. This work continues to investigate achievable location accuracy by developing the triangulation method and applying it to the three candidate systems and taking scanning into account.

C.4 Isolation between radar transmitter and receiver

C.4.1 Introduction

In narrow band CW radar, there is the problem of isolation between the transmitter and receiver. A single antenna can be used for both transmission and reception with the separation between the two signals being achieved by the Doppler shift. However, in practice, it is not possible to eliminate fully the transmitter signal leaking into the receiver with a single transmit/receiver antenna. One solution is to use two antennas that will achieve the largest isolation.

In the proposed narrowband system, there is likely to be a number of bistatic system pairs in a multistatic arrangement. This implies separated transmitter and receiver antennas. It is very important to determine how much isolation is required between each transmitter/receiver pair to give the required sensitivity. Having done that, the next step was to determine how much isolation can be achieved.

C.4.2 Analysis

The analysis begins with the example of the ATC radar examined previously. The receiver sensitivity was set by the noise power in the receiver and was given by:

$$N = kT_0BF = 1.38\times10^{-23}\times300\times18\times2 = -188 \text{ dBW in 18Hz bandwidth}$$

Equation C4-1

Thus the signal leaking directly from the transmitter to the receiver should sit below this level if genuine targets are to be detected. This will lead to slightly pessimistic results, as the true compression gain that may be accrued is not factored into these calculations. The amount of leakage will also set the dynamic range. I.e. if the transmitter leakage is above the noise floor of the receiver, increasing the transmit power will only increase the leaked signal and no benefit will accrue.

It is assumed that power from the transmitter can only reach the receiver via transmit and receive antenna sidelobes. The magnitude of leakage will be fundamentally determined by the transmitter power, the sidelobe levels and the physical separation of the transmitter and receiver. However, this can be further reduced by placing nulls in the antenna beam pattern, dynamic cancellation of any breakthrough and Fourier processing.

Consider, for example, a case with the following parameters to illustrate the requirements to maintain leakage below the noise floor of the receiver. The transmitter power is taken to be 100W and it may be assumed that transmit and receive antenna sidelobes have an effective gain of -10dB. If the transmitter and receiver are separated by 1km then this provides of the order -70dBW of the received signal strength. Thus, a further 120dBs of cancellations are necessary. This might be achieved using combinations of techniques. For example, creating nulls using auxiliary antennas on both transmit and receive may provide another 60dBs of attenuation. Dynamic cancellation of the breakthrough signal can yield up to a further 50dB of attenuation. Lastly Fourier processing could yield a further 30dBs of isolation (except in the DC term in the case of Fourier processing). Thus, the leakage signal should appear at a level of some 20dBs below the receiver noise floor.

In the specific case of marine radar, taking the same approach as for ATC if an average power of 10w and a transmitter-receiver separation of 100m are assumed and the rest the same then the signal is about 10dB below the receiver noise floor. If the average power is reduced to 1W this goes to 20dB's below the receiver noise floor. If the separation is only 10m then, at 10W the signal power is 10dB above the receiver noise floor and for 1W equal to it. i.e. you would need extra suppression.

For the weather radar with an average power of 1kW then for a separation of 1km the signal will be 10dB's below receiver noise. Here, the separation could be bigger with hills in the way improving things quite a bit further.

However, it remains to be seen in practise as to the degree of direct signal attenuation that can be achieved and at what cost. However, if achievable, these figures indicate that up to 10kW of average power could be transmitted, or the transmitter to receiver baseline could be reduced to 10m. An examination into some of the design restrictions and freedoms can now begin. Further research is required to determine exact levels of isolation achievable and these will subsequently fix the range of design freedoms.

C.5 Hardware

C.5.1 Introduction

The calculations so far have suggested that a CW radar alternative looks feasible. However, for a system to be realised and implemented suitable hardware must be available. One important issue is the availability of high linearity transmitters that can deliver high CW power. For the ATC radar scenario, if the peak power of a CW system is 360W i.e. average power of pulsed system, then such power should be delivered by already available equipment. An example of such equipment is shown below.

If the ultra narrow band technology is proven to be feasible then there is a complex economic and logistic issue concerned with the replacement of existing systems. Therefore, not only does equipment have to be available but also the impact of the existing systems on the phasing in of new ones needs to be evaluated.

C.5.2 Example of CW transmitter

Applied Systems Engineering, Inc makes the Model 500W CW TWTA. A summary of the specifications of this transmitter is shown below.

Figure C5-1 Model 500 CW TWTA

Duty Cycle	CW
Spurious Outputs	-50 dBc, Maximum
Phase Noise	Less than ±1°
Amplitude Variation	0.1 dB, Maximum
Gain	35 dB, Minimum
Noise Figure	35 dB, Nominal
RF Connectors	Type SC Coax or Waveguide
Primary Power	3 Phase, 208 VAC ±10%, °50/60 Hz
Operating Temperature	0 to 50°C
Weight	175 lbs, Nominal
Dimensions	21×19×30(in.)

Table C5-1 Specification for Model 500 CW TWTA

Features:

- Frequency 1-18GHz Octave / Multioctave
- Low Spurious Outputs
- Phase and Amplitude Stability
- Complete TWT Protection
  - Helix Overcurrent
  - Cathode Over/Undervoltage
  - Collector Overvoltage
  - Filament Low Voltage
  - Over temperature
  - Input Energy Limit
  - Reverse Power Monitor
- Custom Requirements
- Solid State Except for the TWT
- Front Panel Voltage Adjustments
- Front Panel Fault Isolation
- Modular Construction
- DC TWT Filaments
- Four Line Display
  - Operating Mode
  - Cathode Voltage
  - Collector Voltage(s)
  - Helix Current
  - Filament and Operate Time

Standard Equipment:

- Input isolator
- Filament/Operate Time
- IEEE-488 Remote Interface
- Reverse Power Monitor

Options:

- Driver amplifier
- Higher CW power
- Extended frequency coverage
- RF sample ports
- RS-232/422 Remote Interface
- Other primary power
- Outdoor Enclosure
- RF connectors on front panel
- Harmonic filters

C.6 Conclusions and recommendations of WP4

The findings of this study so far are summarised as:

- Ultra narrow band distributed CW radar can achieve similar sensitivities to a pulsed counterpart.
- The location accuracy of ultra narrow band distributed CW radar is superior in angle and inferior in range. Further work is required to fully evaluate performance and assess the impact of the difference in the form of measurement of location accuracy.
- The potential saving in spectrum is up to approximately 90% of that occupied by current pulsed radar systems.
- Ultra narrow band radar techniques may be applied to Marine Navigation as a replacement for the current magnetron based systems. Indeed the relatively low average powers of this class of radar make some of the problems such as isolation and sensitivity somewhat simpler. However, there are two principle aspects of performance that must be addressed. The first is the restricted sighting of the multiple receivers that will determine the accuracy to which targets can be located. A simple calculation based upon a monopulse antenna with beamwidth of 2 degrees gives an angular location accuracy of approximately 35m. If two antennas are spaced by 50m then this can be further improved using the interferomtric technique there will still be little in the way of separation in range. A more promising approach is to use a combination of angle and velocity to distinguish the differing targets. However, further work is required to evaluate whether or not targets can be sufficiently accurately located and differentiated from one another in all cases. The second problem is that of sea clutter. Little data exists for narrow band CW radar and hence it is not possible to predict with confidence the influence this may have. One area of particular concern is that of close in clutter returns ambiguously competing with target returns which consequently reduce overall system sensitivity. Distinguishing targets from sea clutter on the basis of velocity will generally be aided in the CW case but the achievable performance remains to be clarified. It should also be noted, however, that the Phillips PILOT CW radar has been used in a maritime environment successfully and therefore offers some evidence that sea clutter problems are surmountable (N.B. this radar is an FMCW system and hence has range resolving ability).

This is very much a preliminary feasibility study and much work remains to be done to fully validate this approach. In particular the following is identified as requiring further attention. Most of these items have been the subject of some initial investigations during the project but all require further detailed analysis:

- Extending and validating the isolation requirements to evaluate the design restrictions and cost issues.
- Validating the weather radar data for extended volume scatterers.
- Effect of moving from monostatic to bistatic to multistatic geometry.
- Target location methods using above geometries including modulation schemes to increase location accuracy without increasing bandwidth excessively.
- Practical implementation issues such as location/site availability.

- Cost benefit analysis including trade off between financial benefits for better spectrum efficiency and capital cost of existing equipment modifications.

C.7 Radiolocation spectrum occupancy in the L, S and C-bands

Frequency	**Application**
960-1215 MHz	Military and civil radionavigation;
1215-1240 MHz	Civil airport radars - 23cm band.
1240-1260 MHz	Civil airport radars - 23cm band.
1260-1350 MHz	UK RADIOLOCATION primary on 1300-1350 MHz Civil airport radars - 23cm band. Possible use of 1270-1295 MHz by wind profiler radars on a case-by-case, NIB, basis.
1350-1375 MHz	Some airport radar use.
2500-2520 MHz	UK Radiolocation secondary on 2500-2600 MHz
2520-2655 MHz	UK Radiolocation secondary on 2500-2600 MHz
2700-2900 MHz	Civil airport radars - 10cm band Defence aeronautical radars and other miscellaneous radars (for weather etc.)
2900-3100 MHz	Civil airport radars - 10cm band Defence aeronautical radars and other miscellaneous radars (for weather etc.) Aeronautical radars limited to Shipborne interrogator-transponder system limited to 2930-2950 MHz - Maritime radionavigation is limited to marine radars.
3100-3300 MHz	Military radiolocation band
3300-3600 MHz	Military radiolocation band
4200-4400 MHz	Radio altimeters and ground proximity warning systems.
5350-5460 MHz	Various radar uses, including airborne weather
5460-5470 MHz	Shipborne and associated land based radars.
5470-5650 MHz	Shipborne and associated land based radars

Table C7-1 Radiolocation spectrum occupancy in the L, S and C bands

C.8 Details of power and bandwidth calculations

C.8.1 Analysis of the current ATC en-route primary radar

The parameters of the current system that will be used for benchmarking are as follows:

Parameter	Value
Radar wavelength (cm)	23
Transmitter power (kW)	150
Antenna gain (dBi, nominal)	36
Pulse length (μs)	3
Pulse Repetition Frequency (Hz)	800
Azimuth beamwidth (deg)	5
Antenna rotation rate (rpm)	15
Signal to Noise Ratio (SNR)-single pulse (dB)	16
Receiver bandwidth (kHz)	333
RCS (m^2)	100

Table C8-1 ATC en-route primary radar parameters

Using a radar equation analysis, the range of the radar can be calculated approximately and compared with the instrumented range. The simple form of the radar equation, for a single pulse, expressed in terms of *SNR* is:

$$SNR_{\min} = \frac{P_t G A_e \sigma}{(4\pi)^2 k T_0 B_n F_n r^4 L}$$

Equation C8-1

Where:

P_t – transmitted power in Watts

G – antenna gain

A_e – antenna effective aperture in squared metres

σ - radar cross section (RCS) in squared metres

k – Boltzmann's constant – 1.38×10^{-23} W/Hz.K

T_0 - noise reference temperature 300°K

B_n – receiver (noise) bandwidth in Hz

F_n – receiver noise figure

L – loss factor >1

r – range

For n pulses, the radar equation is modified with the integration gain factor n as follows:

$$SNR_1 = \frac{P_t G A_e \sigma . n}{(4\pi)^2 k T_0 B_n F_n r^4 L}$$

Equation C8-2

Where:

SNR_1 is the *SNR* of a single pulse to give the required probability of detection and false alarm.

n is number of pulses integrated for a target as the antenna scans through its beamwidth. This is given by:

$$n = \frac{\theta_B PRF}{\theta_S} = \frac{\theta_B PRF}{6\omega_m} = \frac{5\times 800}{6\times 15} = 44.4 = 44 \text{ Pulses}$$

Equation C8-3

Where:

θ_B – antenna beamwidth in degrees

θ_S – antenna scanning rate in degrees/s

ω_m – antenna scanning rate in RPM

PRF -- pulse repetition frequency in Hz

n can also be calculated using dwell time t_d as well as follows:

$$t_d = \frac{\theta_B}{\theta_s} = \frac{5}{90} = 0.0556 \text{ s}$$

Equation C8-4

$$n = PRF.t_d = 800\times 0.0556 = 44.4 = 44 \text{ pulses}$$

Equation C8-5

The receiver bandwidth is matched to the pulse width and the typical RCS of a large passenger aircraft is assumed ($100m^2$). In addition, losses of 15dB are included.

Having determined all the parameters in the radar equation, the values are substituted to give the range as follows:

$$r = \sqrt[4]{\frac{P_t G A_e \sigma . n}{(4\pi)^2 k T_0 B_n F_n SNR_1 L}}$$

Equation C8-6

Using $G = \frac{4\pi A_e}{\lambda^2}$, gives:

$$r = \sqrt[4]{\frac{P_t G^2 \lambda^2 \sigma . n}{(4\pi)^3 k T_0 B_n F_n SNR_1 L}} = \sqrt[4]{\frac{150\times 10^3 .(3981)^2 .(0.23)^2 .100.44}{(4\pi)^3 .1.38\times 10^{-23} .300.333\times 10^3 .2.40.32}} = 530\text{km}$$

Equation C8-7

This is much greater than the instrumented range of just less than 400km. This is expected as the instrumented range is not equal to the maximum detection range. The above calculation assumes that the system is noise limited and does not consider the effect of clutter. It is not within the scope of this work to explore this but it is suffice to say that that instrumented range is at least achievable.

Comparison of current pulsed and alternative CW system

a) Radar equation analysis

In this subsection, a radar equation analysis of the current pulsed system and the proposed CW radar will be performed to determine what the tradeoffs are.

Pulsed radar system (23cm en-route):

The average power P_{av} of the current pulsed transmitter is:

$$P_{av} = P_t \tau PRF = \frac{P_t \tau}{PRI} = 150 \times 10^3 .3 \times 10^{-6} .800 = 360 \text{ Watts}$$

Equation C8-8

Hence, the radar equation shown earlier can now be written as:

$$SNR_1 = \frac{P_{av} G A_e \sigma .n}{(4\pi)^2 k T_0 B_n \tau F_n r^4 L.PRF}$$

Equation C8-9

In terms of energy E_p, the radar equation becomes:

$$SNR_1 = \frac{E_p G A_e \sigma}{(4\pi)^2 k T_0 B_n \tau F_n r^4 L}$$

Equation C8-10

Since:

$$E_p = P_{av} .t_d = \frac{P_{av} .n}{PRF} = \frac{360 \times 44}{800} = 19.8 \text{ J}$$

Equation C8-11

CW radar system:

The radar equation for CW radar is given by:

$$SNR = \frac{P_t G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} F_n r^4 L}$$

Equation C8-12

B_{CW}- is the Doppler filter or speed gate bandwidth

Since the duty cycle of a continuous transmitter is 1, the average power is equal to the peak power:

$$P_{av} = P_t$$

Equation C8-13

Therefore:

$$SNR = \frac{P_{av} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} F_n r^4 L}$$

Equation C8-14

In terms of energy E_{CW}, this becomes:

$$SNR = \frac{E_{CW} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} t_d F_n r^4 L}$$

Equation C8-15

Since:

$$E_{CW} = P_{av}.t_d$$

Equation C8-16

To achieve the same *SNR* in both systems, the radar equations expressed in terms of *SNR* are made equal as follows:

$$SNR_{pulsed} = SNR_{CW}$$

Equation C8-17

$$\frac{E_p G A_e \sigma}{(4\pi)^2 k T_0 B_{nP} \tau F_n r^4 L} = \frac{E_{CW} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} t_{dCW} F_n r^4 L}$$

Equation C8-18

Assuming all the parameters are constant except for receiver bandwidth and transmitter power, this reduces to:

$$\frac{E_p}{B_{nP} \tau} = \frac{E_{CW}}{B_{CW} t_{dCW}}$$

Equation C8-19

$$\frac{19.8}{333 \times 10^3 .3 \times 10^{-6}} = \frac{E_{CW}}{B_{CW} \times 0.0556}$$

Equation C8-20

$$19.8 = \frac{P_{av}}{B_{CW}}$$

Equation C8-21

Hence, the average power of the CW case, when the bandwidth is matched to waveform, i.e. 18 Hz, is:

$$P_{av} = 19.8 \times \frac{1}{t_d} = 19.8 \times \frac{1}{0.0556} = 360 \text{ Watts}$$

Equation C8-22

This is the same as the peak power that would be needed in a CW system, which is approximately a decrease by a factor of 417 compared to pulsed radar. This is consistent with the duty cycle which is 1/417.

The energy in the CW case is:

$$E_{CW} = P_{av}.t_d = 360 \times 0.0556 = 19.8 \text{ J}$$

Equation C8-23

This is identical to the pulsed case so no energy has been lost.

b) CW radar receiver bandwidth

The receiver bandwidth will be more than that due to the waveform alone. The Doppler shift of the received echo for moving targets, fluctuations due to target acceleration and scanning fluctuations will widen the spectrum further.

Using the range of aircraft ground speeds, the Doppler frequency range is:

$$\Delta f_d = \frac{2\Delta v_r}{\lambda} = \frac{2(411-21)}{0.23} = 3391 \text{ Hz}$$

Equation C8-24

Ignoring other factors, the receiver bandwidth to be used should be 3391Hz. From the relationship above, the peak transmitter power required would be 67kW. However, as the receiver is not matched, the SNR will not be maximised.

In practice, the range or aircraft speeds in less, around 250-300kt. This gives a Doppler bandwidth of:

$$\Delta f_d = \frac{2\Delta v_r}{\lambda} = \frac{2(180-129)}{0.23} = 222 \text{ Hz}$$

Equation C8-25

In this case, the receiver bandwidth would be 222Hz. The transmitter power required would be 4.4kW. Again, the *SNR* will not be maximised as match filtering is not used.

The predicted range of the CW system with the calculated parameters can be determined from the radar equation as follows:

P_t = 4.4 kW

σ=100m^2

B = 222 Hz

F= 3dB $\approx$ 2

k=1.38$\times$10^{-23} W/Hz.K

T_0=300 K

SNR=16 dB

L=15 dB

$$r = \sqrt[4]{\frac{P_{av} G^2 \lambda^2 \sigma}{(4\pi)^3 kT_0 B_{CW} F_n SNR.L}}$$

Equation C8-26

$$r = \sqrt[4]{\frac{4.4\times10^{3}.(3981)^{2}.(0.23)^{2}.100}{(4\pi)^{3}.1.38\times10^{-23}.300.222.2.40.32}} = 530\,\text{km}$$

Equation C8-27

Hence, the maximum detection range is identical to that obtained for the pulsed radar system.

c) Radial velocity and azimuthal resolution of CW radar

The radial velocity resolution is given by:

$$\Delta V = \frac{\lambda f_R}{2}$$

Equation C8-28

Where, f_R is the frequency resolution which is simply $1/t_d$. The wavelength can be taken as constant over the resolution band since $f_R << \lambda$.

Therefore, the velocity resolution of the CW system developed above is:

$$\Delta V = \frac{0.23}{2\times t_d} = \frac{0.23}{2\times0.0556} = 2.1\,\text{m/s}$$

Equation C8-29

The azimuthal resolution is simply determined from the beamwidth, which in this case is 5 degrees.

d) Distributed Receivers

The proposed CW system cannot be used to determine the range of the target. The location of target will be determined by the use of the Doppler shift and direction of arrival (DOA) of the target echoes. This means that at least 2 receivers are required to be able to determine the target bearing measurements using interferometry and triangulation. The technique has been used by Howland [1] to detect and track aircraft up to a range of 260km.

For this project, the initial proposed geometry of the alternative CW system is shown in Figure C8-1. It consists of a transmitter at the centre with four receivers arranged closely, say in a 100 metres square perimeter. This implies bistatic or multistatic operation. A target at the top right quadrant will be detected by a receiver at that quadrant. By co-operating with the other three receivers the target's location can be determined to a high degree of accuracy. The exact location accuracy depends on the receiver beamwidths and exact target range. This requires more detailed analysis but initial calculations with a 5 degree beamwidth suggest a potential target location accuracy of under 500 m can be achieved.

Using more than one site implies that the proposed system might be more expensive. However, the purpose of the system is to achieve spectrum efficiency so this can be justified.

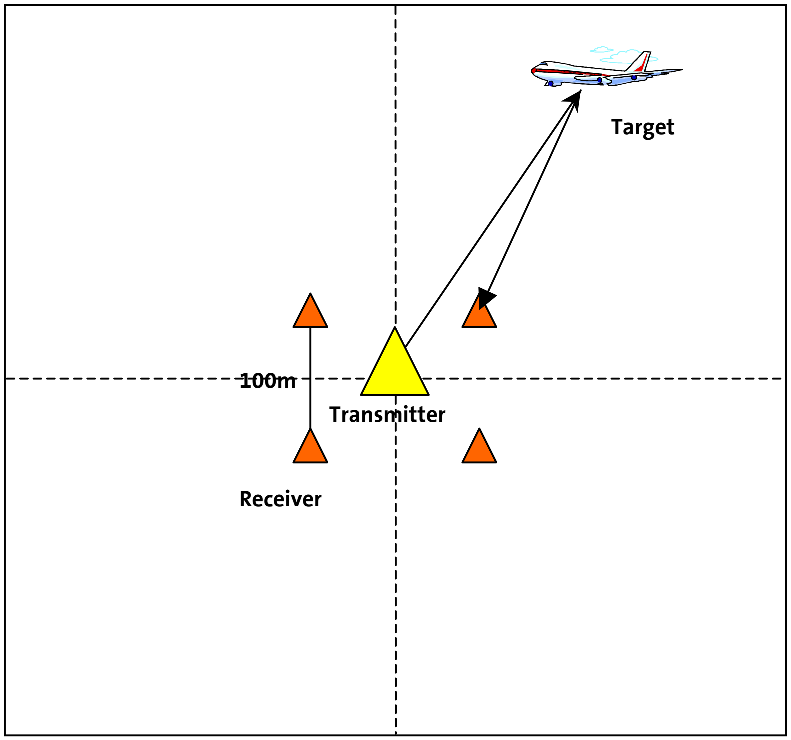


Figure C8-1 Proposed geometry

Consider a transmitter receiver pair as shown above. The distance between the transmitter as and receiver is 71 metres.

With a target at say 300km range, the transmitter receiver separation is insignificant and the system can be treated as a monostatic system, which simplifies the geometry.

The bistatic radar equation is:

$$SNR = \frac{P_t G_t G_r \lambda^2 \sigma_b}{(4\pi)^3 k T_0 B_n F_n r_t^{\,2} r_r^{\,2} L}$$

Equation C8-30

Where:

G_t - transmitting antenna gain

G_r- receiving antenna gain

σ_b - bistatic RCS

r_t - transmitter target distance

r_r - receiver target distance

Because of the very small separation between transmitter and receiver, $r_t \approx r_r = r$. With regards to the RCS, as the bistatic angle at target is very small, the bistatic RCS is equal to the monostatic RCS. In fact, it is a known fact that the monostatic RCS is equal to the bistatic RCS at the bisector of the bistatic angle. In this case, as this angle is very small and insignificant at ranges of interest, both monostatic and bistatic RCS are equal. So the bistatic radar equation reduces to:

$$SNR = \frac{P_t G_t G_r \lambda^2 \sigma}{(4\pi)^3 k T_0 B_n F_n r^4 L}$$

Equation C8-31

This is almost the form of the monostatic equation with the exception of the two gain factors.

If the four receivers are to have smaller antennas, then the gain of the receiving antenna is 10dB less, i.e. 26dBi. For a dish antenna, the dimensions are determined as follows:

$$G = \frac{4\pi A_e}{\lambda^2}$$

Equation C8-32

$$\therefore A_e = \frac{G\lambda^2}{4\pi} = \frac{398.(0.23)^2}{4\pi} = 1.68\ \text{m}^2$$

Equation C8-33

Therefore:

$$A_e = \pi r^2$$
$$\therefore r = \sqrt{\frac{1.68}{\pi}} = 0.731m$$

Equation C8-34

So a 1.5m diameter dish antenna would be required for each receiver.

The beamwidth of this antenna is given approximately by:

$$\theta = \frac{\lambda}{d} = \frac{0.23}{1.5} = 0.153\ \text{rad = 8.8 degrees}$$

Equation C8-35

This is quite poor. But since there is more than one receiver, the overall accuracy in angle of arrival can be improved.

Since the gain of the antenna has decreased, the transmitter power has to increase by 10dB to maintain the *SNR*. Therefore, the power required is 44kW.

This is quite high and at this point it seems that the CW system is not very feasible. However, if a filter bank is used at the CW radar receiver, this can be overcome. Initially the receiver bandwidth would only be matched to waveform, i.e. 18Hz and the transmitter power would be 360W.

Using smaller receive antennas and matched filter processing, the transmitter power that would be required now is 3.6kW. Such powers are achievable. This gives an improvement in transmitter power by almost 42 times and a bandwidth reduction of 18500.

It has been assumed in the above that the transmitter and receiver antennas are fixed. The use of scanning antennas has to be analysed with other issues identified.

This is by no means a complete and accurate analysis. However, it serves as a good starting point outlining the main issues involved.

C.8.2 Analysis of a typical weather radar

Pulsed radar system:

The average power of the transmitter assuming a pulse width of 4.5μs and a PRF of 400Hz is:

$$P_{av} = P_t \tau PRF = \frac{P_t \tau}{PRI} = 750 \times 10^3 .4.5 \times 10^{-6} .400 = 1350 \text{ Watts}$$

Equation C8-36

Hence, the pulsed radar equation can now be written as:

$$SNR_n = \frac{P_{av} G A_e \sigma .n}{(4\pi)^2 k T_0 B_n \tau F_n r^4 L.PRF}$$

Equation C8-37

The number of pulses, *n,* is calculated as follows:

$$n = \frac{\theta_B PRF}{\theta_S} = \frac{\theta_B PRF}{6\omega_m} = \frac{0.95 \times 400}{6 \times 6} = 10.5 = 11 \text{ pulses}$$

Equation C8-38

In terms of energy E_p, the radar equation becomes:

$$SNR_n = \frac{E_p G A_e \sigma}{(4\pi)^2 k T_0 B_n \tau F_n r^4 L}$$

Equation C8-39

Since:

$$E_p = P_{av} .t_d = \frac{P_{av} .n}{PRF} = \frac{1350 \times 11}{400} = 37.1 \text{J}$$

Equation C8-40

CW radar system:

The radar equation for a CW radar is given by:

$$SNR = \frac{P_t G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} F_n r^4 L}$$

Equation C8-41

B_{CW} is the Doppler filter or speed gate bandwidth

Since the duty cycle of a continuous transmitter is 1, the average power is equal to the peak power:

$$P_{av} = P_t$$

Equation C8-42

Therefore:

$$SNR = \frac{P_{av} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} F_n r^4 L}$$

Equation C8-43

In terms of energy E_{CW}, this becomes:

$$SNR = \frac{E_{CW} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} t_d F_n r^4 L}$$

Equation C8-44

Since:

$$E_{CW} = P_{av}.t_d$$

Equation C8-45

To achieve the same *SNR* in both systems, the radar equations expressed in terms of *SNR* are made equal as follows:

$$SNR_{pulsed} = SNR_{CW}$$

Equation C8-46

$$\frac{E_p G A_e \sigma}{(4\pi)^2 k T_0 B_{nP} \tau F_n r^4 L} = \frac{E_{CW} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} t_{dCW} F_n r^4 L}$$

Equation C8-47

Assuming all the parameters are constant except for receiver bandwidth and transmitter power, this reduces to:

$$\frac{E_p}{B_{nP} \tau} = \frac{E_{CW}}{B_{CW} t_{dCW}}$$

Equation C8-48

$$\frac{37.1}{222 \times 10^3 .4.5 \times 10^{-6}} = \frac{E_{CW}}{B_{CW} \times 0.0275}$$

Equation C8-49

The dwell time t_d was calculated as follows:

$$t_d = \frac{n}{PRF} = \frac{11}{400} = 0.0275$$

Equation C8-50

$$37.1 = \frac{P_{av}}{B_{CW}}$$

Equation C8-51

Hence, the average power of the CW case, when the bandwidth is matched to waveform, i.e. 36Hz, is:

$$P_{av} = 37.1 \times \frac{1}{t_d} = 37.1 \times \frac{1}{0.0275} = 1350 \text{ Watts}$$

Equation C8-52

This is the same as the peak power that would be needed in a CW system, which is approximately a decrease, by a factor of 556 compared to pulsed radar. This is consistent with the duty cycle which is 1/556.

The energy in the CW case is:

$$E_{CW} = P_{av}.t_d = 1350 \times 0.0275 = 37.1 \text{ J}$$

Equation C8-53

This is identical to the pulsed case so no energy has been lost, i.e. the detection range of the pulsed and CW systems will be the same if the CW transmit power is 1350 W and the required bandwidth is matched to the transmitted waveform's bandwidth.

As can be seen there is a great reduction in bandwidth and also on the peak transmitter power with a CW radar alternative. There is a reduction in bandwidth by a factor of 6167 and a reduction in peak power by 556.

Note however that the above analysis assumes a point target rather than the extended scatterer usually observed by weather radar. The radar equation in this case is modified to include a reflectivity factor related to the rain rate as follows:

$$SNR = \frac{P_t G_t G_r Z \Delta\Theta \Delta\Phi L}{\lambda^2 R^2}.(2.3x10^{11})$$

Equation C8-54

where:

$$Z = ar^b$$

Equation C8-55

a and *b* are parameters determined by the type of precipitation and *r* is the rain rate in mm/hr. In §C2 values typical of a thunderstorm have been used as follows:

a = 4.86

b = 1.37

r = 10

C.8.3 Analysis of typical marine radar

Pulsed radar system:

The average power of the transmitter assuming a peak power of 10kW, a pulse width of 0.5µs and a PRF of 750Hz is:

$$P_{av} = P_t \tau PRF = \frac{P_t \tau}{PRI} = 10 \times 10^3.0.5 \times 10^{-6}.750 = 3.75 \text{ Watts}$$

Equation C8-56

The number of pulses *n* is calculated as follows assuming a beamwidth of 1.5 degrees and antenna rotation at 20 RPM:

$$n = \frac{\theta_B PRF}{\theta_S} = \frac{\theta_B PRF}{6\omega_m} = \frac{1.55 \times 750}{6 \times 20} = 9.6 = 10 \text{ pulses}$$

Equation C8-57

The pulse energy is given by:

$$E_p = P_{av}.t_d = \frac{P_{av}.n}{PRF} = \frac{3.75 \times 10}{750} = 0.05 \text{ J}$$

Equation C8-57

CW radar system:

In an equivalent CW narrow band radar, the average power is equal to the peak power:

$$P_{av} = P_t$$

Equation C8-58

To achieve the same *SNR* in both systems, the radar equations expressed in terms of *SNR* are made equal as follows:

$$SNR_{pulsed} = SNR_{CW}$$

Equation C8-59

$$\frac{E_p G A_e \sigma}{(4\pi)^2 k T_0 B_{nP} \tau F_n r^4 L} = \frac{E_{CW} G A_e \sigma}{(4\pi)^2 k T_0 B_{CW} t_{dCW} F_n r^4 L}$$

Equation C8-60

Assuming all the parameters are constant except for receiver bandwidth and transmitter power, this reduces to:

$$\frac{E_p}{B_{nP}\tau} = \frac{E_{CW}}{B_{CW} t_{dCW}}$$

Equation C8-62

$$\frac{0.05}{2 \times 10^6 . 0.5 \times 10^{-6}} = \frac{E_{CW}}{B_{CW} \times 0.013}$$

Equation C8-63

The dwell time t_d was calculated as follows:

$$t_d = \frac{n}{PRF} = \frac{10}{0.013} = 0.013$$

Equation C8-64

$$0.05 = \frac{P_{av}}{B_{CW}}$$

Equation C8-65

Hence, the average power of the CW case, when the bandwidth is matched to waveform, i.e. 77Hz, is:

$$P_{av} = 0.05 \times \frac{1}{t_d} = 0.05 \times \frac{1}{0.013} = 3.8 \text{ Watts}$$

Equation C8-66

This is the same as the peak power that would be needed in a CW system, which is approximately a decrease, by a factor of 2632 compared to pulsed radar. This is consistent with the duty cycle which is 1/2632.

The energy in the CW case is:

$$E_{CW} = P_{av}.t_d = 3.75 \times 0.013 = 0.05 \text{ J}$$

Equation C8-67

This is identical to the pulsed case so no energy has been lost, i.e. the detection range of the pulsed and CW systems will be the same if the CW transmit power is 3.8W and the required bandwidth is matched to the transmitted waveform's bandwidth.

As can be seen there is a great reduction in bandwidth and also on the peak transmitter power with a CW radar alternative. There is a reduction in bandwidth by a factor of 25974 and a reduction in peak power by 2632.

C.9 MATLAB simulation

C.9.1 Design and development

The strategy used in the development of the software was to divide it up into three manageable tasks. Each task would implement one of the outputs shown above and would have its own function within the code. The algorithms for each of these tasks are shown next.

Position Plot

This plot allows the user to visualise the geometry of the particular setup in question. If the mode of operation is monostatic a transmitter/receiver and target track would be shown. If bistatic mode of operation is chosen the position of the receiver will be added. The positions of the transmitter and receiver are shown as red crosses and the transmitted beam is shown in blue and the receiver beam is shown in green. The target track is also shown as a series of red crosses. The algorithm used to display this plot is now shown. The input parameters used for this plot are as follows:

Transmitter	Location Beamwidth Broadside angle
Receiver (if bistatic)	Location Beamwidth Broadside angle
Target	Location Speed Direction Flight time
Other	Sampling time

Algorithms

Transmitter (and receiver) plot:

- Check mode of operation.
- Check that relevant inputs are valid. If not display an error message.
- Plot the transmitter (and receiver) location.
- Calculate the broadside angle from the transmitter (or receiver) location.
- Calculate the beam boundaries.
- Draw the beam.
- Label the transmitter (and receiver).

Target track plot:

- Check that the target input parameters are valid. If not, display an error message.
- From initial starting position, calculate bearing.
- For the duration of flight, calculate the new target position at each sampling instant and plot it with a red cross.

Screenshot example for a bistatic mode of operation

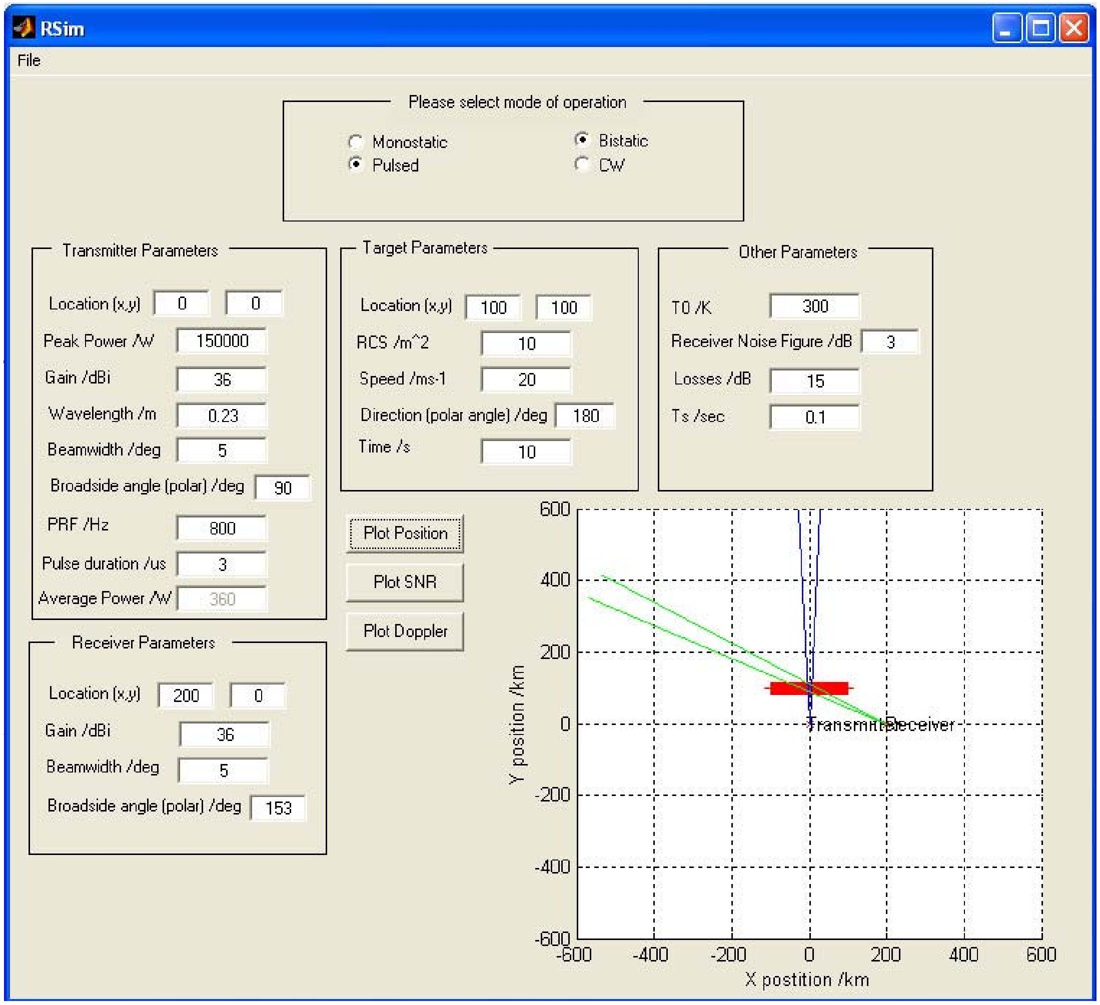


Figure C9-1 Plot of position for a bistatic geometry

Signal to noise power ratio (SNR) plot:

This plot shows the SNR at the receiver as a function of the distance from the transmitter for each target position in the track within the transmitter (and receiver if bistatic) beam. All of the input parameters in the simulation windows are used for this plot unless the system in consideration is monostatic.

Algorithm:

- Check mode of operation and that inputs are valid. If not display, an error message.
- For each target position in the track within the transmitter (and receiver) beam, calculate the distance of the target from the transmitter (and receiver). This is done as follows:
- Set the transmitter as the centre of origin.
- For each target position, determine its quadrant.
- Find the angle of the target position with respect to transmitter.
- If the target angle is within the boundaries of the beam, find the transmitter distance.

- If the system is operated in CW mode, the dwell time is calculated from the sampling time and the number of target samples within the transmitter beam.
- If the system is bistatic, the receiver to target distance is calculated is a similar way.

The *SNR* is determined from one of four different equations depending on mode of operation as follows.

Monostatic and pulsed:

$$SNR = \frac{P_t G_t^{\ 2} \lambda^2 \sigma . n}{(4\pi)^3 k T_0 B_n F_n r^4 L}$$

Equation C8-67

B_n is noise bandwidth given approximately by the reciprocal of pulse duration.

Monostatic and CW:

$$SNR = \frac{P_t G_t^{\ 2} \lambda^2 \sigma}{(4\pi)^3 k T_0 B_{CW} F_n r^4 L}$$

Equation C8-68

B_{CW} is CW matched bandwidth given by the reciprocal of dwell time.

Bistatic and pulsed:

$$SNR = \frac{P_t G_t G_r \lambda^2 \sigma . n}{(4\pi)^3 k T_0 B_n F_n r_t^{\ 2} r_r^{\ 2} L}$$

Equation C8-69

Bistatic and CW:

$$SNR = \frac{P_t G_t G_r \lambda^2 \sigma}{(4\pi)^3 k T_0 B_n F_n r_t^{\ 2} r_r^{\ 2} L}$$

Equation C8-70

Screenshot example for a pulsed monostatic scenario

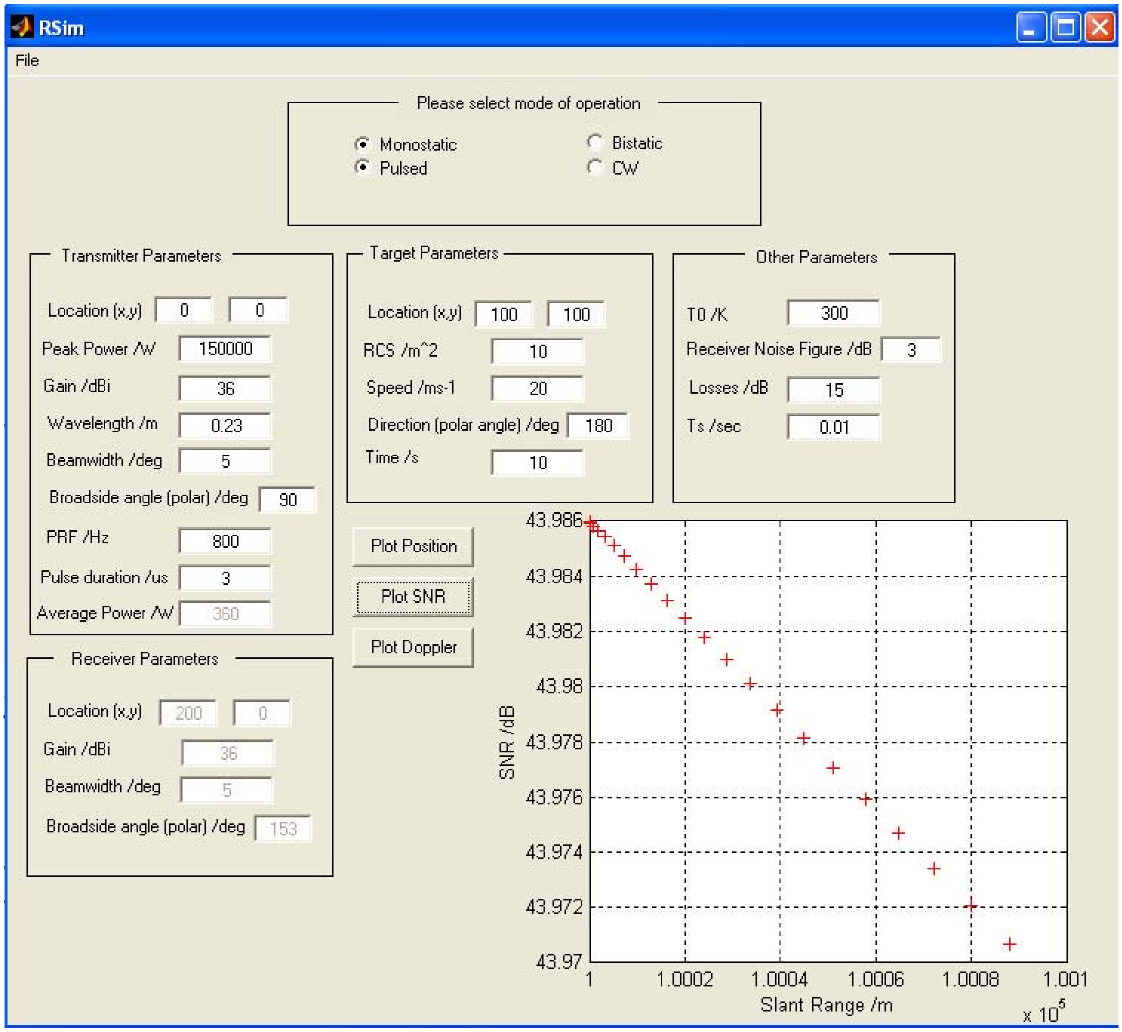


Figure C9-2 Plot of SNR for a monostatic geometry

Doppler frequency shift plot:

This shows the Doppler shift of the transmitted signal because of the target's motion. The plot only shows the Doppler frequency shift for target positions within the transmitter (and receiver) beam. The input parameters used for this plot are as follows:

Transmitter	Location Beamwidth Broadside angle Wavelength
Receiver (if bistatic)	Location Beamwidth Broadside angle
Target	Location

	Speed Direction Flight time
Other	Sampling time

Algorithm

Check that all the input parameters are valid. If not, display an error message.

For each target position in the track within the transmitter (and receiver) beam, calculate the distance of the target from the transmitter (and receiver). This is done as follows:

- Set the transmitter as the centre of origin.
- For each target position, determine its quadrant.
- Find the angle of the target position with respect to transmitter.
- If the target angle is within the boundaries of the beam, find the transmitter distance.
- If the system is bistatic, the receiver to target distance is calculated is a similar way.

The Doppler frequency shift is then determined in two ways depending whether system is monostatic or bistatic as follows.

Monostatic:

$$f_d = \frac{2v_r}{\lambda}$$

Equation C8-71

Where radial velocity, v_r is determined as follows:

At each target position within the beam, the radial velocity is calculated by:

$$v_r = \frac{dR}{dT}$$

Equation C8-72

Where:

$dR = rangenow - rangeprevious$

$dT = timenow - timeprevious$

Bistatic:

$$f_d = \frac{-1}{\lambda} \times \left(\frac{dR_{tx}}{dT} + \frac{dR_{rx}}{dT} \right)$$

Equation C8-73

Where:

$dR_{tx} = rangenowtx - rangeprevioustx$

$$dR_{rx} = rangenowrx - rangepreviousrx$$

$$dT = timenow - timeprevious$$

Screenshot example for a pulsed monostatic scenario

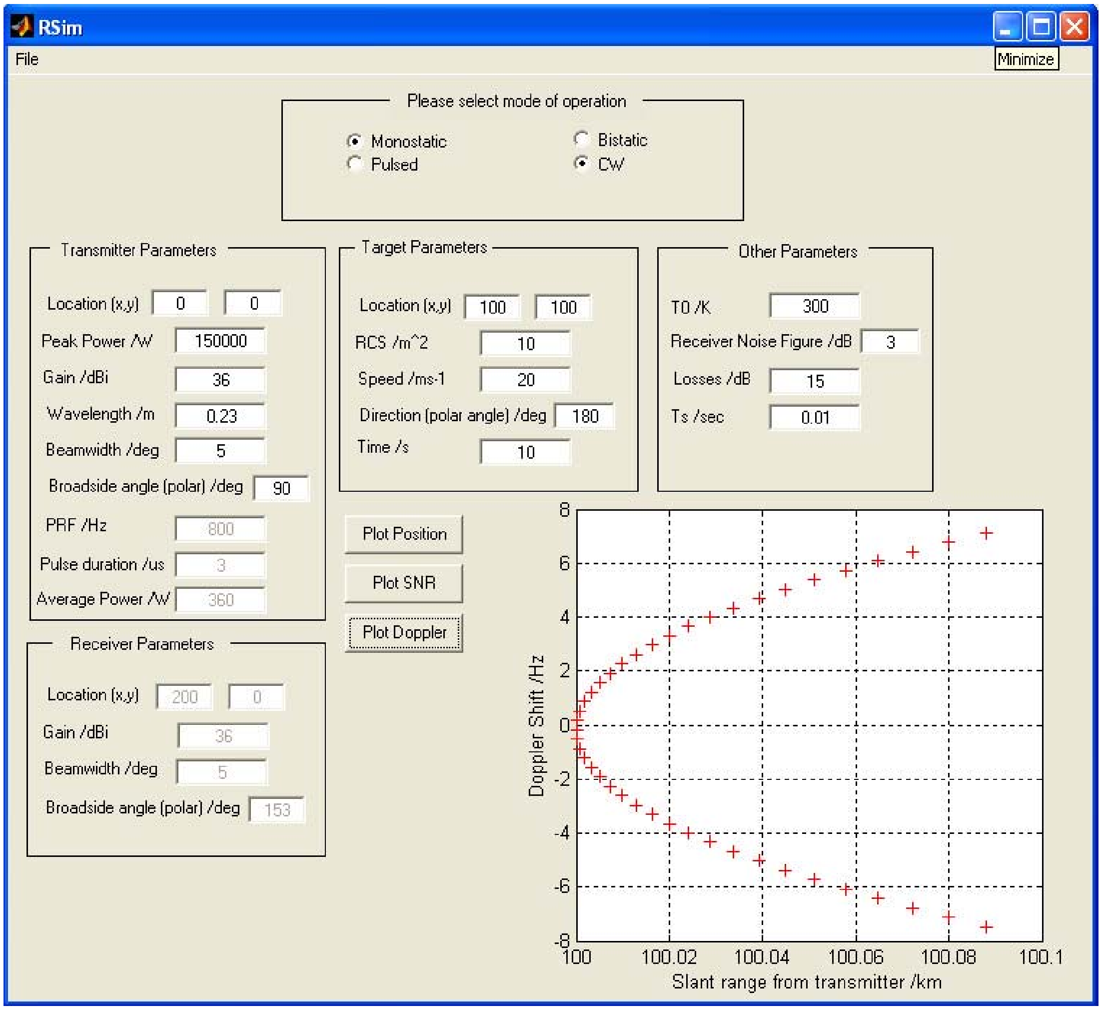


Figure C9-3 Plot of Doppler frequency shift for a monostatic geometry

C.9.2 Validation

The validation of the software tool involved mainly manual calculation to make sure that the results obtained are correct. The first step involved going through the whole program and checking for any mistakes in the code that can give an erroneous output. This was followed by checking the outputs obtained from the program with those obtained manually for different scenarios. The modular design made this process much easier.

C.9.3 User guide

The software simulation tool is very easy to use and is self explanatory. The tool has a clear graphical user interface that allows the user to quickly get acquainted with it and get results.

Starting the simulation:

Before the simulation can be used, the MATLAB files have to be copied to the computer you are using. The files that need to be copied are **rsim.m** and **rsim.fig**. Once that has been done, the simulation can be started by typing rsim at the MATLAB command prompt making sure that the directory path is that where the files are located.

Plotting Position:

1. Select mode of operation: Monostatic or Bistatic and pulsed or CW.
2. Enter all of the parameter values as shown in the graphical user interface making sure to use the correct units.
3. Press the "Plot Position" button and the plot is displayed.
4. The plot is zoom-enabled and you can use the mouse to zoom into any particular region of interest.

Plotting SNR:

Make sure all the inputs are valid and press on the "Plot SNR" button and the plot will be displayed.

Plotting Doppler:

Make sure all the inputs are valid and press on the "Plot Doppler" button and the plot will be displayed.

Exiting simulation:

The simulation can be closed by pressing on the "close button" at the top of the window. Alternatively, it can be closed by selecting close from the File menu in the graphical user interface.

C.10 References/Bibliography

[1] Howland P E, “Target tracking using television-based bistatic radar”, IEE proceedings Radar, Sonar and Navigation, 146(3) 166-174 1999.

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[4] “TD 040 AN Solid State S band Approach Radar”, Airsys ATM Specification

[5] documents, revision 14.

[6] NATS 23 cm radar specifications document.“radar surveillance in en-route airspace and major terminal areas”, Eurocontrol Standard Document, Reference, 006-95, 13 May 1996.

[7] Department of Transport UK, Marine Radar Performance Specification, 1982

D WP6: Waveform design

D.1 Introduction

The objectives of this work package was to demonstrate techniques that allow the mitigation of interference between radars and other service users within a given band by the use of waveform coding and hence allow more users in the band.

This report presents the finding of literature search, study and simulations with such an objective in mind, a partner report (contained in appendix E) will cover the demonstration of some of the techniques discussed herein.

Currently, most civil radar systems use low cost magnetrons that transmit short uncontrolled pulses to achieve adequate range resolution and long range performance. The transmissions from such radars have broad operating bandwidths and suffer from spectral splatter, or spurious emissions. Therefore, the available spectrum is not being efficiently used by these systems. In addition, the number of applications that require use of the electromagnetic spectrum is increasing, which places pressure on existing allocated spectral bands, and their boundaries. With recent advances in technology and demand for bandwidth these pressures have transferred from the below 1GHz bands to all bands between 1-16GHz. This sets the scene for the focus of this study.

The nominal range resolution for a bandwidth *B*:

$$\Delta r = \frac{c}{2B}$$

Equation D1-1

Where *B* is typically the half power bandwidth and termed the range resolution bandwidth.

Basic radars generate bandwidth by transmitting narrow pulses: the nominal range resolution bandwidth is the reciprocal of the pulse-width, this includes most of the civil radar systems under consideration. CW radars have very narrow bandwidths and detect moving targets by the Doppler shifts in the received signal but no range information is obtained. More sophisticated radars employ pulse-compression (PC), where the transmitted waveform is a long modulated pulse that either sweeps the resolution bandwidth e.g. linear frequency modulation (LFM) or have a wide instantaneous bandwidth throughout the pulse e.g. binary coded waveform. Such systems, traditionally, require sophisticated, bespoke waveform generators and receivers. Therefore, currently only the large high cost assets and military systems use such waveform. However, advances in waveform generation technology and digital signal processing now allow PC technology to be accessible to lower cost radar systems.

This work package has examined such sophisticated waveform techniques with an aim to satisfying reduced spectrum and cross system interference objectives, whilst presenting performance issues resulting from use of such waveforms.

In §D2 the collection of a radar database is presented, which is used to act as a focus for the roles that the new waveforms need to replace.

Typical PC waveform and associated theory are then introduced in §D3. §D4 assesses the interference between radar and communication systems operating within a given area, to establish interference performance rejection aims.

The applicability and utility of current regulations are then investigated in §D5.

§D6 introduces orthogonal waveform concepts and discusses notable examples of such waveforms.

§D7 introduces radar waveforms with compact spectrum and methods of operating multiple waveforms in compact bandwidths.

In §D8 the techniques for sharing and operating in congested bands are presented.

However, co-operative band sharing is discussed separately in §D9.

In §D10 the techniques introduced and developed in this study are theoretically applied to reducing usage of three radar frequency allocation bands.

Conclusions and Recommendations are then offered in §D11 and D12.

D.2 Data base of civil radar

D.2.1 Overview

A radar database has been generated to set the scene for the task of designing waveforms that reduce the amount of interference between radars and other users in the same band. Various civil radar systems were investigated; these covered a variety of applications such as marine, meteorological, ATC through to radio astronomy. In general, the data was collected from information available on the Internet or by directly visiting the UK's radar regulators, authorities and companies (e.g. NATS, CAA, Met. Office & MCA). Numerous experts in the field of radar have been consulted and good links have been established with key contacts in the area of wind profiling, weather and ATC radars.

All the data is stored in tabular format in Microsoft Excel and a new worksheet is included for each type of radar to assist categorisation of the data collected. The specification fields were chosen to help calculate waveform details required for this work package, such as: range sidelobes, unambiguous range, range resolution, occupied bandwidth, code length and sensitivity. Approximately 43 parameters are listed for each radar system. However, not all of these values are populated for every radar system.

D.2.2 Findings

The following general characteristics have been noted whilst populating the database for each class of radar system:

Marine radars:

- Utilise magnetrons
- Transmit simple pulses
- Have a pulse width typically between 0.05μs and 1μs
- Pulse repetition frequency (PRF) is typically 1400Hz and 700Hz for commercial radars and 2000Hz and 500Hz for leisure radars
- Duty cycle is 0.007% and 0.07% for commercial otherwise 0.02% and 0.05% for leisure
- Noise figure is approximately 5dB
- Antenna horizontal scan type is continuous (360°)
- Antenna type is normally a slotted array
- Antenna polarisation is normally horizontal
- Antenna pattern type is normally fan

ATC radars:

- Utilise travelling wave tubes (TWTs)
- Transmit NLFM modulated pulses
- Pulse width is typically in the region of 1μs and 50μs
- PRF is normally 400Hz and 1000Hz
- Duty cycle is either 1% or 2%
- Noise figure is approximately 5dB
- Antenna horizontal scan type of continuous (360°)
- Antenna polarisation is normally circular
- Antenna pattern type is cosecant –squared or composite

Meteorological radars:

- Utilise magnetrons
- Transmit simple pulses
- Pulse width is typically 1μs
- PRF is typically 300Hz
- Duty cycle is 0.03%
- Noise figure is approximately 5dB
- Antenna horizontal scan type is either continuous (360°) or stationary
- Antenna type is normally reflector
- Antenna polarisation is normally vertical
- Antenna pattern type is normally a pencil beam

Vessel traffic surveillance (VTS) radars:

- Utilise magnetrons
- Transmit simple pulses
- Pulse width is typically 0.02μs
- PRF is normally 160Hz and 1000Hz or 160Hz and 4000Hz
- Duty cycle is 0.025% or 0.025% and 0.1%
- Noise figure is approximately 5dB
- Antenna horizontal scan type is either continuous (360°)
- Antenna type is normally slotted array
- Antenna polarisation is normally horizontal
- Antenna pattern type is normally fan

D.2.3 Discussion

A vast amount of data has been collected during this study and the database is constantly updated, as new data manifests. Typical values for each type of radar system have been extracted and one user of the database has created a worksheet stating the generic values, for use in calculations. The worksheet is included in this report in §D13. Both OFCOM and consortium have been issued with a copy of the database.

D.3 Pulse compression waveform theory

D.3.1 Introduction

Pulse compression (PC) involves the transmission of a long, modulated pulse, which is subsequently processed to form a narrow pulse with a processing gain given by its pulse compression ratio [1]. For the purpose of this study, pulse compression radar waveforms are discussed as two broad categories: Analogue and Digital. Analogue waveforms may be formed digitally but the phase is continuous, or over-sampled, this tends to make them more spectrally compact. Digital waveforms consist of regular, evenly spaced sub-pulses that are phase-coded according to the selected code sequence.

This section introduces common waveforms and the properties, typically, held by such waveforms.

D.3.2 Analogue waveforms

<u>Linear frequency modulated</u>

Linear frequency modulated (LFM) waveform is the best known and most commonly used analogue pulse compression waveform. It can be formed in a number of ways and its definition in the time domain is given by:

$$S(t) = \begin{cases} e^{j\pi\left[2f_0 + B_c\frac{t^2}{\tau}\right]} & if \qquad |t| \le \frac{\tau}{2} \\ 0 & otherwise \end{cases}$$

Equation D3-1

Where, B_c = the swept bandwidth of the sweep

τ = the pulse width

f_0= the centre frequency of the modulation

The pulse compression ratio is τB_c, which is also known as the time bandwidth product, the instantaneous frequency $f(t)$ is defined as:

$$f(t) = \frac{1}{2\pi}\frac{d\varphi}{dt}$$

Equation D3-2

The waveform has a linear frequency slope with gradient B_c/τ i.e.:

$$f(t) = f_o + \frac{B_c}{\tau}t$$

Equation D3-3

The LFM waveform is a popular choice because it has a number of important advantages, however, it also has a number of disadvantages:

- It is relatively easy to generate by analogue or digital processes and PC processing is simplified by simple “bespoke” radar receiver architectures.
- High compression ratio LFM waveforms have high Doppler tolerance i.e. its SNR and sidelobes are insensitive to Doppler shifts. This is due to its

ambiguity diagram being a sheared ridge[2]. However, this property means LFM suffers from ambiguities between Doppler shifts and range delays, this property is known as range-Doppler coupling.

- The matched filter correlation response is a sinc function with a peak range sidelobe ratio (PSR) of -13dB. However, in practice, the sidelobes are usually suppressed to give a PSR in excess of -40dB by windowing e.g. Hamming or Taylor.

The spectrum of an un-weighted LFM waveform with pulse-width τ of 10μs and a 3dB bandwidth Bc of 10MHz (giving a time-bandwidth product BT of a 100) is shown in Figure D3-1.

The LFM waveform also has a number of important features of significance to this study. LFM waveforms are spectrally efficient signified by the sharp fall off beyond the $f_0 \pm ½Bc$ points i.e. there is little spectral leakage outside the nominal 3dB bandwidth. In addition, the spectrum is contiguous and relatively uniform within the 3dB bandwidth. However, it does have some in-band ripples known as Fresnel ripples. The spectrum becomes increasingly rectangular with large time-bandwidth products.

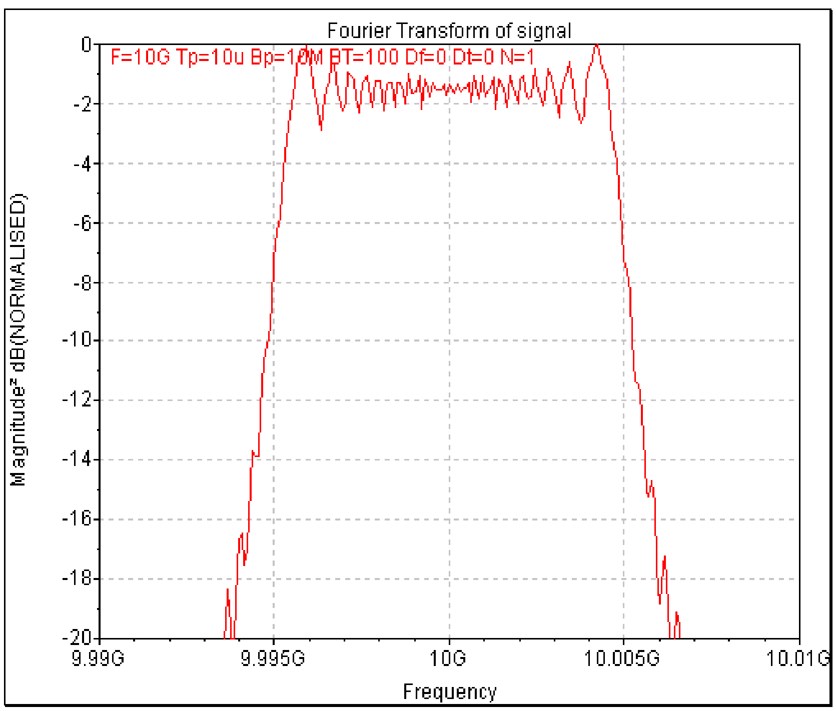

Figure D3-1 Spectrum of a LFM waveform without weighting (T=10μs, B=10MHz).

<u>Non-linear frequency modulated (NLFM)</u>

The near rectangular shape of the LFM spectrum gives high range sidelobes, therefore, as discussed above and weighting is normally applied in the receiver to reduce the sidelobes to acceptable levels. An alternative approach is to form spectral

[2] The ambiguity diagram shows the time and frequency shift performance of a waveform. A ridge in the frequency shift axis indicates poor Doppler resolution, a ridge in the time shift axis indicates poor Doppler frequency resolution. However a sheered ridge, i.e. in both time and frequency shift indicates that targets cannot be resolved if they are time and frequency shifted at a specific time frequency relationship. In addition the ambiguity function shows the range (time shift) and Doppler frequency (frequency shift) sidelobe performance. See reference [2] for further information about the ambiguity diagram.

weighting by applying a non-linear frequency vs. time characteristic to the waveform i.e. Non-linear frequency modulation (NLFM)(Figure D3-2). The spectrum and point spread function (PSF) [3]for the LFM and NLFM waveforms are shown in Figures D3-3 (a) and (b).

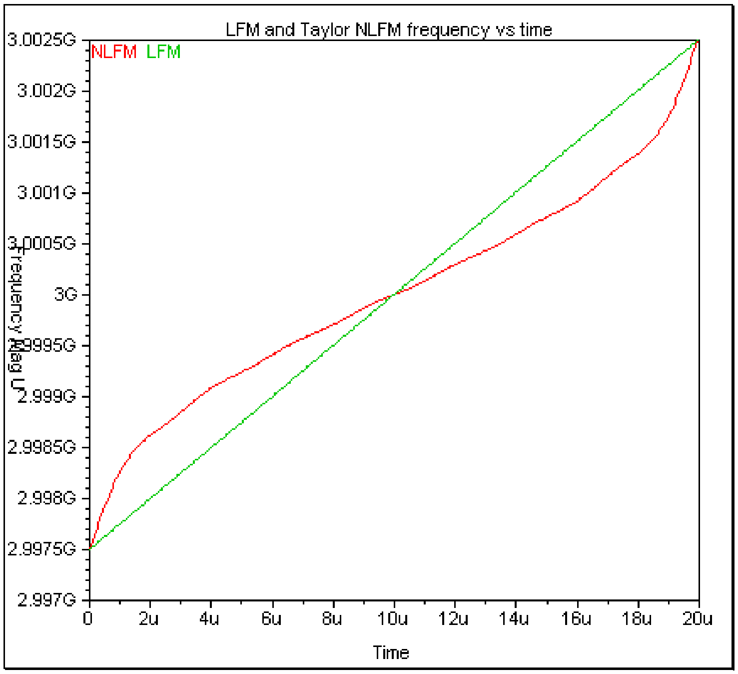

Figure D3-2 LFM and NLFM frequency vs. time.

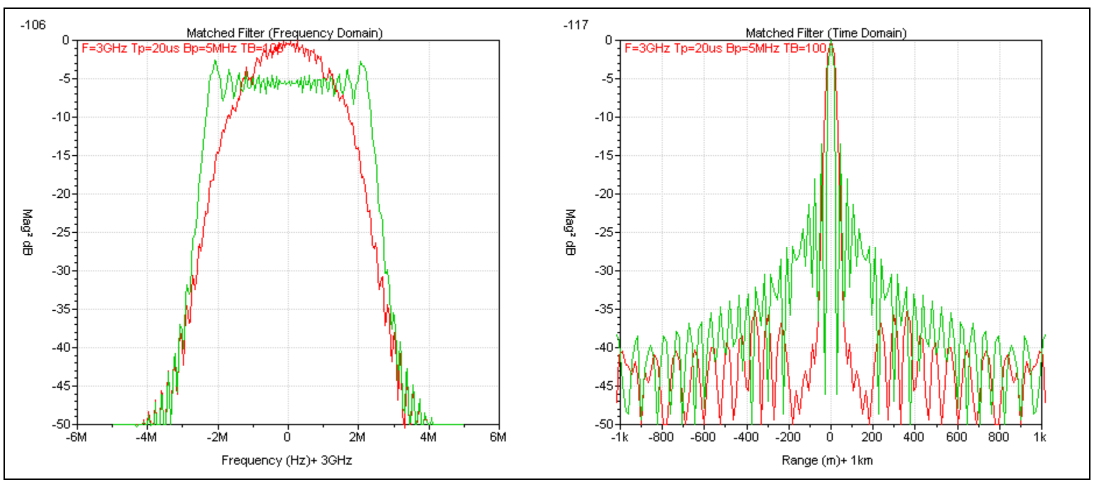

Figure D3-3 LFM vs. NLFM (a) Spectrum (b) PSF.

Whilst, the rectangular pulsed nature of the time response means that there is little difference between the spectra below -30dB in Figure D3-3(a), the NLFM spectrum is narrower above –30dB. The NLFM waveform may be a possible alternative to LFM in constricted spectral environments.

In addition the waveform gives better SNR performance, because no power is lost by windowing in the receiver.

[3] The PSF, mathematically termed the auto-correlation function, shows the time shift performance of a match filtered waveform, it shows the resolution performance from the width of its peak and the range sidelobe performance of the waveform.

Frequency shift keyed (FSK)

There are two reported constructions of frequency shift keyed (FSK) waveforms for radar use where each:

- Pulse of a pulse train is transmitted at a different centre frequency so as to increase the bandwidth of the pulse train [2].
- Pulse is divided into several sub-pulses and each sub-pulse is transmitted at a different frequency.

Pulse to pulse variation of frequency is easiest to generate, but its use is usually for high PRF applications (a category which none of the radars being considered here fall into) or for suppression of ambiguous returns. In addition, it often is combined with other compression techniques, for example LFM, for high-resolution applications.

FSK of sub-pulses is sometimes used as a simple way of producing a LFM ramp, this so called Step FM then has similar properties to LFM. However, its Doppler tolerance is slightly poorer. To completely fill the resolution bandwidth each sub pulse has a frequency step equal to $1/t_s$, where t_s is the length of the sub-pulse and the time bandwidth (TB) product of the waveform is M^2, where M is the number of sub pulses.

Costas Coded waveforms use the sub-pulse approach to FSK, but non-linearly order the frequency steps. This is described further in §D6.3.

D.3.3 Digitally coded waveforms

Digitally coded waveforms typically consist of a train of N sub-pulses or chips of equal duration with each being phase-coded according to a predefined sequence. Examples are binary phase coded (BPSK), where only two-phase states are employed, and Quadriphase, which uses four states. The LFM based Frank and 'P' codes also come in this category, with the number of phase states being governed by their length. The pulse-compression ratio (PCR) of such codes with length τ seconds and N sub-pulses is N, and its nominal range resolution bandwidth is:

$$B = \frac{N}{\tau}$$

Equation D3-4

and hence the time bandwidth product is N. In the simple case coded (phase) modulations have the spectral response given [3] by:

$$S_{rep}(\omega) = \left[\sum_{t=0}^{N-1} \mathrm{e}^{j\phi_t}\, \mathrm{e}^{-j\omega t_c t}\right]\left[\mathrm{sinc}\left(\frac{t_c\omega}{2}\right)\right]$$

Equation D3-5

where ϕ= the phase sequence

t_c=the chip duration

ω= the angular frequency

Equation D3-5 breaks down into a component due to the phase code multiplied by a sinc function envelope, due to the sub-pulse. The phase code component repeats in frequency at $1/t_c$ intervals (see figure). Therefore, the phase coding has little influence over waveforms spectral structure. Although, in §D8.6, an investigation into techniques that are exceptions to this general rule.

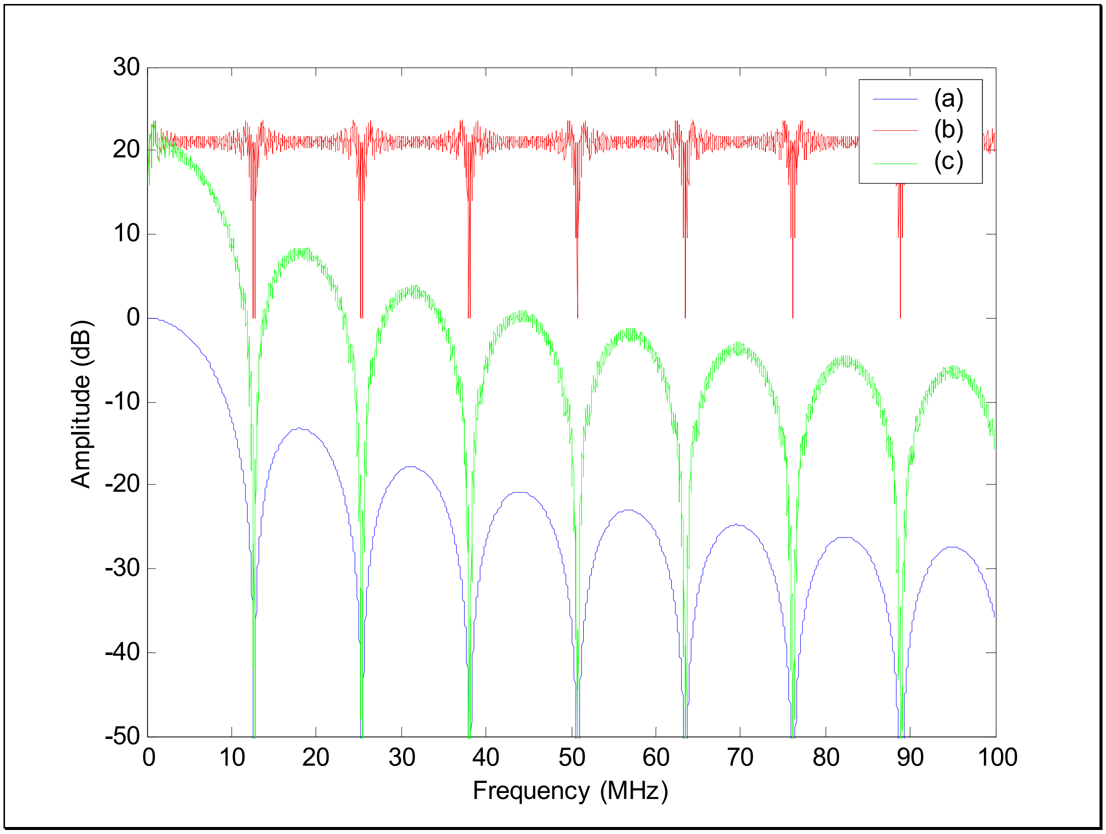

Figure D3-4 showing a coded spectrum of a 10µs pulse phase coded with a 127 length polyphase code (a) sub-pulse component of spectrum; (b) phase code component of spectrum (c) coded waveform spectrum

Phase codes are particularly useful in CW radar modulation, where they have zero sidelobes due to the cyclic nature of their transmission.

In pulsed radar the PSR has an optimum value of $1/N^2$. Binary phase codes exist with this property known as Barker codes [1], but only up to a maximum of N=13. There are codes with more than 2 phases that have this property (known as polyphase barker codes) for values of N up to 36 [4]. Many codes exist with the PSR approaching this optimal value.

By applying weighting functions or filters sidelobes can be reduced subject to small losses and/or extension of the sidelobes in the time shift axis [5,6].

D.4 Interference

D.4.1 Introduction

This study examines the relationship between radar performance, communication systems performance and their ability to operate within a given distance of each other without detectable interference. This definition is less stringent to that given in [7], where the interference to noise ratios of -6dB are considered as being significant. The basic system performances are based upon the following range equations, for radar:

$$SNR_{\det} = \overline{P_r} G_r \cdot \frac{1}{4\pi R_{r\max}{}^2} \cdot \frac{\sigma}{4\pi R_{r\max}{}^2} \cdot \frac{G_r \lambda^2 T_c}{4\pi kTFL}$$

Equation D4-1

which is composed of a transmitted power term, two propagation terms and a receiver term, where:

$\overline{P_r}$ = radar mean transmitted power;

SNR_{det} = threshold for detection;

R_{rmax} = designed maximum radar detection range (for minimum RCS) (10km);

σ = minimum RCS the radar is designed to detect (at the maximum range)(1m^2);

k = Boltzmans constant;

T =system temperature;

F = noise figure;

L= system loss term;

Gr = radar antenna gain (assumed the same for transmission and reception) (30dBi);

λ= illuminating frequency wavelength;

T_c = coherent dwell time, or alternatively the PRI in incoherent systems;

PRI = Pulse Repetition Interval;

and likewise for communication systems:

$$SNR_{rec} = P_c G_c \cdot \frac{1}{4\pi R_{c\max}{}^2} \cdot \frac{G_c \lambda^2}{4\pi kTFLB}$$

Equation D4-2

Which is composed of a transmitted power term a single propagation term and a receiver term, where:

P_c = communication system power;

G_c= communication system antenna gain (same for transmission and reception) (5dBi);

R_{cmax} = designed maximum communications range (10km);

B = bit-rate of the communications system (10^6 bps).

Using these equations, the range separation that represents the threshold at which different systems begin to detect each other or suffer corruption due to interference have been formulated. This Range separation is termed herein the ‘interference range'. At separation ranges below the interference range the systems under consideration suffer from interference. The derivation of the interference range is the aim of the following sections.[4]

D.4.2 Matched radar system to system Interference

The following equation is obtained to calculate the interference range for two radar systems that transmit the same waveform at the same frequency and perform the same receiver function:

$$R_I = \sqrt{I \frac{4\pi R_{r\max}{}^4}{\sigma}}$$

Equation D4-3

In which an interference rejection factor, *I*, is introduced which is used below to examine different system scenarios.

For an example system (using the figures for parameters quoted in §D4.1) Equation D4-3 gives the separation ranges given in Figure D4-1. For a matched system (*I*=1) the interference range is approximately 350,000km! Clearly, the ranges in Figure D4-1 can be scaled to apply to any radar detection characteristic by:

- Accounting for an interference factor *I* of $40\log(K)$ dB per factor *K* in detection range (R_{rmax})[5]
- Accounting for an interference factor *I* of $10\log(K)$ dB per factor *1/K* in RCS (σ)[6]

The interference rejection factor is used to calculate the interference range for situations other than the matched condition. For example, when the main lobes of the two radars do not point towards each other coincidentally then the following may be considered:

- The main to sidelobe ratio of 25dB to be the interference factor giving a interference range of 20,000km;

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[4] Ofcom comment: The analysis of the communication system is greatly simplified, since this analysis is generic and makes no assumptions on the type of communication system used. In particular, it is assumed that:

- The propagation path loss model an inverse square free-space law. At lower frequencies and especially considering cellular systems where multipath exists and the mobile station is low-lying, this may be pessimistic.
- That FEC is not used to make the link more robust; in practice modern telecommunication systems are designed to be error tolerant.
- That the communication system without the radar is noise limited; in practice most systems using intensive re-use patterns are interference limited.
- That no fade margin has been used in the link; typically 10dB of fade margin is used at cellular frequencies which will cushion the cellular system against occasional radar pulses.
- That the communication system is used outside. In many cases, the systems that may co-exist with radar are indoor systems. For example with 802.11a c band there is an additional 20dB of extra isolation assumed for wall penetration losses.
- That many communication systems (e.g. cellular voice) have comparatively low instantaneous bit rate requirements and hence have a high margin to increase power to overcome an intermittent interferer.
- It can be seen that these factors will combine to give the most pessimistic result for interference on the communication system.

[5] If R_{rmax} is increased to 20km *K* is 2 or 40km , *K* is 4 etc.

[6] Similarly if σ is increased to $2m^2$ *K* is 0.5, $10m^2$ *K* is 0.1 etc.

- The sidelobe to sidelobe ratio of 50dB to be the interference factor giving a interference range of 1,000km

In addition, Figure D4-1 can be used to define the required radar performance parameters for a given scenario. For example:

If two similar radar systems are required (with performance defined above) to operate within 3.5km of each other then 100dB interference rejection is required, so that they do not interfere in all but the main beam to main beam scenario. This could be divided into requirements of 30dB main to sidelobe ratio and 70dB in waveform rejection.

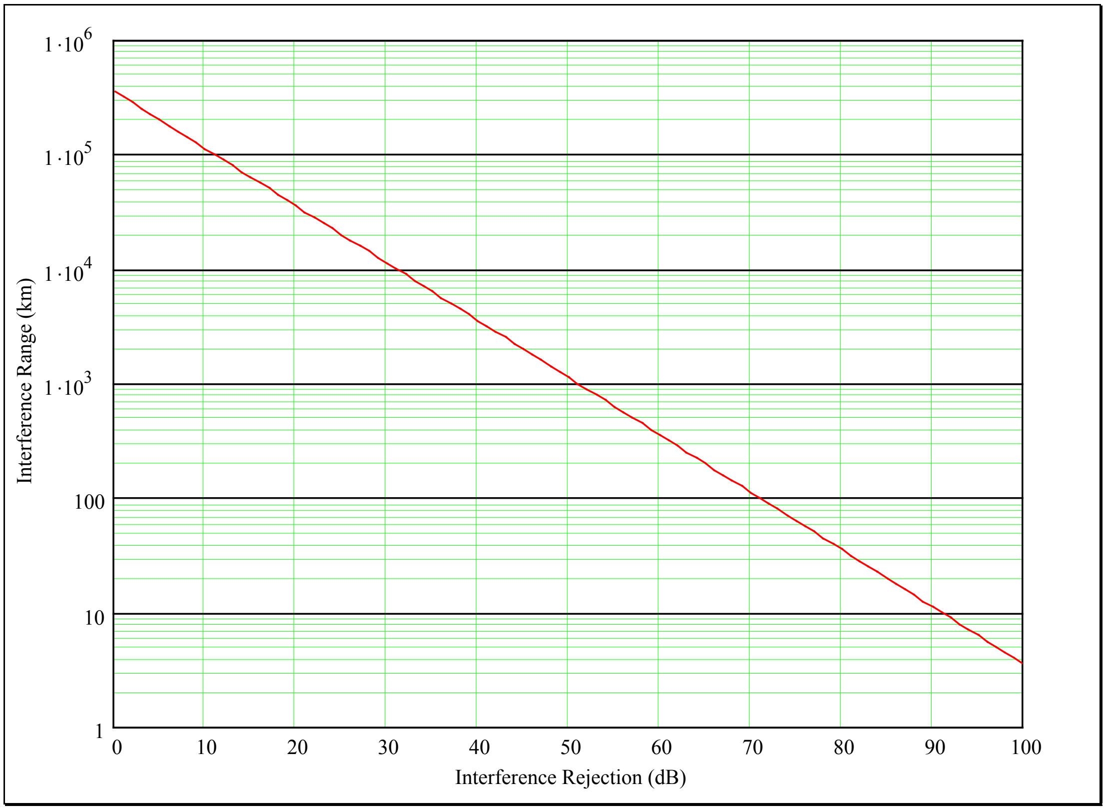

Figure D4-1 Interference range between matched radar systems designed to detect 1m² targets at 10km

D.4.3 Radar system to communications system interference

The equation that calculates the range from a radar at which an unmatched system such as a communications system suffers no interference and assuming they have the same operation frequency, is:

$$R_I = \sqrt{\frac{4\pi R_{r\max}{}^4}{\sigma} \cdot \frac{R_{in}}{R_{\min}} \cdot \frac{SNR_{\det}}{SNR_{rec}} \cdot \frac{G_c}{G_r} \cdot \frac{T_{ce}}{T_c}}$$

Equation D4-4

The first term relates to the radar range performance as in Equation D4-3. The next term is the ratio of the instrumented and the minimum ranges, which transforms the mean transmitted radar power into peak power. The following three terms express the ratios of the SNRs, antenna gains and coherent dwell times (T_{ce}) in the different systems.

Where R_{in} = the instrumented range

$R_{min} = DR_{in}$ = the minimum range

D = Duty ratio

$$T_{ce} = \begin{cases} \dfrac{1}{B} & , \quad \dfrac{1}{B} \le \tau \\ \text{ceil}\left(\dfrac{1}{B \cdot PRI}\right) \cdot \tau & , \quad \dfrac{1}{B} > \tau \end{cases}$$ = the effective coherent dwell time

τ = the pulse width

$$PRI = \frac{2R_{in}}{c}$$ = the radar pulse repetition interval

Figure D4-2 shows the response of this equation, using typical values for those factors that have not been defined (given in caption). It shows that the radar duty ratio makes no difference until the pulse length is greater than the period of a communications bit, after which the interference range reduces. Also, it can be seen that the interference ranges are lower that the matched system ranges in Figure D4-1 due to the mismatched nature.

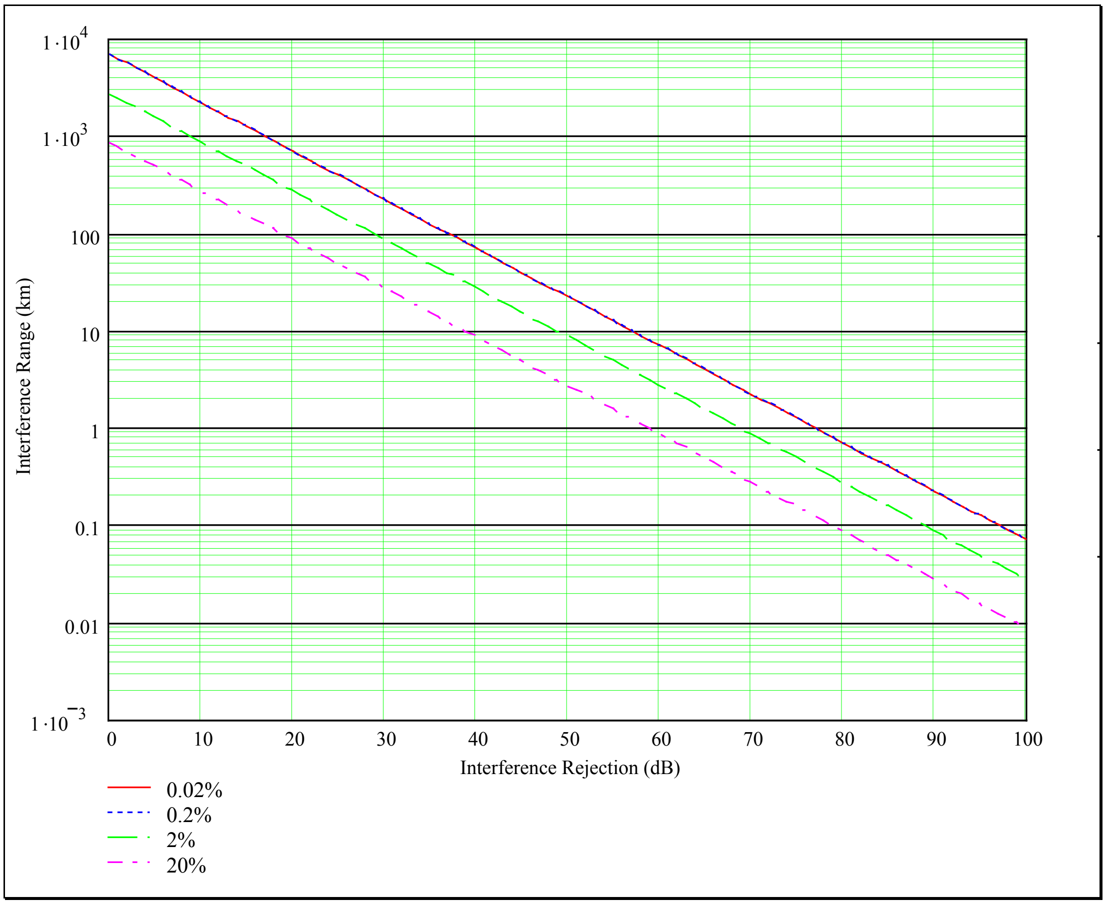

Figure D4-2 Showing the interference range for an instrumented range of 50km and duty ratio from 0.02% to 20%. with T_c= PRI , SNR_{det} =11.3dB and SNR_{rec}=20dB

D.4.4 Communication system to radar system interference

The communication to radar system interference range is calculated in a similar way, but based upon the communications system performance rather than the radar system performance, this gives:

$$R_I = \sqrt{IR_{c\max}{}^2 \cdot \frac{G_r}{G_c} \cdot \frac{SNR_{rec}}{SNR_{\det}} \cdot T_{ce2} B}$$

Equation D4-5

where T_{ce2} accounts for the mismatch between communications waveform and the radar's coherent processing dwell time, which is ideally:

$$T_{ce2} = \begin{cases} \sqrt{N_c} PRI & , \quad \frac{1}{B} \geq \tau \\ \sqrt{N_c \frac{1}{\tau B}} PRI & , \quad \frac{1}{B} < \tau \end{cases}$$

Equation D4-6

where N_c= the number of coherently integrated pulses in a radar system.

A more pessimistic estimate results in $T_{ce2} = T_c$. This results in only small differences in the results for the typical radar systems considered here. Figure D4-3 shows the performance results using the idealised equation, again for a number of different radar duty cycle values.

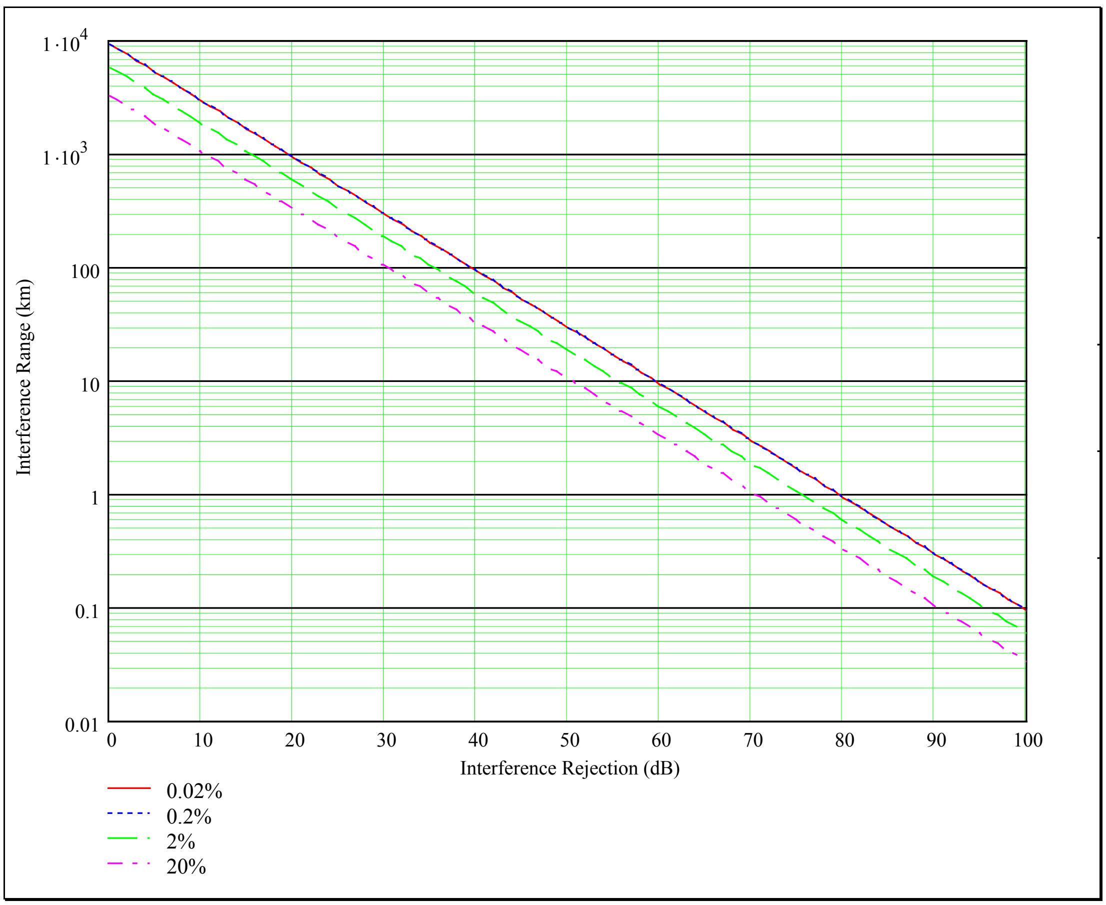

Figure D4-3 Showing interference ranges between a typical communication system and a radar system operating with 50km instrumented range and duty ratios as indicated, SNR_{det} =11.3dB and SNR_{rec}=20dB.

D.4.5 Discussion

The ranges calculated at which interference begins to occur between two matched radar systems is huge and as such likely to be reduced by other factors such as obscuration or earth curvature.

The typical radar systems described in §D2 have similar detection performances to those investigated in this section.

This sets challenging interference rejection targets for like systems operating within a given area. However, it is apparent that the interference ranges are derived from radar performance, which poses the questions:

- Is the performance of current radar systems excessive for purpose?
- Could performance be reduced (power control) in circumstances where there is a high density of systems?
- Is the interference significant to the operation of the radar system?

The answer to the final question for the current generation of radar systems is 'No', as techniques have evolved with the technology to reduce effects of interference and these are discussed in §D8.3. The effect on modern waveforms and processing are also discussed and concludes that the answer to this question may not be as clear.

The challenge is set that if the typical systems described in §D13 are to work within a 3km area the interference rejection needs to be in excess of 100dB. The Antenna sidelobes may be expected to reject 30-40dB of the energy, this leaves greater than 60dB being required of the waveform to reject interference.

The advantage of the high duty ratio radar waveforms is apparent in the interference to and from mismatched systems (Figure D4-2 and Figure D4-3). For example, 20% duty ratio waveforms require between 10dB and 20dB less interference rejection than the lower duty waveforms. Such high duty ratios are only feasible if PC waveforms are implemented to obtain useful range resolution and multi-modal waveforms are implemented to detect targets from close ranges.

The impact of interference on communication systems from current radar systems is recoverable, because correction algorithms work effectively against impulse noise like that from current pulsed radars systems. However, for longer pulses or CW radar it may be less effective, which is an area for further study, simulation and demonstration.

The interference from communication systems on a radar is similar to a noise jammer, where the overall noise/ clutter will be increased, a scenario in which radars show graceful degradation. This results in a reduced detection range performance, which is preferable to multiple target declarations swamping data processing. However, in safety critical applications loss of sensitivity is likely to be unacceptable unless:

- The interference can be controlled to be within an agreed radar coverage arc.
- Or interfering communication systems are co-operative with the radar such that they vary their interference levels with radar operation and detection conditions.[7]

[7] Ofcom comment: It may also be possible for devise mitigation techniques that could limit the interference of the communication systems on the radar. These could range from wholly passive techniques such as greater control of antenna pattern shape at base stations to more active methods such as the communication system learning whether radar systems are active within its bandwidth and either choosing a different frequency, limiting the transmit power

D.5 Regulations and frequency masks

D.5.1 Introduction

There are two sets of regulations governing the spectral extent of signals, those set by the ITU [8,9] and, when these don't apply due to numerous exclusions, those set by the NTIA manual [10] (or the 'redbook').

Both sets of regulations define masks based on the -40dB bandwidth (ITU) or the allowed bandwidth ('redbook'), in which the transmission spectra must lie beneath. These masks, together with spectrum measurement techniques, define the allowed emissions close to the resolution bandwidth. Frequency emissions in remote frequency bands (defined as the spurious domain) are regulated by different regulations [11].

This section examines and critiques the masks and their applicability to the waveform that they regulate. All the radars in the database do not meet the exclusion conditions of the ITU so only the ITU masks are discussed further in this section.

The derivation of -40dB bandwidth and the resulting masks for simple pulsed, LFM pulsed FMCW unmodulated CW and phase coded waveform are reviewed and discussed in a summary.

Following this, a concept based on the mask is described to reduce spectral occupation. In addition, the effects of altering the peak power are discussed in the light of coherent and pulse compression radar.

Finally, the spectral occupation of the typical systems given in §D13 are illustrated based upon the masks. In addition, the effect of changing all radar systems to a more compact waveform and the impact of universally adopting PC is illustrated, in which reductions in spectral emissions are quantified.

D.5.2 Simple pulse

Regulations and texts on pulsed waveforms consider pulses as being trapezoidal, which facilitates numeric analysis. Edges can be formed by a simple convolution (Figure D5-1), where τ is the pulse width and t_r is the rise time. The convolution can be achieved by multiplying the two pulses in the frequency domain. For the basic pulsed case, the trapezoidal pulse is a multiplication of two sinc functions (the sinc function being the Fourier transform of a rectangular pulse). This leads to the spectrums given in Figure D5-2.

Figure D5-2 shows that the ITU –40dB bandwidth figures from [8] match the theoretical spectral responses and the roll off of all the spectra is 20dB/decade or better, as per the mask description.

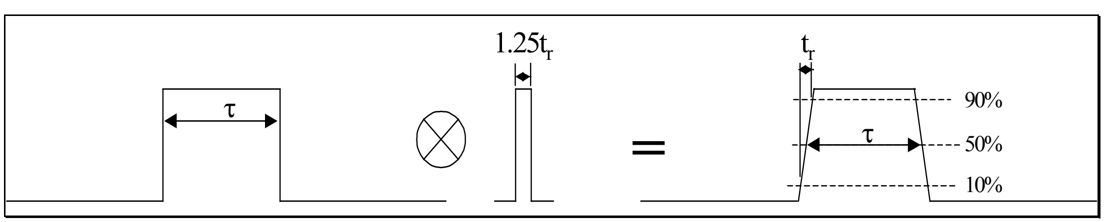

Figure D5-1 Mechanism for generating trapezoidal pulses, by convolution of 2 pulses

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or using time slots when the radar is not active. There is precedence for this approach with DFS in 802.11h systems. Further analysis of coexistence is beyond the scope of this project. However, it is recommended that other studies be undertaken to see what systems could be made to co-habit with radar in suitable frequency bands.

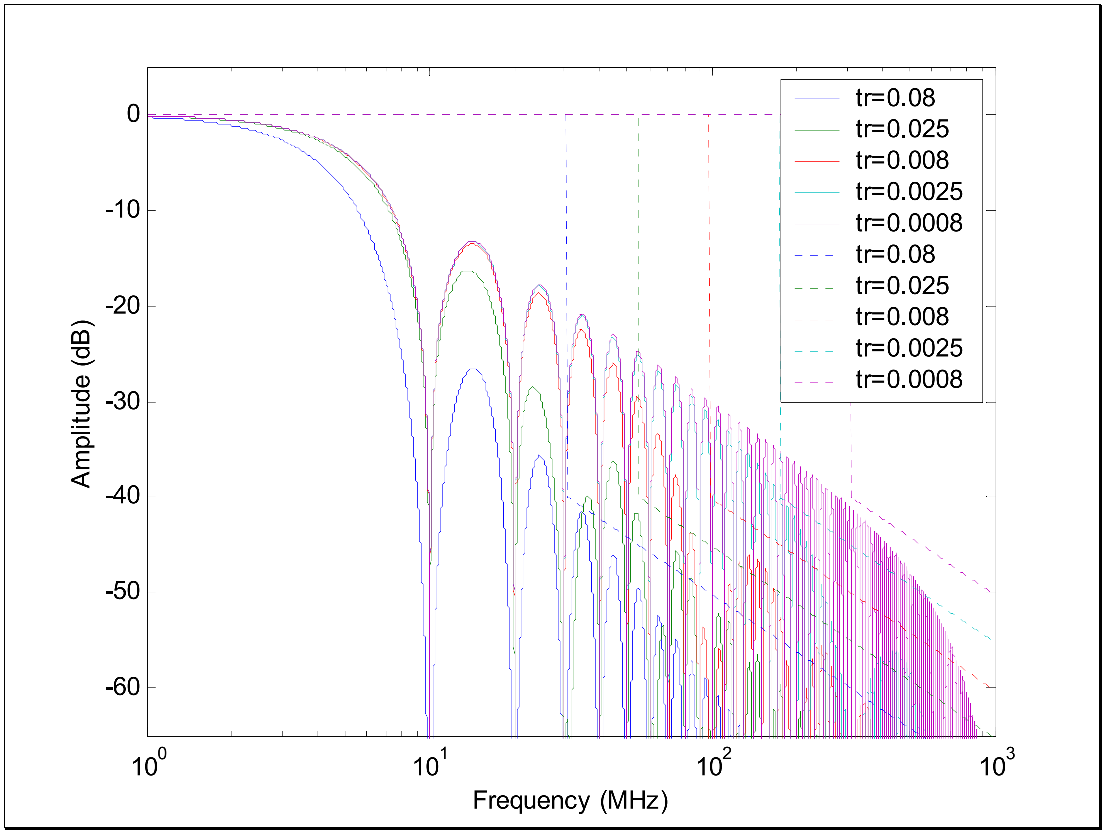


Figure D5-2 Spectrum of a simple trapezoidal pulse (solid), with pulse width 0.1µs and various rise times (all given in µs), also respective ITU-R masks (dotted)

D.5.3 LFM

Definition of the LFM pulse mask is currently under dispute, and hence definition of a new mask description is of current concern to regulatory bodies. Therefore, in this section, the exact derivation of the LFM spectrum is discussed, and compared to other techniques used in recent publications on this contentious issue.

The definition of an ideal LFM pulse is given by Equation D3-1 and the exact solution by solving the Fourier transform of an ideal LFM pulse (for f_0=0) is:

$$\int_{-\infty}^{\infty} S(t)e^{j\omega t}\,\mathrm{d}t = j2\sqrt{\frac{\tau}{B_c}}e^{-\frac{j\tau\omega^2}{4\pi B_c}}\left(\operatorname{erfi}\left(\frac{\sqrt{j\tau}\left(\omega-\pi B_c\right)}{\sqrt{4\pi B_c}}\right)-\operatorname{erfi}\left(\frac{\sqrt{j\tau}\left(\omega+\pi B_c\right)}{\sqrt{4\pi B_c}}\right)\right)$$

Equation D5-1

where *erfi* is the complex error function.

The common technique used in the literature to examine trapezoidal pulsed LFM spectrums is to generate accurate digital time domain representations and obtain their spectrum by digital frequency transform. However, using the exact spectral definition given in Equation D5-1 with the technique explained above for generating trapezoidal pulses negates to need to use the common digital frequency transform technique. As shown by Figure D5-3 the accuracy of the digital transform technique become poor once the frequency is greater than between 10% and 20% of the digitised waveform sample frequency. Therefore, the authors would advise against the use of such digitised techniques.

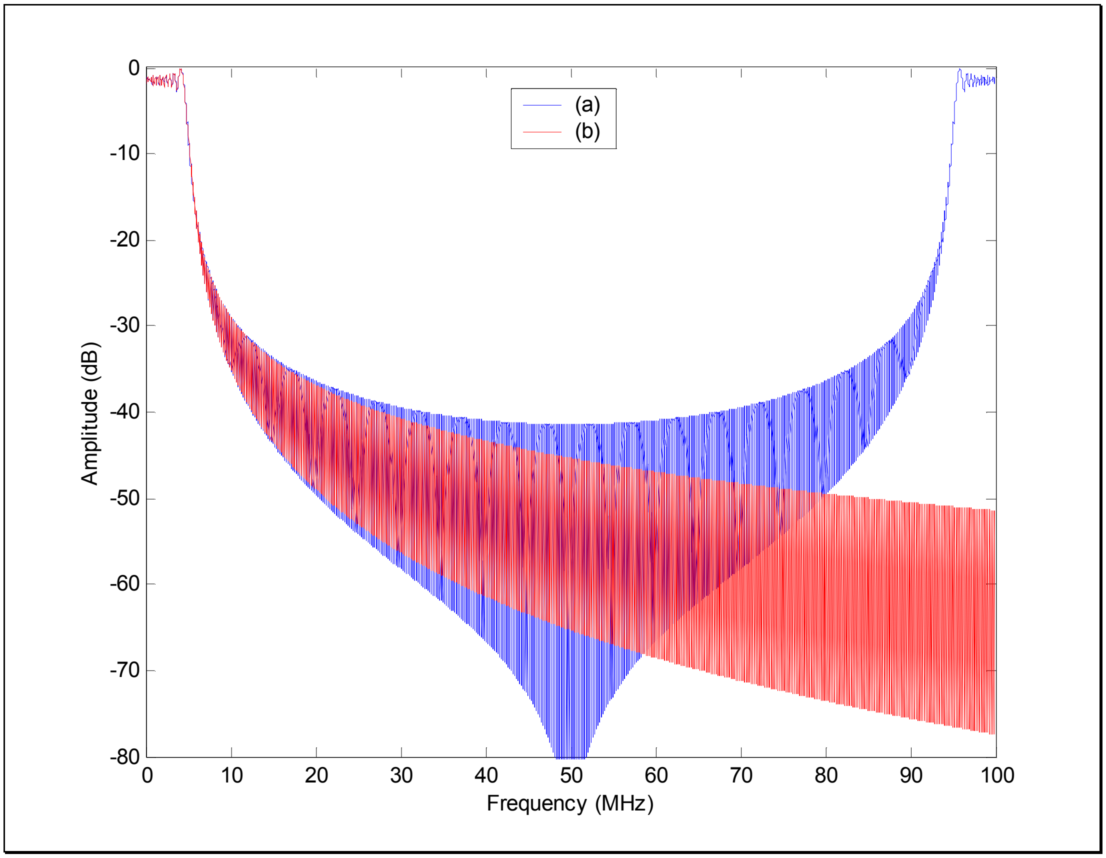


Figure D5-3 LFM spectrum of a 10μs pulse with 10MHz sweep bandwidth. (a) using a digitised approximation (digitised at 100MHz); (b) using the direct exact solution

The exact solution for trapezoidal pulse together with the masks defined by the -40dB bandwidth value given by formulas in reference [8] are given in Figure D5-4. However, because the figures for the pulsed FM -40dB bandwidth masks are in contention [12,13], the current thinking is to redefine the formula to be:

$$BW40 = \frac{k}{\sqrt{\tau t_r}} + X\left(B_c + \frac{A}{t_r}\right)$$

Equation D5-2

where X = 1.2 whereas in reference [8] it is 2.0

A= typically 0.105, as in reference [8]

k = typically 6.2 as in reference [8]

In addition, in 2006, design aims are to be introduced that determine the mask roll off from the -40dB point is to be 40dB/decade. Masks for this stringent description are shown in Figure D5-5.

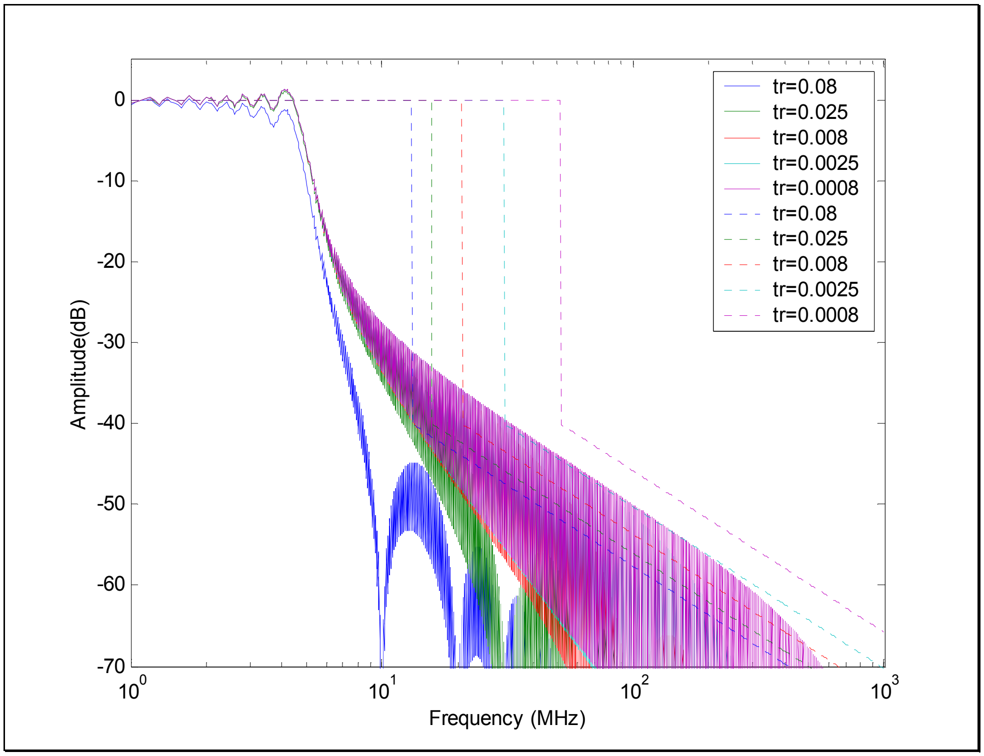


Figure D5-4 Spectrum of a trapezoidal LFM pulse with 10μs pulse width, 10MHz frequency sweep and variable rise times, solid lines show pulse spectrum for different rise times and dotted lines show masks generated ITU-R SM.1541-1 annex 8 definitions.

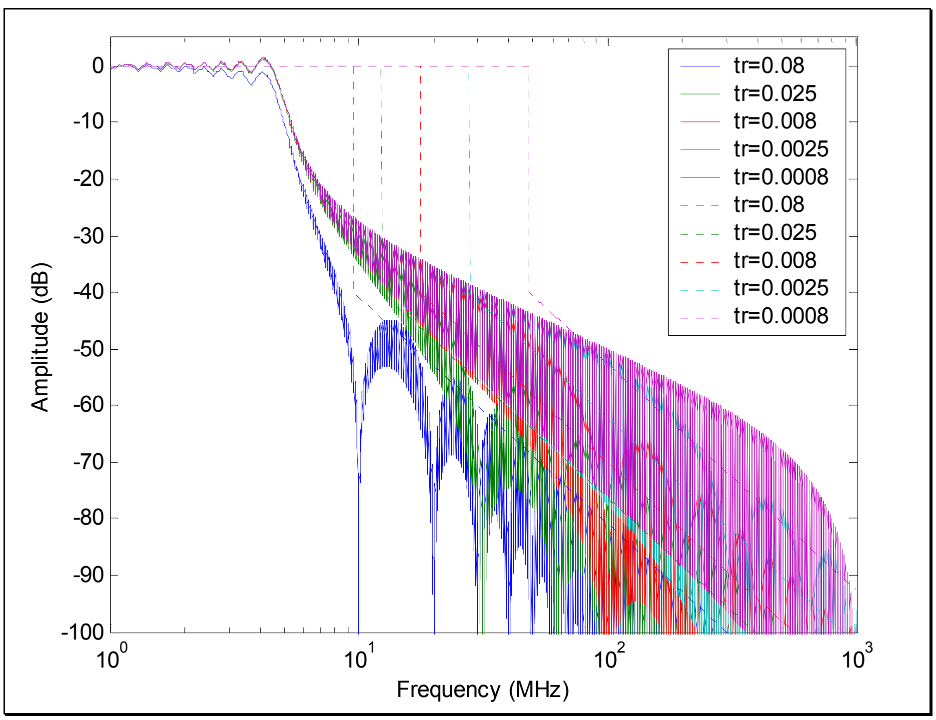


Figure D5-5 Spectrum of a trapezoidal LFM pulse with 10μs pulse width, 10MHz frequency sweep and variable rise times, solid lines show pulse spectrum for different rise times and dotted lines show masks generated using stringent guidelines.

D.5.4 FMCW

The FMCW mask has had little attention and is very simple, the -40dB bandwidth is calculated using:

$$BW40 = 2B_c + 0.0003 f_0$$

Equation D5-3

where f_0 is the centre frequency

The f_0 component is common to the un-modulated CW radar,-40dB bandwidth calculation and presumed to be a source frequency stability component, similar to those given in reference [10]. However, even with this stability component and assuming a perfect source the CW mask it generates is not adequate to contain the FMCW spectral envelope, see Figure D5-6.

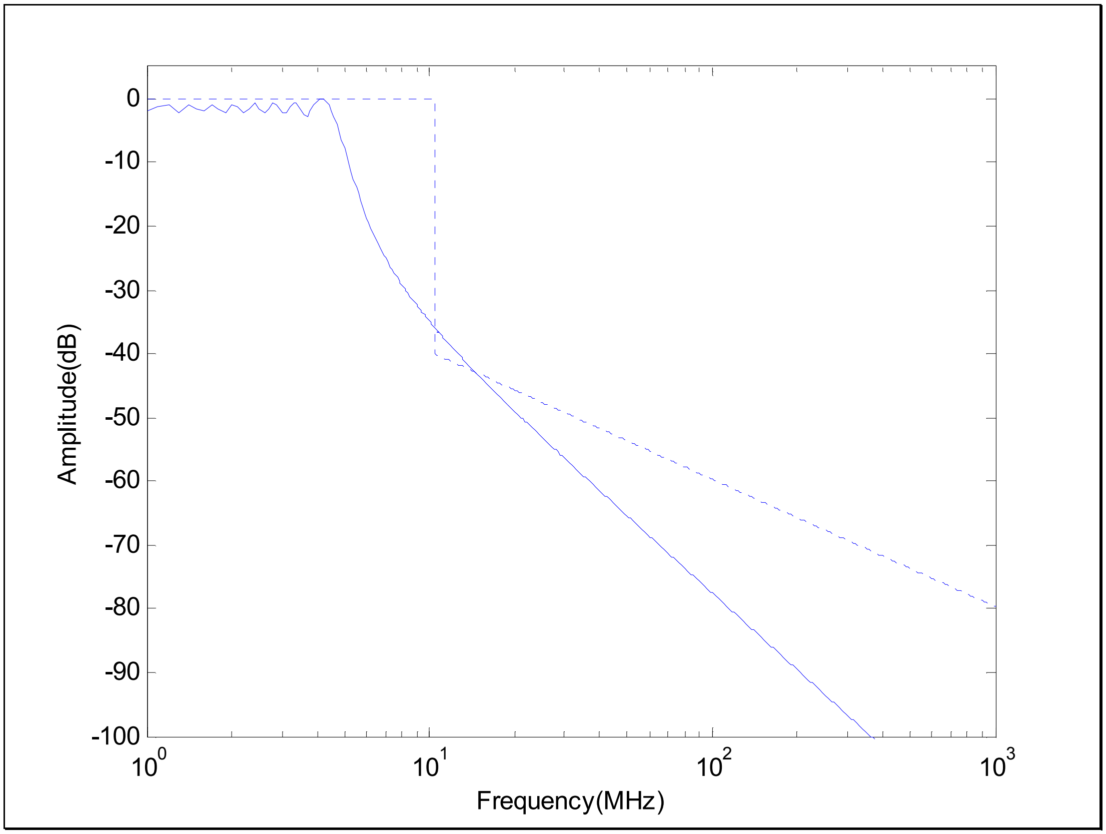


Figure D5-6 spectral envelope of a FMCW signal with a 10MHz sweep (solid) and SM1541 mask (dotted) assuming a 3GHz transmit frequency.

D.5.5 Un modulated CW

Only provisions for 300ppm of the centre frequency, f_0, is given for CW transmissions, which is less than frequency tolerance figures given in the 'redbook' for all transmissions above 470MHz.

D.5.6 Coded modulation

For coded modulated waveforms, the current regulations rely on using the pulsed masks defining the pulse width parameter to be the sub-pulse width. This is in agreement with the coded waveform spectra properties described in §D3.3. However, the definition of sub-pulses could be ambiguous in some waveforms, such as those composed of multiple codes transmitted at different phases (i.e. quadriphase codes, see §D11.7).

D.5.7 Discussion on spectral allocation masks

The spectral allocation masks attempt to represent the theoretical spectrum of different waveforms. Although simple trapezoidal pulses are well defined by the masks, other waveforms are less well defined.

The width of the masks for pulsed FM waveform are narrower than the equivalent resolution pulsed masks. However, the current pulsed FM masks appear to be relaxed. This may be tightened in possible upcoming amendments to the regulations [13]. However, FMCW masks are narrower than pulsed FM masks and appear to be tight considering theoretical responses.

This prompts questions to the motivation of these regulations, do they:

- Regulate efficient use of the spectrum?
- Ensure faithful representation of theoretical waveform performance?

The masks, clearly, do not perform the first task in that they do not penalise waveforms which have broad spectrum (based on the pulse waveform definition) whilst penalising those based on the narrow spectrums (CW definitions). However, they do attempt to satisfy the second objective, this establishes a level-playing field as it is likely that radar waveforms of all descriptions are required to satisfy all the varied tasks performed by the radars under consideration.

Once this level playing field is established, representing theoretical performance, should the radar producers then be allowed an additional comfort zone? It is acceptable that implementation generally leads to excursions from theoretical performance. For example, stability of RF sources appears to be considered for CW waveform, but not other waveforms.

One degree of freedom that is offered to the radar designer by the regulations is by use of the trapezoidal pulses in derivations, which imply transmit chains with only single pole filter performance. Therefore, by careful pulse shaping or transmit waveform filtering, grounds exist to exceed performance requirements.

Pulse shaping is most difficult in high gain-high power TWT amplifiers, where such high Q, high power filtering may not be possible or practical, and amplifiers are ideally driven quickly into saturation. Therefore, rise time control and design of optimal transmit chain gain to achieve such rise times may be an issue for further research.

Finally, under current guidelines loopholes exist, where the worst-case mask defines operating masks. The worst case mask is generally defined by the mode with the shortest, lowest mean power pulse, which could never be used or only in occasional circumstances (e.g. when a ship docks). Whereas, higher mean power modes actually could operate in a smaller masks. This issue is addressed, in part, in the following sub-section.

D.5.8 Matched transmitted waveform concept

Matching receiver response to waveform response is a familiar technique in radar systems engineering that is used to optimise radar detection performance. Under these conditions, the receiver bandwidth is such that it is just sufficient to achieve its required range resolution. For simple pulsed radars, this gives a bandwidth of approximately the reciprocal of the pulse width ($1/\tau$).

However, radar transmitter bandwidths are generally considerably greater than what is required for their range resolution performance. This results in fast pulse rise time figures, broad spectra and the large allowed bandwidths under current legislation.

The concept of matched transmitted waveforms is thus to limit the transmitters 3dB bandwidth to that required by the radar range resolution bandwidth. In a practical system, this could be achieved by pulse rise time control or by transmitter filters. The filter technique is likely to require high Q fixed characteristic filters, and the poor stability of the RF source leads to a requirement that the filter is larger than the ranging waveform 3dB bandwidth or that the filters are (rapidly/actively) tuneable.

Rise time control is likely to lead to better overall system spectral responses in multi-modal radars, where the pulse widths, 3dB bandwidths and rise times, could be different in the different modes of operation.

This technique is demonstrated in §D5.10.

D.5.9 Reduction in peak power

The interference between mismatched systems (between radars with different waveform characteristics or between radars and communications systems) is derived from their peak power, and degree of match. Radars that operate with PC waveforms achieve equivalent mean powers to a simple pulsed radar equivalent by using *N* times higher duty waveforms, and *N* times lower peak powers, where *N* is the pulse-compression ratio (PCR). Therefore, radars with pulse compressed waveform interfere with other systems to a lesser degree. This property is apparent in the interference study, Figure D4-2 and Figure D4-3. The peak power reduction factor is given by:

$$K_{pc} = \frac{1}{N}$$

Equation D5-4

where *N* is the pulse compression ratio (PCR).

In addition to pulse compression, modern radar designs can operate with reduced peak power by reducing receiver losses and increasing the receiver efficiency. These can result from digital signal processing (DSP) that is now achievable at a relatively low cost. Such digital techniques can improve receiver matching and can perform coherent processing. For coherent integration of multiple pulses, the reduction in peak power due to higher efficiency is at best:

$$K_{coh} = \frac{1}{\sqrt{M}}$$

Equation D5-5

where *M* is the number of pulses integrated over a coherent dwell time.

For typical simple pulsed magnetron radars in the database, peak powers could be reduced by of order 20dB due to pulse compression 1 to 5 dB due to matching, and 10dB due to coherent processing. Thus typical radar peak transmitted power could be reduced by typically 30dB with corresponding reductions in peak emissions across the entire spectrum. Interference between such systems is then influenced by their degree of match, as discussed in §D6.

D.5.10 Spectrum of typical radar systems

The pulse mask and the stringent FM mask formulae give us a method of representing the current spectral usage by the typical radar systems listed in §D13. Assuming that they adhere to the equations (i.e. they are compliant with the ITU regulations) the current equivalent radiated peak powers are given in Figure D5-7, when measured in a 1MHz bandwidth.

Figure D5-7 shows that pulses from the same radar (e.g. Marine-commercial S and marine Commercial S(2)) have different measured peak powers due to different bandwidths, but eventually follow the same spectral sidelobe mask due to having the same rise time.

Figure D5-8 shows the effect of implementing the matched transmitter concept by rise time control on all pulses from the exemplar radar systems. Here, the long pulses have tighter spectrums, and overall spectral occupation is reduced.

The -40dB bandwidths of the spectrums are reduced by factors of two for short pulses and up to factors of ten for long pulses.

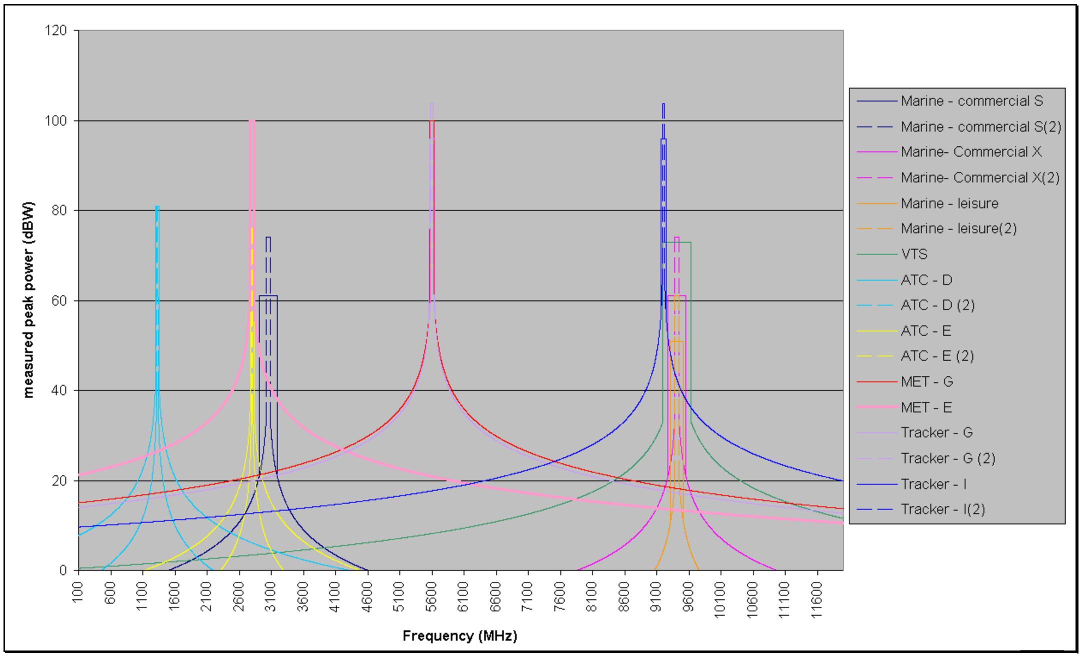


Figure D5-7 Spectral occupation of typical systems with bandwidth defined by relevant BW40 mask equations.

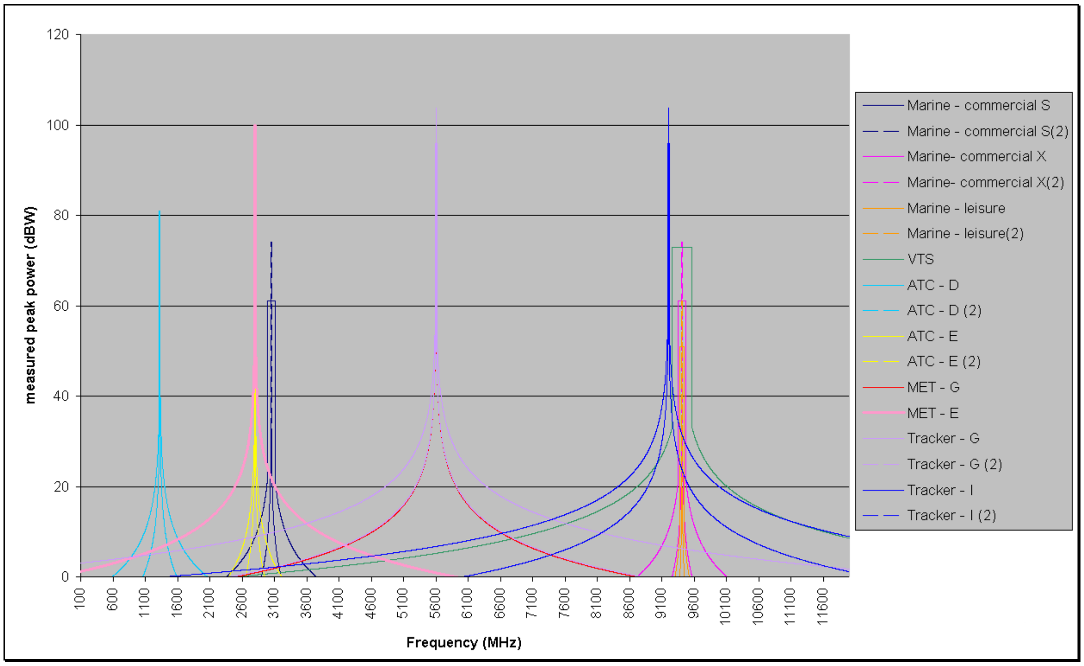


Figure D5-8 Spectral occupation of typical systems with BW40 calculated assuming matched transmitter bandwidth.

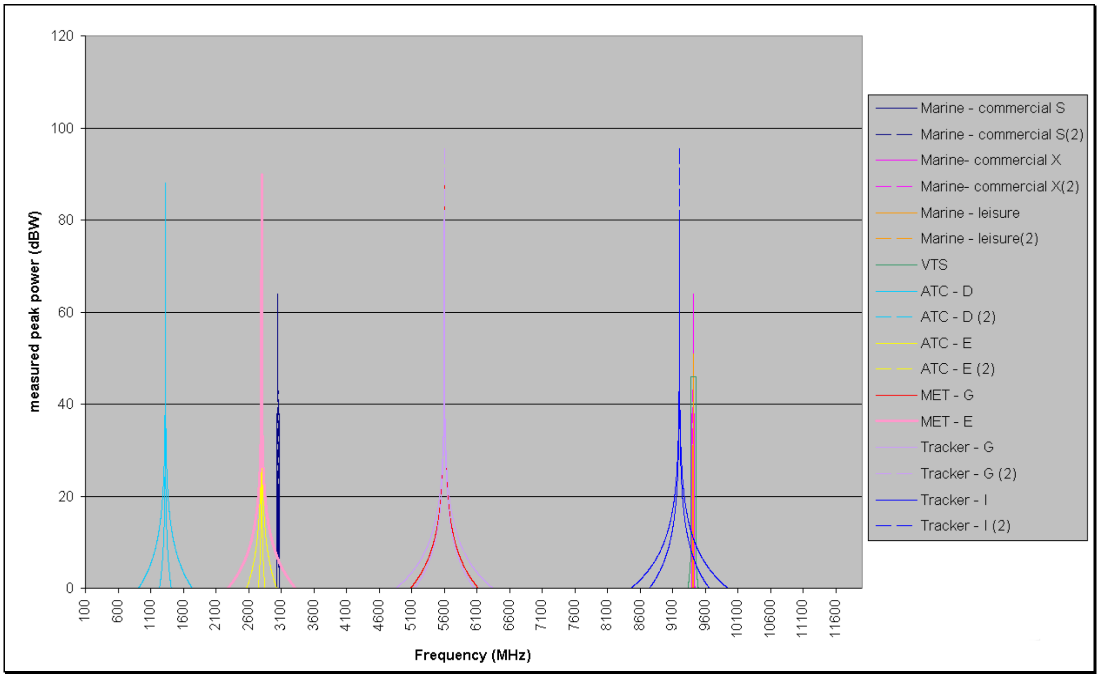


Figure D5-9 Spectral occupation of typical systems when all waveforms use 10μs FM pulses with matched transmitter bandwidths.

Finally, the effect of implementing FM pulse compression with 10μs pulses and B_c determined by the same range resolution bandwidth of the current system's pulse is given in Figure D5-9. The impact of narrower spectral masks is clear to see. The -40dB bandwidths of the spectrums are reduced by factors of between ten to twenty.

D.5.11 Summary of regulations and frequency masks

The ITU masks have been shown to give faithful representations of theoretical waveform performance. Therefore the current spectral usage of a subset of typical radar systems has been calculated, assuming their waveforms are compliant with the ITU masks. The band between 100MHz-11.6GHz has been shown to suffer from emissions across its entirety of above 20dBW mainly due to the spectral spreading of the radar waveforms.

Two simple techniques have been devised for reduction in the spectral responses of the current radar waveforms:

- Limiting the rise time of the pulse to that which is needed to obtain the radars designed range resolution.
- Implementing pulse compression to lower the entire peak power amplitude of a mask, and in the case of FM pulse compression, this improves the compactness of the spectrum.

In addition, receiver efficiency and coherent processing has been discussed, which can further lower the peak power level.

This investigation revealed that rise time reduction can reduce the mask widths of the typical radar systems by factors of between 2 and 10, whilst FM pulse compression can reduce mask widths by a further 2 to 5. In addition, the entire masks may be lowered by typically 30dB for the typical radar system parameter sets.

Power levels of emissions over the spectral band have been shown to be considerably improved by rise time control and FM pulse compression.

D.6 Orthogonal waveform

D.6.1 Introduction

Ideally, a radar's matched filter would give no response to codes other than its own. However, in-band energy received from interfering sources has to go somewhere and in the ideal case it will appear as an amorphous response distributed over all the range bins. The best that can be expected is a rejection of approximately *1/N*, where *N* is the radar's time-bandwidth (TB) product i.e. 21dB for a 128-chip code. This level of rejection is difficult to achieve in practice but it is clear that rejection increases with longer codes (i.e. with large *N*). Longer codes mean larger time-bandwidth products (TB) i.e. longer pulse lengths and/or wider bandwidth.

The problem of forming truly orthogonal waveforms can be shown by considering the radar's matched filter operation in the frequency domain:

$$Z_{MF}(f) = Z_R(f)\, Z_T^*(f) \qquad where|\, f\,| < B/2$$

Equation D6-1

The receiver output is the product of the received signal Z_R and the complex conjugate of the transmitted signal Z_T at each point within the bandwidth *B*. Any interference signal coinciding with regions of finite Z_T will therefore be passed on to the receiver. As broadband radar spectra tend to be continuous within the main pass band (see Figure D5-2 to Figure D5-6), any interfering signal within its pass band will also be accepted. The best that can be achieved with this approach is that the interfering signal does not correlate i.e. its energy is dissipated over all the range bins whereas the wanted signals will correlate and be concentrated into single range cells.

This rejection property is, commonly, described as an orthogonal property, whereas mathematically the property is one of independence, rather than orthogonality. However, whilst inaccurate, the orthogonal term is widely adopted and is be used in this report.

The second aim for such waveforms is that they deliver their primary role, which is to perform as good as PC waveforms. Such waveforms are ones that have low range and Doppler sidelobes. The lower limit of range sidelobes for pulse compression is $1/N^2$ as discussed in §D3.3. Such performance is critical in high dynamic range environments, where high clutter levels are expected or large targets within close range separation with the smallest targets of interest. This is the case with many of the radars being considered by this report. However, the issue of waveforms with low cross-correlation and low auto-correlation sidelobes has been previously reported by Sawarte [14], in which these metrics are reported to be dependant on each other. As such, the optimum PC waveforms for radar applications are unlikely to have optimum cross-correlation performance and vice-versa. Therefore, both correlation metrics need to be considered in consideration of candidates for spectral sharing radar waveforms.

One can conceive of truly, orthogonal waveforms e.g. a gapped LFM where the two waveforms are the spectral complement of each other. Mutual rejection would be reasonable in the zero Doppler case, but Doppler shifts would allow the spectrum of one to encroach on the other thereby reducing rejection performance. With regular or irregular gaps in the spectrum, the sidelobe performance would be poor and significant signal processing effort would be required to remove the gaps (see §D8.5).

Orthogonal waveforms and their issues will be illustrated in the following sections, which examine implementations of different classes of waveforms that are designed to meet the orthogonal waveform properties described above.

D.6.2 Maximum length sequences

The pseudo-random nature of maximum length sequence (MLS) codes would appear ideal orthogonal waveform candidates as entirely different codes can be transmitted by the selection of different coefficients.

Considering the worst case, where two radars are transmitting codes of the same length in the same bandwidth. Figure D6-1(a) and (b) shows the response of the radar to a 128 chip quintic (5 phase states) code drawn from the same and different MLS sequence respectively. For MLS and other pseudo-random codes, the radar's response to these foreign codes is generally of the same order as its range sidelobes i.e. between $1/N$ and $1/\sqrt{N}$. In this case, the cross-correlation rejection is about 15dB, where the optimum level for such a time bandwidth product is 21dB.

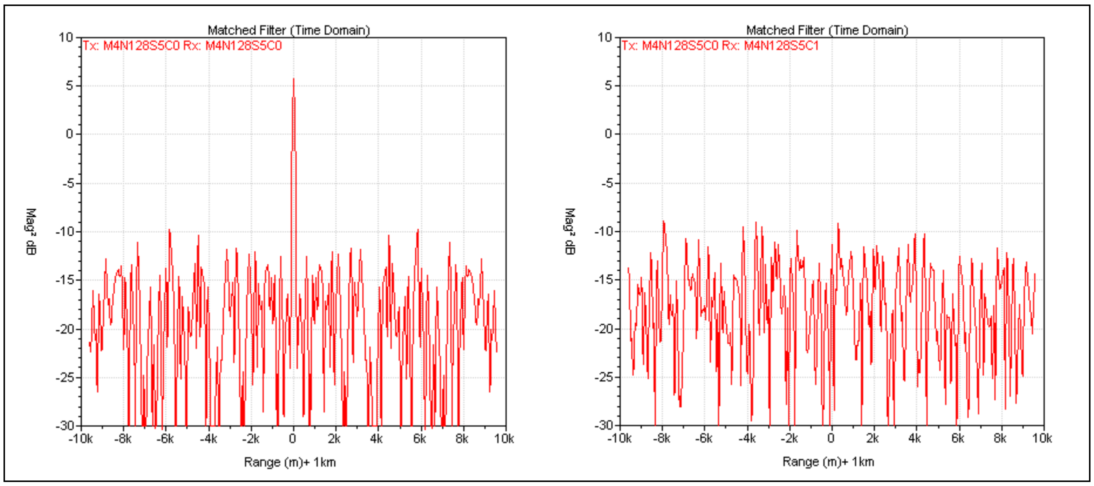


Figure D6-1 128-element MLS Quintic code PSF (a) same code (b) different code.

D.6.3 Costas codes

Costas codes (sometimes called Costas arrays or Welch codes) were designed for frequency shift keyed waveform (FSK) waveforms so that they have minimal correlation in the delay (range) and frequency shift (Doppler) axis. Costas waveforms require M frequency steps of $1/t_s$ where t_s is the time duration of each frequency sub-pulse, so the TB product is approximately M^2. The waveform ambiguity response aspires to the thumb tack response[8]. Typically the power sidelobes are less than $1/M^2$ of the peak power. However, there are some exceptional sidelobes that can be 4 times (6dB) higher than this level. These large sidelobes tend to be close to the main lobe, so they may not have the same impact as if they were else where [15]. The auto-correlation function of a pulse with a Costas waveform is given in Figure D6-2 to illustrate these points.

There are Costas codes with TB products (M^2) for almost all values of M and for large M the number of different codes for each TB product becomes large. Different Costas codes are known to have orthogonal properties [16]. This is shown by the cross correlation of two pulses with Costas waveforms, given in Figure D6-3. However, the peak sidelobe is considerably higher than the ideal orthogonal waveform level.

The spectrum of a Costas coded pulse is given in Figure D6-4, its form is similar to that of a LFM pulse.

[8] The thumbtack ambiguity response has a large central spike, at zero delay and frequency shift, and a low plateau of sidelobes at all other positions of time and frequency shift, thus resembling a thumbtack.

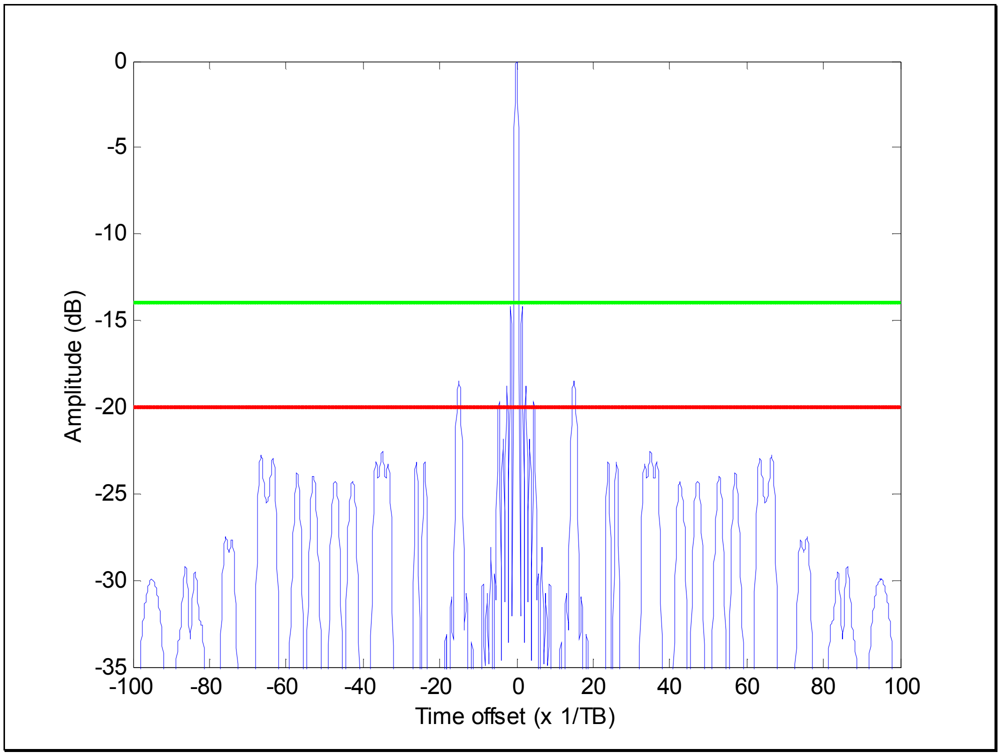


Figure D6-2 Auto-correlation function of an M=10 Costas waveform, with typical Minimum sidelobe ($1/M^2$) (red) and maximum sidelobe ($4/M^2$)(green) levels indicated.

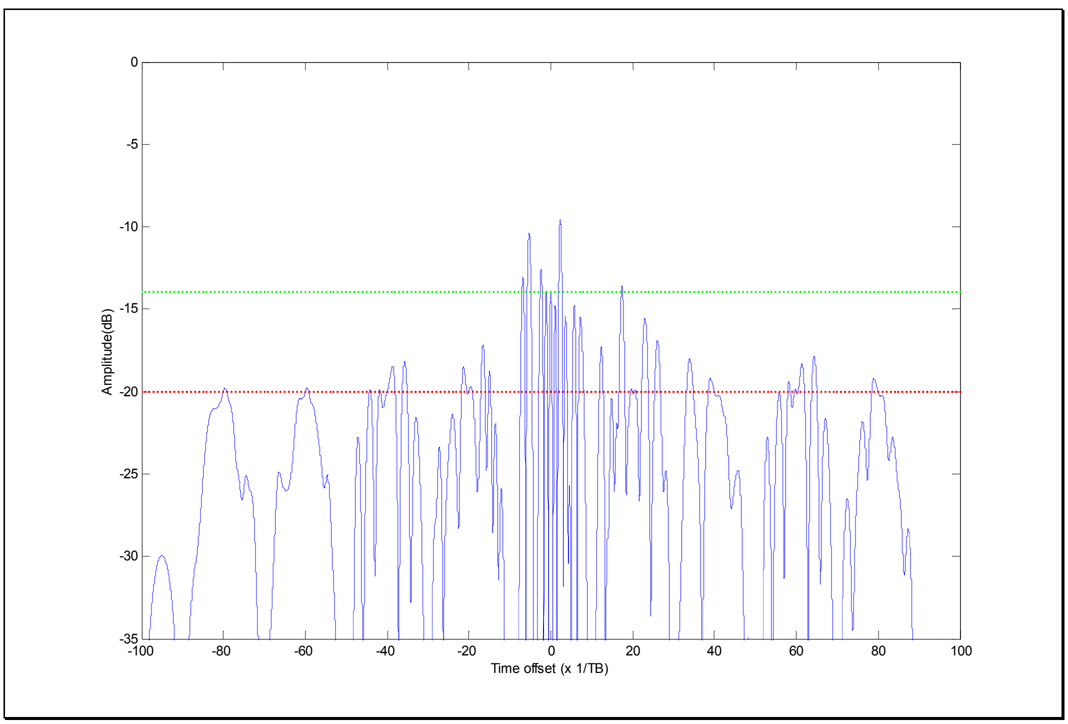


Figure D6-3 Cross correlation response of two different M=10 Costas waveforms, with an ideal orthogonal waveform sidelobe level ($1/M^2$) and maximum auto correlation sidelobe level ($4/M^2$) indicated.

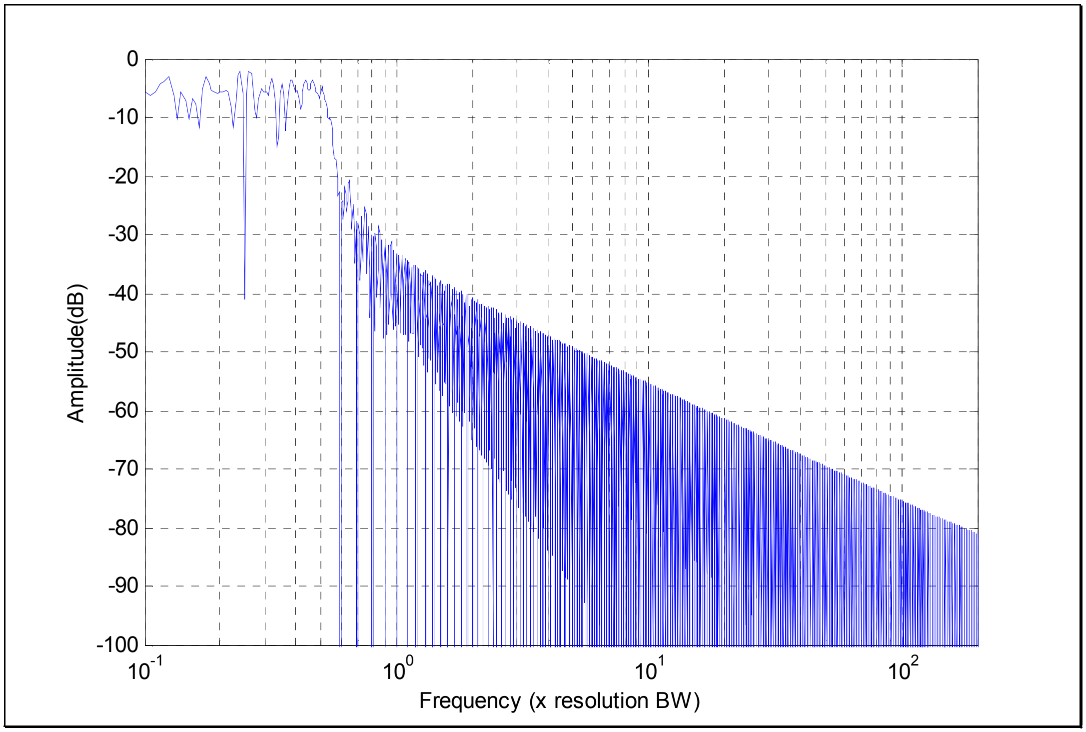


Figure D6-4 Showing the frequency response of a M=10 waveform, where the range resolution bandwidth is defined as approximately M times the frequency step (1/t_s)

D.6.4 Linear frequency modulation (LFM)

A pair of LFM waveform can be generated to be orthogonal to each other by using opposite frequency gradients, i.e. with reference to Equation D3-1 one with positive B_c and the other with negative B_c.

This gives cross correlation response that meets the theoretical minimum for an orthogonal signal. By example, Figure D6-5 shows the correlation performance of FM chirps with time bandwidth products of 100 and maximum cross correlation values of -20dB (i.e. 10log(1/100)).

Unfortunately, the set of orthogonal waveforms cannot be extended to greater than 2 for LFM waveform, thus limiting their utility in dense environments. However, the degree of coupling of slopes of different gradients may give larger sets of candidates if non-linear FM waveforms were considered. This is an area for further investigation.

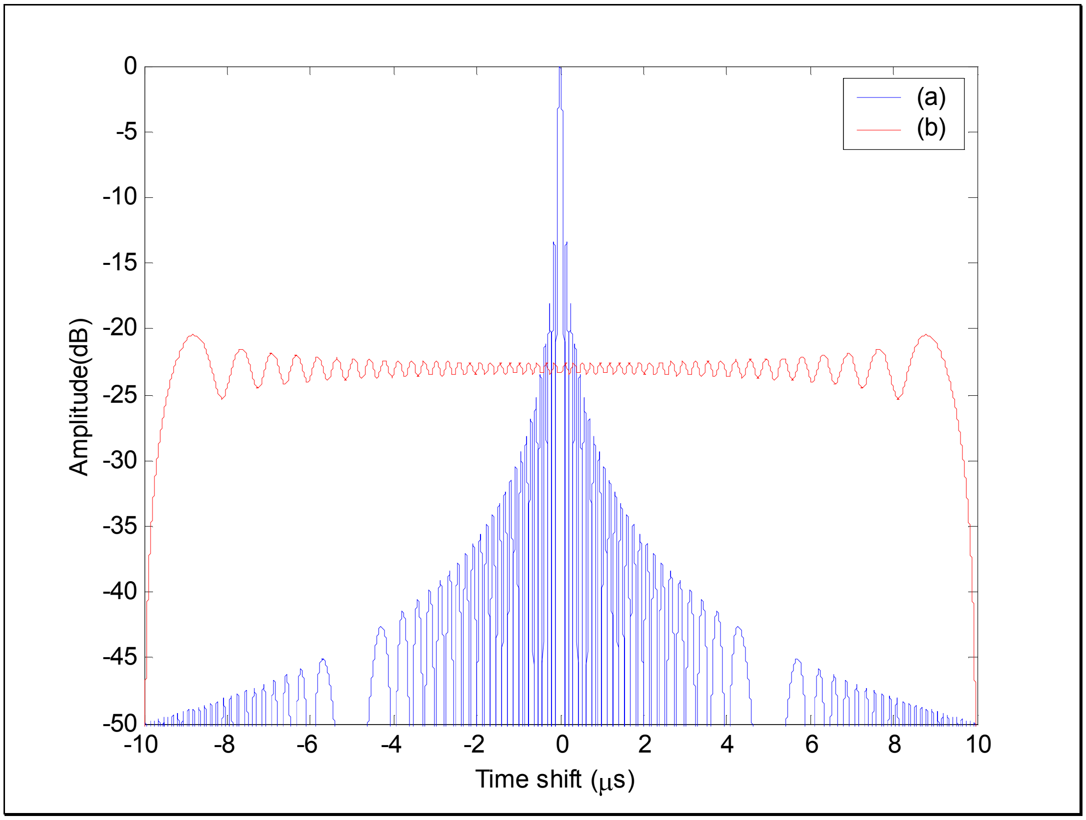


Figure D6-5 Correlation functions of FM chirps, with 10μs pulse width and 10MHz swept bandwidth (a) matched sweep gradients (b) opposite sweep gradients.

D.6.5 Polyphase

Many methods of producing orthogonal phase codes exist, when the phase is not constrained to strict subsets of values, termed here 'polyphase'.

FM based polyphase waveforms have 2 to 3 times worse cross-correlation levels than the analogue FM equivalents [17]. Also, like analogue FM, they suffer the problem of being members of sets of only two.

Nunn and Welch presented a noteworthy paper [18] in 2000, which generated phase codes by optimal search. Results are presented showing eight codes that are N=512 length sequences, all of these codes have approximately -40dB auto-correlation sidelobes $10Log(5/N)^2$, and cross correlation peaks of between -17dB and -20dB (10Log (5/N)). The paper claims that larger sets (thousands) are possible that exhibit good auto-correlation and cross-correlation performances. In addition, because of the large numbers of candidate codes new members of the set can be produced from exemplar members. These new members generally have good properties with respect to the entire set.

To investigate these claims, an optimal search program has been written to find pairs of polyphase codes with the dual aim of a low peak sidelobe in auto-correlation and cross-correlation functions. Results giving maximum sidelobe levels are given in Table D6-1. The optimum results are given in brackets for each value calculated. The examples of cross and auto-correlation function of one pair are given in Figure D6-6. The results look encouraging for consideration as a low interference waveform. The results show a trade-off where 1dB greater auto-correlation sidelobe levels can result in up to 7dB improvement cross-correlation performance. The results obtained do not give as low auto-correlation sidelobes as in the paper, especially for larger TB products. This is probably due to the optimisation program only allowing 256 discrete phases.

Time bandwidth product (Code length)	Code pair No	Peak range sidelobe #1 (dB)	Peak range sidelobe #2 (dB)	Peak cross-correlation level (dB)
64	1	27.5(36)	27.6(36)	11.6(18)
	2	27.3(36)	26.5(36)	17.6(18)
	3	26.5(36)	25.9(36)	17.1(18)
128	1	29.6 (42)	28.5(42)	14(21)
	2	28.5(42)	27.5(42)	20.8(21)
256	1	31(48)	30.6(48)	18(24)
	2	30.1(48)	30(48)	23.1(24)

Table D6-1 Results of polyphase code pair performance, generated by optimal search.

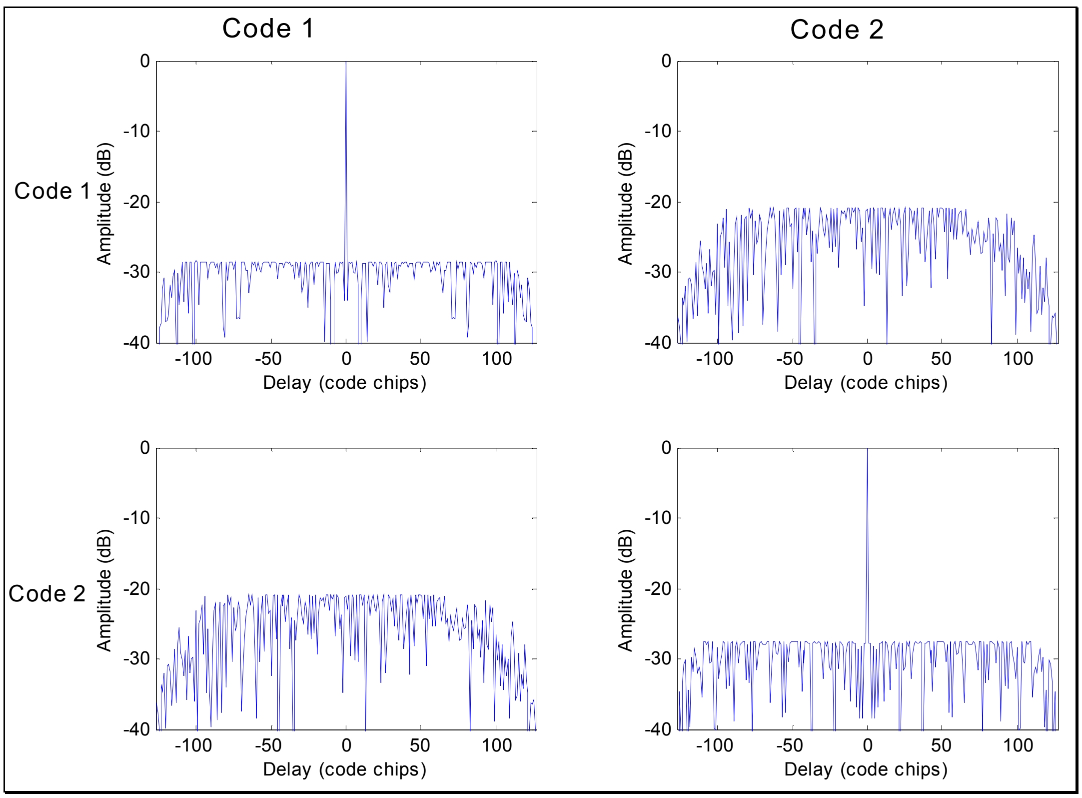


Figure D6-6 Polyphase code pair auto-correlation and cross-correlation performance for two optimised 128-length polyphase codes

D.6.6 Summary of orthogonal waveforms

This section has investigated waveforms with low cross-correlation properties to limit interference between like systems. The practicality of orthogonal has been discussed with PC waveforms, normally giving rejection of the order between *1/TB* and *1/√TB*. To achieve performance levels required for interference rejection, outlined in §D4, would require TB products of order 1×10^{10}, which is impractical, due to time and bandwidth restrictions. Therefore, orthogonal waveforms can only be considered a single measure to be used with other techniques, such as those are described in the following sections of this appendix.

LFM and digitally derived equivalents were found not to offer sufficient orthogonal variants for multiple radar systems operating in a confined area. Polyphase coded and Costas coded waveforms offer the best candidates for operation of a large number of radar systems, in a shared spectrum environment.

Polyphase coded waveforms offer the best option with the ability to optimally manage the balance between interference performance and PC performance. However, TB products of order 100 to 500 are required to achieve range sidelobe levels, of order 40dB. This is considered to be adequate for the majority of the radar applications being considered. Such TB products will give approximately 20dB interference rejection performance. Higher TB products could be considered. However, this is subject to the investigation of the systematic problems that high TB products produce, i.e.

- Large range swaths of range sidelobe clutter if the pulse period is high, Increasing self interference.
- High resolution and larger resolution bandwidths, if the bandwidth is high, Increasing spectral usage and processing requirements.

D.7 Spectrally compact waveforms

D.7.1 Introduction

This section looks at generating waveform with greater spectral compactness by investigation of purer RF sources, waveforms shaping or by waveform filtering.

The benefit of producing spectrally compact waveform is that waveforms from multiple systems can operate in spectrally closer proximity without interference.

The received interference level between two waveforms is typically defined as the integral of the product of the overlapping spectra, and forms the motivation for investigating waveforms with compact spectra. Therefore, if two waveforms are spaced such that their centre frequencies differ by greater than their –40dB bandwidth then the correlation between them will be below -40dB compared to the matched condition (in centre frequency and waveform).

Clearly, the shape of the whole waveform spectrum influences the actual cross interference level. In addition, the use of different waveforms (especially orthogonal waveforms) requires further analysis. Therefore, in addition to their spectral response, the maximum correlation level of classes of waveforms operating on different centre frequencies are presented in this section. This serves as a reference for calculating the number of waveform that can be packed into a given radar band of interest.

D.7.2 Magnetron

The low-cost of magnetrons make them the source of choice for the majority of simple radars. Improvements in magnetron design e.g. the coaxial magnetron, has improved their reliability and frequency stability.

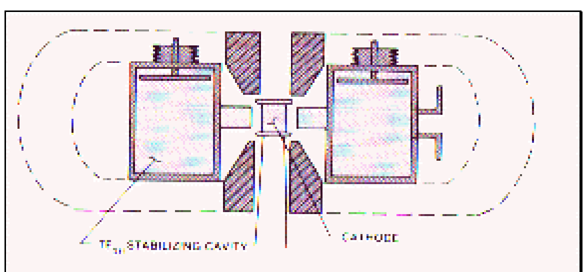


FIGURE 5. COAXIAL MAGNETRON

Figure D7-1 Coaxial magnetron

The distinguishing feature of the coaxial magnetron is the presence of a high Q stabilising cavity between the anode and the output waveguide. The coaxial stabilising cavity significantly increases the purity of the magnetron RF emission. However, the time taken for a magnetron to oscillate at its design frequency is higher in high Q devices, therefore the spectra from short pulses may be less pure when compared to a standard magnetron spectra.

Careful design of the magnetron modulator is considered by magnetron manufacturers [19] to be the best method to reduce unwanted non-harmonic spectra. In addition, control of the current and voltage can negate effects of ageing in modern magnetrons.

Injection locking of magnetrons is possible to enable coherent operation. This could result in cleaner spectra by reducing initial oscillation modes and enable techniques that reduce the peak power requirement (see §D5.9). However, the sum cost of injection locking components is likely to be several times the costs of current radar magnetrons.

An investigation is required into whether modern magnetrons are amenable to tighter spectral specification that would allow more radar systems to occupy the same region of the spectra with a minimal interference.

D.7.3 Constrained FM

Naturally, the spectrum of FM (linear or non-linear) pulses are relatively well constrained and a technique for reducing the spectral spreading of LFM waveforms in response to tight emission restrictions in the HF radar band is discussed in reference [20].

A rectangular spectrum with the correct quadratic phase profile for a LFM waveform is formed in the frequency domain. This is then converted to the time domain for transmission. It successfully narrows the spectrum, resulting in near-perfect rectangular spectra (Figure D7-2b), at the expense of transferring the spectral shape into the time-domain i.e. pulse shape (Figure D7-3).

Radar power amplifiers generally operate in saturation and these amplitude variations will be distorted by the non-linear nature of the power amplifiers. The authors employ a feedback loop to adjust the waveform weights to minimise the spectral spreading.

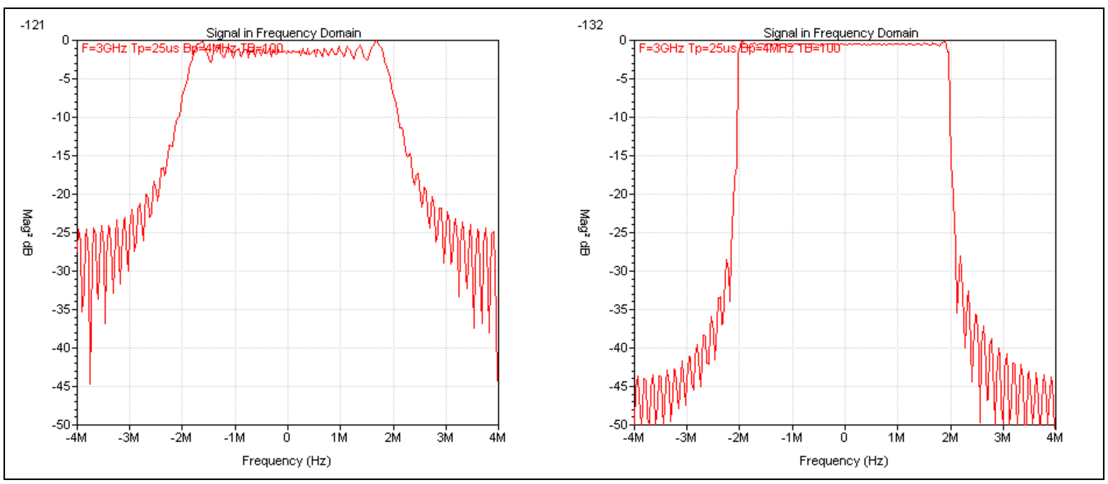


Figure D7-2 LFM spectrum (TB=100) (a) normal (b) shaped

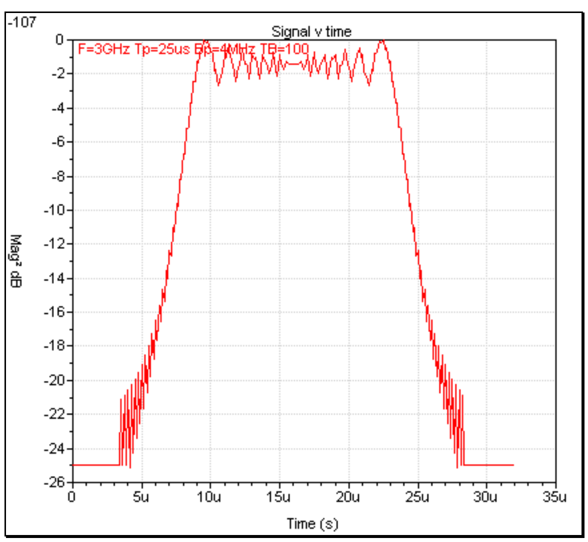


Figure D7-3 Shaped LFM pulse amplitude

D.7.4 Taylor quadriphase

An important exception to the typical spectrum obtained from phase coded waveform (see §D3.3) is the novel form of quadriphase coding introduced by J.W. Taylor [21] to meet the requirements of the US Office of Telecommunications policy limiting the spectral extent of radar emissions.

The Taylor quadriphase (TQ) code comprises of a pair of binary phase coded signals in quadrature. Its make up is unique for coded waveforms in that each sub-pulse is not rectangular but a half-cosine of twice the width of the nominal sub-pulse width. This limits this technique's applicability to Quadriphase codes where the bit-to-bit phase shifts are $\pm\pi/2$. Unlike their rectangular sub-pulse counterpart, where the phase jumps sharply between sub-pulses, the TQ has a continuous phase profile, which is the basis of its spectral performance.

It should be noted that, despite the cosine shaping of the sub-pulses, the TQ waveform as a whole is not amplitude modulated.

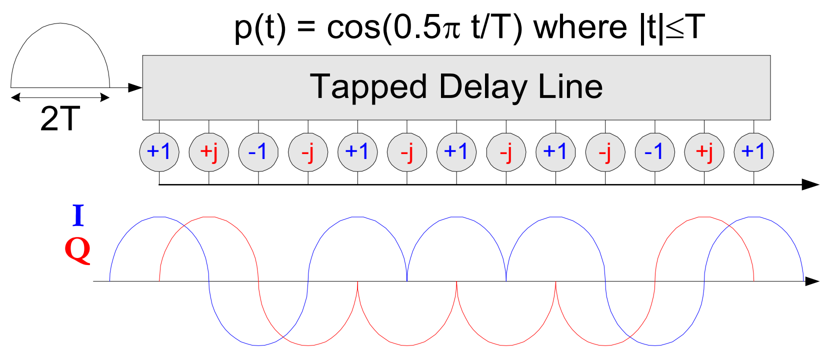


Figure D7-4 Taylor Quadriphase (Barker 13) code generation.

Figure D7-4 shows how a 13 sub-pulse TQ code could be generated by passing a cosine shaped pulse through a tapped delay line.

The TQ achieves a significant reduction in spectral extent: a comparison of the spectrum of a 13 element, 1μs sub-pulse, rectangular and TQ code is shown in Figure D7-5. As can be seen from Table D7-1, the TQ reduces the -40dB spectral width by a factor of almost thirteen.

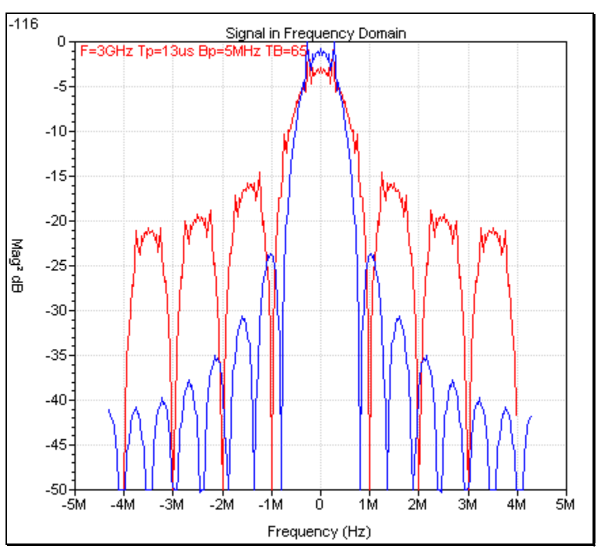


Figure D7-5 Quadriphase spectrum (red) Taylor quadriphase spectrum (blue)

	Rectangular	Taylor
-40dB spectral width	64/*N*	5/*N*
Fall off	6dB/octave	12dB/octave

Table D7-1 Spectral width of Rectangular and Taylor Quadriphase.

Taylor's paper describes its use with short minimum-sidelobe codes with a maximum of 28 elements. Most phase codes are much longer than this and its applicability to longer codes has been investigated. Figure D7-6 shows the spectrum of a 128-element maximum length sequence (MLS) code in rectangular and Taylor forms. The improvement in spectral performance is similar to that given in Table D7-1.

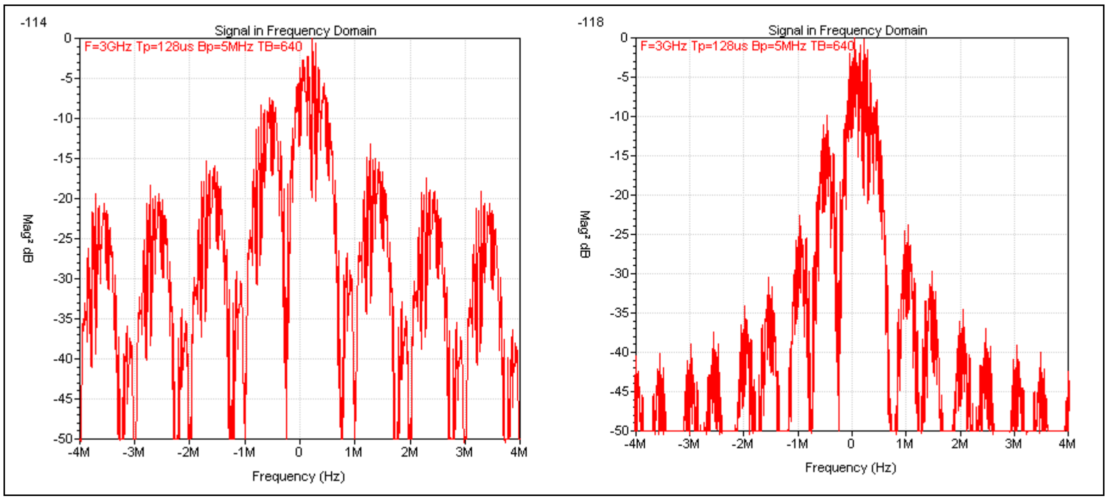


Figure D7-6 128-element MLS quadriphase spectrum (a) rectangular (b) Taylor

The range domain point spread functions (PSF) (Figure D7-7) are different, but the sidelobe performance of TQ is never significantly worse than its rectangular counterpart, and in fact frequently better.

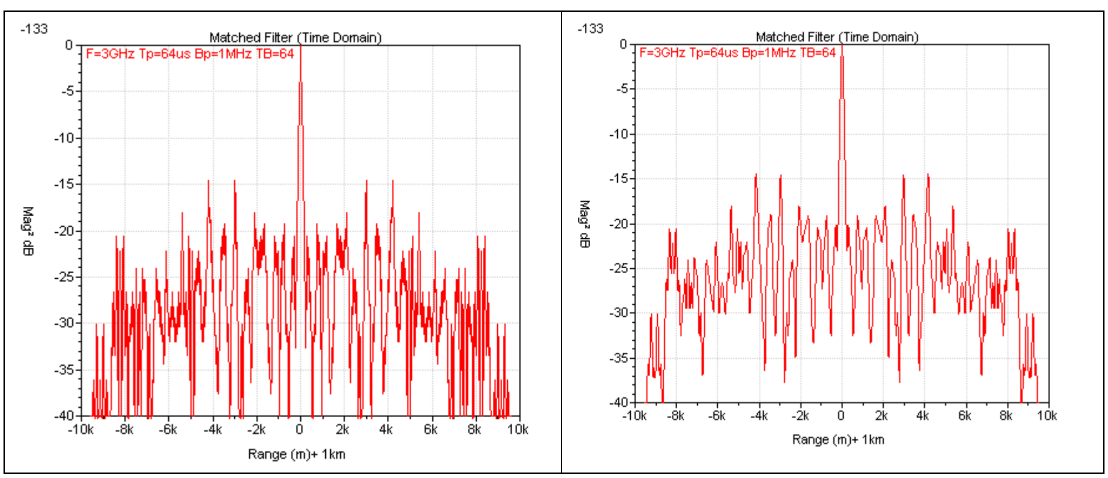


Figure D7-7 128-element MLS quadriphase PSF (a) rectangular (b) Taylor

From a spectral occupation viewpoint, the TQ looks an attractive alternative to bi-phase coding and rectangular quadriphase. By having less power in the sidelobes, its performance is also more resilient to anomalous effects subsequent band limiting filters.

TQ principles are not applicable to the polyphase codes, because the phase states are defined by a formula and are not arbitrary.

D.7.5 Filtering waveform

Most of the energy of coded waveforms lies within the $1/\tau$ first nulls: the extraneous parts of the spectrum can be discarded without significant detriment to the waveform's target point response.

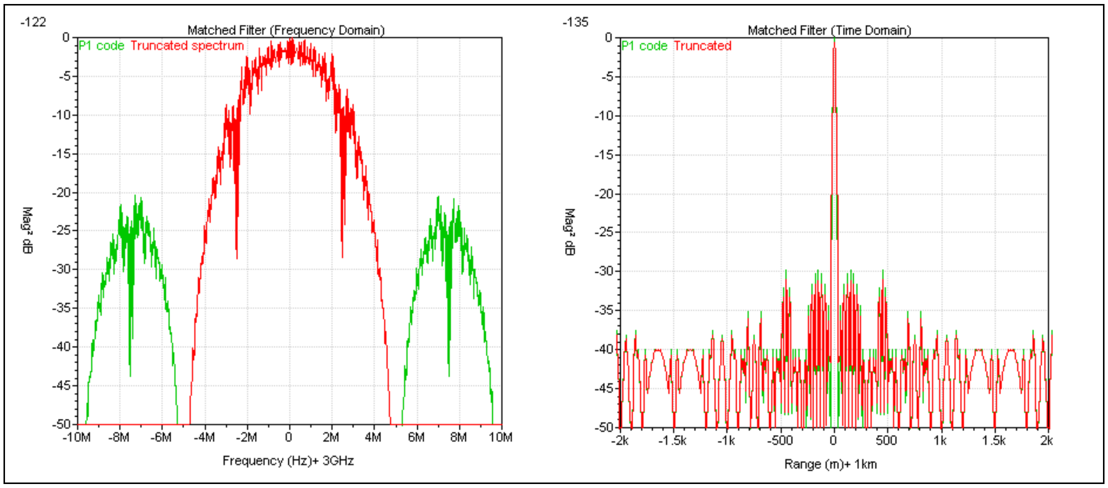


Figure D7-8 100-element P1 code (a) spectrum (b) PSF full truncated

An example of this is shown in Figure D7-8. This shows the spectrum and range-domain response of a 100 element P1 polyphase code without (green) and with (red) filtering. The consequence is a slight broadening of the main lobe, but the sidelobes have been reduced somewhat in this example.

D.7.6 Closely spaced waveforms

Introduction

§D6 examined the response of waveforms with low cross correlation, when they are deliberately mismatched (or are orthogonal). Whereas in this section above, waveforms that have a low spectral leakage is discussed. The aim of the low spectral leakage is to minimise the interference between waveforms operating with close, but not identical centre frequencies. Orthogonal waveforms, on the other hand, assume that they are operating on identical centre frequencies i.e. are locked to each other. One conclusion of §D6 is that, additional measures must used with orthogonal waveforms to achieve the level of interference rejection required. One such technique could be by operating on different frequency offsets.

This section considers the impact of spectral leakage, by calculating peak correlation levels between waveforms operating at different centre frequencies i.e. waveforms that have a frequency offset to one another.

Waveforms considered, in this section, include frequency shifted identical waveform as well as orthogonal waveforms. Specifically, the responses of Costas, LFM and MLS pulsed waveforms are considered.

The results from this section are presented in different formats. This is due to the nature in which the ambiguity functions are produced, therefore the summary helps to form comparisons of the following sub-sections.

Costas coded waveform

The maximum correlation return when Costas coded waveforms are frequency offset from each other is given in Figure D7-9. For the identical Costas coded waveforms (Figure D7-9(b)) the figure shows:

- that the central correlation peak is very sensitive to frequency offset.
- the peak sidelobe plateau's at a level of approximately *1/M* for frequency offsets when the range resolution bandwidth overlaps (less than $\pm 10/T$) (also see Figure D6-4).
- finally, the peak sidelobes tail off in a similar manner to the frequency skirts of the Costas waveform spectrum for comparison again see Figure D6-4.

When using different (orthogonal) Costas coded waveforms (Figure D7-9(a)), the figure shows:

- There is no correlation peak, at zero frequency offset (by design).
- The peak sidelobe has a plateau, for frequency offsets such that the range resolution bandwidth overlaps and the level is slightly higher than that encountered in the identical waveform case.
- Finally, the peak sidelobes tail off, similarly, to the sidelobes in the identical case.

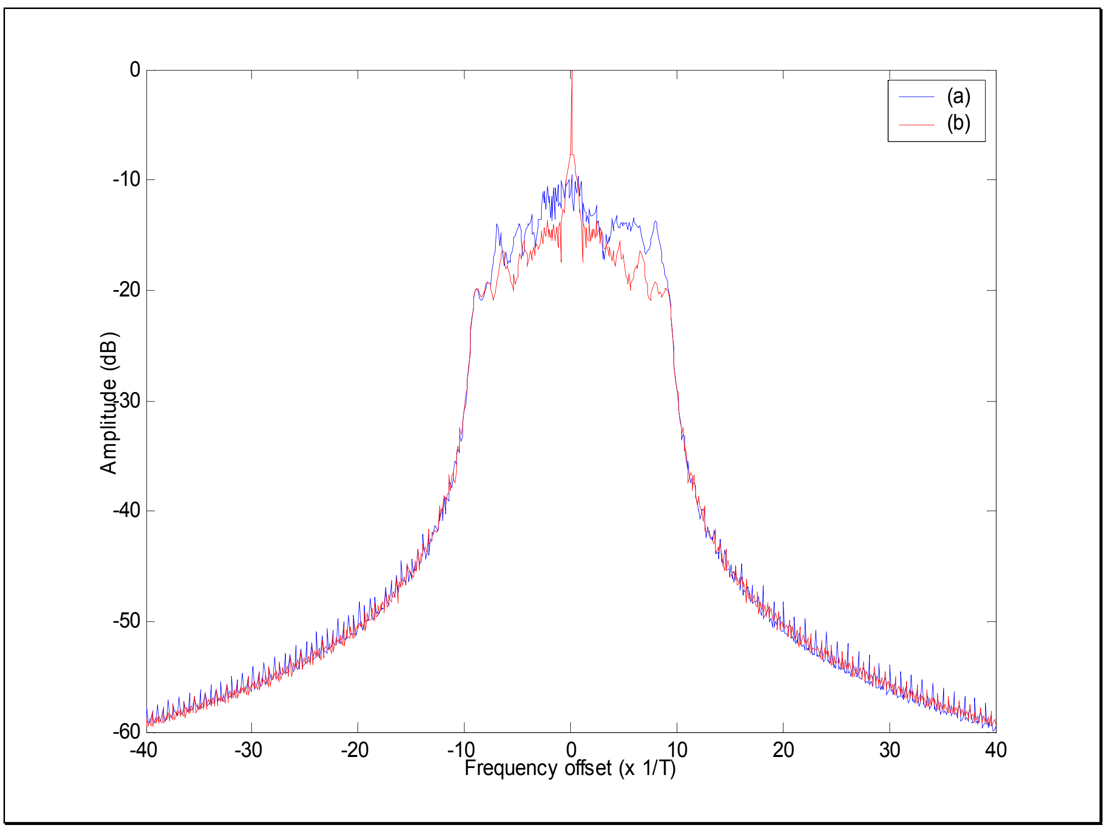


Figure D7-9 Showing the maximum sidelobe return when M=10 Costas waveforms are frequency offset, (where 1/T is the frequency step used to construct the Costas code) for (a) different Costas coded waveform (b) Identical Costas coded waveform.

Pulsed LFM

Doppler tolerance is a key benefit of LFM waveforms, but a consequence is that neighbouring returns have high correlation over significant frequency offsets. This can be seen in the Figure D7-10, where the peak amplitude of the correlation peak is high for frequency offsets up to the chirp bandwidth *Bc*. After this heavily correlated region the peak sidelobe response falls off in a similar manner to the shape of the spectral sidelobes of the FM waveform (referring to Figure D5-3(b)).

Figure D7-11 shows the peak sidelobe response between two orthogonal pulsed LFM waveforms. Here, the correlation peak is approximately 10log(TB), where the frequency offset is less than the chirp bandwidth. This is the same value as found for the orthogonal FM pulses with no frequency offset see §D6.4. However, in the region where the frequency offset is greater than the chirp bandwidth (>10MHz), it has similar correlation peaks to that observed for identically chirped waveforms. This is a similar conclusion to that was found for Costas coded waveform.

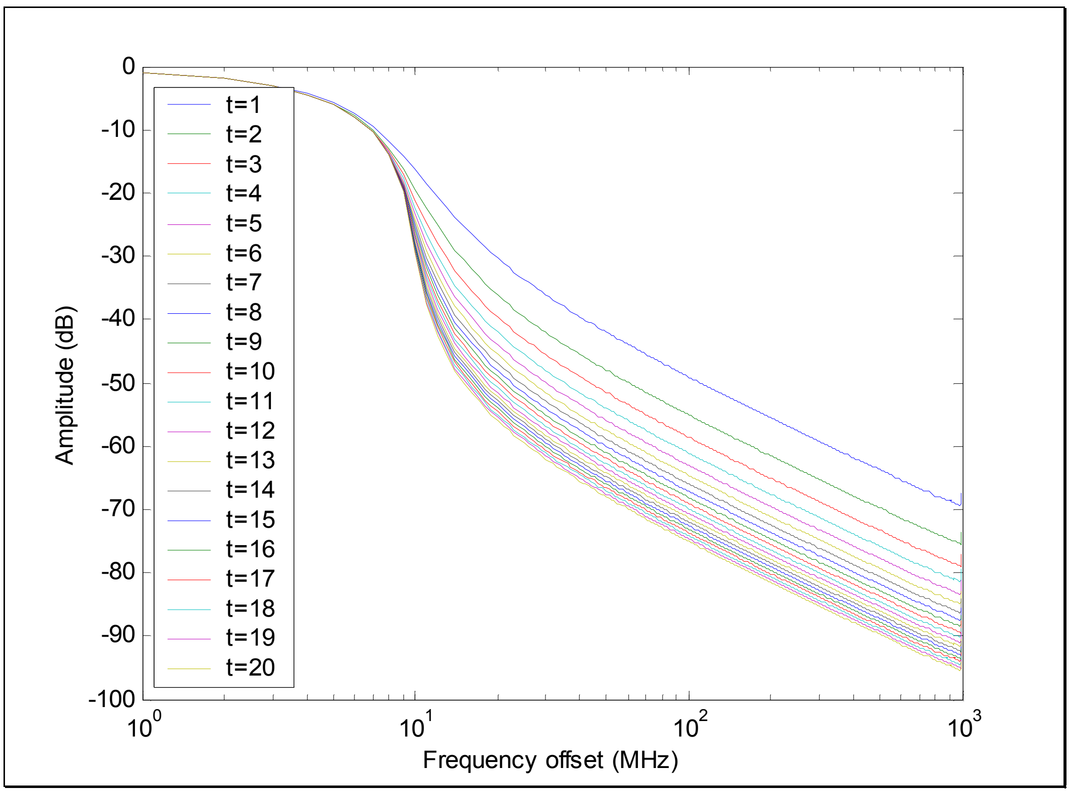


Figure D7-10 Peak sidelobe response as an 10MHz FM chirp is frequency shifted for different pulse widths, t (t given in microseconds)

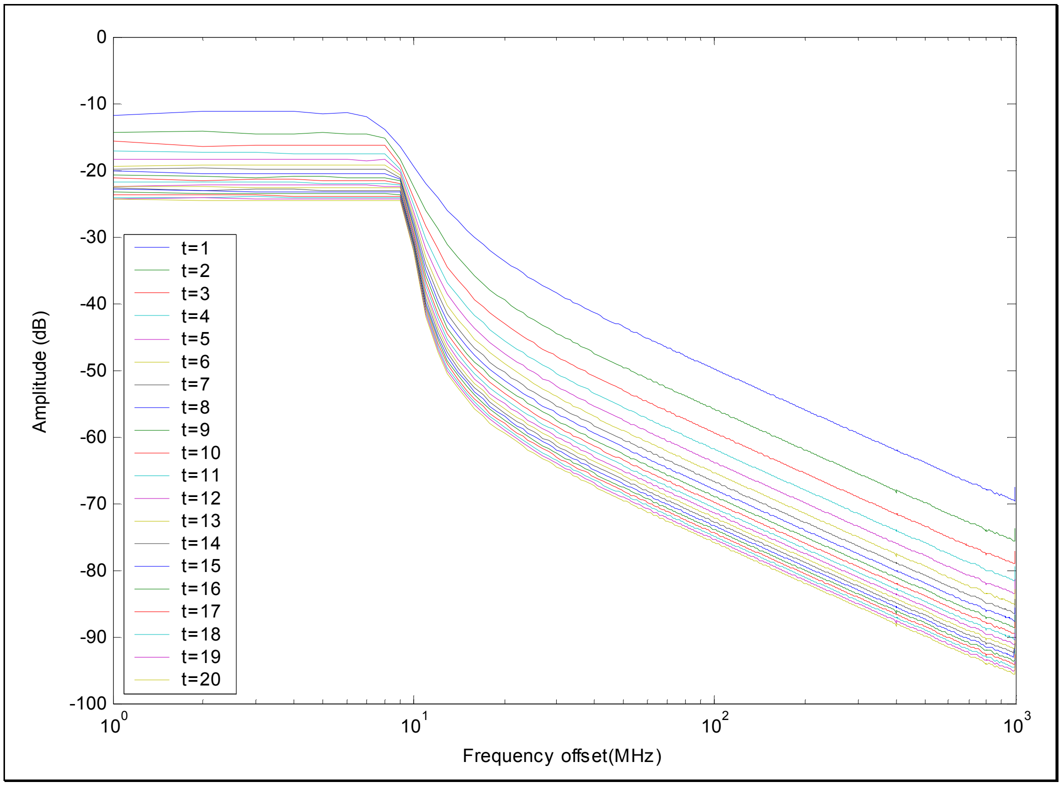


Figure D7-11 Peak sidelobe cross correlation response as two 10MHz FM chirps, which have opposite slopes, are frequency shifted with respect to each other for different pulse widths, t (t given in microseconds)

<u>Phase coded waveform</u>

The correlation peaks, at different frequency offsets, for identical and orthogonal phase coded waveforms are shown in Figure D7-12. The identical case (Figure D7-12(b)) heavily correlates at regular frequencies due to the spectral repeating property of coded waveform explained in §D3.3.The amplitude of the peaks approximately follow the sinc function spectral response. However, the orthogonal correlation functions remain orthogonal over all frequency offsets, which for this 127-length code, is approximately 14dB below the matched response.

Furthermore, §D7.5 showed that filtering coded waveform is effective. Therefore, the use of transmitter filters will reduce the outer spectral response to levels that rival analogue waveform spectrum whilst the orthogonality will reduce correlation levels further.

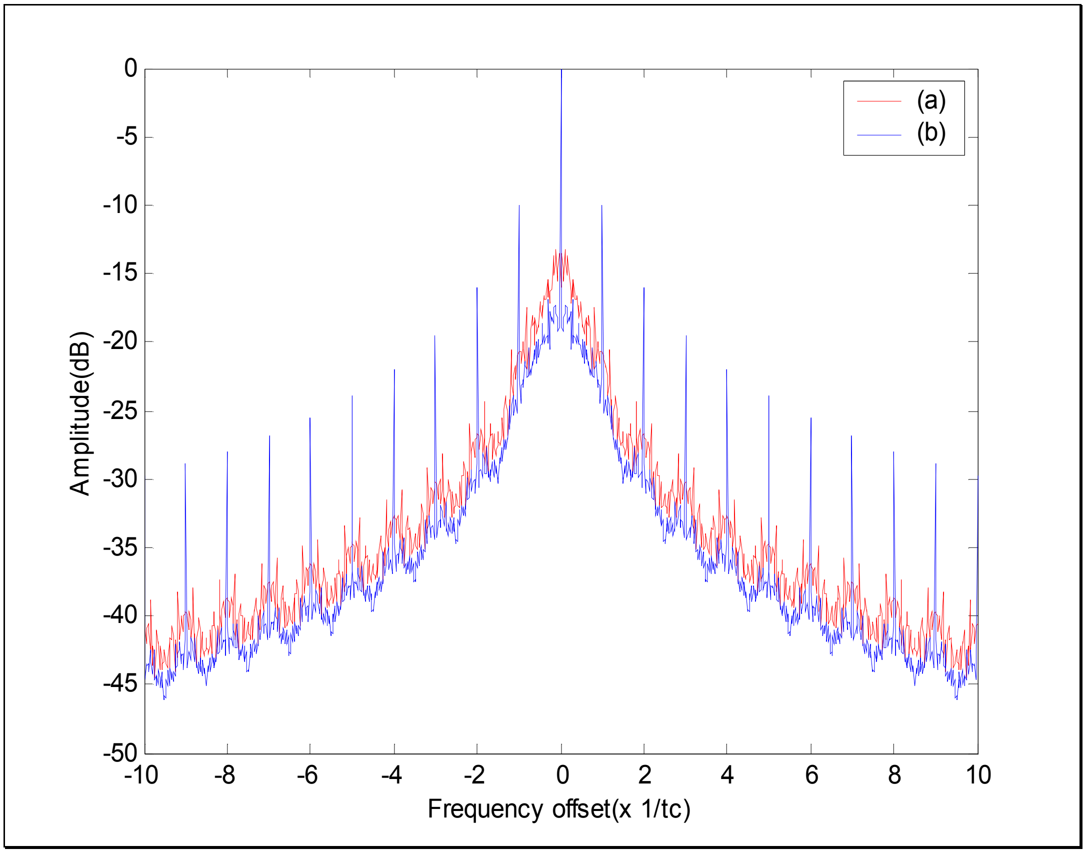


Figure D7-12 showing the correlation peak of MLS 127 length binary phase code tc= chip time = the resolution bandwidth (a) identical code correlation (b) orthogonal code correlation.

D.7.7 Summary of spectral compact waveforms

The correlation performances of frequency offset waveforms have been presented, in which identical and orthogonal waveforms have been investigated.

The correlation of Costas codes and pulsed LFM sidelobes fall off the sharpest with increasing frequency offset values, which is in line with their spectral responses. However, at frequency offsets of greater than the range resolution bandwidth the correlation performance of identical and orthogonal waveforms is virtually identical. This is the most likely region that such a technique will be used for reduction of interference between radar systems as §D4.5 stated that interference rejection levels of in excess of 60dB is likely to be required of the waveform. Therefore, use of these orthogonal waveforms for radar systems is of limited use, considering the moderate rejection achieved when the operating frequencies are close and extra complexity involved in controlling orthogonal waveform and they give no advantage over identical waveform in the most likely area of operation.

Frequency separation of less than the bandwidth can result in reduced interference for identical and orthogonal Costas codes, orthogonal pulsed LFM waveform, mismatched phase coded waveform and matched phase coded waveform. The matched Costas waveform actually gives better interference rejection than the mismatched waveform for all, but when the frequencies are matched.

Phase coded waveform retain their orthogonal nature in the sidelobes. Therefore, the use of orthogonal codes when combined with frequency offsets between radar systems give interference rejection performance, that rivals more spectrally compact analogue waveforms such as LFM. Filtering improves the performance figures, resulting in superior performance to enable reduction in frequency separations of waveforms.

For comparison, the frequency offsets required for a number of peak cross-correlation performances have been estimated for different waveforms in Table D7-2. However, it is likely that results reported with greater than 100 times the resolution bandwidth are likely to be irrelevant (shown in red). This is because the responses of system components are likely to influence the spectra at these high separation frequencies.

Waveform	**Peak cross correlation level**				
	-20dB	**-40dB**	**-60dB**	**-80dB**	**-100dB**
Pulsed, rise time = $\tau/20$	*6.5xB*	20x*B*	80x*B*	210x*B*	800x*B*
Pulsed, matched rise time	*1.25xB*	4x*B*	11x*B*	40x*B*	120x*B*
Identical Polyphase codes	*3xB*	50x*B*	1000x*B*	>1000x*B*	>1000x*B*
Orthogonal Polyphase codes	0	3*B*	50x*B*	1000X*B*	>1000x*B*
Filtered Identical codes	<*B*	2.4x*B*	4x*B*	6x*B*	16x*B*
Filtered Orthogonal codes	<*B*	1.2x*B*	1.7x*B*	3x*B*	6x*B*
LFM	*B*	1.3x*B*	5x*B*	30x*B*	300x*B*
Orthogonal LFM	0	1.1x*B*	3.5x*B*	30X*B*	300X*B*
Identical Costas	*B*/10	10x*B*	4.5x*B*	60x*B*	>1000x*B*
Orthogonal Costas	<*B*	1.2x*B*	4.3x*B*	55x*B*	>1000x*B*

Table D7-2 Shows the frequency offset required to achieve different interference rejection targets, for different waveforms. All waveforms are assumed to have TB products of approximately 200.

D.8 Techniques for improving band sharing

D.8.1 Introduction

This section examines waveform design and processing measures that can be taken to address the problem of in-band interference. This situation is inevitable if the frequency band is shared with other similar radar systems or with dissimilar systems operating in the same area.

D.8.2 Interference signal rejection

An active approach is to remove the interference signal before it enters the radar-matched filter. This can be achieved in a variety of ways e.g. the adaptive filter model shown in Figure D8-1 [22]. The interference signal is subtracted from the radar's signal when the adaptive filter has minimised its output power. The signal is then passed to the matched filter as normal.

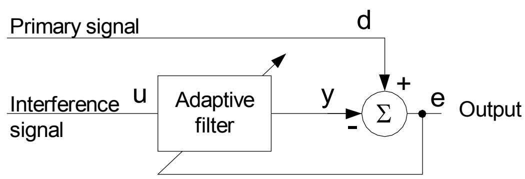


Figure D8-1 Adaptive filter for interference removal.

The use of an active filtering method with the same waveforms as presented earlier has been examined. Figure D8-2 shows the matched filter response for a MLS coded waveform (a) without and (b) with interference. The signal has been swamped by an interference signal with five times its amplitude. A copy of the interference signal allows its amplitude and phase to be measured in the radar signal. The interference can then be removed prior to the radar's matched filter to give a near-perfect response (Figure D8-2c).

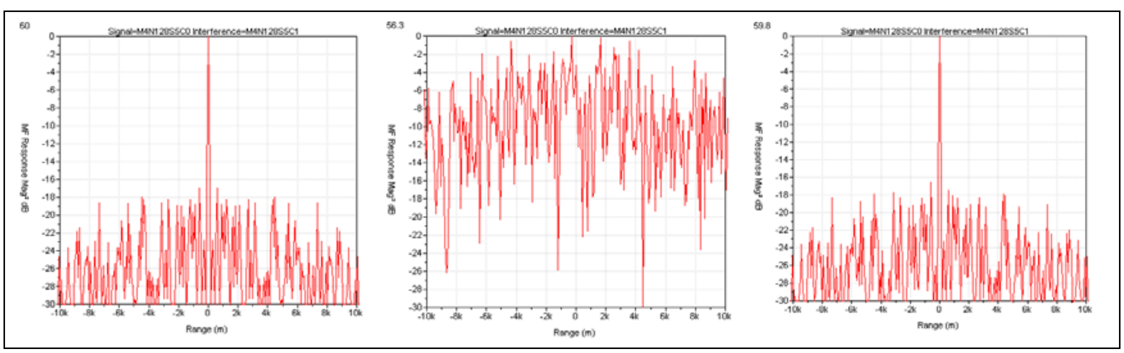


Figure D8-2 Interference: (a) without (b) with (c) removed

This particular type of system requires a copy of the interference signal that is an accurate representation of the signal as received by the radar. This should not present a problem however as this information can be gathered by setting the radar to a passive i.e. listening mode. The technique could be used to remove direct transmission between radars but the removal of multiple returns or those affected by Doppler requires more sophisticated processing. In practice, the interference is likely to be time-varying, so the learned interference signal would have to be probed periodically. The period of validity of the interference signal would have to be determined experimentally in further tests.

Reduction of interference is only limited by the dynamic range of the receiver, however,, because high interference to noise ratio (INR) is required to achieve high suppression ratios. Thus, performance is limited to high INR signals.

The residual INR, after the filter, is likely to be relatively constant (of order 20dB) due to either undetected interferers or errors in the rejection filter.

D.8.3 Impulse interference rejection

Radars have evolved methods to reject transmissions from other like radar systems when pulses of received energy may cause an invalid detection .

A typical approach is to vary the PRI between pulses (termed staggered repetition period) so that the likelihood of consecutive pulses occurring after the same delay after transmission is reduced. This concept can be extended so that a detector would only declare a detection given m detections in every n pulses. The probability that erroneous detections would occur from matched radar systems is given in Figure D8-3, where the number of free range bins express the degree that the PRI is varied. A range 'bin' is the common term for the processed return from all range returns within one range resolution.

For example: if the PRI values were allowed to vary on average by 2% more than defined by an Instrumented range of 50km and the range resolution were 0.1km the total Instrumented range bins would be 50/0.1=500 and the free range bins would be 2% of 500 = 10.

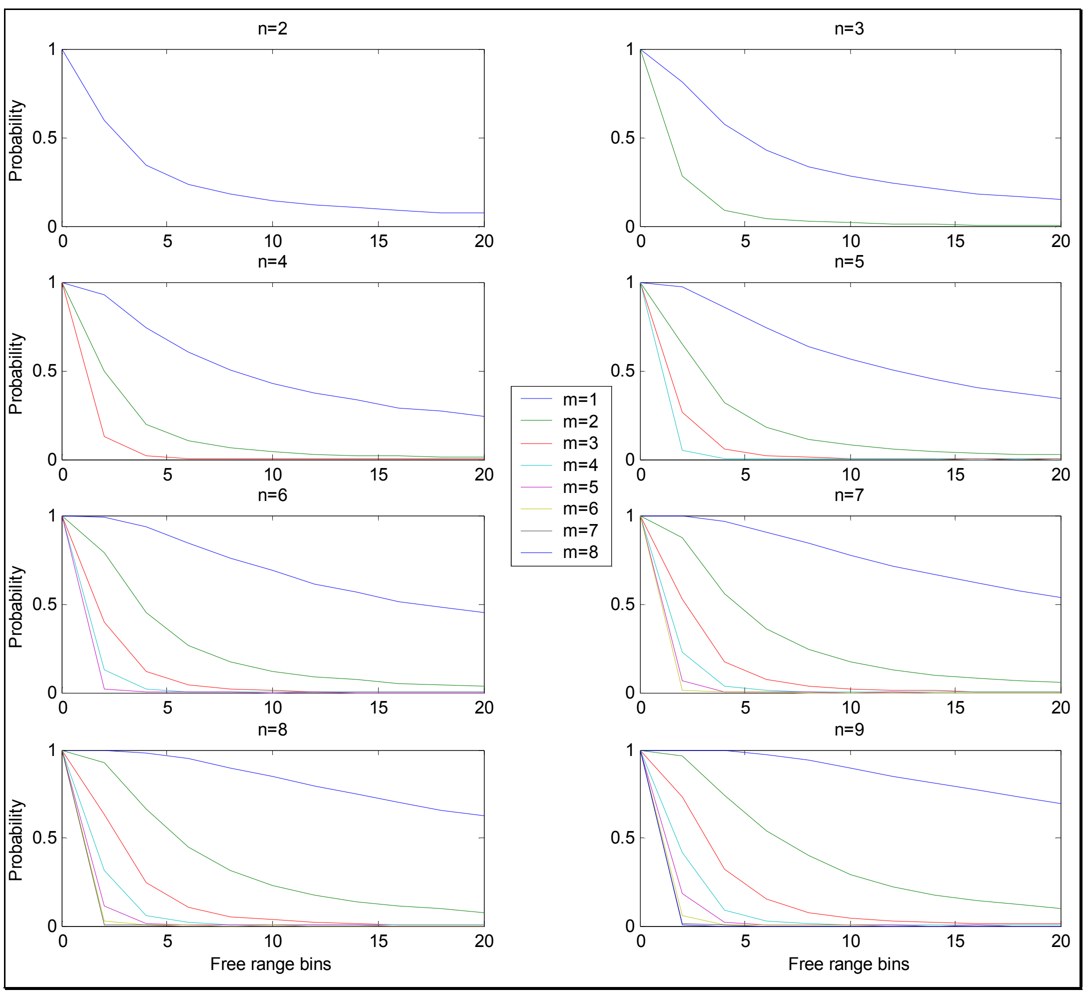


Figure D8-3 Interference probability between two radar systems that employ m from n detection logic and staggered pulse periods.

Analysis of the effect of multiple systems and systems with unmatched average PRFs and pulse widths requires more in depth simulations, which is left here for further work.

Pulse integration techniques also exist to reject interfering sources with unmatched PRFs, and one of these is detailed in reference [23].

However, these techniques are designed to work with radars that incoherently integrate pulses, therefore they are not relevant for radars that implement Doppler and MTI techniques. Such processing techniques achieve gains in signal to noise and clutter ratios through constant PRI, which gives orthogonal Doppler filtering and effective MTI filtering. In addition, schemes that limit amplitudes to remove excessive interference (such as in reference [23]) destroy the ability for coherent processing techniques to extract signals from noise.

Although linked to waveform PRF, incoherent and coherent processing techniques are outside the remit of this study so further investigation is needed for further work.

D.8.4 Narrow band IF

Unlike phase coded waveforms, LFM has a narrow instantaneous frequency with a well-defined slope i.e. B_c/τ. This is most clearly seen in a time-frequency analysis (TFA) [24] plot, which shows how the instantaneous frequency changes over the duration of the pulse. The difference between in the LFM and phase coded waveforms, in Figure D8-4 is very evident.

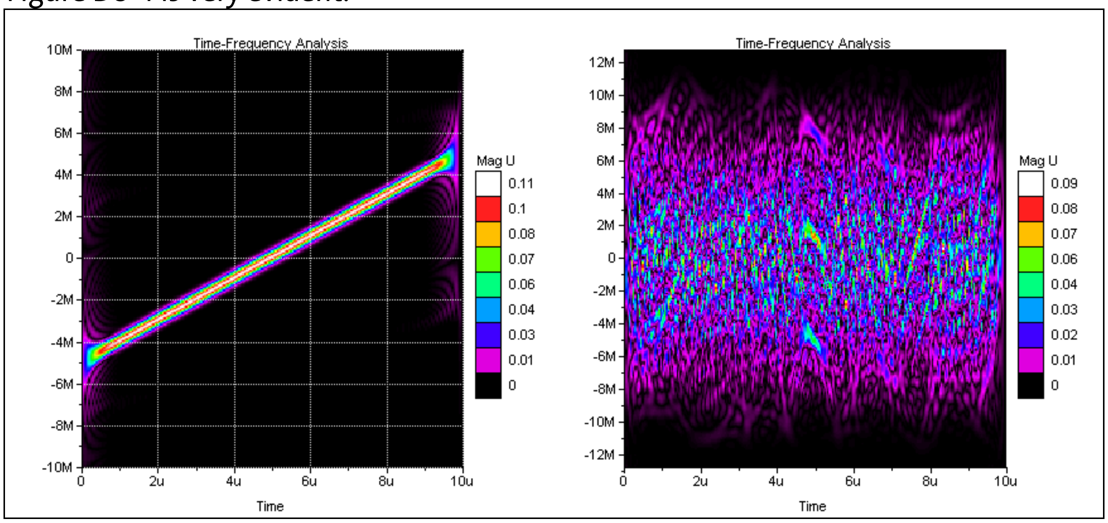


Figure D8-4 Time-frequency plots (a) LFM (b) MLS Quintic code

The mutual rejection of two LFM waveforms occupying the same band if one has the reverse slope of the other has been investigated. Instead of a MF, stretch processing (SP) mixes a delayed copy of the transmitted signal with the returned signal (Figure D8-5). The resulting signal, after low-pass filtering (LPF), consists of a number of tones corresponding to the relative range of the targets. A range profile is obtained by transforming the signal to the frequency domain.

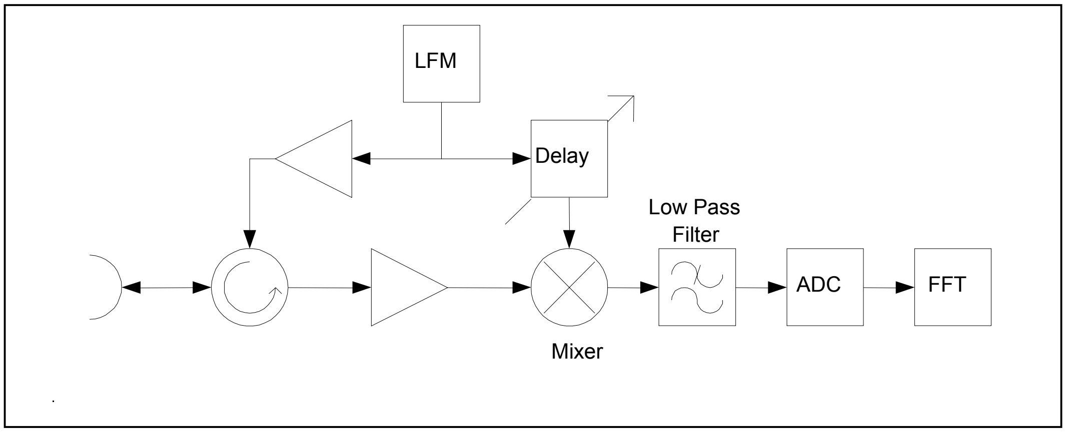


Figure D8-5 Stretch Processing

A key benefit of SP is that the IF bandwidth is defined by the LPF, which can be significantly less than the RF bandwidth. A lower IF bandwidth reduces the minimum sampling rate of the ADC at the expense of reducing the range swathe captured by the radar.

Where, two LFM waveforms occupy the same band, but with opposite slope, only a fraction of the foreign signal goes through to the ADC. If the RF and IF bandwidths are B_{RF} and B_{IF}, the rejection is twice the ratio of the IF/RF bandwidth:

$$Rejection = \frac{2B_{IF}}{B_{RF}}$$

Equation D8-1

For a B_{RF} of 500MHz and B_{IF} of 1MHz, the rejection is about 24dB, giving a valuable degree of rejection at the expense of a reduced range window width.

D.8.5 Bandwidth extrapolation/interpolation and spectral gaps

<u>Introduction</u>

Bandwidth extrapolation (BWE) is a well-established technique that applies Linear Prediction principals to increasing the radar's range (and angular) resolutions. A model is formed from the known spectrum, which is then used to extrapolate the spectrum by a factor of two or more to give a commensurate increase in the range resolution. The resolution enhancements of a factor of two can be made without prejudicing the original data. Significant extension beyond this factor should be made with care, because the model will dominate the original data [25,26,27].

The use of this technique is not without penalty, as it requires high SNR and the spectra of single targets need to be as smooth and flat as possible.

The technique to be discussed next is the allied method that can be used to interpolate the spectrum in order to fill gaps in the spectrum e.g. from interference blanking or due to a gap in its frequency coverage.

<u>Bandwidth interpolation/extrapolation</u>

BWE is based on linear prediction (LP) filters, which work on the principle that the sum of a set of equi-spaced signal samples, multiplied by a set of complex weights, can be used to predict the next sample. The nth sample y_n is then related to its M predecessors (y_{n-1} to y_n-M) by the linear expression:

$$y_n = \sum_{j=1}^{M} a_j y_{n-j} + x_n$$

Equation D8-2

The prediction coefficients a_j for this M pole LP filter are obtained from the known data by a number of techniques e.g. the Burg's Maximum Entropy algorithm. Differences between the data and the model end up in the residual or noise term x_n. The fitting algorithm finds the coefficients that give the minimum RMS residual [25].

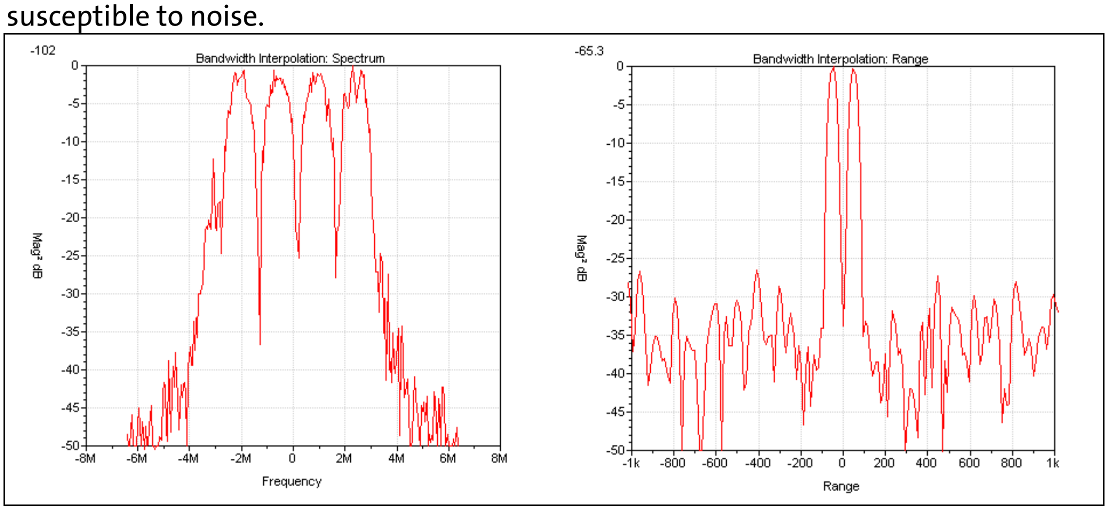
The BWE employs the linear prediction in the frequency domain (FD) and works best when all the spectral components have the same weights [26]. This is rarely the case with radar waveforms and any spectral weighting must be removed prior to the application of this technique. Further corrections can be applied e.g. the removal of the Fresnel ripple in LFM or the sinc function in coded waveforms. However, these additional corrections will only apply to the zero Doppler case, and the corrections become errors as the spectrum shifts with the target induced Doppler. This may degrade the performance of the waveform and BWE technique when Doppler is involved.

To illustrate this technique, a large 2MHz (~30%) gap is inserted into an LFM spectrum with centre frequency f of 4GHz, pulse-length, τ, of 20μs and a bandwidth B of 6.4MHz (TB=128). This gap is filled by bandwidth extrapolating in the frequency domain (FD) from both sides, the missing data is the weighted sum from the two contributions. For increased realism, thermal noise has also been added to the data.

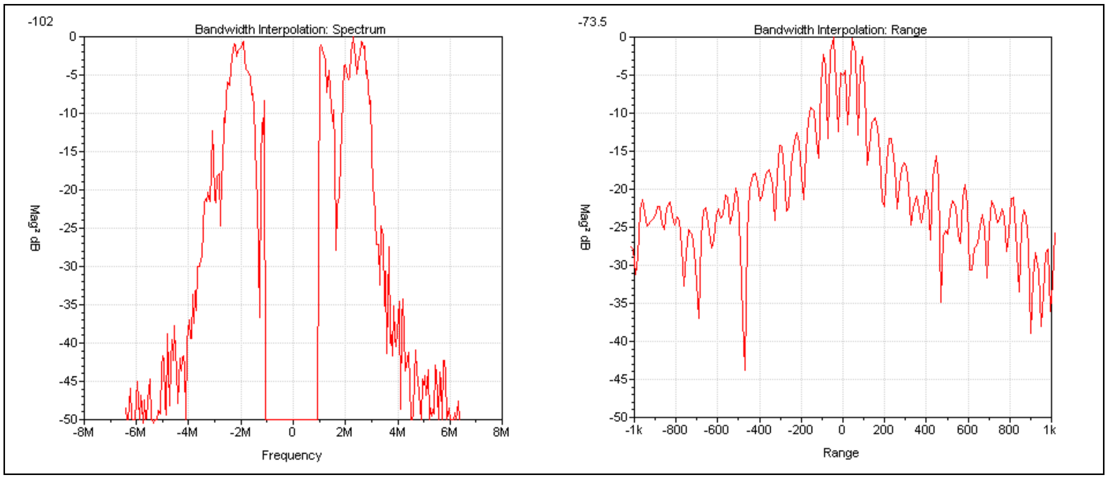
Figure D8-6 shows the LFM spectrum and its time domain (TD) equivalent for a pair of targets at $\pm$50m from the reference range without a spectral gap. Figure D8-7 shows the detrimental effect of the 30% spectral gap.

Figure D8-8 shows the effect of bandwidth interpolation with a 12-pole LP filter. Its success in reconstructing the missing features of the spectrum is remarkable and, whilst there is inevitably some corruption in the target profile, but most of the salient structure has been restored.

The technique works very well with the LFM waveforms due to their relatively clean and flat spectral structure but also gives good results with the coded waveforms. The spectra of these waveforms tend have a noise-like structure superimposed on a sin(x)/x base. This increases the demand on the BWE process and consequently requires a higher order LP filter. This requires more known data and makes it more susceptible to noise.



Figures D8-6 LFM waveform: no gaps (a) FD (b) TD



Figures D8-7 LFM waveform: 20% gap (a) FD (b) TD

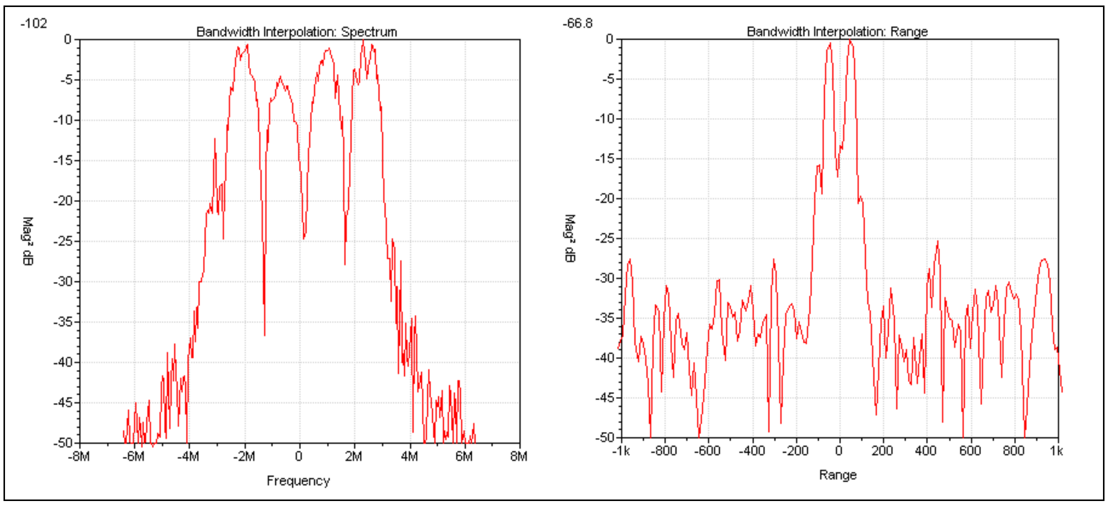


Figure D8-8 LFM waveform: 20% gap filled (a) Frequency Domian (b) Time Domain

D.8.6 Generating spectral nulls

<u>Introduction</u>

As discussed in §D6.1, the radar's matched filter relies on the magnitude of the waveform’s spectrum to maximally detect its signal and reject interference. Interference rejection would therefore be much improved if nulls could be inserted into the waveform’s spectrum to coincide with the frequencies of foreign signals. An additional benefit is that interference by the radar to the signal source would also be reduced. Two methods of generating such nulls are discussed.

In both methods it is possible to define the depth and width of each null, however, decreasing the spectral occupation and the consistency of the main spectra inevitably leads to worse range resolution and larger range sidelobes.

<u>Gerlach method</u>

The technique reported in reference [28] places nulls in the spectrum of an ultra wide band (UWB) radar waveform to reduce its interference with other users of the spectrum. It draws on theory developed for null steering antenna beam patterns and is based on the well known stepped frequency (SF) waveform.

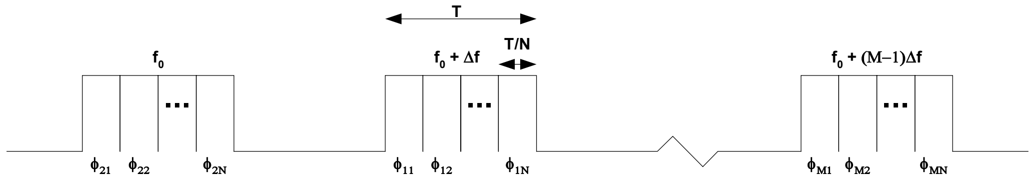


Figure D8-9 Stepped Frequency Polyphase Coded Waveform

Whilst the SF waveform consists of a series with *M* bursts and a frequency step *Δf*, the stepped frequency polyphase coded (SPFC) waveform adds additional coding to the waveform by splitting each pulse into *N* sub-pulses (Figure D8-9). These sub-pulses are amplitude and phase coded in such a way as to null the spectrum at desired frequencies. Up to *N/2* nulls can be placed in the spectrum – but crucially they cannot be placed within the main spectral lobe. The response to interference lying within the main lobe is to skip that particular frequency– with the consequential degradation in range sidelobes.

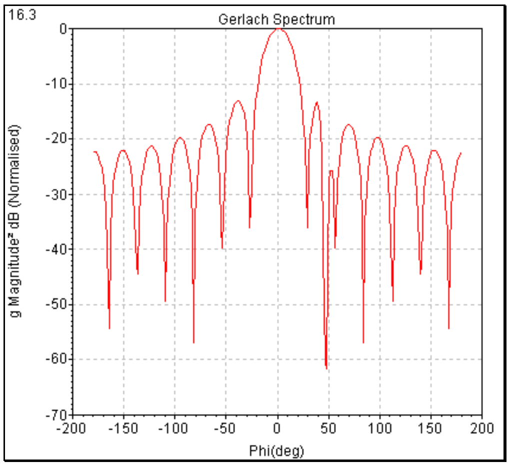


Figure D8-10 Spectrum with null at 48°.

The phase coding is kept to the minimum necessary for null generation and consequently the waveform remains a SF waveform. As the nulls are generated by small phase perturbations to the underlying waveform, it is likely that the technique can be applied to any waveform such as binary phase codes (BPSK). Each BPSK chip would have to be sub-divided in to *N* sub-pulses, which would be subjected to the same phase perturbations as before.

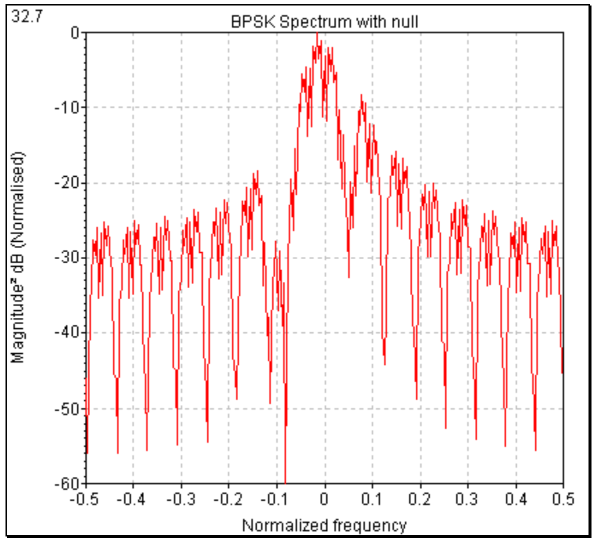


Figure D8-11 BPSK Spectrum with null.

Its stated inability to insert nulls in the main spectral lobe reduces the utility of this technique for RF radar use. In addition, the nulls are generated at some cost because the waveform has to be over-sampled by a factor *N* in order to modulate the sub-pulses. The receiver also has to over-sample by the same factor to benefit from the gaps in the spectrum and filter out interfering signals.

<u>Spectral nulls by code</u>

Spectral nulls can also be generated in phase coded waveforms by careful choice of phase sequence and values. This is primarily applicable to MLS based codes i.e. where the code sequence is based on a pseudo-random code (and therefore includes Taylor quadriphase). Small changes to the code values can be made to form nulls of the desired depth and position within the main spectral lobe without seriously degrading sidelobe performance.

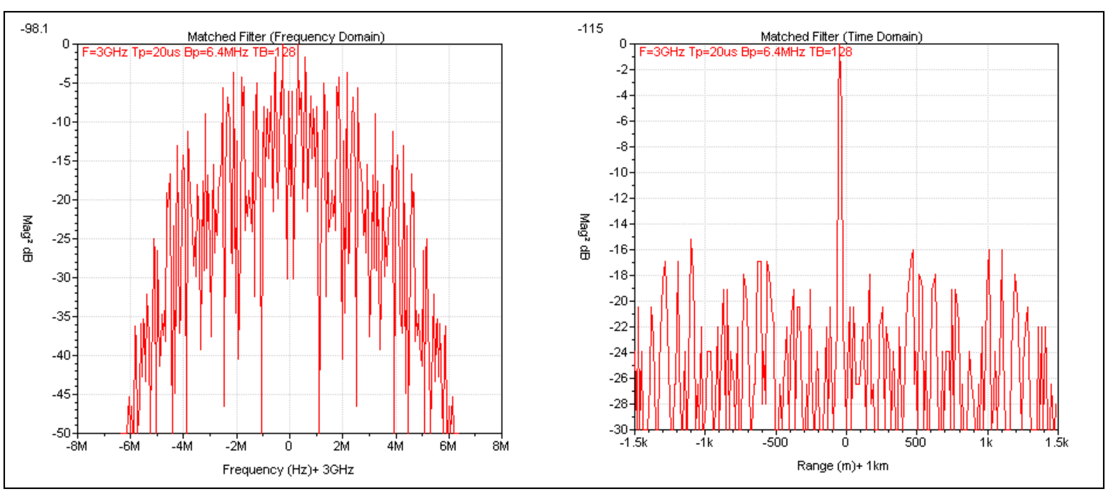


Figure D8-12 MLS code (a) spectrum (b) PSF

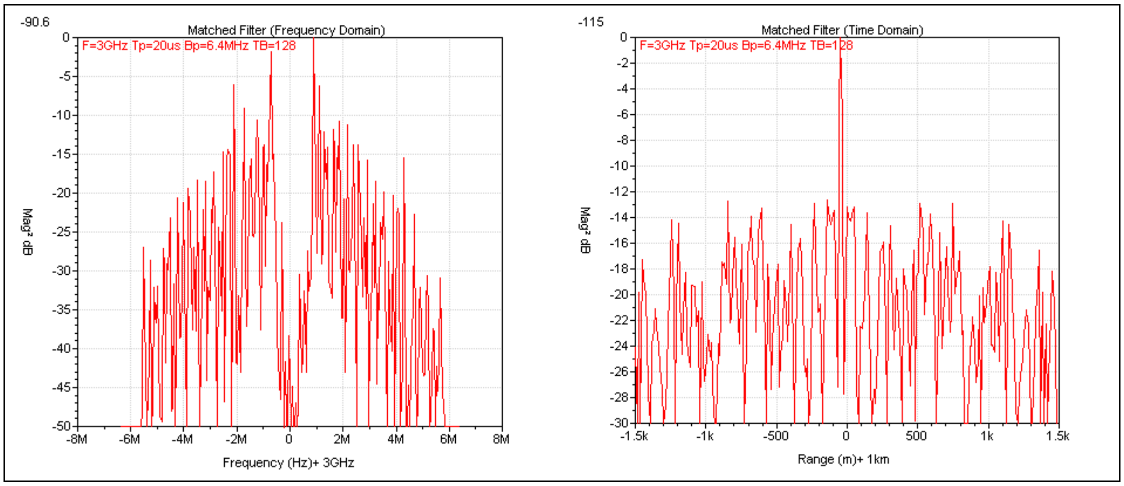


Figure D8-13 MLS code with null (a) spectrum (b) PSF

The spectrum and PSF of a typical binary MLS phase coded waveform is shown in Figure D8-12. The same waveform with a broad null inserted in the centre of its spectrum and its PSF is shown in Figure D8-13. There is inevitably some distortion to the PSF, but this is likely to reduce in further development of technique. Note that no amplitude modulation is employed and no power is lost by the waveform.

D.9 Co-operative waveform

D.9.1 General

Co-operation between radar systems operating in the same geographic area allows the ability of the radars to have mutually advantageous parameters and waveforms. This could be in PRF selection, pulse width, power level, central operating frequency or compression waveform. The number of possibilities is large in division of the available time/ frequency/coding resources.

Co-operation could be achieved via communication between radar systems or in an unassisted approach, where each radar monitors the activities of other radars in their band so that, by appropriate adjustment, mutually compatible waveforms could evolve.

One particular implementation is given in the following sub section, which suffers no loss in instrumented range (the time domain) and efficiently uses the spectrum, given the range resolution that is achievable.

D.9.2 Stepped chirp

An example of where waveform choice and co-operation allows radars to occupy the same bandwidth without significant interference is the stepped chirp (SC) waveform. It splits the full bandwidth LFM chirp into a sequence of overlapping, narrow-band sub-chirps, which may overlap in frequency. The radar transmits and receives on one frequency at a time. It then adjusts both transmitter and receiver to operate at the next frequency and transmits the next sub-chirp (Figure 9D-1) – and so on.

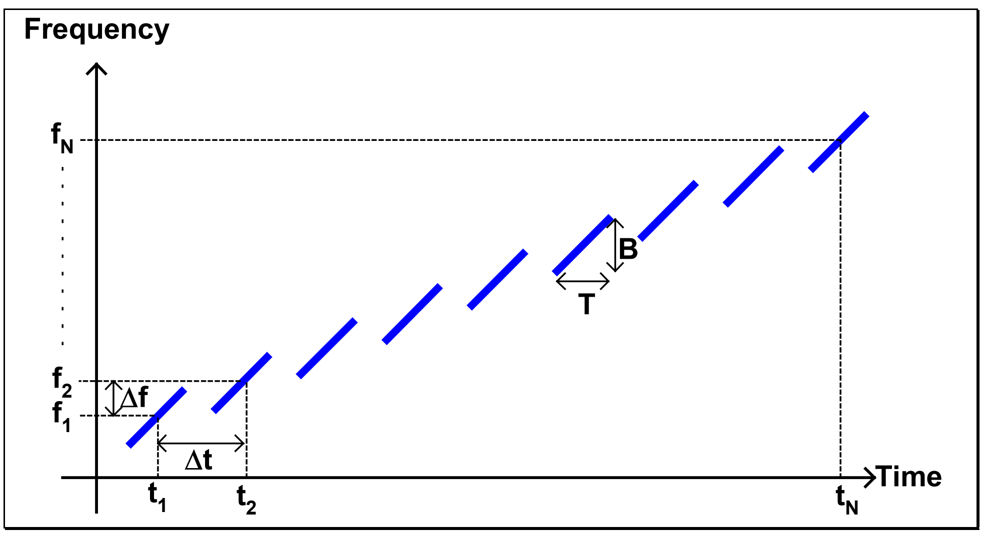


Figure D9-1 Stepped Chirp.

Thus, giving an overall bandwidth of $B+(N-1)\Delta f$. For example, with 128 sub-chirps of 5MHz bandwidth and a frequency step Δf of 3.2MHz will yield an overall bandwidth of 411MHz. A wide bandwidth is obtained with an inherently narrow-band system. In theory, *N* radars could operate on the same band provided that they are co-operating and synchronised. However in reality frequency and timing synchronisation between radar systems will not be perfect, therefore, the actual number would be reduced.

The illustrated example of the pulse train transmits one pulse at each frequency, but this need not be the case. A series of pulses can be transmitted at a high PRF to improve the SNR by increasing the overall pulse length.

This waveform gives a useful flexibility in that a high range resolution (HRR) image can be formed by single-sample-stepped-chirp (SSSC) processing or by frequency domain concatenation (FDC).

SSSC creates a HRR profile by applying a Fourier transform to the collection of single low-resolution target range cells from each frequency sample. Its main advantage over FC is its simplicity - only the range cells of interest need be analysed. The main drawback of this technique is its limited HRR range window, with the additional problem of targets outside the window aliasing into the HRR image.

The HRR range window is set by the frequency step interval Δf:

$$\Delta R = \pm \frac{c}{4\Delta f}$$

Equation D9-1

Which puts, constraints on the frequency step because the range window must be wide enough to cover the target object extent. The degree to which objects outside the window interfere with the image is determined by the low-resolution range response, which is determined by the windowing function used in the pulse compression window. It should also be noted that windowing also attenuates the in-region response, increasingly so as the object gets closer to the edge of the range window.

The sub-chirps need not be transmitted in order (or Costas sequenced), however this makes processing more difficult. In this case, each sub-chirp could be transmitted opportunistically when its sub-band is free.

For this efficient approach, providing all systems are operating with a different PRI period (t_1 to t_N), no time or bandwidth is lost in multiplexing with other radar systems, because:

- The realisable resolution is given by the total 3dB bandwidth.
- The spectral response is as compact as a single FM chirp with properties given by Δf and T.
- The ambiguous range is derived from the pulse repetition interval Δt.

D.10 Implications of waveform design on band reduction

D.10.1 Introduction

This section quantifies the effect on spectrum efficiency that the waveform design techniques, explored in this report, are likely to have upon three selected radar bands. These three bands and their primary users are:

- 1.215-1.350GHz with only ATC being the primary user.
- 2.700-3.100GHz where ATC radars are a primary user throughout the band, meteorological (Met) radars are a primary user below 2.9GHz and marine radars are a primary user above 2.9GHz.
- 15.40-15.70GHz where airport surface detection equipment (ASDE) radars are a primary user.

The guidance rules for this project are to not consider changing Marine radars, because of the number of systems and the International implications.

D.10.2 Radar parameters used for this exercise

The isolation that is required between radar systems operating in the same band, is explored in §D4 and was found to be derived from range between radars and their detection performance.

The specifications of the current, in-service, radars have been assessed to obtain their detection performance. In addition, their geographical locations have been analysed to obtain separation range statistics and the number of systems operating within an area that could interfere with each other. From this, the required waveform interference rejection figures have been derived for mutual operation of the radar systems. In addition, the number of systems requiring such a rejection level is also derived based on the geographical distribution. The parameters for the systems that are derived are given in Table D10-1. These are used in analysis of the different candidate solutions below.

	Interference rejection	**Range resolution bandwidth (MHz)**	**Number of interfering systems**
L-band ATC	-90	2.5	5
S-band Met	-90	1	1
S-band ATC	-90	2.5	5
Ku-band ASDE	-60	25	5

Table D10-1 Parameters for calculation of band usage requirements

D.10.3 Candidate solutions

The following solutions have been selected as promising candidates from the earlier sections of this report.

<u>Matched transmitted pulses:</u>

This solution is to apply the concept of matching pulse rise times to the resolution bandwidth as described in §D5.8. To enable Interference rejection between multiple radar systems operating in the band, it is assumed the central operating frequencies are different. Frequency separations for different Interference rejection figures are given in Table D7-2.

Matched transmitted pulses + interference rejection:

To apply the concept of matching pulse rise times to the resolution bandwidth as described in §D5.8. To enable Interference rejection between multiple radar systems operating in the band, it is assumed the interference rejection techniques described in §D8.3 are used.

Frequency modulation:

To use FM waveforms for all radars, using their compact nature. Many radar systems already use similar waveforms, therefore these results serve as a method of evaluating the theoretical bandwidth of current radar systems. Again, it is assumed the central operating frequencies are different, and frequency separations for different interference rejection figures are given in Table D7-2.

Filtered orthogonal polyphase codes:

To use orthogonal polyphase waveforms as described in §D6.5 with time bandwidth products of approximately 200 for all radars. Again it is assumed the central operating frequencies are different, and frequency separations for different interference rejection figures are given in Table D7-2.

Filtered orthogonal polyphase codes + interference rejection:

This is similar to the above, however interference rejection is achieved by active cancellation, as described in §D8.2. The active cancellation is assumed to reduce interference to below 20dB INR, the orthogonal nature of the codes reduce the residual Interference to below the noise floor, this concept requires clarification through further work.

Co-operative:

The step FM co-operative scheme is described in §D9. This scheme is practical in these radar systems being considered because of the low number of radar systems that are within a significant geographical area (Table D10-1). The waveform on each pulse is assumed to be an FM Chirp, and hence the total spectral usage is likely to resemble that of N LFM waveforms with a range resolution bandwidth given by the *B, as* indicated in Figure D9-1. For this study a value of $N = 10$ was used.

D.10.4 Results of solutions applied to bands of interest

The solutions described above are applied to the three bands being considered in this study. These result in the total bandwidth requirements given in Table D10-2, ensuring no interference with radar systems in neighbouring bands. In Table D10-2, provisions have been included to allow for frequency shifts in radars that use frequency diversification techniques, indicated by $B_{s,}$ the shift bandwidth.

Clearly, the results for required bandwidth of the techniques exceed the available bandwidth (in red), this is due to unreliable values in Table D7-2, which is explained in §D7.6. In these cases the techniques are likely to improve current use of the spectrum, but demonstration and accurate measurement are required to assess improvements.

Table D10-2 shows that considerable improvements can be made in all the bands of interest, particularly in the cases, where filtered waveforms and co-operative step chirp FM waveforms are used.

Band	1.215-1.35GHz	2.79-3.10GHz	15.4-15.7GHz
Available bandwidth (MHz)	135	400	300
Technique	Required bandwidth (MHz)		
matched transmitted pulses	1500	1800	1100
Matched + interference rejection	300	200	275
FM	3700	4200	625
Filtered orthogonal polyphase codes	$75+B_s$	$281+B_s$	$212+B_s$
Filtered orthogonal polyphase codes + interference rejection	$10+B_s$	200	$100+B_s$
Co-operative	$4+B_s$	$204+B_s$	$32+B_s$

Table D10-2 Bandwidth requirements utilising the investigated techniques in the three bands of interest, NB the marine radar band is untouched, thus the minimum possible bandwidth in S band is 200MHz

D.11 Conclusions of WP6

D.11.1 Introduction

Methods for reducing a radar's spectral occupancy, whilst retaining its performance have been investigated. Methods for mitigating the effects of interference from other radar and communication systems are also considered.

D.11.2 Interference between radar systems

This study shows that for a typical system to work within a nominal 3km of a radar or communication system with similar bandwidth, the interference rejection needs to be in excess of 100dB. Whilst, the antenna sidelobes may be expected to reject 30-40dB of the energy, this leaves a challenging 60dB or more to be dealt with by waveform design and signal processing.

D.11.3 Radio regulations

In line with current thinking, some of the current regulations have been shown to be flawed. In addition, the regulations were found not to significantly prejudice waveform with theoretically noisy spectra, and in some cases perform the opposite. This is summarised by:

- The practice of adjusting waveform characteristics to give increased allowed bandwidths, which is common for radar manufactures to achieve type approval.
- Allowing spurious signals levels to be relative to peak envelope power rather than absolute levels.

D.11.4 Simple spectral noise reduction

A matched transmitter technique based on closing the loopholes in the regulations was presented which was shown to be effective in cleaning the spectral noise produced by current radar systems. This technique constricts the transmitted waveform to the minimum required to maintain its range resolution, typically its 3dB bandwidth. This concept could form the basis of future regulations. In addition, peak power reductions due to pulse compression and coherent processing were shown to be effective in reducing noise over the entire spectrum.

Reductions of allowed bandwidths of factors between 10 and 20 were produced over the current systems, if these techniques are implemented.

D.11.5 Orthogonal waveforms

The practicality of rejecting interference by employing orthogonal PC waveforms was investigated. These would normally give a rejection of the order of $1/TB$ and $1/\sqrt{TB}$, where TB is the time-bandwidth product. To achieve more, one would require complementary spectra, but with a major degradation in sidelobe and Doppler performance.

LFM and digitally derived equivalents were found not to offer sufficient orthogonal variants for multiple radar systems operating in a close proximity. Polyphase coded and Costas coded waveforms offer the best candidates for operation of a large number of radar systems, in a shared spectrum. However, large *TB* products are required to achieve adequate range sidelobe levels, and an unfeasibly large *TB* product would be required to meet the interference rejection targets.

It has been presented that when analogue orthogonal waveforms are used in conjunction with the technique of operating on different centre frequencies, the use of orthogonal waveforms gives no significant advantages. Whereas combining the techniques of frequency separation with coded orthogonal waveforms can enable interference performance as good as analogue waveforms, which have more compact spectrum.

D.11.6 Magnetron radars

Modern coaxial magnetrons, magnetron modulation and injection locking have been discussed to offer more frequency stability and less noise than current magnetrons.

This may enable tighter packing of such radars into a given frequency band with reduced harmonic emissions and lower power levels.

D.11.7 Pulse compression waveform with narrow spectrum

The radar's spectral footprint can be reduced by careful selection of waveform or by modifying its existing waveform. The ideal would be, where all the transmitted energy lies between the range resolution (half-power) bandwidth, which is essential for maintaining the range resolution.

Taylor quadriphase:

Taylor quadriphase is a variant of the four-phase coded waveform that can reduce the wide spectral splatter normally associated with coded waveforms, whilst maintaining the performance of its conventional counterpart. Its spectrum falls off at a much steeper 12dB/octave, compared to 6dB/octave for conventional waveforms, reaching -40dB in a 1/13th of the bandwidth. In addition, as there is much less energy in the spectral sidelobes, its spectrum can be truncated at the first nulls without a significant performance degradation.

Frequency modulation:

Given a reasonably high time-bandwidth product, the near rectangular spectrum of LFM waveforms means that they are naturally more spectrally efficient than coded waveforms. It can be improved however, and a number of techniques for reducing its spectral impact were also investigated.

Non-linear frequency modulation (NLFM) can significantly narrow the spectral width, but only about -30dB below peak. Below this level, the LFM spectral skirt fall-off is set by the sharp edges of the transmitted pulse. It is an attractive option, because it removes from the transmitted spectrum the part that would have been removed in the receiver by windowing. Furthermore, NLFM improves receiver SNR performance as no power is lost by windowing and the non-linear slope does not significantly affect its Doppler tolerant performance.

A somewhat drastic method of generating LFM waveforms with much reduced out-of-band emission was also examined. It has a considerable potential for spectral cleanliness, but as the familiar LFM spectral shape as now been transposed to the pulse envelope, it has the disadvantage that it requires a degree of linearity in the radar’s power amplifiers. It is an effective technique, however when the maximum bandwidth is required within a constrained spectral environment.

Frequency separation:

Frequency separation of waveforms to reduce interference between standard analogue and digital waveforms has been examined. Predictably it were found that waveforms with compact spectra can operate with closer frequency separations for the given interference targets. For example, LFM waveform operating on frequencies separated by only five times their range resolution bandwidth can reject each other by greater than 60dB.

However, orthogonal digital waveforms were found to be superior in some aspects when frequency shifted, especially if combined with waveform filtering.

D.11.8 Spectral nulls

One solution to interference rejection is to generate nulls in the waveform's spectrum that coincide with the interference signals. Interference by the radar on the signal source would also be reduced along with the sensitivity of the matched filter (MF) to those signals. Two methods of generating such nulls were discussed.

Such methods could also be used to form sets of waveforms with non-overlapping spectra, giving truly orthogonal waveforms. This possibility needs further work.

D.11.9 Interference filtering/rejection

The ideal solution to interference is to remove it prior to the matched filter (MF) and a method for so doing was shown to be effective. Rejection methods have potential, where a degree of co-operation between the radar systems is present. However, simple interference rejection methods, currently used in incoherent radar, were thought to be less effective for pulse to pulse coherent modern radar, because of the need for a staggered PRI.

This area of interference filtering/rejection outside the remit of this report. However, it has been covered briefly, because when used in conjunction with radar waveform techniques presented, they make the discussed techniques and waveform viable. For example, interference rejection techniques require that the interfering signal has high signal to noise(>20dB), and the interference may not be completely removed, whereas the waveform techniques reduce interfering signals at any signal to noise, often by smaller amounts than the interference rejection. Therefore, the dual use of interference filtering/rejection and the waveform techniques can satisfy interference rejection targets given in conclusion D.11.2. This is considered to be as an area in which further study will be required.

D.11.10 Stepped chirp and FSK waveforms

The stepped chirp waveform splits the full bandwidth LFM chirp into a sequence of overlapping, narrow-band sub-chirps, which may overlap in frequency. As the receiver is only tuned to the narrow-band sub-chirp it offers high rejection to other radars transmitting at other sub-bands.

The sub-chirps need not be transmitted in order (i.e. Costas waveform), but this adds to the complexity of processing. However, in this case, each sub-chirp could be transmitted opportunistically when its sub-band is free.

D.11.11 Bandwidth extrapolation /interpolation for gap filling

Bandwidth extrapolation is a well-established technique that applies linear prediction (LP) principals to increasing the radar's range (and angular) resolutions. A model is formed from the known spectrum, which is then used to extrapolate the spectrum by a factor of two or more to give a commensurate increase in range resolution. The resolution enhancements of a factor of two can be made without prejudicing the original data.

This technique has been applied to the allied problem of interpolating the spectrum in order to fill gaps in the spectrum. Gaps could be due to null forming in the transmitted waveform, or due to interference blanking in the frequency coverage. The technique has been found to be successful, but requires good SNR to be effective.

D.11.12 Application of waveform techniques to bands of interest

Application of a selection of the waveform concepts to three exemplar bands of interest exposed a need for assessment of the waveform spectrums through demonstration.

However results shows that considerable improvements can be made in all the bands of interest, particularly in the cases where filtered waveforms and co-operative step chirp FM waveforms are used. Reductions of the selected radar bands are calculated to less than half of the current radar frequency allocations.

D.11.13 Summary

A number of waveforms, schemes and techniques have been explored, which have the potential to reduce the spectral occupation of radars. In addition, the problem of interference between radar systems has been addressed. Whilst no single panacea was found that fully met the requirements for band sharing in all circumstances, a combination of techniques could be effective.

Such combinations will form complex systems that require design, simulation, demonstration and evaluation. As such this will be a topic for further work.

D.12 Recommendations of WP6

D.12.1 Demonstration and optimisation of identified waveforms and techniques

This study covered theoretical aspects of waveform design and the practicalities of implementing the waveform in civil systems need further examination. In addition, spectral spread and hence rejection of waveforms with different frequency shifts and bandwidths is likely to be equally dependent of the radar system as the waveform. Therefore, measurement of implemented waveforms on representative systems is essential for assessment of spectral spread and interference.

In addition, because of the broad nature of this study, it is recommended that further in-depth study and optimisation of the most promising techniques identified in this report and investigation into the effects or combination of these techniques be carried out.

D.12.2 Interference signal rejection techniques

This study briefly reviewed the benefits of a small selection of interference rejection techniques, and found them useful in support of efficient waveform design. Therefore, it is urged that a further broad investigation into interference rejection techniques be made to fully assess the assumptions made in this report and to investigate further benefits.

D.12.3 Co-operative/sympathetic radar techniques

A co-operative radar technique has been explained, which show the benefits of co-operation. In addition, several other techniques have been introduced that can be used to adaptively use available bandwidth (or segments of bandwidth), and hence reduce interference to and from radar systems. The subjects of co-operation, dynamic radar bandwidth allocation and adaptation will need further investigation and study, which could enhance constrained frequency bands usage by multiple radar users.

D.12.4 Support for changes in radio regulations for radar

In examination of the mask definitions, it has been found that pulsed FM mask formulae do not accurately reflect theoretical pulse FM spectrums. Therefore, in common with current thinking it is recommended that the regulations be changed.

In addition, it is also recommended that regulations be tightened so that they encourage spectral efficiency in waveform design.

D.13 Typical radar systems

	A	B	C	D	E	F	G	H
1	**CHARACTERISTICS**	Marine - commercial S	Marine - commercial S(2)	Marine- Commercial X	Marine- Commercial X(2)	Marine - leisure	Marine - leisure(2)	VTS
2	**Platform type**	Mobile(Ship) based		Mobile(Ship) based		Mobile(cruiser) based		Ground fixed
3	**Operational freq.**	3050	3050	9410	9410	9410	9410	9410
4	**Modulation**	pulse		pulse		pulse		pulse
5	**Tx Power into Antenna (kW)**	25kw		25kw		4kw		25kw
6	**Pulse width (usec)**	0.05	1	0.05	1	0.1	1	0.02
7	**Pulse rise/fall time (usec)**	0.01	0.01	0.01	0.01	0.01	0.01	0.01
8	**PRI (pps)**	700	1,400	700	1,400	500	2000	500
9	**Duty Cycle (%)**	0.007	0.07	0.007	0.07	0.02	0.05	0.008
10	**Chirp bandwidth(MHz)**	20	1	20	1	10	1	50
11	**Phase coded sub-pulse width**	N/a		N/a		N/a		N/a
12	**Compression ratio**	N/a		N/a		N/a		N/a
13	**RF emission bandwidth (-20dB)**							
14	**(-3dB)**							
15	**Rotation rate(rpm)**	24		48		24		20
16	**Output device**	Magnatron		Magnatron		Magnatron		Magnatron
17	**Output power (dBW)**	74	74	74	74	61	61	90
18	**Antenna pattern type**							
19	**Antenna type**							
20	**Antenna Azimuth beamwidth (deg)**	1		1		5		0.4
21	**Antenna polarisation**	H		H		H		H
22	**Antenna mainbeam gain (dBi)**	30		30		25		45
23	**Antenna elevation beamwidth (deg)**	25		25		25		2
24	**Antenna horizontal scan rate (deg/sec)**	144		288		144		120
25	**Antenna horizontal scan type**	rotating		rotating		rotating		rotating
26	**Antenna Vertical scan rate (deg/sec)**	N/a		N/a		N/a		N/a
27	**Antenna Vertical scan type**	N/a		N/a		N/a		N/a
28	**Antenna sidelobe levels 1st SLS and remote SLS**							
29	**Antenna height (m) above ground**	40		30		20		100
30	**Rx If bandwidth (3dB)**							
31	**Rx noise figure (dB)[2]**	5		5		5		5
32	**Min discernible signal (dBm)**							
33	**Rx front end 1dB gain compression point**							
34	**Rx on -tune saturation level Power density @ Antenna W/m^2**							
35	**RF rx 3dB bandwidth**							
36	**RX IF and IF saturation levels and recovery times**							
37	**Doppler filtering bandwidth**	Na		Na		Na		Na
38	**Interference rejection features**							
39	**Geographical distribution**							
40	**Fraction of time in use**	100%		100%		100%		100%
41	**Comments**							
42								

Figure D13-1 Typical radar systems, extracted as exemplars from the data base

	A	I	J	K	L	M	N	O	P	Q	R	S
1	CHARACTERISTICS	VTS(2)	ATC - D	ATC - D (2)	ATC - E	ATC - E (2)	MET - G	MET - E	Tracker - G	Tracker - G (2)	Tracker - I	Tracker - I(2)
2	Platform type		Ground fixed		Ground fixed		Ground/mobile/ship	Ground/mobile/ship				
3	Operational freq.	9410	1325	1325	2800	2800	5600	2800	5600	5600	9200	9200
4	Modulation		NLFM		NLFM		pulse	pulse	pulse		pulse	
5	Tx Power into Antenna (kW)		160		60		500kw	500kw	500kw		500kw	
6	Pulse width (usec)	0.02	1	50	1	50	1	1	0.25	1.5	0.25	1.5
7	Pulse rise/fall time (usec)	0.01	0.02	0.02	0.02	0.02	0.01	0.01	0.02	0.02	0.02	0.02
8	PRI (pps)	4000	1000	400	1000	400	300	300	1000	160	4000	160
9	Duty Cycle (%)	0.001	0.1	2	0.1	2	0.03	0.03	0.025	0.025	0.1	0.025
10	Chirp bandwidth(MHz)	50	2.5	2.5	2.5	2.5	1	1	4	0.666666667	4	0.66666667
11	Phase coded sub-pulse width		N/a		N/a		N/a	N/a	N/a		N/a	
12	Compression ratio		5:1 & 100:1		5:1 & 100:1		N/a	N/a	N/a		N/a	
13	RF emission bandwidth (-20dB)		2-20MHz		2-20MHz							
14	(-3dB)		2.5MHz		2.5MHz							
15	Rotation rate(rpm)		10		10		1	10				
16	Output device		TWT		TWT		Magnatron	Magnatron	Magnatron		Magnatron	
17	Output power (dBW)	90	85	85	80	80	100	100	102	102	102	102
18	Antenna pattern type											
19	Antenna type											
20	Antenna Azimuth beamwidth (deg)		1.5		1		1	1	1		1	
21	Antenna polarisation											
22	Antenna mainbeam gain (dBi)		35		35		43	43	45		45	
23	Antenna elevation beamwidth (deg)		5		3		1	1	1		1	
24	Antenna horizontal scan rate (deg/sec)		60		60		6	60				
25	Antenna horizontal scan type											
26	Antenna Vertical scan rate (deg/sec)											
27	Antenna Vertical scan type											
28	Antenna sidelobe levels 1st SLS and remote SLS		25 & 35		25 & 35		25 & 30	25 & 30				
29	Antenna height (m) above ground		35		10							
30	Rx If bandwidth (3dB)											
31	Rx noise figure $(dB)^2$		2		2		5	5	5		5	
32	Min discernible signal (dBm)											
33	Rx front end 1dB gain compression point											
34	Rx on -tune saturation level Power density @ Antenna W/m^2											
35	RF rx 3dB bandwidth											
36	RX IF and IF saturation levels and recovery times											
37	Doppler filtering bandwidth											
38	Interference rejection features											
39	Geographical distribution											
40	Fraction of time in use		100%		100%							
41	Comments											
42												

Figure D13-2 Typical radar systems, extracted as exemplars from the data base

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E WP640: Demonstrating waveform designs

E.1 Introduction

The objective of this work package was to demonstrate the autocorrelation and cross correlation properties of representative pulse compression (PC) waveforms in order to validate the analysis in WP630. The demonstrations were carried out using a coherent L-band radar at CCLRC's (RAL) Chilbolton Observatory.

Three sources of interference were simulated:

- Trans horizon main beam to sidelobe interference from an in-band radar, using the L-radar receiver path and a separate test transmitter.
- Line of sight interference from the sidelobes of a nearby communications transmitter to the radar's main beam using a test transmitter simultaneously sending radar and communications waveforms.
- Illumination of ground and rain clutter in a scattering volume visible to the interfering and victim radars, using the L-band radar transmitter and receiver but with orthogonal codes in the receiver correlator.

E.1.1 L-band radar

This instrument, still to some extent under development, is a coherent, Doppler radar designed for observations of echoes from clear air, but as it has a flexible, programmable arbitrary waveform generator and signal processor, it can readily be adapted for testing PC techniques.

It should be noted that this instrument is not normally used with long PC waveforms and has not been optimised for low sidelobe levels.

Figure E1-1 L-band radar - Pulse generator and signal processor rack.

E.1.2 L-band radar specifications

Antenna type	Prime-focus fed parabolic dish
Diameter	25 m
Gain	47.5 dBi
Beamwidth	0.66° (FWHM; -3 dB, 1-way)
First sidelobe level	-20 dB (1-way, estimated)
Scan rate	Typically 0.2° / second in azimuth and elevation
Far-field distance	5.3 km
Receiver	Single-conversion super-heterodyne, 60MHz IF
Noise figure	2.5 dB excluding duplexing / miscellaneous losses
IF type	Linear amplifier with I/Q detector
IF bandwidth	4 MHz
Video bandwidth	2 MHz
Dynamic range	72 dB
Transmitter	Travelling wave tube amplifier (TWTA)
Peak power	2 kW at TWTA output
Pulse-width	6.4 μs, typical
Pulse repetition frequency	3125 Hz, typical
Pulse-coding	8-bit BPSK complementary codes, typical
Data acquisition / processing system	Pentium PC plus custom ADC / timing cards
Number of channels	4 (of which 3 are in use)
Number of bits per channel	12
Sampling rate	2.5 MHz
System timing / clock frequency generation	Derived from crystal-controlled reference

Table E1-1 L-band radar specification

E.1.3 Test transmitter

This is similar to the radar exciter, consisting of a Lecroy LW120 arbitrary waveform generator generating baseband I and Q components of the required waveform, modulating an Agilent ESG D4000A signal generator. For simulations of interference from communications transmitters, a Rohde and Schwarz SMIQ03B signal generator, programmed with a number of communications waveforms, is coupled to the test transmitter antenna.

The AWG is triggered by the transmit pulse of the L-band radar, which for the interference measurements serves only to synchronise the test transmitter. The transmitter is set up 500m from the radar, so a near field correction must be applied to the radar antenna gain;

it was found necessary to elevate the antenna by 1º and to delay the test transmitter pulse by 150μs in order for the returns to clear local clutter.

Due to the small separation of the transmitter and receiver, no RF power amplifier was needed and a dipole antenna was sufficient.

In the interests of realism, the frequency references of the test transmitter and receiver were not synchronised, resulting in a frequency offset of a few 10's of Hertz, timing jitter in the AWG trigger of up to 18ns and a mismatch between the transmitter and receiver sampling clock transitions.

Figure E1-2 L-band radar seen from the test transmitter

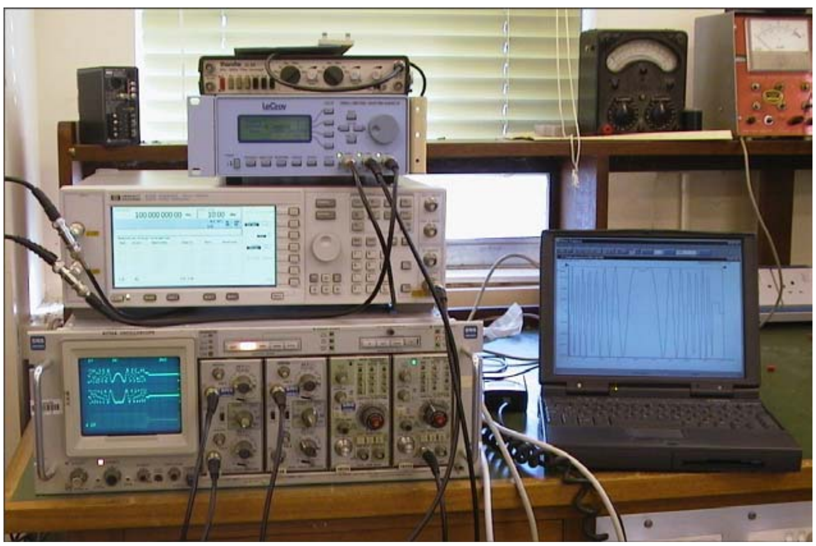


Figure E1-3 Test transmitter showing baseband and RF chirps

E.2 Interference from an in-band radar

The test transmitter power level was set to simulate main beam to sidelobe trans-horizon interference from a distant radar, assumed to have the following characteristics:

Peak power, P_t	100 kW
Antenna gain, G_t	40dBi
Distance, R	100 km
Victim sidelobe gain, G_r	0dBi

The path loss, L_p, is highly dependent on topography, but a typical value would be about 200dB; the received power in this case would be:

$$P_r = P_t G_t G_r L_p = \text{-80dBm}$$

Equation E2-1

To calculate the power required to produce this power from a test transmitter at a distance of 500m:

$$P_r = P_t G_t G_r \lambda^2 / (4\pi R)^2$$

Equation E2-2

In this case,

P_r = -80dBm,

G_t = 0dBi (this is the gain of the test transmitter antenna)

G_r = 29dBi (this is the main beam gain of the L-band radar, corrected for elevation and for the near field)

λ= 0.235m

and P_t is then -20dBm.

The measurements were repeated with a power level of +10dBm in order that the cross correlation sidelobes could be recorded uncontaminated by noise.

E.2.1 Test results

Several chirp waveforms, polyphase codes and quadriphase codes were transmitted at power levels of –20dBm and +10dBm, and each pulse was cross correlated with the transmitted waveform and also with reverse slope chirps and orthogonal codes.

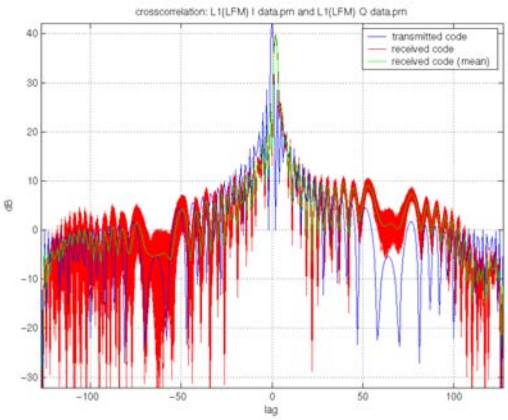
In all the following plots, the "blue trace" is the autocorrelation function of the transmitted waveform, the "red trace" is the superposition of the cross correlation functions of 4096 received pulses and the "green trace" is the average of the 4096 cross correlation functions. In the case of the chirp waveforms, the "magenta trace" is the cross correlation of the received chirp with a reverse slope chirp.

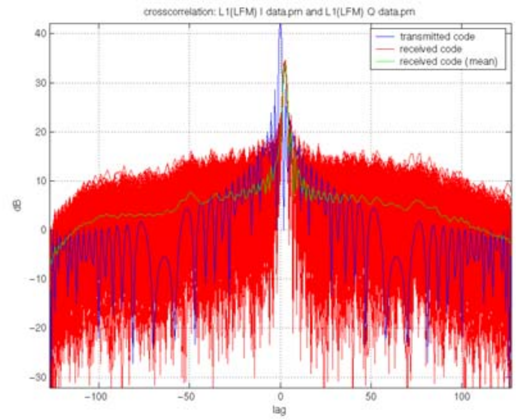
The high sidelobe level on the right hand side of several of the chirp plots is due to a fault in the manufacturers AWG control software. The effect of this was to replace an apparently random number of chips at the end of the code with a DC level, causing a peak in the cross-correlation function at a lag of around half the chirp length. The effect of this on the polyphase codes is to modify the sidelobe structure and to increase the average sidelobe level. This is being investigated, but has not been resolved at the time of writing.

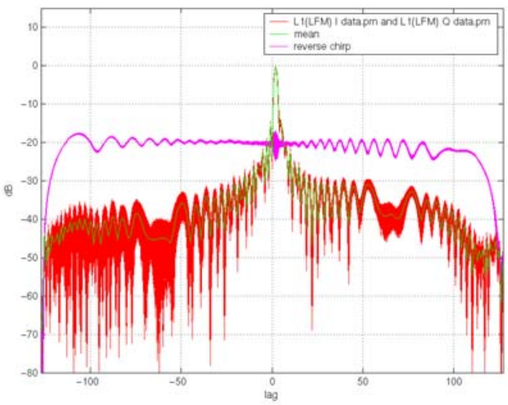
E.2.2 Test results: Chirp waveforms

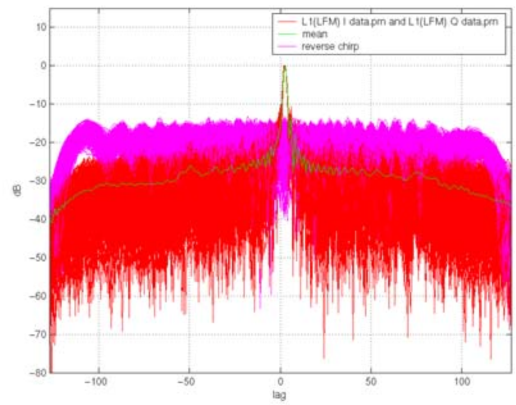
In each example, the cross-correlation function of the received and transmitted chirp at high and low signal to noise ratios (SNR) is plotted with the autocorrelation function of the transmitted chirp for comparison, followed by the cross correlation function of the received chirp and a reverse slope chirp with the normal cross correlation function for comparison.

<u>L1(LFM) Linear FM:</u>

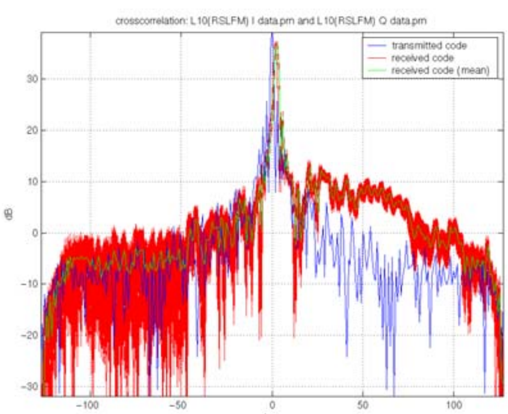


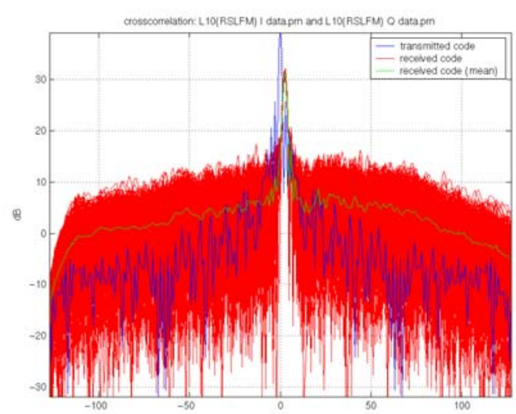


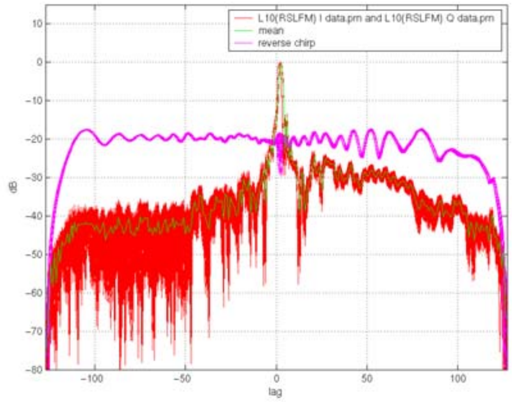


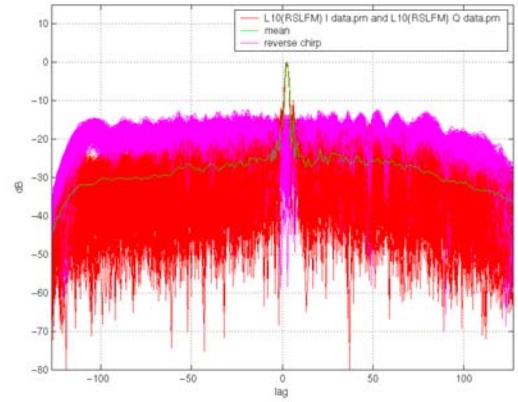


<u>L10(RSLFM) non-linear FM:</u>

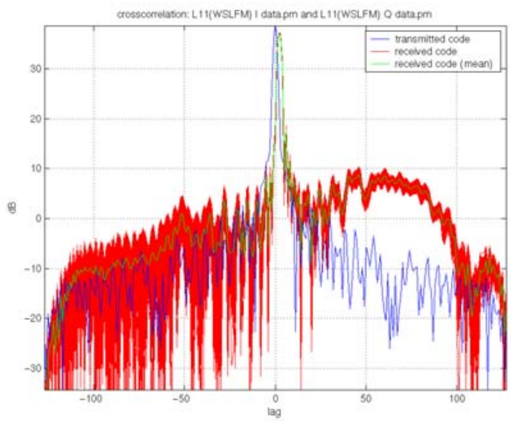


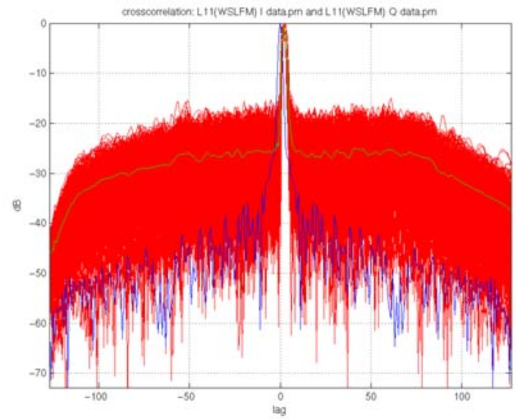


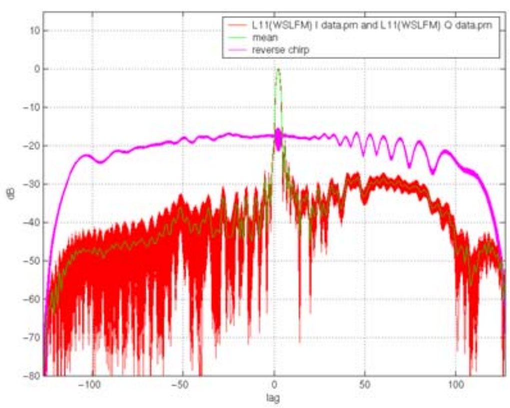


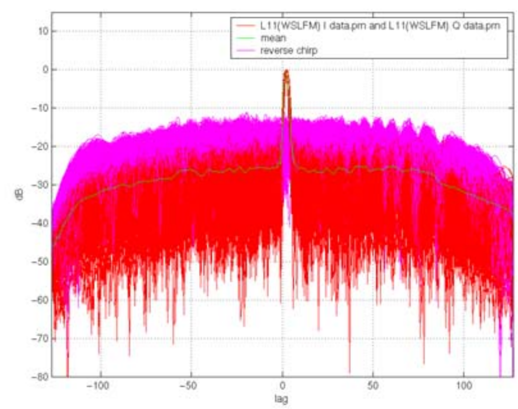


<u>L11(WSLFM)) non-linear FM:</u>

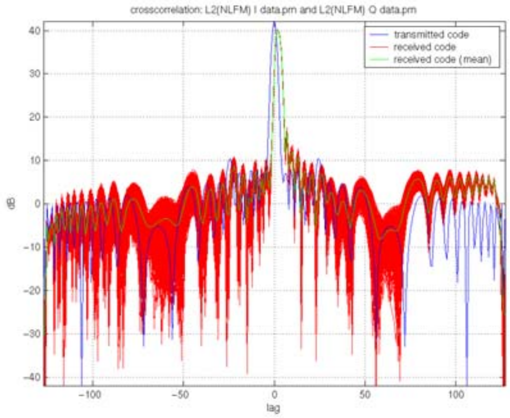


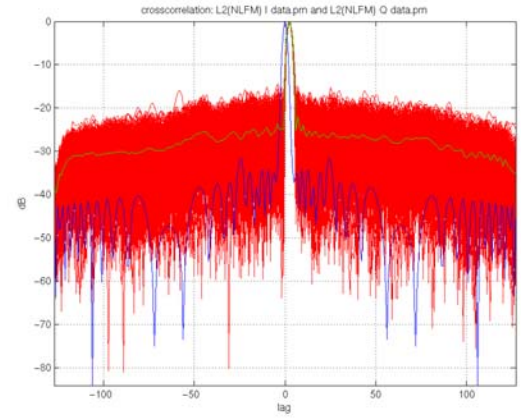


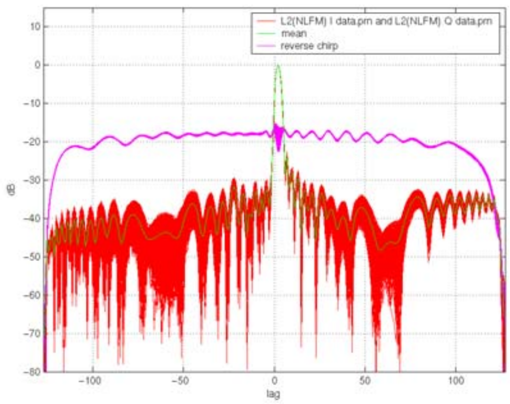


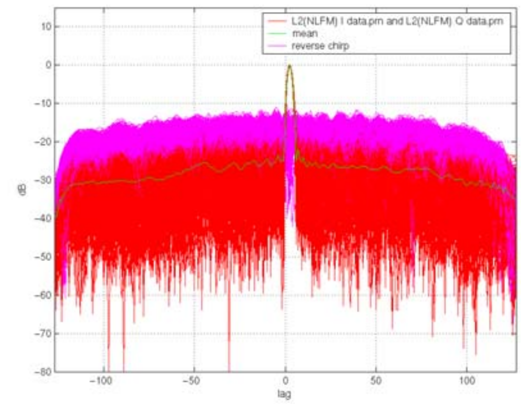


L2(NLFM)) non-linear FM:

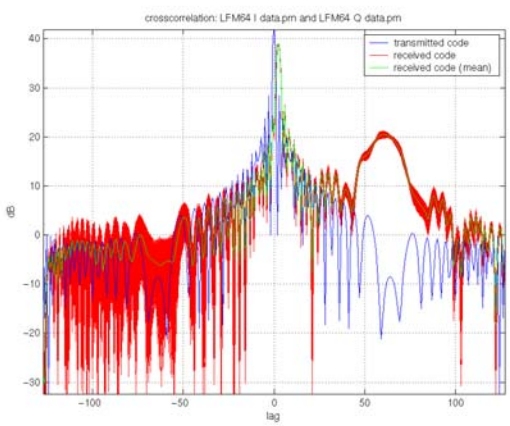


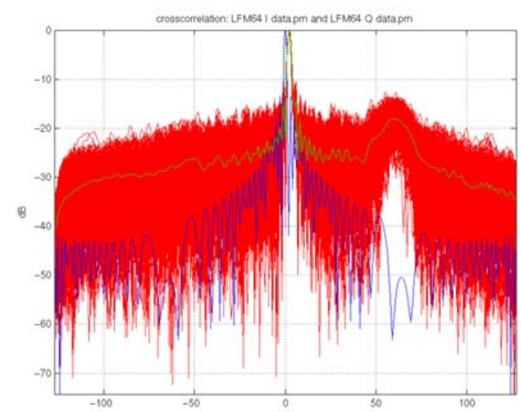


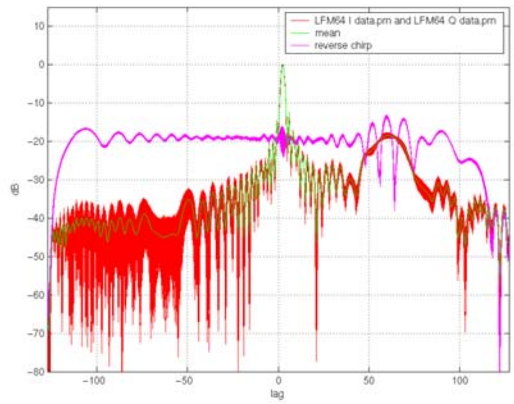


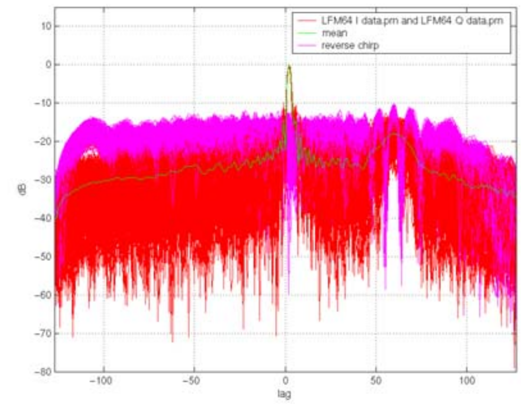


LFM(64) linear FM:

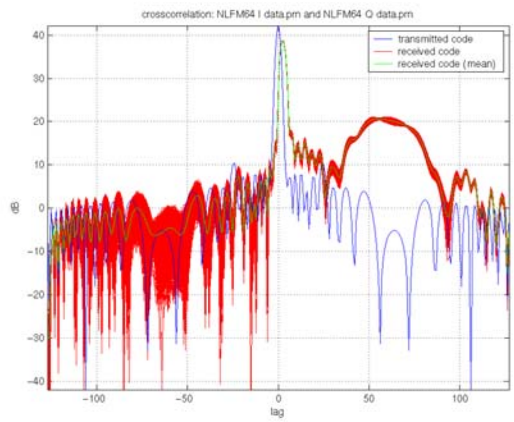


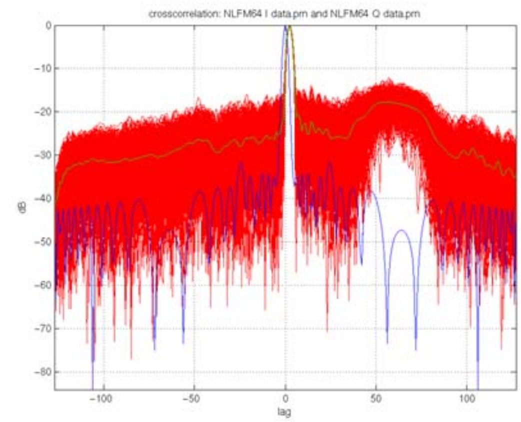


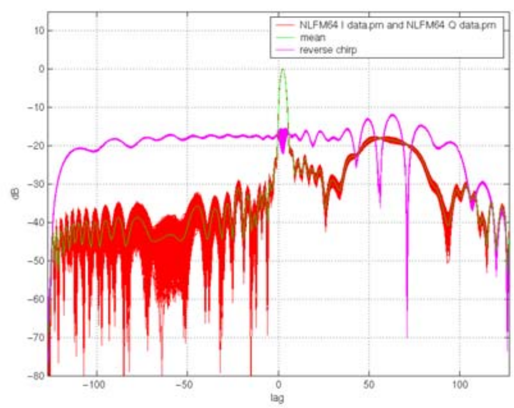


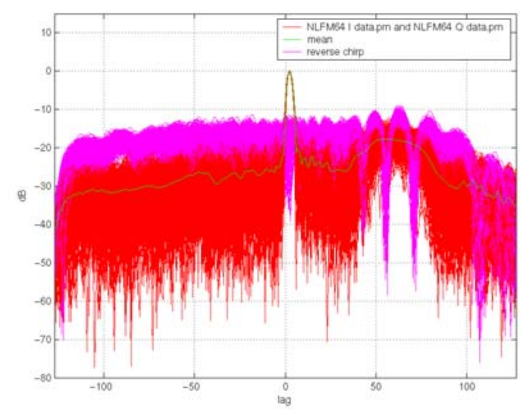


<u>NLFM(64) non-linear FM:</u>





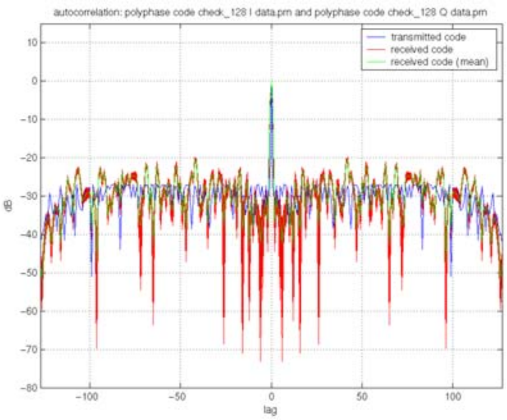


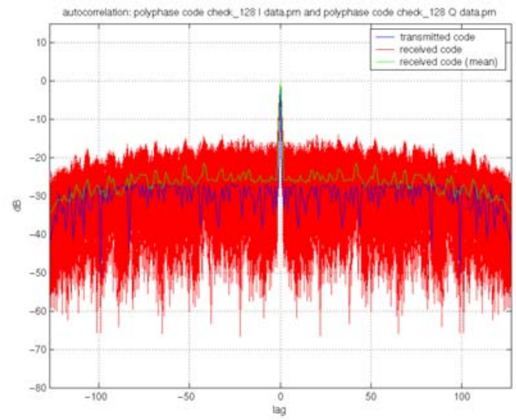


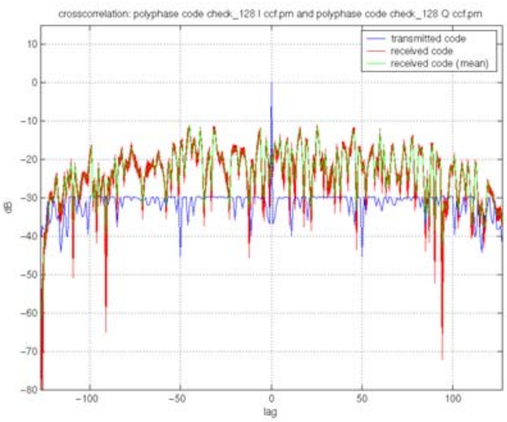
E.2.3 Test results: Polyphase codes

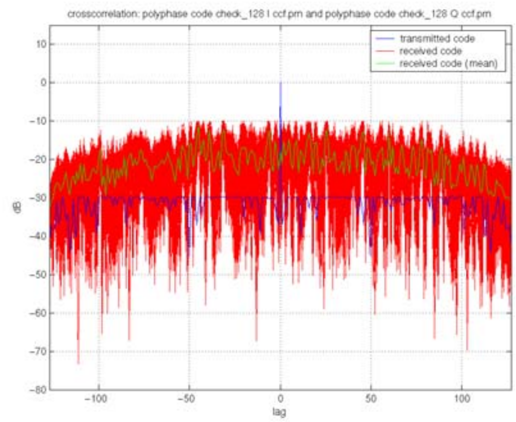
The plots show each received code correlated with the transmitted code at high and low power levels (labelled data.prn), followed by the cross-correlation function of the received code and an orthogonal code designed for optimum cross-correlation properties (labelled ccf.prn), again at high and low power.

<u>Polyphase code check 128:</u>

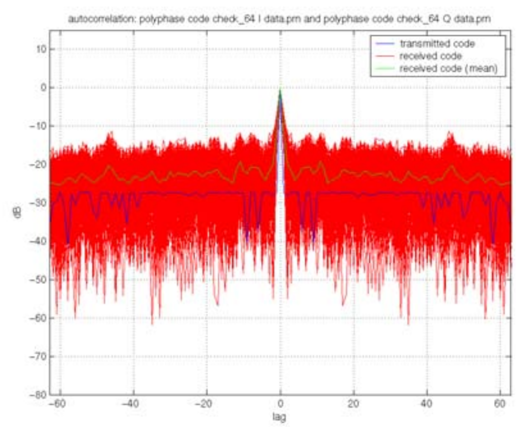


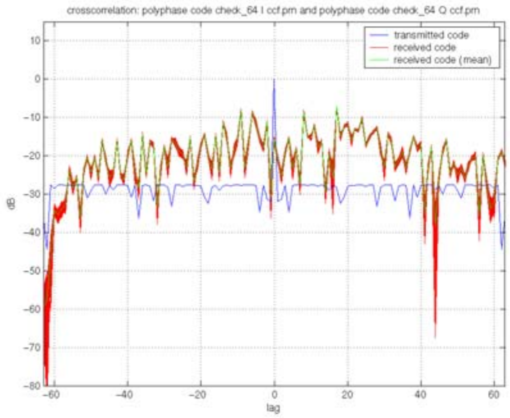


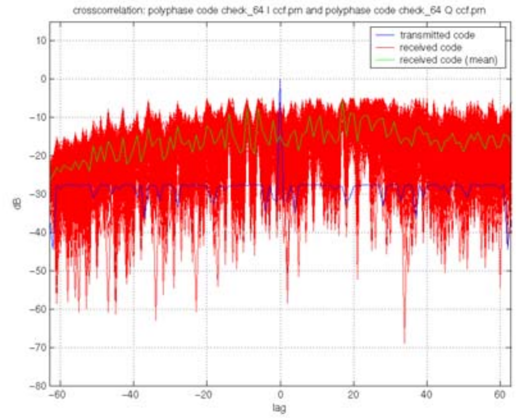




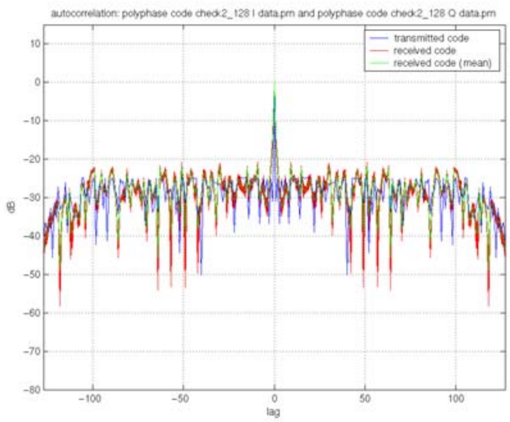
<u>Polyphase code check 64:</u>

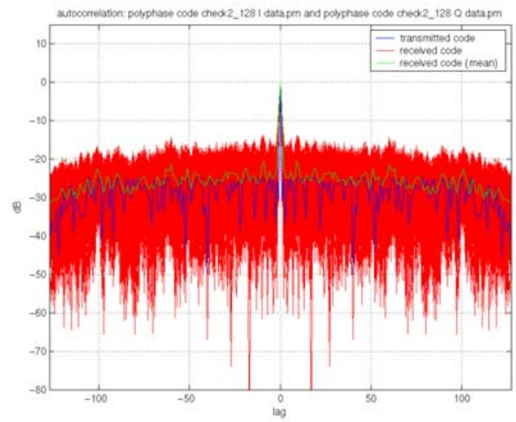


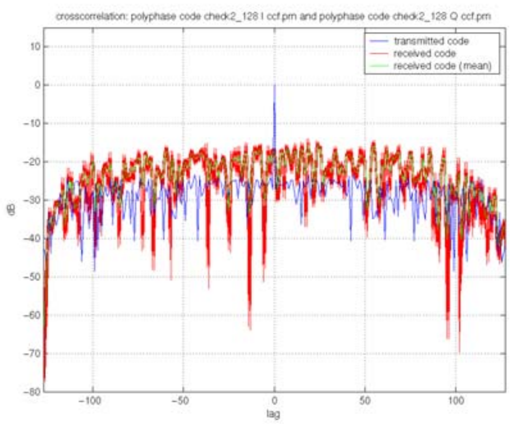


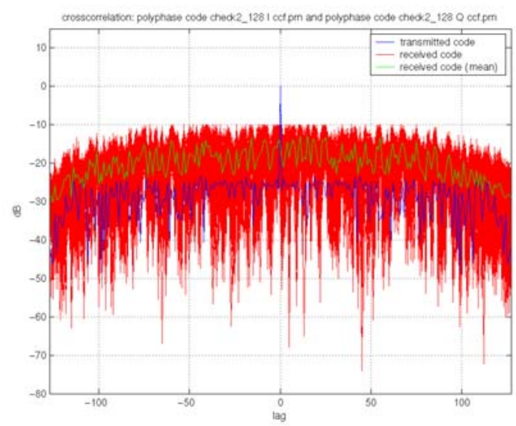


Polyphase code check2 128:

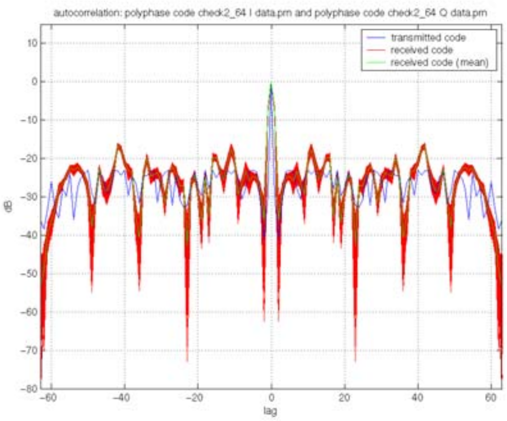


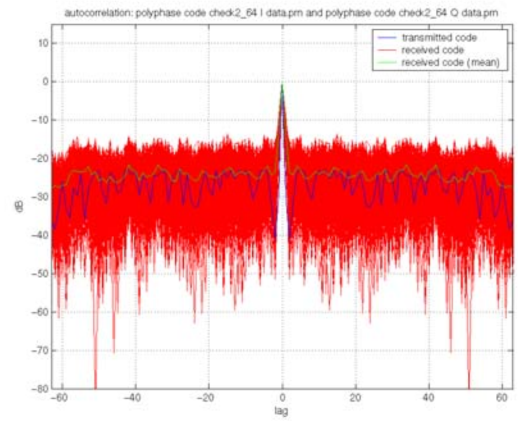


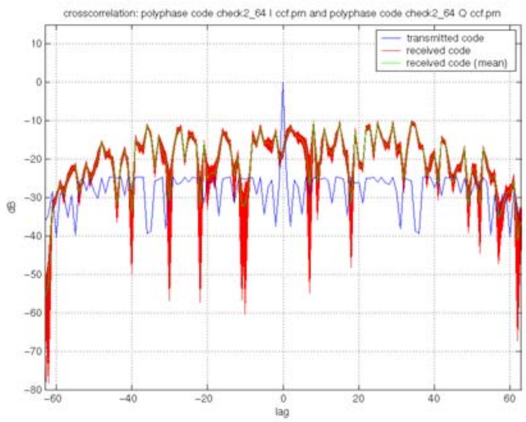


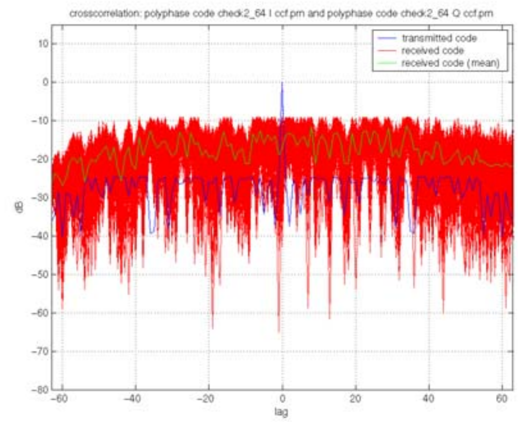


Polyphase code check2 64:



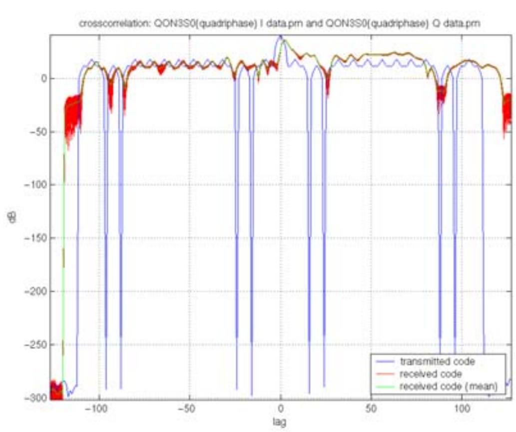


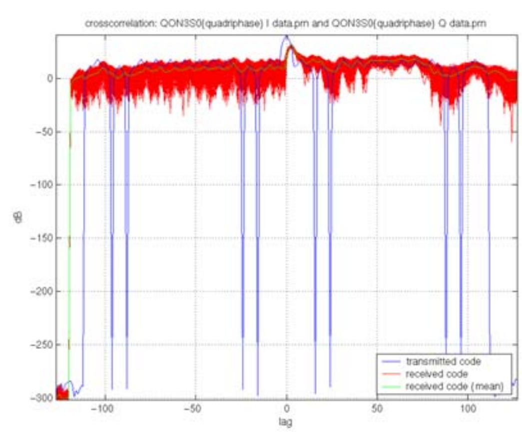


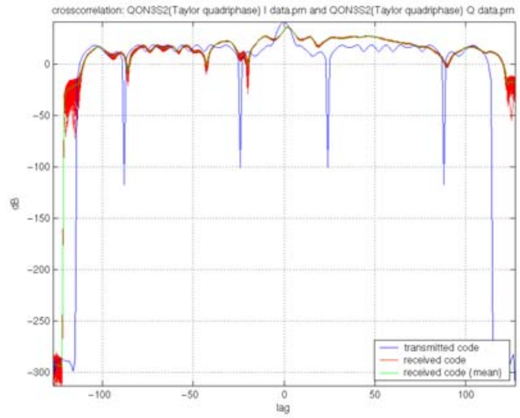


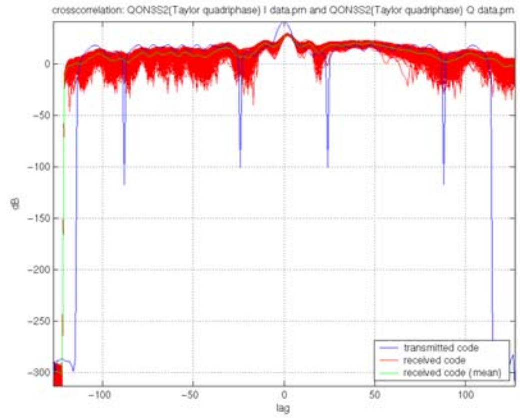
E.2.4 Test results: Quadriphase codes

The following plots show the cross correlation functions of a quadriphase code followed by a Taylor quadriphase(TQ) code, both at high and low power levels. The fault in the arbitrary waveform generator software has a particularly bad effect on these waveforms, resulting in a lower and broader correlation peak in addition to higher sidelobe levels.









E.2.5 Conclusions

These results probably contain more information about the shortcomings of the test system than the waveforms themselves. In the case of the polyphase codes, where as a result of the truncation of the waveform, the codes and orthogonal codes are no longer optimised for minimum sidelobe levels. However, the results do show that the waveforms tested behave as expected in practice and that sidelobe levels of better then -40dB can readily be achieved, even in a system that has not been optimised for long pulse lengths or very low sidelobes.

E.3 Interference from communication sources

E.3.1 Test transmitter

This was the same as the test transmitter in §E1.3, but with a second signal generator for the communications signals coupled to the antenna.

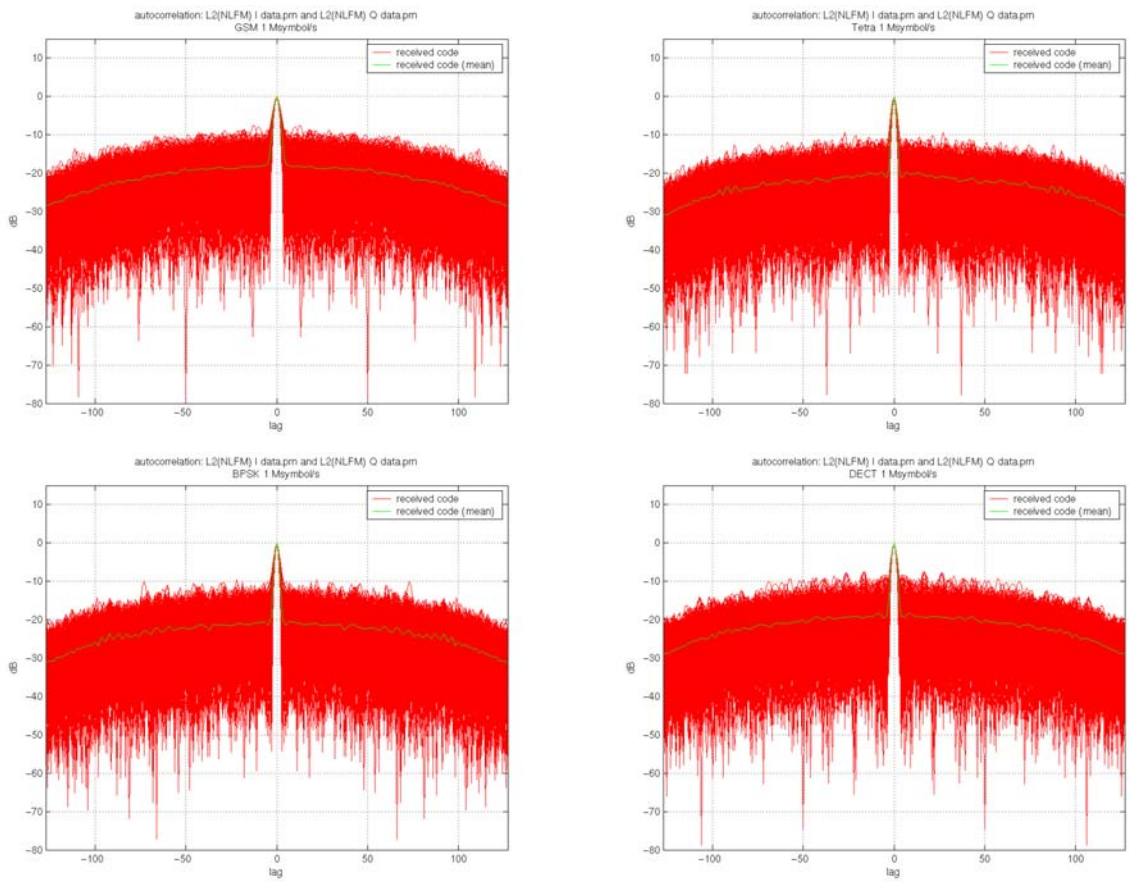
The communications signal generator is a Rhode and Schwarz SMIQ03B, programmed with a number of communications waveforms. Examples of waveforms commonly used in terrestrial transmissions were used in the tests; BPSK, GSM, TETRA and DECT, all modulated at 1Msymbol/second.

The communications code and a chirp (L2(NLFM) or polyphase code (polyphase code check_128) were transmitted simultaneously from the same antenna, both at a mean power level of +10dBm, in order to give a direct measurement of the suppression of the communications code in the receiver.

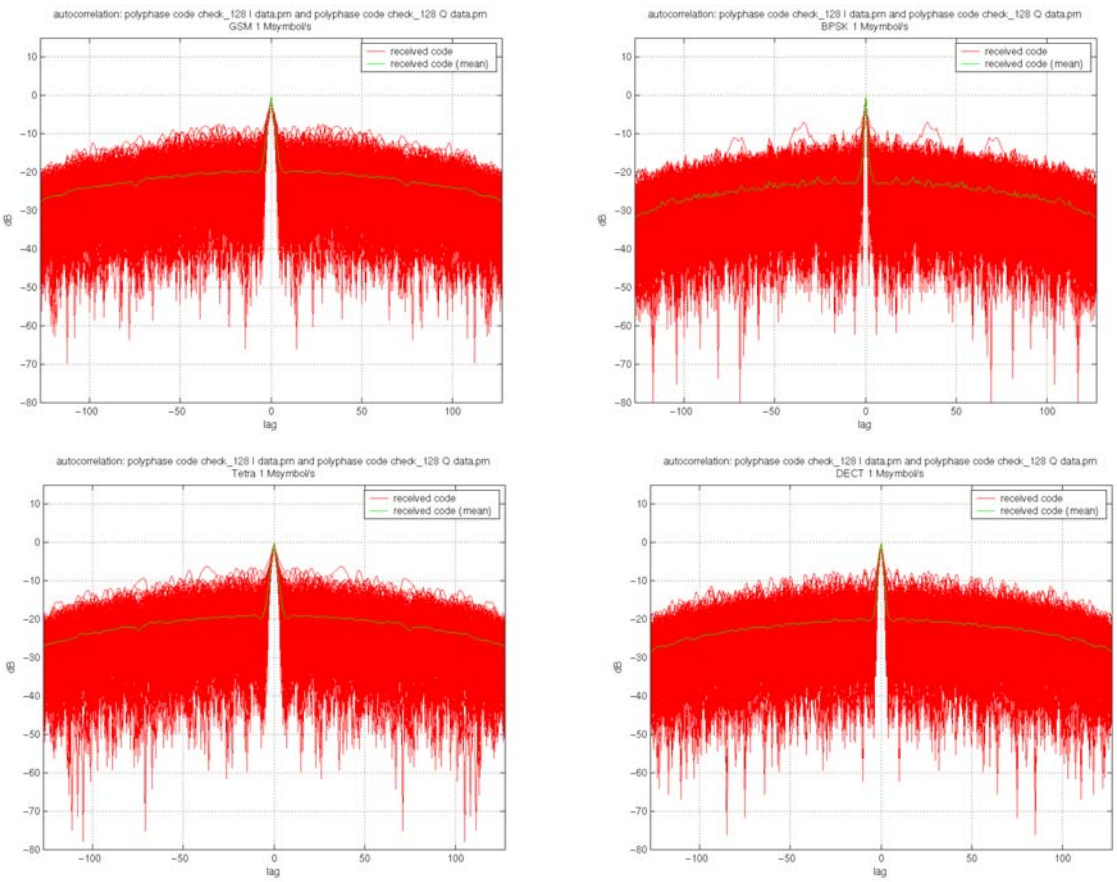
E.3.2 Test results

The following plots show the cross correlation functions of the transmitted chirp or polyphase code and the received signals consisting of the chirp or polyphase codes and a communications code at the same mean power level.

<u>Correlation of communications codes and L2(NLFM) chirp:</u>



Correlation with polyphase code check_128:



E.3.3 Conclusions

All of the communications codes tested have similar levels of suppression and there is a little structure in the sidelobes, suggesting that their effect is purely to raise the post-correlation noise floor of the receiver.

E.4 Summary of chirp and polyphase code cross-correlation properties

Table E4-1 summarises the levels of rejection achieved by correlating the received codes with reverse slope chirps or orthogonal polyphase codes and by correlating received communications codes with chirp and polyphase codes.

Transmitted code	**Correlated with**	**Mean rejection (dB)**
Radar codes:		
L1(LFM)	Reverse slope chirp	20
L10(RSLFM)	Reverse slope chirp	19
L11(WSLFM)	Reverse slope chirp	17
L2(NLFM)	Reverse slope chirp	18
LFM64	Reverse slope chirp	19
NLFM64	Reverse slope chirp	17
Polyphase code check_128	Orthogonal code	17
Polyphase code check_64	Orthogonal code	14
Polyphase code check2_128	Orthogonal code	19
Polyphase code check2_64	Orthogonal code	15
Communication codes:		
GSM	L2(NLFM)	17
TETRA	L2(NLFM)	19
BPSK	L2(NLFM)	18
DECT	L2(NLFM)	18
GSM	Polyphase code check_128	18
TETRA	Polyphase code check_128	18

Table E4-1 Summary of level of rejection by correlating various codes

E.5 Clutter illuminated by an interfering radar

The most probable source of interference from one radar to another is from clutter in a common scattering volume. To simulate this, the radar is used as the interfering transmitter and also the victim receiver; the fact that this has monostatic rather than bistatic geometry is not a significant source of error in the case of Rayleigh scattering in rain.

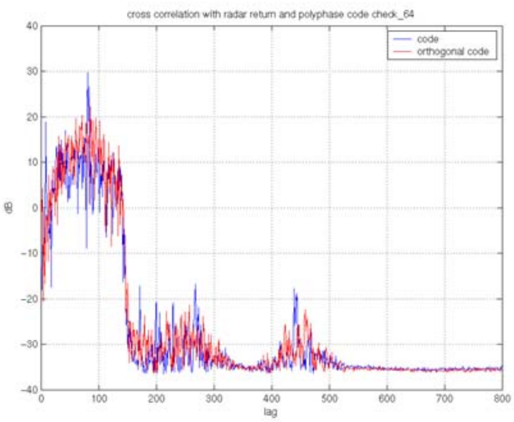
There is a limited amount of data in this set of measurements as, due to pulse length limitations in the transmitter, only 64 chip codes could be used.

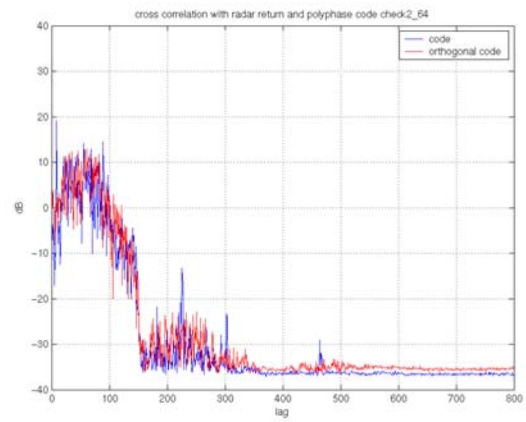
Returns from ground clutter and rain were correlated with the transmitted code and with orthogonal codes optimised for minimum cross correlation response.

In each plot, the blue trace is the cross correlation function of the return signal with the transmitted code and the red trace with the orthogonal code. The returns at a lag of 400 to 500, corresponding to a range of 24 to 3 km, are from rain, earlier returns are from ground clutter.

E.5.1 Test results

The following plots show returns from clutter and rain, correlated with 64 bit polyphase codes and the corresponding orthogonal codes. The units of lag are 0.4 ms; returns at lags up to 300 are from ground clutter and those at 400 to 500 are from rain.





E.5.2 Conclusions

The results from §E2 suggest that for point targets, the rejection of the orthogonal 64 chip codes is approximately 15dB. The results shown above suggest that this can be achieved if the return approximates to a point target. For example, the clutter at a lag of 220 in the second plot, but this does not appear to be generally the case for distributed targets.

F List of radar industry participants

Civil Aviation Authority (CAA)

National Air Traffic services (NATS)

Maritime and Coastguard Agency (MCA)

Met Office

DRS Technologies

Thales Sensors

Raytheon Systems

AMS

Sperry Marine

E2V Technologies

BSC Filters

Microwave Marketing

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Abstract This report has been prepared for Ofcom under study contract AY4490 by QinetiQ with contributions from its partners through the 2003-04 Spectrum Efficiency Scheme (SES) initiative. The primary requirement was to investigate methods that reduce interference between radars and other services and considers the practical and theoretical possibilities of sharing radar spectrum with other radio services in the frequency band 1-16GHz. The report has shown the value of the techniques investigated by applying them to exemplar bands within the spectrum and a cost benefit analysis has been produced to show the potential benefit of implementing such techniques. Recommendations have been presented which offer a strategy for taking credible options forward.			
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