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York EMC Services Ltd

Further work into the potential effect of the use of Dithered Clock Oscillators on Wideband Digital Radio Services

(Copy 1 of 5)

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LIST OF TERMS AND ABBREVIATIONS

AGL	Above Ground Level
ASIC	Application Specific Integrated Circuit
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BIOS	Basic Input Output System
CDMA	Code Division Multiple Access
COFDM	Coded Orthogonal Frequency Division Multiplexing
CW	Continuous Wave (conceptually a single frequency)
DAB	Digital Audio Broadcasting
DCO	Dithered Clock Oscillator
DTTV	Digital Terrestrial Television
DVB-T	Digital Video Broadcasting – Terrestrial
EMC	Electro-Magnetic Compatibility
EUT	Equipment Under Test
IC	Integrated Circuit
IFFT	Inverse Fast Fourier Transform
ISI	Inter-Symbol Interference
OFDM	Orthogonal Frequency Division Multiplexing
PCB	Printed Circuit Board
PAL	Phase Alternating Line
PAL-I	Version of PAL used in UK and Ireland
PCI	Peripheral Component Interconnect bus
PLL	Phase Lock Loop
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase Shift Keying
SIR	Signal to Interference Ratio
SNR	Signal to Noise Ratio

SSC	Spread Spectrum Clock equivalent term to Dithered Clock Oscillator
UKAS	United Kingdom Accreditation Service
UMTS	Universal Mobile Telecommunications System
UMTS-FDD	Universal Mobile Telecommunications System – Frequency Division Duplex
UTRA	UMTS Terrestrial Radio Access
UMTS-TDD	Universal Mobile Telecommunications System – Time Division Duplex
VCO	Voltage Controlled Oscillator
W-CDMA	Wideband - Code Division Multiple Access
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EXECUTIVE SUMMARY

Recent years have seen the manufacturers of digital electronic products increasingly utilising dithered clock oscillators (DCO) in their equipment, primarily as a means of reducing the peak levels of emissions seen by the detectors used in EMC compliance tests. The study described in this report modelled a digital broadcast receiver to determine its immunity to DCO generated interference. The modelled levels were compared with practical test results obtained from typical digital receivers.

It has been concluded that DCO enabled equipment has a smaller margin of Electromagnetic Compatibility between it and broadcast services (analogue or digital) than non-DCO enabled equipment. Furthermore it was found that the digital receivers investigated were less immune than might be expected from the modelling: their immunity to both DCO and CW interference is considered to be dependent on the tuner design. Two possible approaches to improving the EMC margin have been suggested.

1. The existing EMC limits could be reduced for DCO enabled equipment. This is highly practical in terms of being able to perform the test with existing methods and equipment. The proposed reductions would give a significant improvement to the EMC of DVB-T and DAB receivers. However, it is not known how practical it is to build IT equipment with these more stringent limits. This would have to be investigated before the proposed limits could be implemented.
2. An alternative approach to improving the EMC of both DVB-T and DAB receivers is to investigate improving the immunity of the receivers to DCO interference. The immunity levels of the receivers tested were not as good as the modelled value. This suggests that the tuner design might be improved to give superior immunity to DCO noise. If this approach proved practical then an in-band DCO immunity test would have to be developed and standardised so that receivers can be tested to ensure maximum immunity.

1 INTRODUCTION

This report investigates the effects of DCO generated interference with respect to their likely effect on digital broadcasting services. In particular DVB-T and DAB systems are investigated. The investigation involves both computer modelling and measurements on commercial DVB-T and DAB receivers.

The investigation has been divided into four work packages and the report describes each of these areas of activity in sections 2 to 5. The contents of the four sections are summarised as follows:

- Section 2 describes a literature survey. The survey researches the schemes currently being used to produce Dithered Clock Oscillators (DCOs). Based on the research, a high EMC threat scenario is identified. It is parametised for input to the modelling stage of the project described in section 3.
- Section 3 describes the modelling of a DVB-T receiver. The model establishes expected levels of susceptibility to DCO noise for DVB-T receivers. The model is also run for QPSK modulation to give an expected immunity level for a DAB receiver. To ensure the integrity of the model, susceptibility to white noise is also investigated.
- Section 4 describes practical susceptibility measurements on two DVB-T set top boxes, a PAL television (for comparison) and a DAB receiver. The receivers were subject to controlled levels of DCO and CW interference and the results observed.
- Section 5 draws together the results from sections 3 and 4 and applies them to possible 'real world' interference scenarios. This gives an indication of the circumstances in which interference is likely to occur. Idealised EMC test methods are discussed and the limitations of existing EMC measurement methods considered.

Based on section 5 the Conclusions section suggests new EMC test limits that would give significantly improved protection to digital radio services whilst being practical with existing UKAS accredited EMC test facilities.

2 LITERATURE REVIEW

This section of the report details the results of a literature survey, which has been performed with the aim of determining and characterising the schemes currently being used to produce dithered clock oscillators, in order to permit realistic simulations of the potential effects on radio services. These schemes are readily characterised by three parameters of most interest: the spreading range, the modulation rate, and the shape of the frequency against time profile of the clock signal produced.

An introduction to the concept of dithered clocks is given first, with a discussion of the various factors affecting the choice of spreading range, modulation rate and frequency-time profile used. A survey of available devices that can provide dithered clocks is then given. Finally, a set of recommendations for suitable models to use for simulating dithered clocks is given, based on common practice in the industry.

2.1 Introduction to Dithered Clocks

Dithered clock oscillators (DCOs) (more commonly known as spread-spectrum clocks (SSCs)) are most simply produced by locking a phase lock loop (PLL) to a crystal oscillator, and then adding a small modulating signal to the input of the voltage-controlled oscillator (see Figure 2.1 below). When the frequency of the oscillating voltage is greater than the bandwidth of the PLL's loop filter, but within the bandwidth of the voltage controlled oscillator (VCO), the resultant output frequency becomes modulated by the signal.

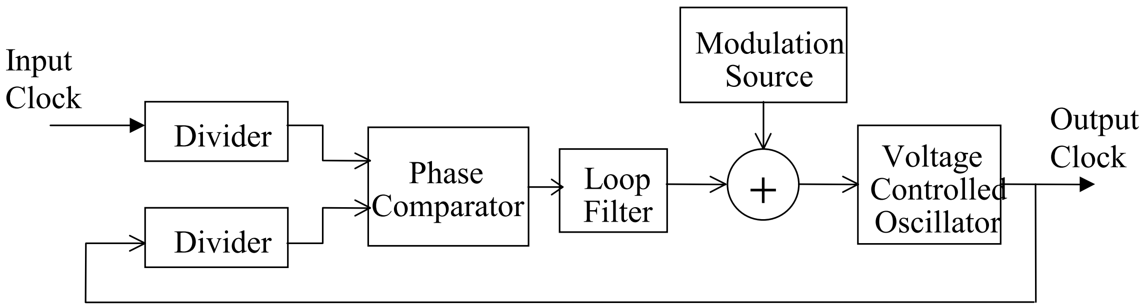


Figure 2.1: Schematic of Basic Clock Dithering PLL

Such solutions are invariably performed by integrated circuits (ICs) specifically designed for the purpose. There are two types of IC available to perform this function: one performs only the function of adding dithering to a pre-existing clock without changing the frequency of the clock (for example the P2010 from PulseCore), others include this function in an existing clock generator IC which provides a range of output clock frequencies from a single input clock (for example the CDC930 from Texas Instruments). These latter ICs are often designed to be the central clock generator chip for a complex system such as a modern personal computer (PC); the former ICs are more usually intended for lower-frequency clocks, in general digital equipment such as printers, scanners, televisions and other appliances.

Many systems also implement dithered clocks as part of application specific integrated circuits (ASICs) developed specifically for the target systems. There is much less information publicly available about the characteristics of these implementations since they are not advertised, however it can be safely assumed that they will be broadly similar to those available from dedicated ICs. ASIC implementations now account for the majority of products sold with the optimal profile (see section 2.1.3) as licensed from Lexmark.

2.1.1 Clock Spreading Ranges

Most of the ICs available provide a choice of different spreading ranges and frequency, and provide both centre-spreading (varying the output frequency around the nominal frequency) and down-spreading (varying the output frequency between the nominal frequency and some frequency below nominal, so that in this case the average frequency decreases due to the introduction of the modulation).

Neither technique is without disadvantages: using centre-spreading results in some clock periods being shorter than nominal, with the result than timing margins in the rest of the design may be squeezed, and the design may become less reliable. Down-spreading results in the average clock frequency decreasing, and hence a slight loss in system performance. Also any sub-system relying on the average system clock frequency close to a nominal frequency (for example a communications link) may become less reliable as the average deviation from the nominal frequency for the same spreading factor is increased.

As a result, system designers will typically choose the minimum amount of spreading that their system requires to pass the tests specified in relevant EMC standards, and this can only be done during the final testing phases of their products. In some cases, this is done in software as part of the BIOS set-up, although this feature is not usually accessible to the final user (it is unlikely that any final user would want to configure a piece of equipment to give either worse performance, or fail the applicable EMC standards).

This being the case, it is difficult to know what values of spreading are being used in any given application without extensive measurements of a wide range of equipment (the use of dithered clocks is very common in modern high-speed digital equipment of all descriptions); although maximum and minimum values are available from the relevant device data-sheets, and a first estimate might suggest that an average of such parameters would be the mean value available from each device. Another clue is available from the fact that Intel themselves give examples using a total range of modulation of both 0.5% and 1.0% [1].

However, a good indication of common spreading ranges can be deduced from those ICs that only offer one choice of modulation depth, and in this case, a value of +0/-0.5% might be regarded as typical, at least for the case of personal computers. (For the particular case of personal computers, down-spreading is more common than centre-spreading, due to the tight timing requirements specified by the Intel processors in most of these computers.) Alternatively, a figure of ±0.5% is suggested by at least one manufacturer of variable-spreading clock generators for PCs in their data and application sheets as the most common used by their customers [2], and other reports indicate that this is a common range for personal computers [3].

For other devices which do not include such tight timing margins, or have other phase lock loops trying to track the output of the system clock (as is the case with all modern PCs), a wider range of frequency deviation can be used, up to 3% in some cases. When there is no performance implications for these wider frequency ranges, there is often no reason not to employ them, as the wider the spreading range used, the greater the advantage in terms of lowering the peak emissions around the clock frequencies.

For the purposes of investigating the emissions resulting from systems using dithered clocks, the choice of centre-spreading or down-spreading is not usually important, since the nominal frequency of the system clock is not known in advance; this is particularly true for the case of systems other than personal computers, which can use a very wide range of different clock frequencies.

2.1.2 Clock Modulation Rates

Several of the currently available solutions also allow a variable modulation rate. Again the choice of ideal modulation rate is a ‘trade-off’. Too slow a modulation rate (in the audio band) can results in audible interference due to demodulation by FM radio receivers. Too high a modulation rate, and the cycle-to-cycle jitter (the difference between the period of two consecutive clock waveforms) can become too great. This is a key parameter in the specification of the input clocks of many current high-performance microprocessor systems. This is particularly true of systems employing chips with PLLs included in the processor chip to allow a faster internal clock speed than the system clock can provide: this includes almost all current personal computers. Indeed it has been claimed that it is impossible to use any spread-spectrum clock without invalidating the specification of the PCI-bus.

Most manufacturers recommend (or only provide) a modulation frequency slightly above the audio spectrum, typically between 20 kHz and 50 kHz, with 31.25 kHz being a commonly chosen modulation rate. This is safely above the frequency that can result in audio interference in frequency-modulated radio systems, while still being low enough to allow other phase lock loops in system designs to maintain lock with the system clock. The only exception to this rule is the Fairchild IC, which has been designed with a much lower modulation rate (3 kHz) in order to limit the cycle-to-cycle jitter in the output dithered clock [4].

2.1.3 Modulation Profiles

The shape of the modulation profile (the curves of frequency against time) also varies between manufacturers, with two basic schemes being widely used: a triangular or step-wise triangular profile, and an optimised profile known variously as the "Lexmark shape" (after the company that invented it), also known as a "Hershey Kiss" after the sweet, or the "optimal" or "non-linear" solution, see Figure 2.2. (The first experiments in spread-spectrum clocks were made using sinusoidal frequency modulation, but this was soon found not to be optimal [5].)

The Lexmark shape is optimal in the sense that it results in a frequency spectrum for the spread clock (as measured by a standard detector as used for EMC compliance measurements) which is almost flat across the frequency ranges occupied by the dithered clock; and thus results in the maximum reduction of unwanted emissions for any given spreading rate and range [6]. It has the form (for the first quarter of the waveform) of:

$$Frequency\ Deviation = k\left(0.45\,(4t)^3 + 0.55\,(4t)\right) \qquad 0 \le t < 0.25$$

Equation 2.1

where k is a constant related to the range, and the time t is normalised so that an entire cycle of the modulating waveform lasts from t = 0 to t = 1. The remainder of the shape can be generated by reflecting this quarter-wave [7].

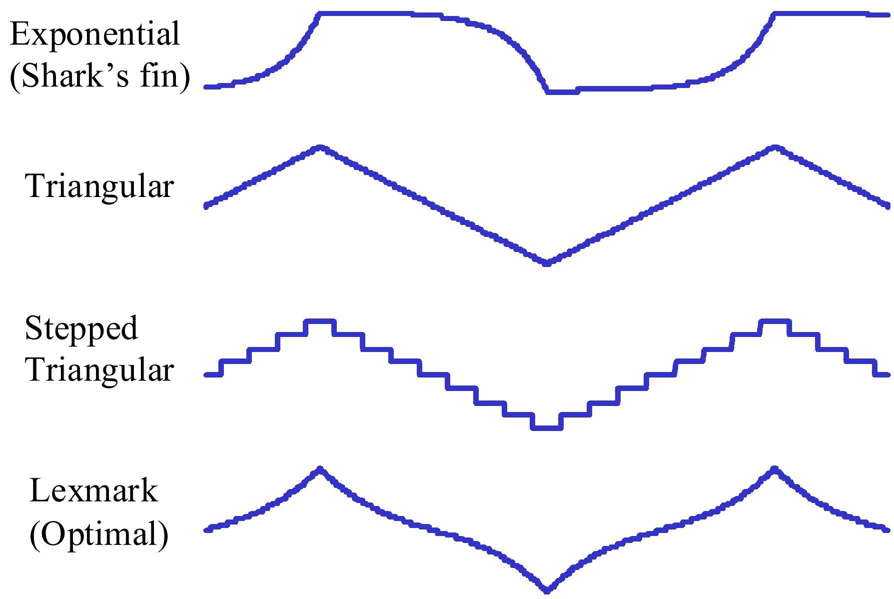


Figure 2.2: Modulation Profile Waveforms

Use of this Lexmark shape requires licensing from Lexmark [7], as Lexmark holds a patent for this optimal modulation profile, whereas use of any of the three sub-optimal profiles does not: this accounts for the popularity of the exponential, triangular and stepped-triangular shape with some manufacturers. The difference in reduction of emitted radiation available from the use of the optimal profile is dependent on the application and the harmonic number, but has been estimated to be between 2-3 dB [8].

The exponential profile is the simplest profile to generate: a capacitor can be charged and discharged repeatedly from a fixed voltage reference, adding an offset to the input to the VCO, which varies the output frequency from the loop. This is not the most controllable method for producing a fixed frequency deviation however, since any variation in the voltage to frequency characteristic of the VCO will result in a variation in the modulation depth between individual ICs. It also provides the least gain in terms of reduction of the size of emission peaks, due to the comparatively long times the output frequency remains at the extremes of the frequency ranges.

The triangular profile provides better performance, and can be conveniently generated by charging and discharging the capacitor from a current source rather than a voltage source; however this method also suffers from the variability caused by manufacturing tolerances in the voltage-to-frequency characteristics of the VCO.

The stepped triangular waveform is usually generated by a slightly different method to that shown in Figure 2.1. Another way to modify the output frequency is to step the division ratio of the clock-divider in the PLL through several consecutive values: this provides a very accurate and repeatable modulation depth immune to variations in the characteristics of individual ICs [9]. However the output frequency can now change between several discrete step frequencies, rather than being smoothly varied. Careful selection of the bandwidth of the loop filter and the number of different discrete frequencies used can help to ameliorate this effect, however.

Use of the optimum "Lexmark" shape requires the payment of a license fee to Lexmark; this is the main reason that this shape is not universally used. Although most manufacturers of ICs do use this shape, the largest volume manufacturer of standard clock generation chips for personal computers in the world (ICS) does not use this shape, but instead uses a stepped-triangular shape, with eight steps, as shown in Figure 2.2 [10].

2.2 Survey of Integration Solutions

A survey was performed of vendors who offer standard ICs for producing dithered clocks, and the results are shown in summary form in Table 2.1. This table includes most of the clock suppliers for personal computers: although it should be noted that it is likely that a majority of the systems that currently use dithered clocks generate their clocks through the use of mixed-signal ASICs, and these are not reflected in Table 2.1.

(In those cases where there is insufficient data in the data sheets and the suppliers have not responded to requests to supply further information, the entries have been marked with a "???")

Manufacturer	*Example Part Number*	*Spreading Waveform*	*Spreading Range*	*Spreading Frequency*	*Centre/Down Spreading?*
American Microsystems	FS6251	Lexmark	+0/-0.5%	31.25 kHz	Down only
Cypress	CY25568	Lexmark	±0.3% to ±1.4% variable +0/-.5% to +0/-3% variable	31.25-62.5 kHz, dependent on freq in.	Both available
Integrated Circuit Systems	ICS9148	Triangular	±0.5% or ±1.5% +0/-1% or +0/-3%	50 kHz	Both available
Fairchild	RC7100	Triangular	+0/-.5%	3 kHz	Down only
Philips	PCK2010	???	±0.6% or +0/-0.6%	???	Both available
PulseCore	P2160	Lexmark	±0.3% to ±2.5%, variable	20-72 kHz, variable	Centre only
Texas Instruments	CDC930	???	+0/-0.6% fixed	???	Down only

Table 2.1 - Manufacturers of Spread-Spectrum Clock Generator ICs

Integrated Circuit Systems claim to be the world's largest volume supplier of PC clock generator ICs [11], they produce ICs that produce the entire range of clocking frequencies required by most modern computer systems; and they suggest that most of their customers use +/- 0.5% centre-spreading, with a triangular modulation profile, and 50 kHz modulation rate. The spread spectrum options in these chips are somewhat limited, however.

Cypress produce two ranges of ICs, due to acquisitions of two companies that pioneered the production of spread-spectrum clock generators: IMI and ICworks. Cypress chips use the optimal Lexmark profile and provide a wide range of modulation depths both centre and down-spread.

PulseCore specialise in producing small, low-cost devices that spread in frequency an existing clock, without changing the fundamental frequency. A range of devices is available to accommodate frequencies from 10 to 166 MHz. All their chips are available with centre-spreading only, and provide a rate of spreading related to the clock frequency being used. It is also interesting to note that PulseCore produce a software tool that can estimate the degree of emissions reduction through the use of their chips: "EMI-lator" [12].

The other manufactures mostly produce standard clock generator circuits for Intel PC-based applications, and have added limited spread-spectrum capability into existing clock generator designs.

2.3 Recommendations for Modelling Systems

Personal computers almost all use a frequency spread of between +0/-0.5% (down-spreading only), with a modulation rate of around 31.25 kHz; however the modulation profile is commonly either a stepped triangular shape, or the Lexmark (optimal) shape. Other systems with less strict requirements on their system clocks (printers, televisions, etc) often use much wider spreading ranges of up to 3%, but often use similar modulation rates, and again either the simpler and cheaper stepped triangular or exponential shapes, or the optimal Lexmark shape are in common use.

In one sense, the use of the optimum profile can be regarded as the worst case from the point of view of simulating the potential effects of dithered clocks. Since this profile produces the flattest frequency spectrum for the dithered clock, it can provide the greatest reduction in emissions at any single frequency (with a 120 kHz measurement bandwidth typically used for EMC measurements above 30MHz). This margin can then be used for cost-reduction methods, such as the use of only double-sided printed circuit boards (PCBs) or reduced screening in the enclosures used. Assuming that these cost saving have been maximised, then the EMC limits will be approached over a wider range of frequencies with the use of the Lexmark shape than with any other modulation profile.

For a worst-case analysis, therefore, it is recommended that the Lexmark (optimal) shape should be assumed; along with a 31.25 kHz modulation rate in all cases; and a +0/-0.5% frequency range for personal computer use, and +0/-3% frequency range for all other applications.

3 THEORETICAL ANALYSIS AND COMPUTER MODELLING

The main focus of this work is the influence of interference from dithered clock oscillators (DCO) on digital terrestrial broadcast systems. The main systems involved are Digital Audio Broadcasting (DAB) [13] and Digital Video Broadcasting - Terrestrial (DVB-T) [14], both of which employ Coded Orthogonal Frequency Division Multiplexing (COFDM), and for this reason in this section COFDM is considered in the context of DCO interference. The main principles of COFDM are introduced first, and the effect of interference in general, and then DCO interference in particular.

To back up the more general theoretical analysis of this, a computer simulation model of a DVB-T system in the presence of both white Gaussian noise and DCO interference has been considered. The strategy behind this model is describedand give some results given. Finally the likely effect of DCO interference is discussed with respect to some other possible victim systems, notably those based on Code Division Multiple Access (CDMA), including third generation mobile systems such as UMTS [15], since different principles apply here from both COFDM and existing narrow-band systems.

3.1 Introduction to COFDM

COFDM is designed to overcome several degradations that occur on the wireless channel, and thereby to create a transmission format that is highly robust, while retaining optimum efficiency. This is particularly important in broadcast systems where reception may be required in poor conditions, and there is no two-way link to ensure correct reception. The main degradations involved are frequency-selective fading and dispersion due to multipath, and narrow-band interference. The multipath effects are considered first: interference is considered in the next sub-section.

Multipath is a general problem in wireless systems, occurring wherever multiple transmission paths exist due to reflections, etc, of the radio signal. These interfere with one another to cause fading, which may be frequency-selective because the exact phase relation between the multipath components depends on the transmission frequency. Some components may suffer a time delay relative to others, and this causes dispersion of the transmitted signal, which may in turn give rise to intersymbol interference (i.s.i). In many wireless systems this is the factor that limits the maximum data rate. The rate of several Mbit/s required for digital broadcasting, and especially digital video broadcasting (DVB), is not achievable on the terrestrial broadcast channel with conventional single carrier transmission. An effect very similar to multipath is created by quasi-synchronous transmission, in which, in order to save transmission channels, the same signal is transmitted on the same frequency at several adjacent broadcast transmitters. This, however, is not currently implemented in the U.K.

The principle of OFDM [16,17] is to split the data stream into parallel sub-channels, each of which is modulated onto one of a set of carriers, chosen so that the modulated signals are orthogonal to one another. In this way the symbol rate on each sub-channel is greatly reduced, and hence the effect of intersymbol interference (i.s.i.) due to channel dispersion is reduced. This is illustrated in Figure 3.1, which shows the increased symbol period on each sub-channel, which will reduce the intersymbol interference caused by the dispersion. Figure 3.2 shows the conceptual structure of the transmitter and receiver, although in practice these would be implemented using fast Fourier transforms, rather than multiple separate modulator/demodulators.

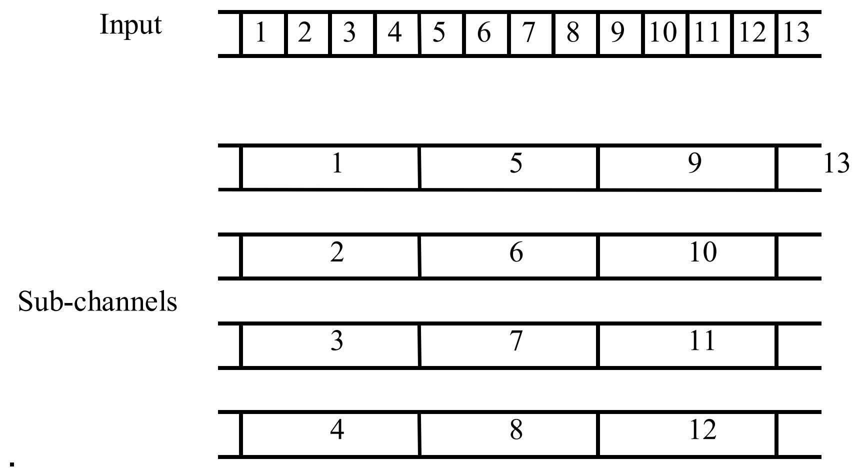


Figure 3.1: Principle of OFDM: division into sub-channels

If the channel spacing is exactly equal to the symbol period, the sub-channels will be *orthogonal* to one another, which allows the signals to be completely separated from one another at the receiver. Figure 3.3 shows the spectrum of the sub-channel signals. Note how the frequency of one sub-channel falls into the spectral null of the adjacent channels (and all others), which maintains orthogonality despite the substantial overlap between adjacent spectra. This obviates the need for guard bands between the sub-channels, and means that the overall spectrum efficiency remains high.

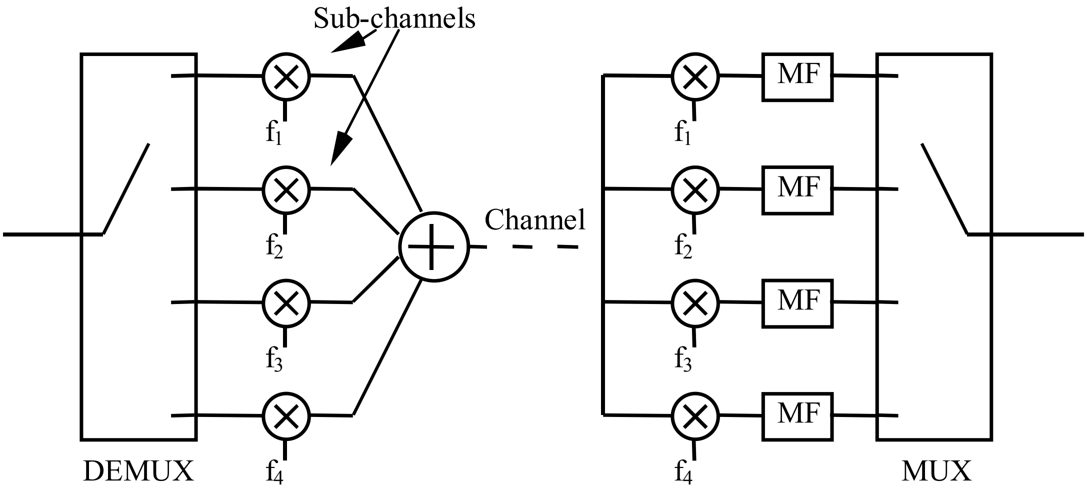


Figure 3.2: Conceptual structure of OFDM transmitter/receiver

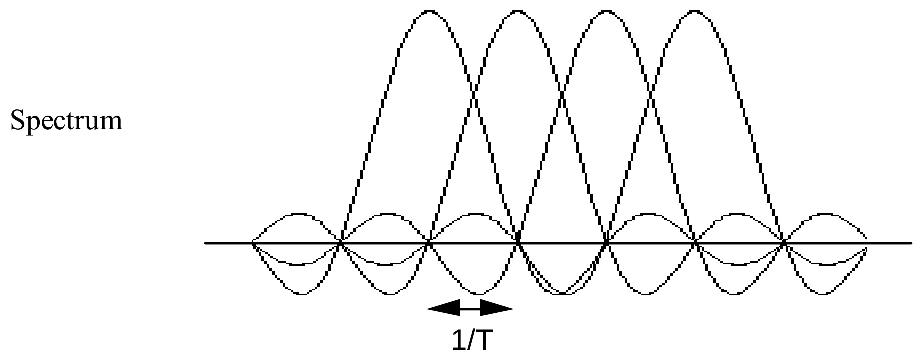


Figure 3.3: Spectrum of sub-channels

Another important feature of OFDM is the *cyclic extension*, or *guard period*, which enables the receiver to reject multipath dispersion. This is illustrated in Figure 3.4(a), which shows the signal for two adjacent sub-channels. Over the first period, of length T, the signals are orthogonal (there is an integer number of cycles). An extension of length T_g is then added which is a cyclic repetition of the beginning of the symbol, so that the signal is continuous. This means that the signals are orthogonal over any period T within the extended symbol period. Figure 3.4(b) shows the situation at the receiver in the presence of multipath dispersion. The receiver applies a window of length T at the end of the symbol period, as shown. Provided any delayed component is delayed by less than T_g, it will remain orthogonal to all other sub-channels, and so multipath dispersion will have no effect.

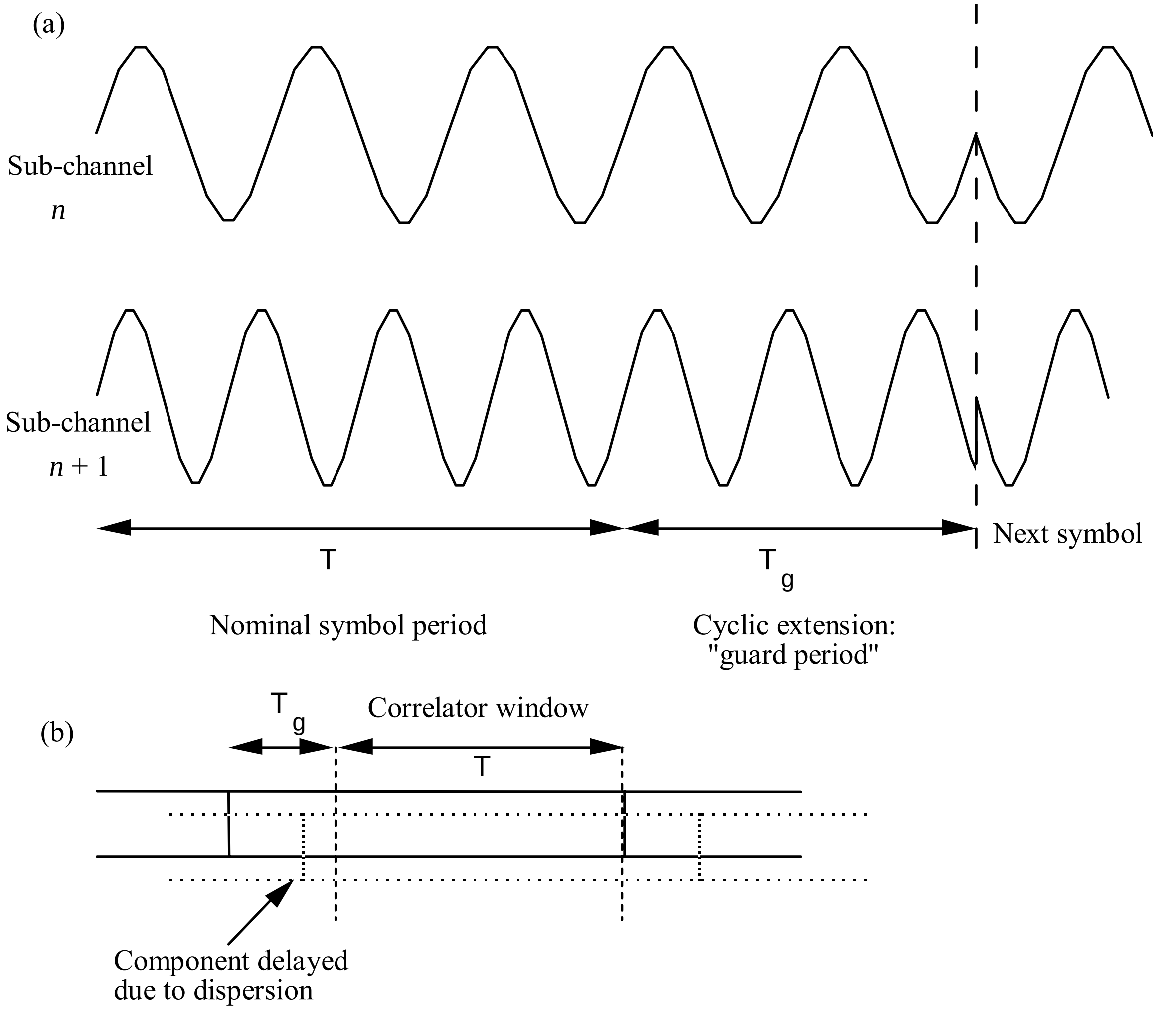


Figure 3.4: Cyclic extension and effect of multipath dispersion

Even with the guard period, multipath will still cause narrow-band fading, because the signals in some sub-channels will be received in opposite phase and will therefore partially cancel. This is overcome in digital broadcast systems by the use of error-correcting coding, along with interleaving to distribute the effect of fading through the code stream. Hence the use of the term Coded OFDM (COFDM) above. Where multilevel modulation schemes are used, which transmit several bits on one modulation symbol (as in DVB-T), the code bits are also interleaved across different symbols, so that adjacent code bits are not transmitted on the same symbol. Figure 3.5 shows a possible scheme for illustrative purposes only (it does not resemble a scheme used in any digital broadcast system) in which QPSK is used, transmitting two code bits/symbol. The squares on the diagram represent symbols on each sub-channel: the two triangles into which they are divided represent the two bits transmitted on each. The numbers are index numbers in the code stream. It is observed that no two adjacent bits in the code are transmitted on the same symbol, nor on adjacent sub-channels. This enables the system to tolerate fading of several sub-channels without loss of data. It can be characterised as a diversity advantage, with diversity order approximately equal to the minimum Hamming distance of the code.

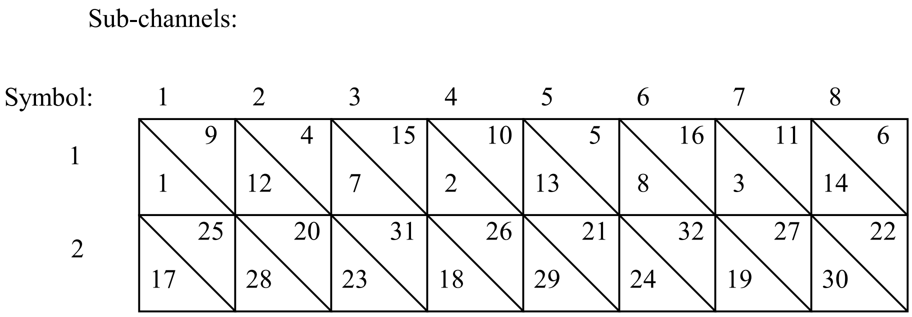


Figure 3.5: Illustrative interleaving scheme for digital broadcasting using QPSK

For optimum reception the receiver requires knowledge of the frequency response of the channel. This is firstly because multipath can cause phase errors as well as fading, which will cause significant errors in the demodulator if not corrected. Secondly, knowledge of the fading may allow the decoder to ignore (treat as *erasures*) the code bits received in sub-channels with very low signal to noise ratio, which are more likely to be erroneous. This can improve the error-correcting power of the code. In digital broadcast systems this information is obtained by transmission of known *pilot symbols* within the OFDM multiplex. For example, Figure 3.6 shows the distribution of pilot symbols in the DVB-T multiplex.

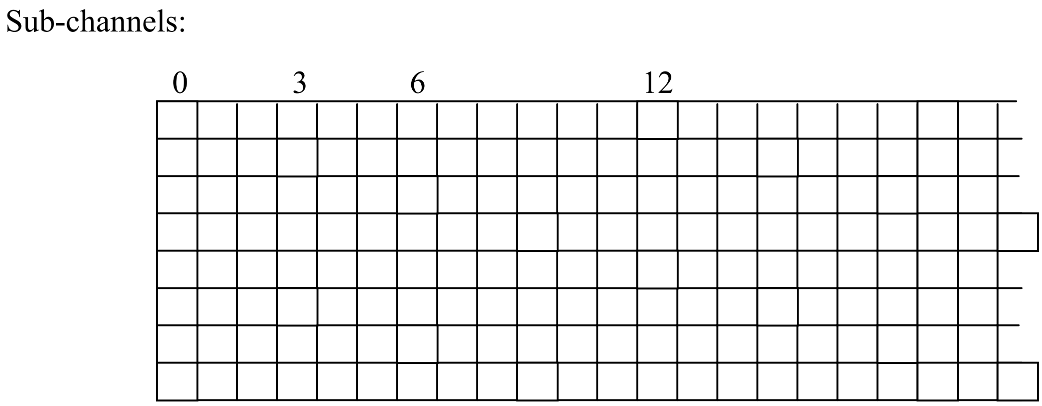


Figure 3.6: Pilot symbol distribution in DVB-T (pilot symbols shown shaded)

3.2 Effect of narrow-band and wideband interference

The properties of COFDM that make it robust to narrow-band fading are also effective against narrow-band interference. Narrow-band interference which affects only a few sub-channels will cause only a few errors in the code, which will be widely distributed by the de-interleaver and can therefore be corrected.

In fact the DVB-T interleaver distributes bits transmitted on the same symbol so they are at least 120 places apart in the code stream, and therefore will not affect one another. Hence it should be able to cope with interference in at least $d_{min}/2$ sub-channels, where d_{min} is the minimum distance of the code. For example, for the rate 2/3 64-state convolutional code used for DVB-T in the U.K. the minimum distance is 6, and so it will always cope with 3 sub-channels affected by interference.

In practice it will usually cope with many more than this, because most combinations of interferers will affect widely spaced code bits. If it is assumed that the interleaver scatters the bits affected by the interferers randomly, and that the interference affects N_{int} out of the total N_c carriers, then the probability of code bit error is approximately:

$$p = \frac{N_{int}}{N_c}$$

Equation 3.1

Using the method of [18] together with Table 7.4 of [19], we can approximate the error probability of the decoded data as:

$$P_e = \frac{1}{2}\left(3p^3 + 422p^4 + 1705p^5 - 6803p^6 + 10208p^7 - 4819p^8 + \ldots\right)$$

Equation 3.2

Solving Equation 3.2 to find p to give the maximum allowed error probability at the input of the outer decoder, namely $P_e = 2 \times 10^{-4}$, we obtain $p = 0.029$. For $N_c = 1512$ (as in the '2k' version of DVB-T used in the U.K.) this gives $N_{int} = 44$: i.e. 44 sub-channels may be affected before the system is likely to fail.

Note that the power of these narrow-band interferers could in principle be unlimited: provided only a limited number of sub-channels is affected, the interference power will have no further effect. Note also that the above figure of 44 channels is only an average: in some cases fewer than this may cause errors, as for example if a set of sub-channels were affected whose data happened to interleave to positions adjacent in the code stream.

If the receiver has knowledge of the frequencies of narrow-band interferers, sub-channels affected can be erased in the same way as described above for the case of narrow-band fading, with concomitant performance improvement. This information could in principle be obtained from the pilot symbols, but since the effect of interference would be to increase the signal in the pilot symbol, in contrast to the effect of fading, a different algorithm would be required to implement this. Hence we have not considered this further in this report.

If, on the other hand, the interference is wideband, affecting a larger number of sub-channels than we have suggested above, or the whole multiplex, the situation is different. In this case the interference is probably best treated as an additional noise floor, reducing the signal to noise ratio in a substantial proportion of the sub-channels. Errors will then begin to occur if the signal to noise ratio per sub-channel drops below the level required to maintain a sufficiently low BER at the output of the inner decoder.

For example, suppose that an interferer has total power P_{int} and bandwidth W_{int}. For an OFDM symbol period (excluding guard interval) T the effective noise bandwidth is approximately $1/T$, and hence the signal to noise ratio per sub-channel:

$$SNR_{sc} = \frac{\frac{S}{N_c}}{P_{int}\frac{1/T}{W_{int}}} = \frac{STW_{int}}{P_{int}N_c}$$

Equation 3.3

where S here is the total signal power of the OFDM signal. For a system using QPSK with the same rate 2/3 64-state convolutional code as used above, the decoded BER for given SNR can be estimated using the technique described in [19]. We adapt equation (7.9) from [19]:

$$P_b = \sum_{d=d_{free}}^{d_{max}} e(d)Q\left(\sqrt{d\,SNR_{sc}}\right)$$

Equation 3.4

where $e(d)$ is the *error-weighted distance spectrum* obtained from Table 7.4, along with d_{free} and d_{max}. Again solving this for $P_b = 2 \times 10^{-4}$ we find $SNR_{sc} = 5.1$ dB. This can be used along with Equation 3.3 to find the maximum signal to interference power for a given interferer bandwidth - although it should be noted that DVB-T uses 64-QAM rather than QPSK.

A further possible effect of interference follows from the use of pilot symbols, as noted above. If the interference affects these, it may cause the channel to be wrongly estimated, which could result in erroneous reception. However this will not be considered further here, since its effects will depend on the precise algorithm used for channel estimation. Moreover, its effects are not likely to be serious until the signal to noise ratio in the pilot symbol is quite low, at around the level at which the BER is in any case likely to be excessive.

3.3 Effect of DCO Interference

The nature of the interference generated by dithered clock oscillators (DCO) has been described in section 2. There are a variety of types of DCO in use, but they have quite similar characteristics, in which the clock frequency is swept up and down over a frequency range close to, but usually below, the nominal clock frequency. The profile of the frequency sweep is usually something close to a triangular wave. In section 2 the so-called "Lexmark" profile was identified as a suitable typical profile on which to base further investigation. It is designed to yield a spectrum as flat-topped as possible, and so can be considered as the ultimate example of a dithered clock signal. Any other profile is likely to yield results nearer to the conventional non-dithered clock.

For half the cycle the frequency of the Lexmark profile is given by:

$$f(t) = f_c + k\, f_{rep}\left(0.55\left(4t\, f_{rep}\right) + 0.45\left(4t\, f_{rep}\right)^3\right) \quad -0.25 < t\, f_{rep} \leq 0.25$$

Equation 3.5

where f_c is the centre frequency (which may not be the nominal clock frequency if the dithering is downward only), $k\, f_{rep}$ is the peak frequency deviation, and f_{rep} is the repetition frequency of the dithering profile. For the other half cycle the profile is the mirror image of this, descending in frequency, as shown in Figure 3.7.

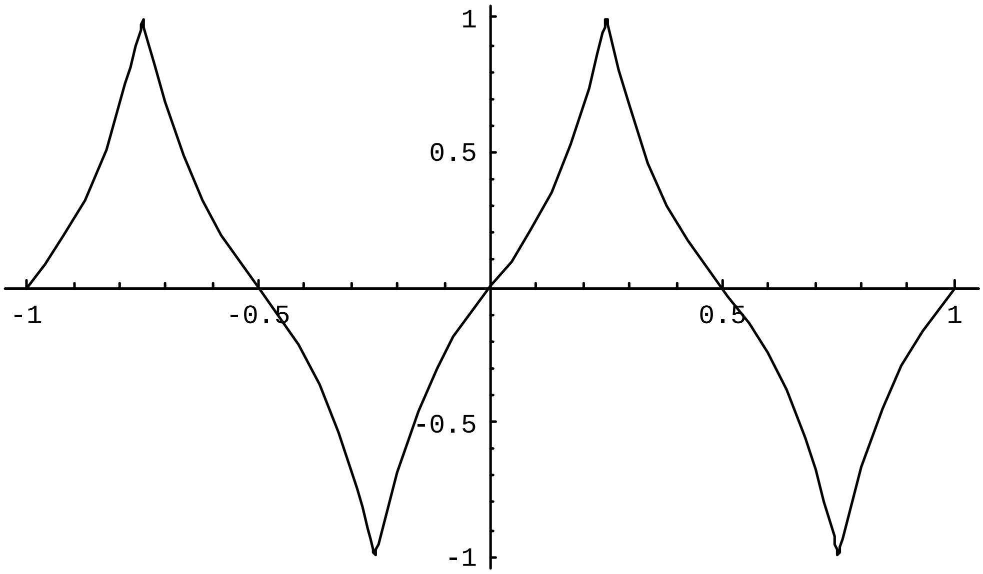


Figure 3.7: "Lexmark" frequency profile, for k = f_{rep} = 1

The DCO signal can then be modelled as a sine wave phase modulated with phase given, for one half cycle, by:

$$\phi(t) = \int f(t)\, dt = 2k\pi\left(1.1\left(t\, f_{rep}\right)^2 + 7.2\left(t\, f_{rep}\right)^4 - 0.096875\right) \quad -0.25 < t\, f_{rep} \leq 0.25$$

Equation 3.6

For the other half cycle the phase is the inverse of this.

Section 2 has recommended the use of 31.25 kHz as a typical value for f_{rep}. (It is noted below that the precise value does not have a very great effect). Note however, that for one manufacturer (Fairchild) the dithering frequency was much less than this, and might result in different effects, as we will consider briefly below. For the maximum frequency deviation, there are two recommended cases: PCs tend to use a clock variation of +0/-0.5%, while other devices (which are less sensitive to clock variations) use +0/-3%. Clock frequencies and harmonics thereof are considered that fall within the broadcast bands, and focus, for example, on the upper end of the TV broadcast band, say around 800 MHz. At this frequency a variation of 0.5% corresponds to 4 MHz (narrower than a DVB-T multiplex), while 3% corresponds to 24 MHz (much wider). Note that although the clock waveform itself is not a sine wave, here we are concerned only with the harmonic that falls within the signal band, which we model as a sine wave.

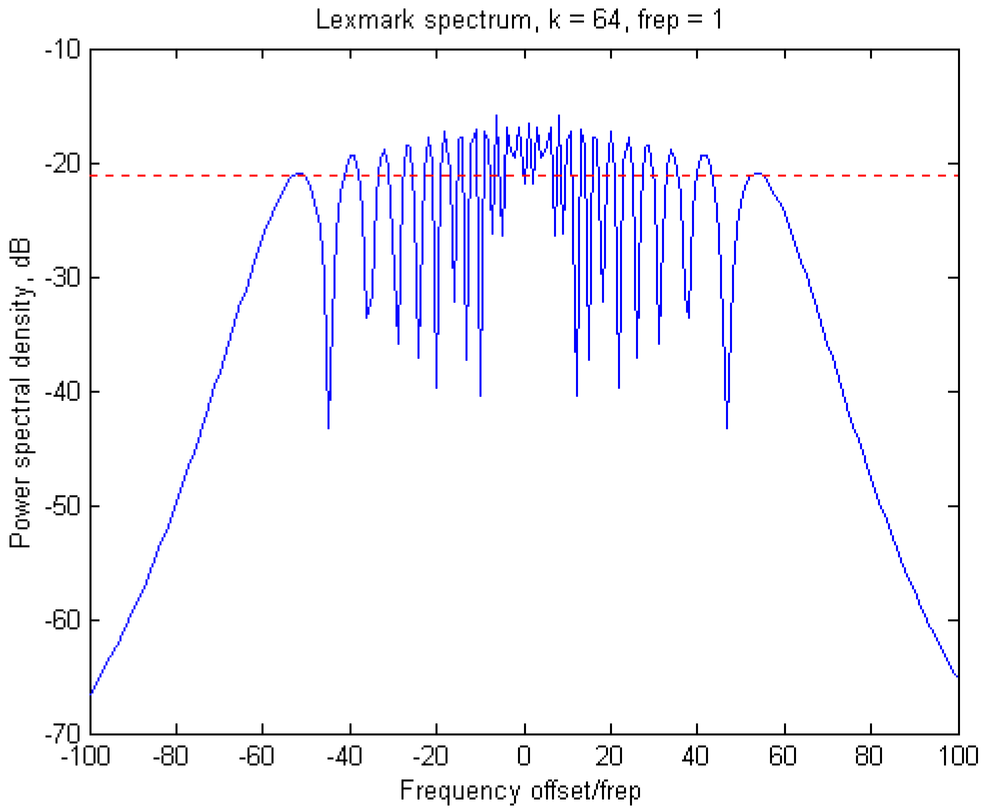


Figure 3.8: Spectrum of "Lexmark" frequency profile, for k = 64 and f_{rep} = 1

Figure 3.8 shows the spectrum of the clock signal for k = 64, shown as an offset from the centre frequency. This corresponds to the narrower bandwidth signal (4 MHz) discussed above. In the plot the repetition frequency is taken as unity, which does not affect the shape of the spectrum, but the frequency axis should be scaled by the factor f_{rep} to give an accurate indication of the bandwidth. The y axis is power spectral density in dB relative to $1/f_{rep}$ W/Hz, and should similarly be scaled to reflect the actual repetition frequency. The dotted line represents the power spectral density of a signal with uniform power spectral density over the frequency range $-k$ to $+k$: we observe that the DCO spectrum is quite close to this over that range, and that its bandwidth is close to $2k$.

The DCO signal is thus evidently a wideband signal in the terms of the previous section. The narrowest bandwidth signal will encompass many more than 44 sub-channels of the OFDM multiplex. Thus we are concerned with the interference power in each sub-channel, and should ensure that it is low enough to allow an adequate signal to noise ratio.

An important feature of the spectrum is, however, not obvious in Figure 3.8. Since the DCO signal is repetitive, its spectrum must be a line spectrum, with harmonics at multiples of its repetition frequency. (In Figure 3.8 the spectrum has been calculated from one clock cycle, and so the lines do not appear). Since the line spacing, f_{rep}, is significantly greater than the OFDM symbol rate, sub-channels near these harmonic frequencies will suffer much more interference than others. However, the number of sub-channels affected by this greater interference level is still likely to be many more than 44, so the conclusion of the previous paragraph still applies.

The ratio of the repetition frequency of the DCO signal and the OFDM symbol rate is also important. If the clock frequency varies slowly, it might affect only a few sub-channels in any one symbol period. However, the OFDM symbol rate is around 4 kHz (nominal symbol period T = 224 μs for DVB-T), while the DCO repetition frequency is 31.25 kHz, and hence it is to be expected that all sub-channels will be affected in all symbol periods. This will apply even for the Fairchild clock, whose repetition frequency is 3 kHz.

Since the interference is wideband, we might use the approximation of Equation 3.3, treating the interference as white Gaussian noise. This, however, would ignore the effect of lines in the DCO spectrum, as described above. It is instructive, therefore, to calculate the signal in a typical OFDM sub-channel due to the DCO signal. The OFDM demodulator calculates the correlation of the received signal with the reference carrier over the OFDM symbol period, as shown in Figure 3.2. Then the demodulator output for a sub-channel at frequency f:

$$y(f) = \int_{t_0}^{T+t_0} x(t)\exp(j2\pi f t)dt$$

Equation 3.7

where $x(t)$ is the signal at the input to the demodulator, and t_0 is the timing offset of the start of the symbol period. This will be random in relation to the period of the DCO signal, as will the phase of the reference signal. Now for the DCO repetition frequency it has been chosen to model, f_{rep} = 31.25 kHz, the OFDM symbol period T = 224 μs happens to be an integer number of DCO cycles. In this case the demodulated signal $y(f)$ is independent of t_0, as the integral always includes an integer number of cycles. Since this is a special case, and this relationship cannot be guaranteed to always apply accurately, a slightly different value of f_{rep}, which gives $T f_{rep}$ = 6.7 cycles, will be considered.

Figure 3.9 shows the signal at the OFDM demodulator in each sub-channel due to a DCO signal. The line structure of the DCO signal is clearly visible, resulting in interference which is small in most sub-channels, but much larger where a sub-channel frequency coincides with a spectrum line. Figure 3.10 shows the variation with t_0 more clearly, by plotting sub-channel power against t_0 for eight sub-channels, which coincide with spectral lines.

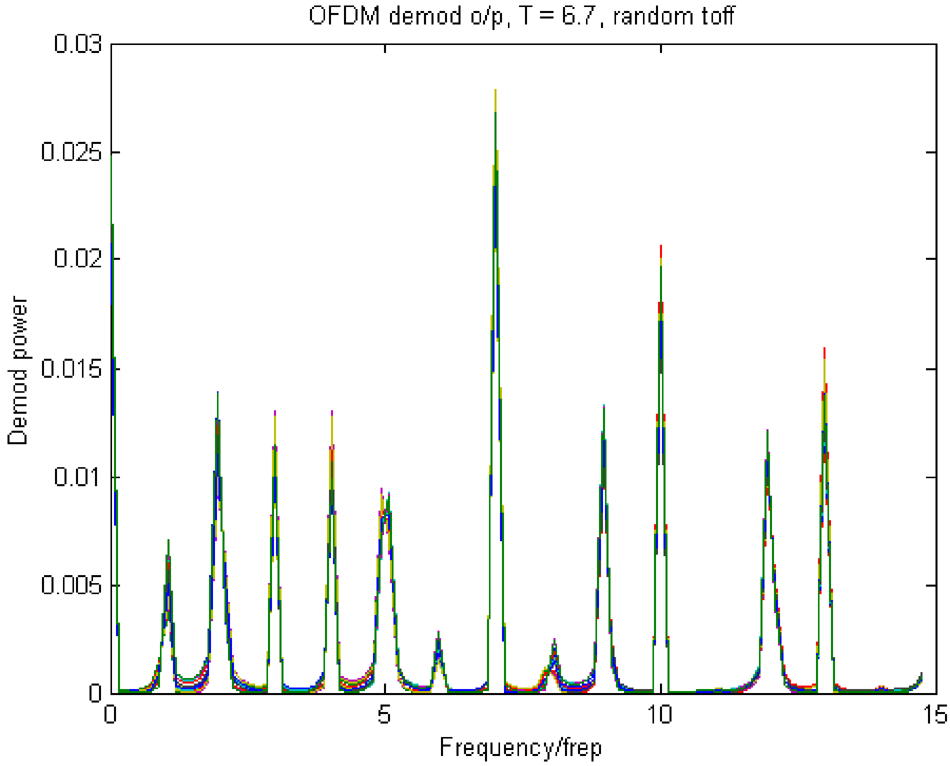


Figure 3.9: Demodulated signal in OFDM sub-channels, for 16 randomly-chosen time offsets

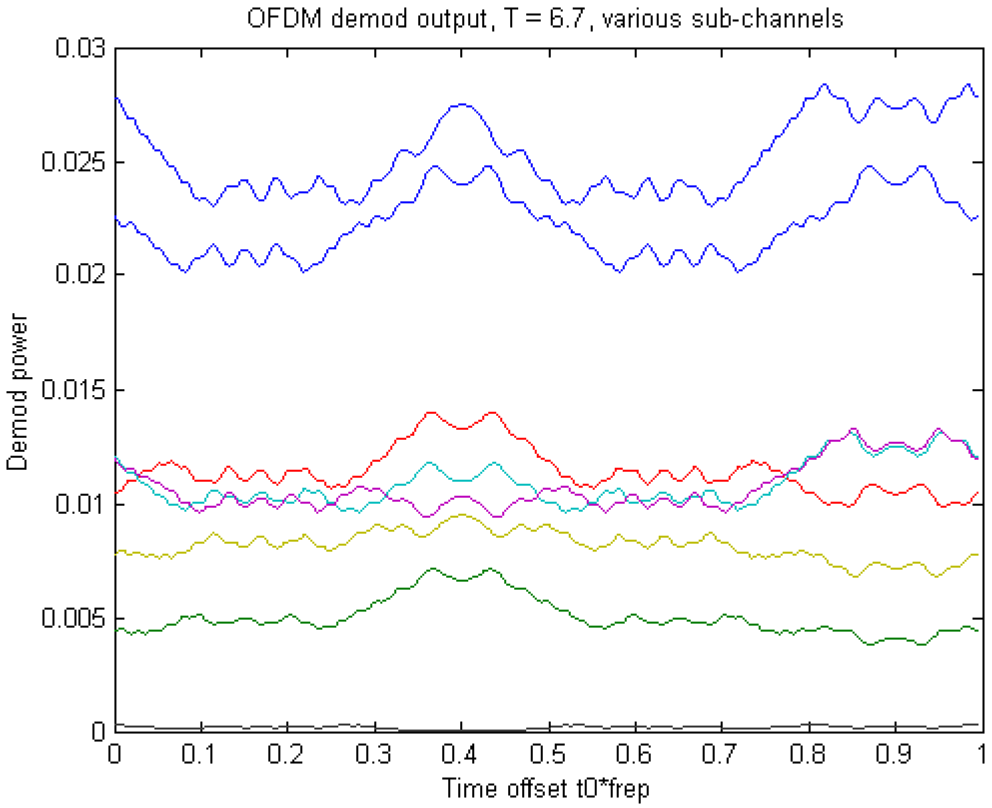


Figure 3.10: Demodulated signal in eight OFDM sub-channels, plotted against time offset

The effect of the de-interleaver in the receiver of a digital broadcast system is to randomise the order in which the demodulated decoder is presented to the decoder. Hence the interference power presented to the decoder can be treated as a random variable the distribution of whose power is the same as that within the sub-channels plotted in Figure 3.9. Figure 3.11 shows the cumulative distribution function (c.d.f) of the amplitude of this interference (the phase will be uniformly distributed on 0 - 2π). The plot assumes that the time offset t_0 is random in each sub-channel. This c.d.f. represents a heavy-tailed, or even multimodal distribution: it has a maximum near zero amplitude, representing the majority of sub-channels which do not lie near to a spectral line, and a long tail representing the sub-channels containing interference from a line. There is also a distinct maximum interference (unlike noise distributions). The mean square amplitude (i.e. interference power per sub-channel) is 0.0014, which is in fact consistent with the unit power of the DCO signal being distributed over all the sub-channels with frequencies between +/-k:

$$\overline{I_{sc}} = \frac{I}{2k\, f_{rep}T}$$

Equation 3.8

where I is the total DCO interference power. The mean power in the "tail" of the distribution (that is, in those sub-channels affected by the DCO lines) is 0.0907, which is larger by approximately the factor $f_{rep}T$:

$$\overline{I_{asc}} = \frac{I}{2k}$$

Equation 3.9

The maximum interference amplitude, from Figure 3.11, is approximately 0.17, which is about 13 dB greater than the mean interference.

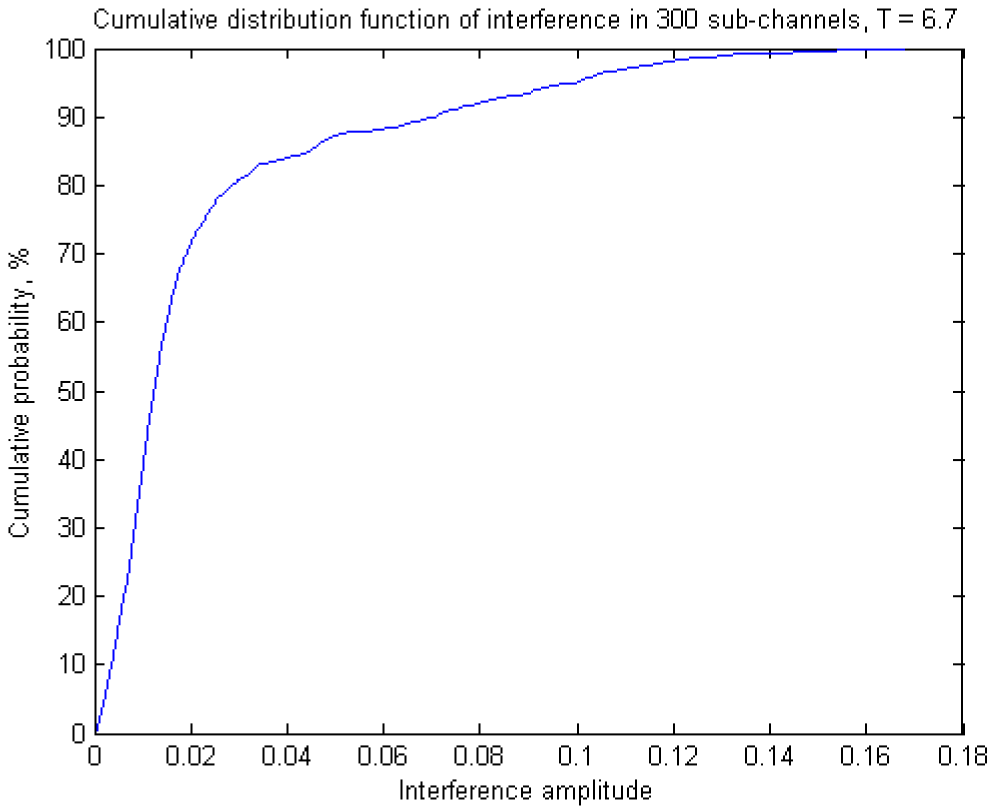


Figure 3.11: Distribution of amplitude of interference in sub-channels after de-interleaving, within band affected by DCO

Note that if the DCO repetition frequency were an exact multiple of $1/T$ (which would be the case for the repetition frequency chosen by the majority of DCO manufacturers, taken with the DVB-T symbol period) the interference in the majority of sub-channels would be precisely zero.

The exact effect of this interference on the BER will depend on the extent to which the code correct a stream with every $f_{rep}T^{\text{th}}$ symbol in error, before de-interleaving. This in turn will depend on the exact operation of the interleaver. However, to ensure that the DCO interference does not affect the signal it would be best to ensure that the signal to interference ratio in the affected sub-channels is adequate.

3.4 Modelling strategy and validation

To back up the above analysis and to obtain estimates of the effect of DCO interference on the digital broadcast signal computer modelling has been performed of a DVB-T system and the dithered clock signal. Figure 3.12 shows a block diagram of the simulated system. It includes the inner convolutional encoder of the DVB-T system [14], the inner interleaver (both bit and symbol) and the modulator, and, of course, the inverse functions in the receiver. The inner encoder is fed from a random data source, and the decoder feeds an error counter.

There is no need to simulate the outer (Reed-Solomon) encoder/decoder since the standard [14] defines the required BER of 2×10^{-4} at the output of the inner decoder.

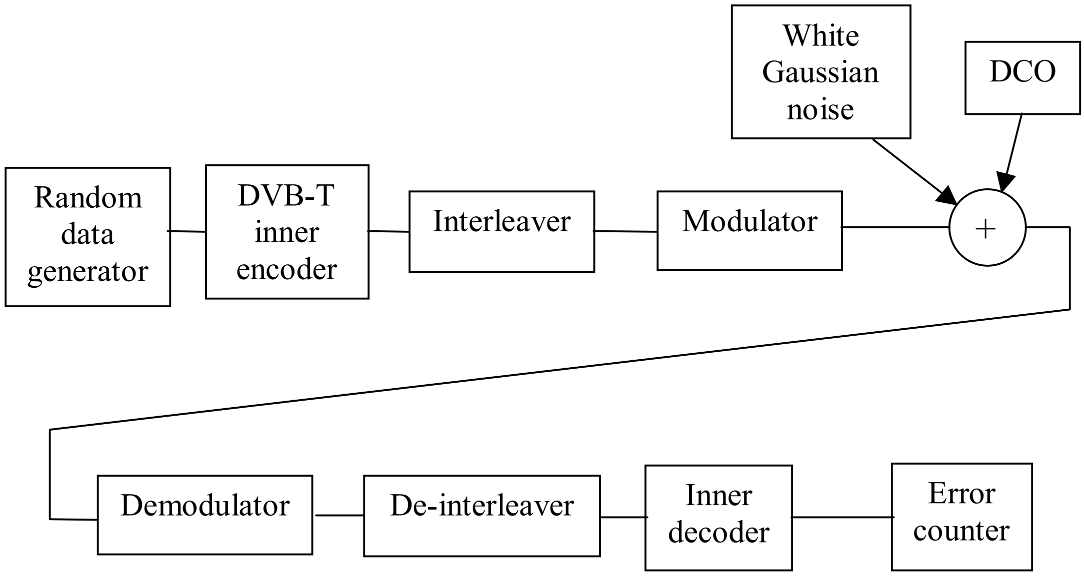


Figure 3.12: Simulation of DVB-T system with DCO interference

The modulator is shown in detail in Figure 3.13. It includes OFDM multiplexing functions as well as constellation modulation, including the inverse fast Fourier transform (IFFT) which effectively performs the multi-carrier modulation, and cyclic prefix insertion. The output of the interleaver consists of K separate streams, where $K = \log_2(M)$, with M the size of the constellation. In the U.K. 64-QAM is used, so $K = 6$. Reference (pilot) symbols are also inserted in the multiplex, with appropriate power assigned, but these are not used in the demodulator. The so-called ‘2k’ DVB-T system was simulated, which is the one used in the U.K, with a nominal 2048 carriers, although only 1705 of these are used, the remaining sub-channels being padded with zeros. However the channel portion of the simulation operates at an over-sampling ratio of 2 (i.e. 2 samples per sub-channel per OFDM symbol). Hence the IFFT length is doubled to 4096.

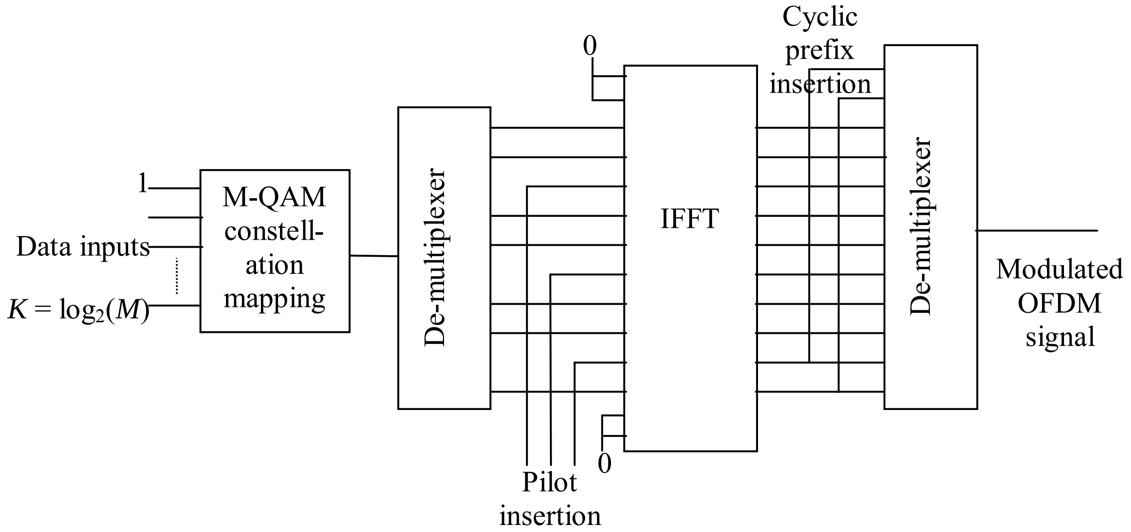


Figure 3.13: Structure of modulator

At the output of the OFDM modulator white Gaussian noise or DCO interference is added. The noise is represented as an uncorrelated random variable added to each sample, which represents noise which is white over a bandwidth equal to the sample rate. The DCO is represented as a sine wave modulated with the phase given by Equation 3.6 and amplitude determined by the signal to interference ratio. The level of interference is calculated from the desired signal to interference ratio and the average signal power, which is calculated from the sampled signal.

The simulation has been validated with white noise by comparison with theory, for a series of comparatively simple cases. Figure 3.14 shows the results for coded and uncoded QPSK. They are compared with simple theoretical estimates; for uncoded QPSK this is [20]:

$$P_b = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

Equation 3.10

while for coded QPSK it is given by Equation (7.9) of [19]. These equation describe the case when no guard period and no reference (pilot) symbols are used: they can be adjusted by an allowance given by $10\log_{10}\left(1705\left(1+T_g/T\right)/1512\right)$ when these items are present, since both require additional power by do not contribute to data transmission. The bit energy to noise density ratio can be calculated from the standard deviation of the simulated noise:

$$\frac{E_b}{N_0} = \frac{ST_S/M}{N/f_s} = \frac{ST_S f_s}{2\sigma^2 M}$$

Equation 3.11

where S/N is the signal to noise power ratio (within the whole simulated bandwidth), σ^2 is the variance of each of the real and imaginary parts of the simulated noise, so that $N = 2\sigma^2$, $T_S = T$

$+ T_g$ is the total OFDM symbol period, f_s is the sampling frequency, and M is the total number of information bits per OFDM symbol.

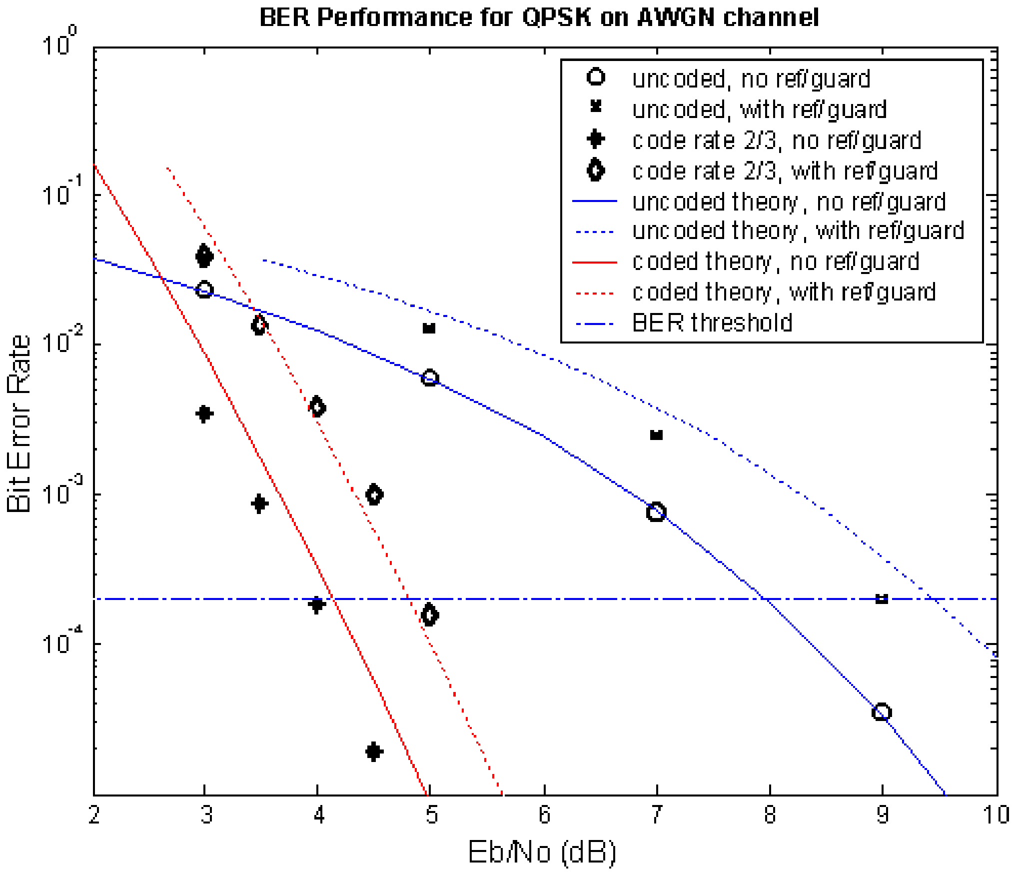


Figure 3.14: Modelling results for QPSK on AWGN channel

Figure 3.14 shows results for these cases, giving generally good agreement. Note that Equation 3.4 on which the theory for coding is based is strictly an upper bound. Figure 3.15 shows the results for the 64-QAM modulation used in DVB-T in the U.K. The theoretical result for the uncoded modulation is obtained from [20]:

$$P_b = \frac{7}{12} Q\left(\sqrt{\frac{2}{7}\frac{E_b}{N_0}}\right)$$

Equation 3.12

Again quite good agreement is observed for the uncoded case. No theoretical result was available for the coded system, but the coding gain obtained is consistent with that for QPSK. These results for known cases give some confidence of the validity of the results for the unknown case of DCO interference presented in the next section. These results will also be examined for their consistency with the theory of the effect of DCO interference, described above.

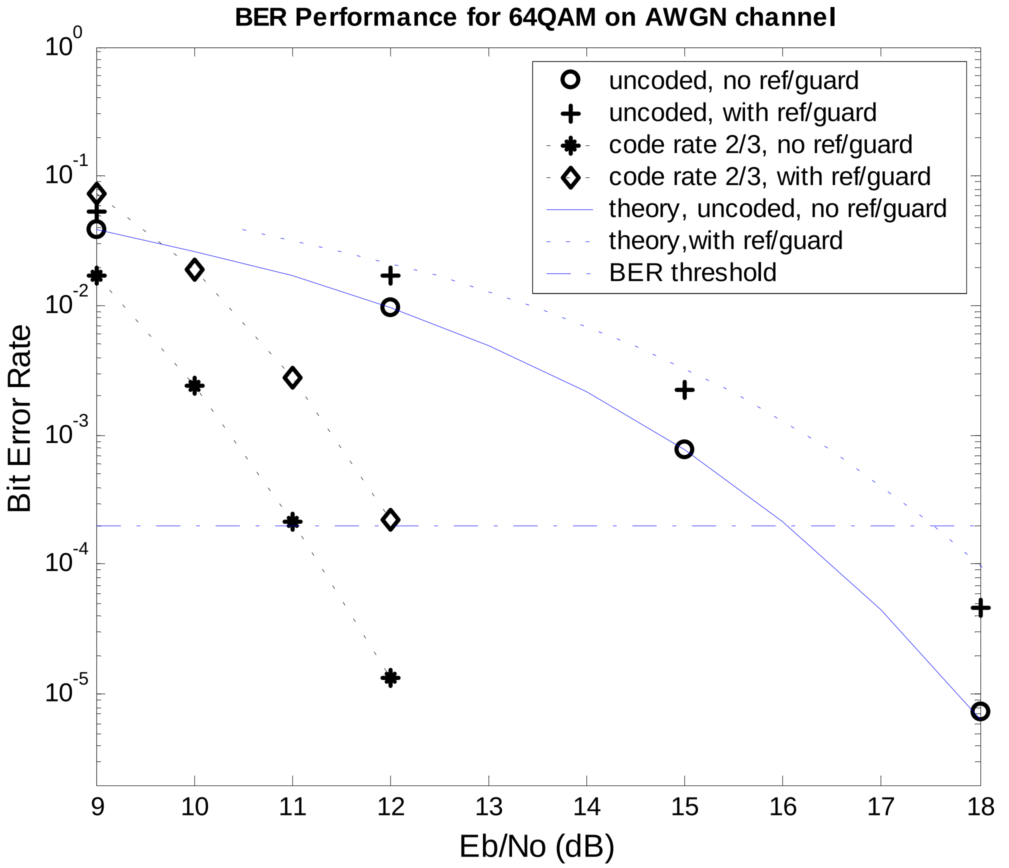


Figure 3.15: Modelling results for 64-QAM on AWGN channel

3.5 Results

The results for DCO interference will be evaluated in terms of signal to interference ratio, rather than bit energy to noise density ratio, as has been considered for the white noise results. This is because the bandwidth of the DCO signal is not well defined. It is instructive to compare the results for DCO with those for white noise, and for this purpose they must be converted to signal to noise ratio. For this purpose Equation 3.11 is used, but amended so as to include only the noise power that falls within the signal bandwidth:

$$\left(\frac{S}{N}\right)_{OFDM} = \frac{S}{N_0\, n_c/T} = \frac{STf_s}{2\sigma^2 n_c} = \frac{E_b}{N_0}\frac{MT}{n_c T_S}$$

Equation 3.13

where n_c = 1705 is the total number of carriers in the OFDM multiplex. M is given by the number of information symbols (1512), times the number of bits per symbol (k = 2 for QPSK, 6 for 64-QAM), times the code rate. The theoretical results for AWGN can be obtained by adapting Equation 3.10 and Equation 3.12, giving:

$$P_b = Q\left(\sqrt{\frac{S}{N}}\right)$$

Equation 3.14

for QPSK and:

$$P_b = \frac{7}{12} Q\left(\sqrt{\frac{1}{42}\frac{S}{N}}\right)$$

Equation 3.15

for 64-QAM.

The simulation results have been calculated for f_{rep} = 30 kHz rather than 31.25 kHz as had previously been suggested in section 2, for reasons which will be explained below. The results are given in Figure 3.17 for QPSK and 64-QAM respectively.

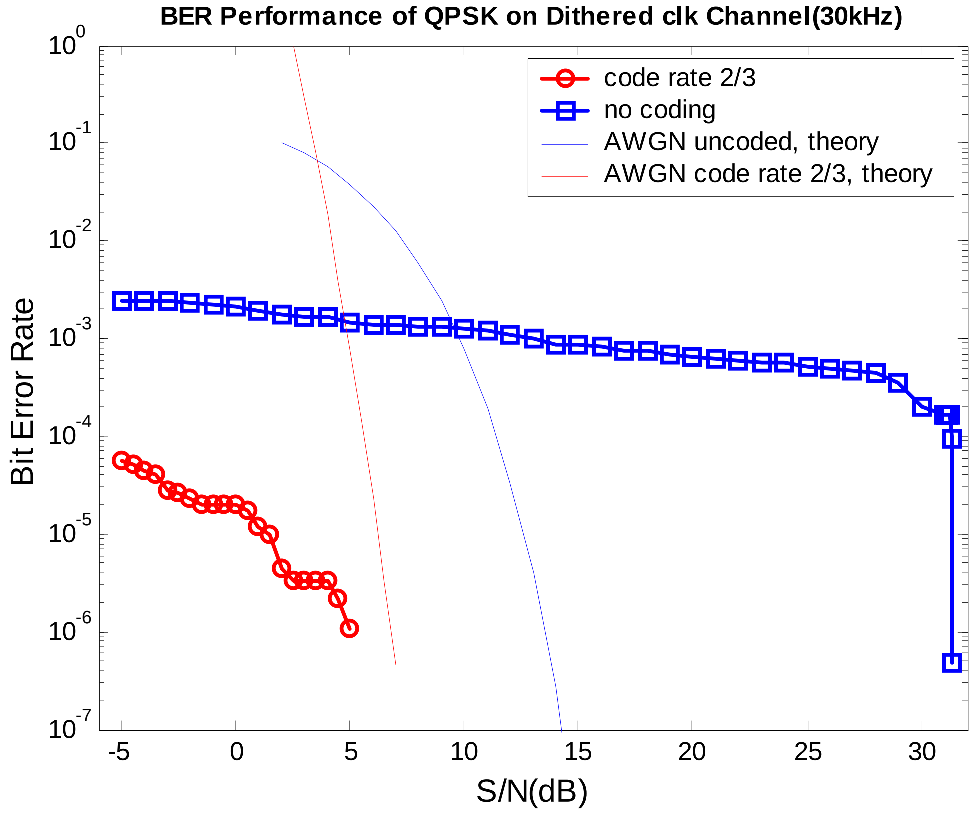


Figure 3.16: Modelling results for QPSK with DCO interference, compared to theoretical results for AWGN, for f_{rep} = 30 kHz

For the uncoded case it is noted that the required signal to noise ratio is much larger than for Gaussian noise. This is partly because the bandwidth of the noise is approximately twice that of the DCO signal, and hence the noise per sub-channel is approximately 3 dB less on average. More significantly it is due to the "heavy-tailed" distribution of the interference, as demonstrated in Figure 3.11, which means that in a number of cases the signal must be a great deal larger than average in order to overcome the interference levels falling in the "tail" of the distribution. Note also that above a certain signal to interference ratio the BER falls abruptly to zero, in contrast to the result for noise, where it continues to fall in the characteristic "waterfall" curve, never strictly reaching zero. This is because, as also noted from Figure

3.11, there is a finite maximum interference amplitude, and if the signal to interference ratio is such that this amplitude can never cause an error, the BER will drop to zero.

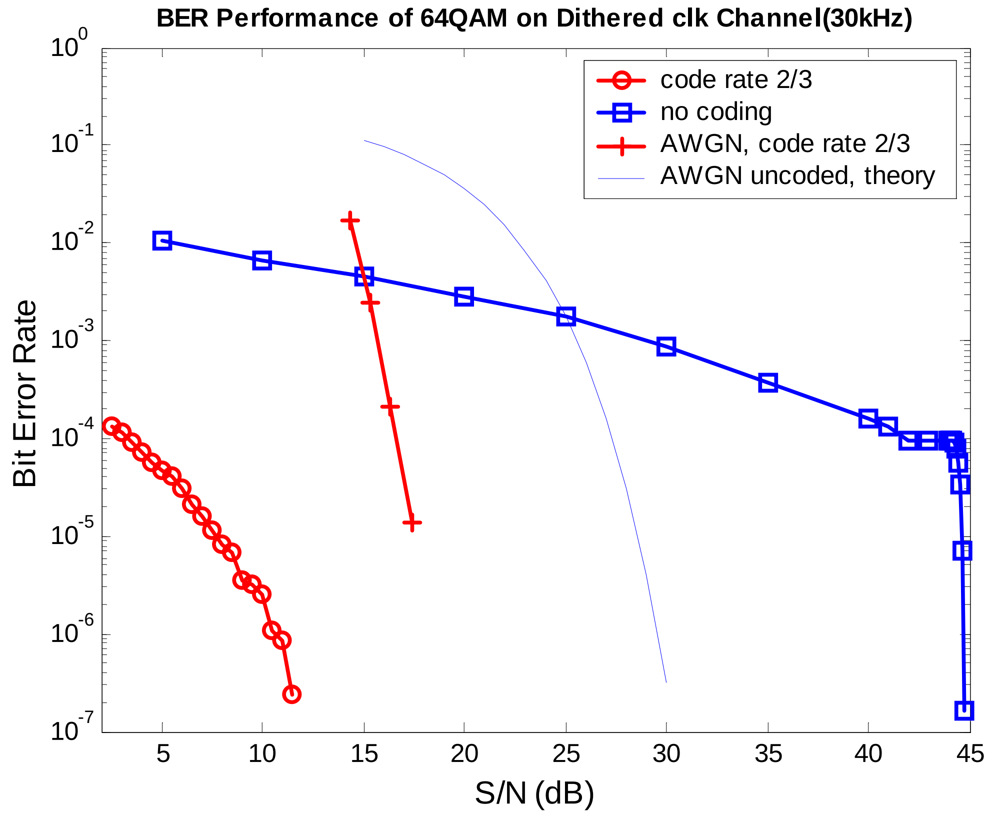


Figure 3.17: Modelling results for 64-QAM with DCO interference, compared to theoretical results for AWGN, for f_{rep} = 30 kHz

The BER performance for the coded case, however, is significantly better than for Gaussian noise. In fact the simulation was first carried out for f_{rep} = 31.25 kHz, with a frequency offset of zero between the DCO centre frequency and the frequency of the nearest OFDM sub-channel, and it was found that the BER was zero for almost any signal to interference ratio. It was noted in section 1.3 that if $f_{rep}T$ is an exact integer, there is no interference at all except in every $f_{rep}T^{\text{th}}$ sub-channel. For f_{rep} = 31.25 kHz and T = 224 µs, $f_{rep}T = 7$. It now appears that the inner encoder and de-interleaver are able to correct errors in every seventh sub-channel, provided there is no interference elsewhere, whatever the level of the interference in those sub-channels.

Thus the case f_{rep} = 31.25 kHz is far from a worst case for DCO interference. Although we note that this effect does not occur if there is a frequency offset, it is probably best to avoid this frequency as it is sensitive to the precise DCO centre frequency. For that reason results given here are for the slightly different value f_{rep} = 30 kHz (which gives $f_{rep}T = 6.7$). This, then, is the value used in Figure 3.16 and Figure 3.17.

Note that the result for coded systems is closer to the result for Gaussian noise, while still being significantly lower. This is because the effect of coding is to combine signals from several sub-channels (according to the interleaving) in order to make the decision. Because of the Central Limit Theorem the resulting total interference is nearer to the Gaussian distribution.

3.6 Other victim systems, including CDMA

In this section the effect of DCO interference is considered very briefly on other possible victim systems. However, conventional narrow-band systems will not be considered, since these have been considered more thoroughly in [10], and are relatively well understood.

In the foregoing, which has considered digital broadcast services based on OFDM, the concentration has been on DVB-T, which has enabled a detailed simulation model of the system to be established. Here some of the conclusions will be extended to the other main digital broadcast standard, DAB. It has been observed that some features of the DVB-T standard have a significant effect on its performance in DCO interference; it will be considered whether DAB may be different.

Digital services based on CDMA will also be considered, which primarily means third generation mobile systems using wideband CDMA (W-CDMA). In Europe, the main example of such a standard is UMTS-FDD [15], but most of the conclusions will also be applicable to UMTS-TDD and to other similar standards elsewhere in the world, including cdma2000.

Commencing with digital broadcast services using DAB (noting that although the standard was designed for audio transmission, there are now proposals that it be used for general data transmission) [13]. It was noted in section 3.3 that the effect of DCO interference depends on the ratio of the OFDM symbol period to the dithering period. DAB uses a range of symbol periods (depending on transmission mode: 125 µs, 250 µs, 500 µs and 1 ms). The highest frequency this corresponds to is 8 kHz, which is still far below the usual dithering frequency of 31.25 kHz, and hence it can be expected that dithering will affect the whole dithering bandwidth in every symbol period. The only exception to this might be the single instance of dithering at 3 kHz mentioned in section 2. However even here the dithering could be expected to affect at least half the affected bandwidth, so there would be little practical difference.

Further it was noted that if this ratio was precisely an integer (say p), and there was no frequency offset between the DCO centre frequency and a sub-channel frequency, then the DCO signal would cause interference only in every p^{th} sub-channel. It was noted above that in DVB-T this seems to result in no errors at all in the coded system in this exact case. However, assuming dithering at 31.25 kHz this condition would not be met in DAB for any transmission mode, and therefore the question does not arise. The results given above for dithering frequency of 30 kHz would therefore probably be close to the actual result for DAB. Note also that DAB uses only QPSK transmission, with variable code rate. Thus the effects will be more similar to the results given above for QPSK than for 64QAM.

W-CDMA operates on a completely different principle to OFDM, and its implications for emission limits are in many ways the opposite of OFDM. Essentially the data signal is spread in frequency by multiplying it by a pseudo-random sequence, called the *spreading sequence*, operating at a symbol rate (usually known as the *chip rate*), which is much higher than the data rate. The chip rate in most third generation mobile standards, including UMTS-FDD, is such as to give a spread signal bandwidth of around 5 MHz. In the receiver the received signal is multiplied by a replica of the spreading sequence, operating in synchronism with the sequence at the transmitter. The effect of this is to *de-spread* the wanted signal, and regenerate the original data signal, which can then be extracted from the interference using a narrow-band filter. However any interfering signal which is not correlated with the spreading sequence, will be spread in bandwidth, and will be largely, but not entirely, rejected by the narrow-band filter. The spread bandwidth of the interference will also be around 5 MHz, regardless of its original bandwidth. However an interferer, which is sufficiently powerful can still disrupt the wanted signal, especially if the signal is small. This is called the *near-far problem*.

Thus if the spectrum of a DCO signal falls entirely into the bandwidth of a single W-CDMA signal, its effect will be the same as that of a single, unspread carrier of the same power. Clock dithering gives benefits only insofar as it spreads the interference signal beyond this 5 MHz bandwidth. While dithered bandwidths may in some cases be greater than 5 MHz, it will clearly require a very wide bandwidth to gain a significant advantage. Thus in the UMTS bands care must be taken in devising emission limits in order to protect these services. The preferred approach is to limit the total emission within each 5MHz band: it would be difficult to base a reliable limit on measurements made in a significantly narrower measurement bandwidth.

4 EXPERIMENTAL WORK

This section describes the equipment and methods used to determine the interfering signal power required to cause an off-air DVB-T signal to fail when subjected to both unmodulated and frequency modulated interfering signals. A comparison is made with the effect caused by this interference on terrestrial analogue PAL signals.

4.1 Test setup

The following equipment was used to perform the tests:

- Antiference DX8 Digital high gain (14.2dBi) aerial – Group B
- Triax TA25W 25dB wideband UHF masthead amplifier
- Triax TP1285 power unit
- Marconi Instruments model 2019 signal generator
- York EMC Services programmable wave generator (PWG)
- Farnell LFM3 signal generator (master clock for PWG)
- RS 610-578 frequency meter
- Farnell TOPS3D power supply (power for PWG)
- Mini-Circuits BMP-5075 50/75Ω matching pad
- HP 8594E spectrum analyser
- Maxview MAS2 2-way & MAS4 4-way splitters
- Set-top box #1
- Set-top box #2
- Sony KVM1602U television
- Panasonic NV-HS800BY video cassette recorder
- Toshiba VTV1402S television / video cassette recorder

The equipment was connected up as shown in Figure 4.1. The off-air signal was received and amplified before being applied to one port of the 2-way splitter which was being used as a combiner. The output of the Marconi Instruments signal generator was converted from 50 to 75Ω before being applied to the other port of the 2-way combiner. The combined signal and interference was applied to a 4-way splitter which feeds the two set-top boxes, the Sony television and the spectrum analyser. The modulation generated by the programmable wave generator was applied to the Marconi Instruments signal generator. The master clock for the programmable wave generator was taken from the Farnell signal generator, the frequency monitored by the RS frequency meter. Set-top box #1 was connected to the Sony television via the Panasonic video recorder using SCART cables. Set-top box #2 was connected to the Toshiba television / video recorder using a SCART cable.

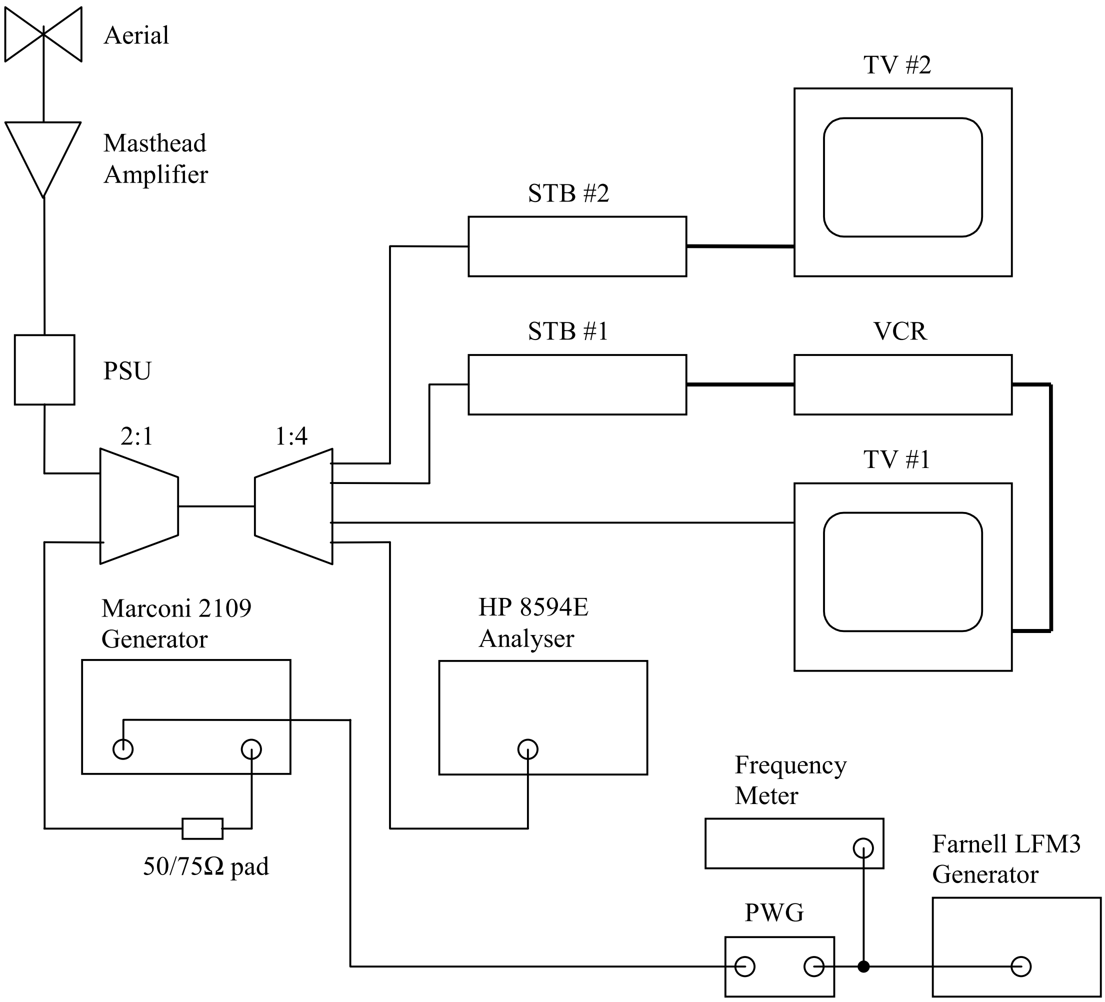


Figure 4.1: Test Setup

4.1.1 Signal source

The off-air signal shown in Figure 4.2 was taken from Emley Moor with the spectrum analyser connected directly to the feed from the aerial and mast-head amplifier:

- BBC2 analogue on channel 51 (710-718MHz) with vision at 711.25MHz, FM sound at 717.25MHz and Nicam sound at 717.80MHz. Peak amplitude 82dBµV into 75Ω.
- BBC multiplex on channel 52 (718–726MHz) + 0.4MHz offset. Average amplitude in a peak detector of 42 dBµV into 75Ω, signal-to-noise ratio > 20dB.

Figure 4.2: Off-air signal (measured at RF feed from mast-head amplifier – reduced by 11.8 dB for spectrum at STB)

The spectrum analyser was set up as follows:

- Centre frequency 718MHz
- Span 20MHz
- Resolution bandwidth 100kHz
- Video bandwidth 10kHz
- Input impedance 75Ω
- Input attenuation 0dB
- Measurement units dBμV
- Reference level 90dBμV

Note that splitter/combiner introduced a loss of 11.8dB.

The peak amplitude of BBC2 analogue fed to the television was nominally 70dBμV in 100 kHz and 71 dBμV in 300 kHz. For PAL-I the signal level is usually quoted as the RMS value at sync tip which is the same as the peak value.

The average amplitude of the BBC multiplex fed to the two set-top boxes was nominally 30 dBμV, as measured in 100 kHz peak detector. This corresponds to –65 dBm RMS power in the multiplex.

4.1.2 Interference source

The Marconi Instruments signal generator output was converted from 50Ω to 75Ω with a Mini-Circuits matching pad. Measuring the output of the pad and taking into account the splitter/combiner loss gives the following conversion factor:

-40dBm ≡ 50dBμV into 75Ω ≡ -58.8 dBm

Therefore the total RMS interference power fed to the TV receiver is 18.8 dB lower than the power indicated on the signal generator.

The level of the DCO interference corresponding to a given RMS level in the entire interference bandwidth was measured using a number of detectors. The results are shown in Table 4..

	RMS **Full Bandwidth**	**Peak** **120 kHz**	**Peak** **100 kHz**	**Quasi-Peak** **120 kHz**
CW	0	0	0	0
Lexmark, **30 kHz, 0.5 %**	0	-8.5	-8.2	-9.4
Lexmark, **30 kHz, 1 %**	0	-12.8	-13.1	-13.9

Table 4.1: Suppression factors for DCO interference measured using different detectors and bandwidths relative to the RMS level in the entire bandwidth of the interference

4.1.3 Test methods

Two sets of tests were carried out, one to determine the susceptibility to an unmodulated interference source, the second to determine the susceptibility to a frequency modulated interference source.

In both cases the interference level was adjusted until the set-top boxes started to exhibit gross errors (display freezing or many macroblock faults). For the frequency modulated interference the carrier was switched from the digital multiplex to the analogue channel and the interference level adjusted until the interference (typically diagonal patterning) became invisible.

4.1.3.1 Unmodulated interference

The set-up and test was as follows, the results are shown in Table 4.2:

- Set carrier to 722.4MHz (centre of BBC multiplex)
- Set modulation OFF
- Set level to –40dBm
- Set Δ carrier to 200Hz
- Set Δ level to 1dB
- Make measurements
- Step to next frequency
- Repeat until frequency is 722.405MHz

Frequency (MHz)	Failure level (dBm)	
	STB #1	STB #2
722.4000	-43	-57
722.4002	-42	-57
722.4004	-40	-56
722.4006	-38	-55
722.4008	-36	-51
722.4010	-35	-42
722.4012	-38	-52
722.4014	-40	-55
722.4016	-42	-57
722.4018	-43	-59
722.4020	-42	-59
722.4022	-44	-58
722.4024	-45	-58
722.4026	-45	-58
722.4028	-46	-56
722.4030	-46	-55
722.4032	-47	-51
722.4034	-48	-51
722.4036	-47	-53
722.4038	-45	-55
722.4040	-44	-57
722.4042	-43	-57
722.4044	-42	-57
722.4046	-41	-58
722.4048	-40	-57

Table 4.2: Unmodulated interference results

4.1.3.2 Frequency modulated interference

The setup and test was as follows, the results are shown in Table 4.3:

- Set carrier to 722.4MHz (centre of BBC multiplex)
- Set FM modulation to 0MHz external (no deviation)
- Set level to –40dBm
- Set Δ carrier to 8.4MHz (allows quick switch to centre of BBC2 analogue)
- Set Δ modulation to 3.6MHz (allows quick switch to ~0.5% & ~1% deviation)
- Set Δ level to 1dB
- Make measurements
- Set FM modulation to 3.6MHz external

- Set modulation rate to 30kHz
- Load Triangle modulation waveform
- Make measurements
- Load Stepped Triangle modulation waveform
- Make measurements
- Load Lexmark modulation waveform
- Make measurements
- Set FM modulation to 7.2MHz external
- Make measurements
- Repeat Lexmark modulation waveform measurements at 3, 4, 31.25kHz rates

Modulation type	**Deviation**	**Rate**	**Failure Level (dBm)**		
			STB #1	**STB #2**	**Analogue**
None	None	None	-43	-57	-68
Triangle	0.5%	30kHz	-61	-62	-68
Stepped Triangle	0.5%	30kHz	-60	-61	-67
Lexmark	0.5%	3kHz	-62	-62	-69
Lexmark	0.5%	4kHz	-61	-62	-69
Lexmark	0.5%	30kHz	-59	-62	-69
Lexmark	0.5%	31.25kHz	-59	-61	-68
Lexmark	1%	3kHz	-62	-62	-67
Lexmark	1%	4kHz	-61	-62	-67
Lexmark	1%	30kHz	-61	-62	-69
Lexmark	1%	31.25kHz	-61	-62	-70

Table 4.3: FM interference results

4.2 Interference generator validation

The following equipment was used to validate the interference generator:

- Anritsu MS2667C spectrum analyser
- Rohde & Schwarz power meter
- York EMC Services envelope detector
- Kikusui COS6100A oscilloscope

The interference generator was connected in turn to the spectrum analyser, power meter and the envelope detector / oscilloscope to measure the spectrum, the total power produced and any ripple (amplitude modulation) present respectively.

This was carried out as described in the following sections at modulation rates of 3, 4, 30 and 31.25kHz for Triangle, Stepped Triangle, Ramp, Sine and Lexmark waveforms with both 0.5% and 1% deviations for the power and ripple measurements, and a subset of these at various resolution bandwidths for the spectrum measurement.

4.2.1 Power measurement

- Set carrier frequency to 722.4MHz
- Set output level to –30dBm
- Set FM modulation OFF
- Measure power
- Load modulation waveform
- Set modulation rate to 3kHz
- Set FM modulation to 3.6MHz external
- Measure power
- Set FM modulation to 7.2MHz external
- Measure power
- Repeat for 4, 30, 31.25kHz rates
- Repeat for all waveforms

In the worst cases the measured power was +0/-0.15dB with respect to the unmodulated signal.

4.2.2 Ripple measurement

- Set carrier frequency to 722.4MHz
- Set output level to +5.5dBm
- Set FM modulation OFF
- Measure detector DC output
- Load modulation waveform
- Set modulation rate to 3kHz
- Set FM modulation to 3.6MHz external
- Measure detector peak-to-peak ripple

- Set FM modulation to 7.2MHz external
- Measure detector peak-to-peak ripple
- Repeat for 4, 30, 31.25kHz rates
- Repeat for all waveforms

In all cases the detector ripple was ~3% of the detector DC output for 0.5% deviation and ~6% for 1% deviation.

4.2.3 Spectrum measurement

- Set carrier frequency to 722.4MHz
- Set output level to –30dBm
- Set FM modulation OFF
- Set centre frequency to 722.4MHz
- Set span to 20MHz
- Set resolution bandwidth to 100kHz
- Measure spectrum
- Load Triangle modulation waveform
- Set modulation rate to 30kHz
- Set FM modulation to 7.2MHz external
- Measure spectrum
- Repeat for Stepped Triangle modulation waveform
- Repeat for Lexmark modulation waveform
- Set FM modulation to 3.6MHz external
- Measure spectrum
- Set modulation rate to 3kHz
- Set FM modulation to 7.2MHz external
- Measure spectrum
- Set modulation rate to 30kHz
- Set resolution bandwidth to 300kHz
- Measure spectrum
- Repeat for resolution bandwidths of 1MHz and 3MHz
- Set FM modulation OFF

- Set resolution bandwidth to 300kHz
- Measure spectrum
- Repeat for resolution bandwidths of 1MHz and 3MHz

These measurements are shown in Figure 4.3 to 4.14.

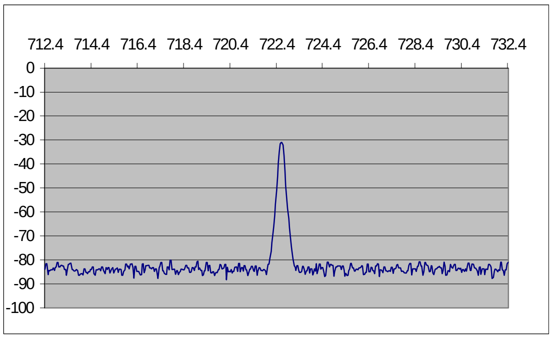


Figure 4.3: Unmodulated, 100kHz RBW

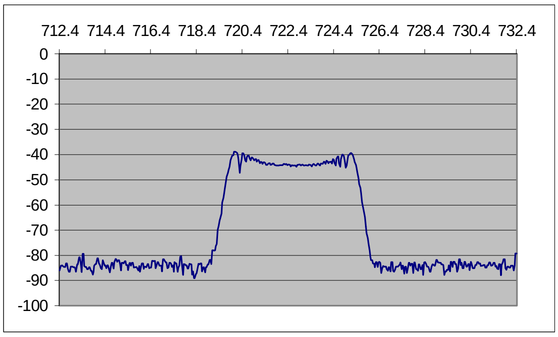


Figure 4.4: Triangle, 1%, 30kHz, 100kHz RBW

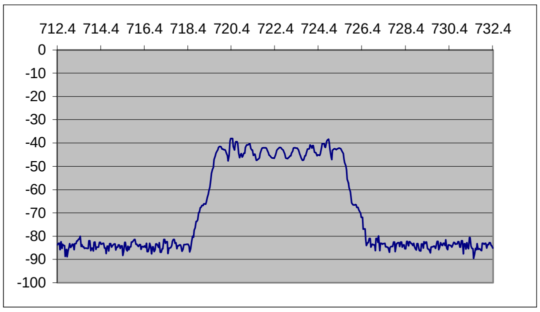


Figure 4.5: Stepped, 1%, 30kHz, 100kHz RBW

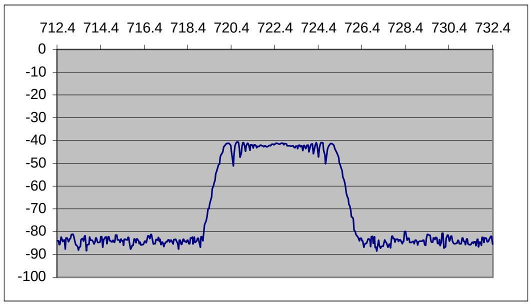


Figure 4.6: Lexmark, 1%, 30kHz, 100kHz RBW

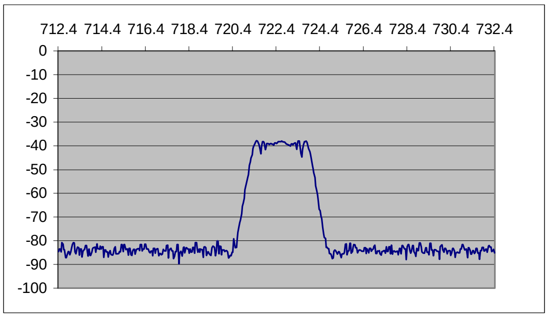


Figure 4.7: Lexmark, 0.5%, 30kHz, 100kHz RBW

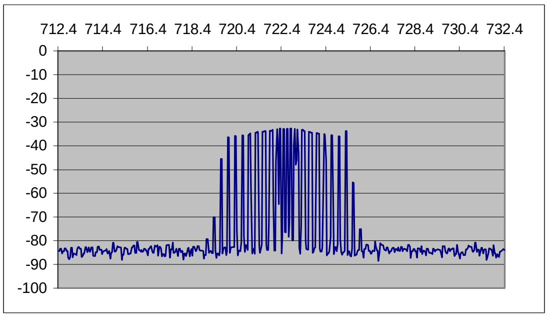
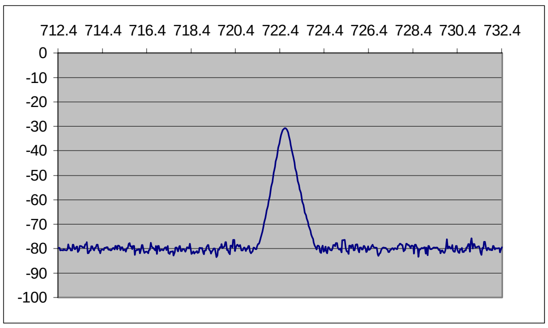


Figure 4.8: Lexmark, 1%, 3kHz, 100kHz RBW



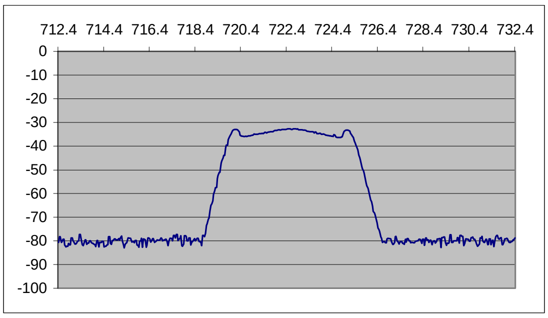
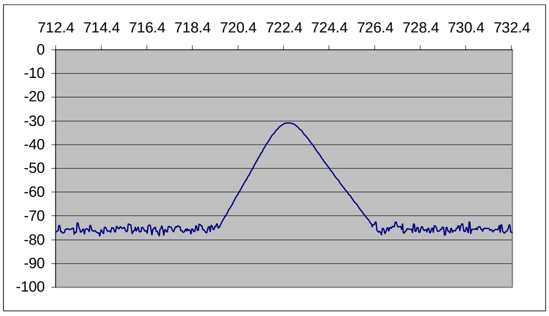


Figure 4.9: Unmodulated, 300kHz RBW

Figure 4.10: Lexmark, 1%, 30kHz, 300kHz RBW



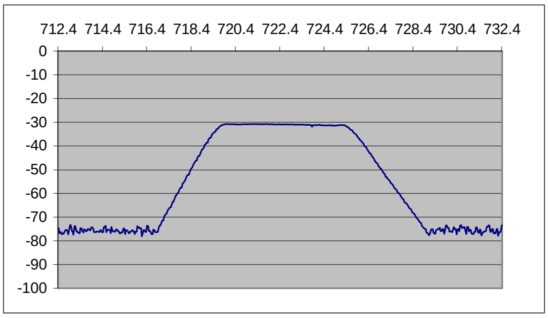
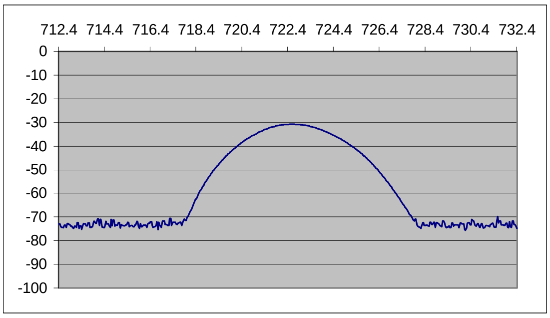


Figure 4.11: Unmodulated, 1MHz RBW

Figure 4.12: Lexmark, 1%, 30kHz, 1MHz RBW



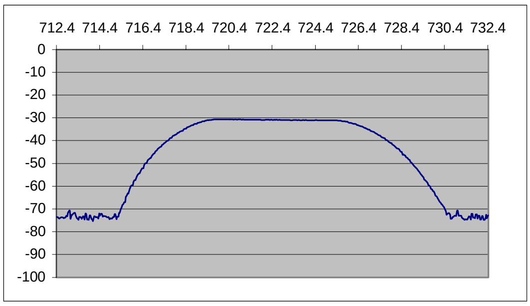


Figure 4.13: Unmodulated, 3MHz RBW

Figure 4.14: Lexmark, 1%, 30kHz, 3MHz RBW

4.3 Immunity tests on DAB receiver

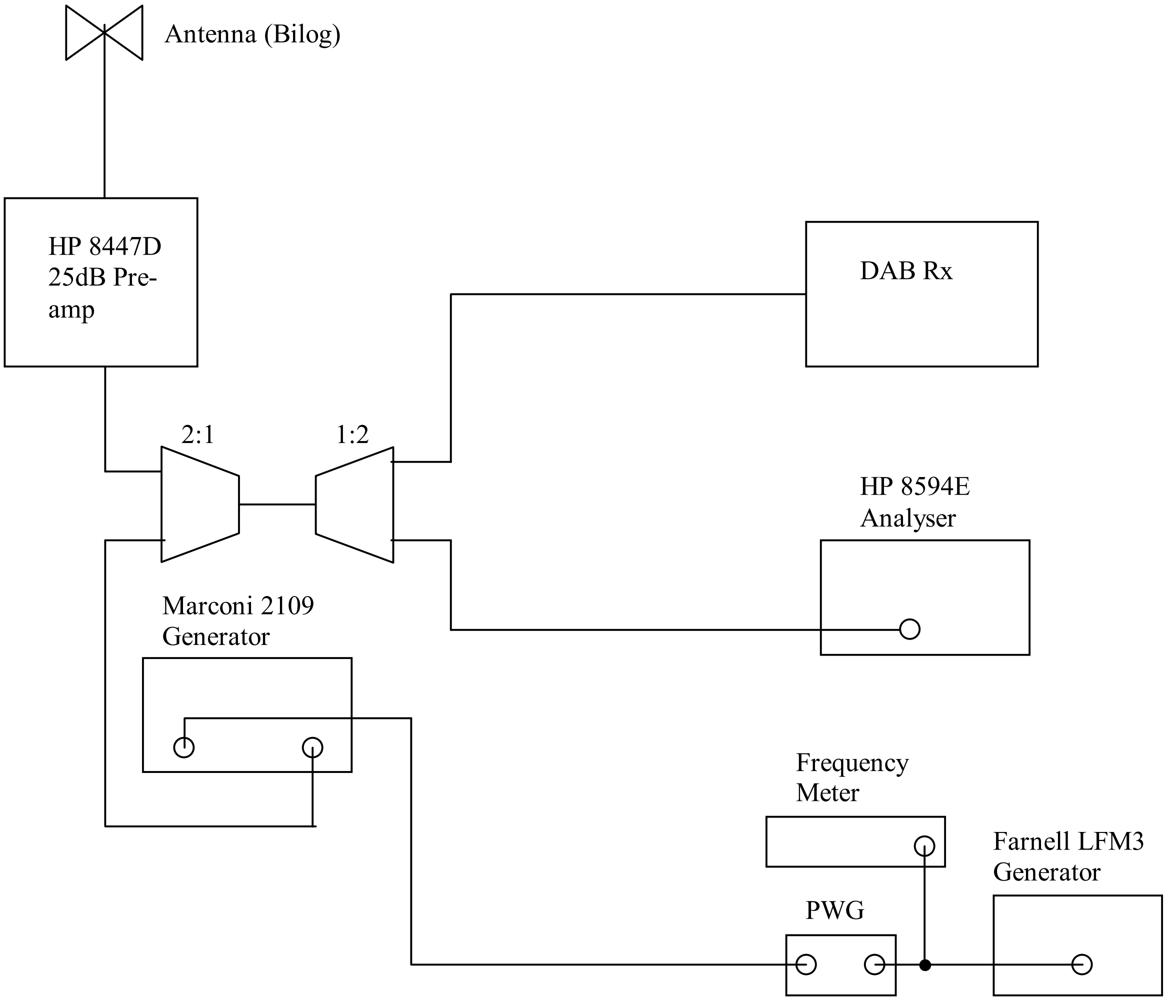


Figure 4.15: Equipment layout for testing DAB receiver immunity

In the same way that the immunity of DVB-T was investigated DAB reception was investigated also. The test set up was broadly similar to the DVB-T one, with the simplification that the DAB receiver tested had a 50Ω antenna impedance. Also, only one receiver was tested. The test set up is shown in Figure 4.15.

The equipment used for the test was as follows:

- Marconi Instruments model 2019 signal generator
- York EMC Services programmable wave generator (PWG)
- Farnell LFM3 signal generator (master clock for PWG)
- RS 610-578 frequency meter
- Farnell TOPS3D power supply (power for PWG)
- HP 8594E spectrum analyser
- Hewlett Packard HP 8447D 25dB pre-amplifier

- Bilog antenna and tripod
- Mini circuits ZFSC-2-5 splitter/combiner
- Mini circuits ZSC-2-1 splitter/combiner
- ARCAM Alpha 10 Digital Radio Tuner (DAB receiver under test)

The DAB receiver tested was of high quality and had both a “user mode” and an “engineering mode”. The engineering mode allowed the receiver to be interrogated for details of the selected multiplex and the quality of the reception. Without any interfering signal applied the tuner was set to receive Virgin Radio and the following “engineering mode” information recorded.

- Multiplex ensemble 222.063MHz
- Selected – Virgin Radio Data Rate 160kbit/s
- Virgin Radio StartCU:174 Size:116
- Errors/CIF: 0 CRC, 4 Viterbi
- Signal Strength (AGC) – 14 bars out of 20
- Transmitter ID 1/1, Main: 53, Sub: 4, FS: 1
- Virgin Radio joint stereo

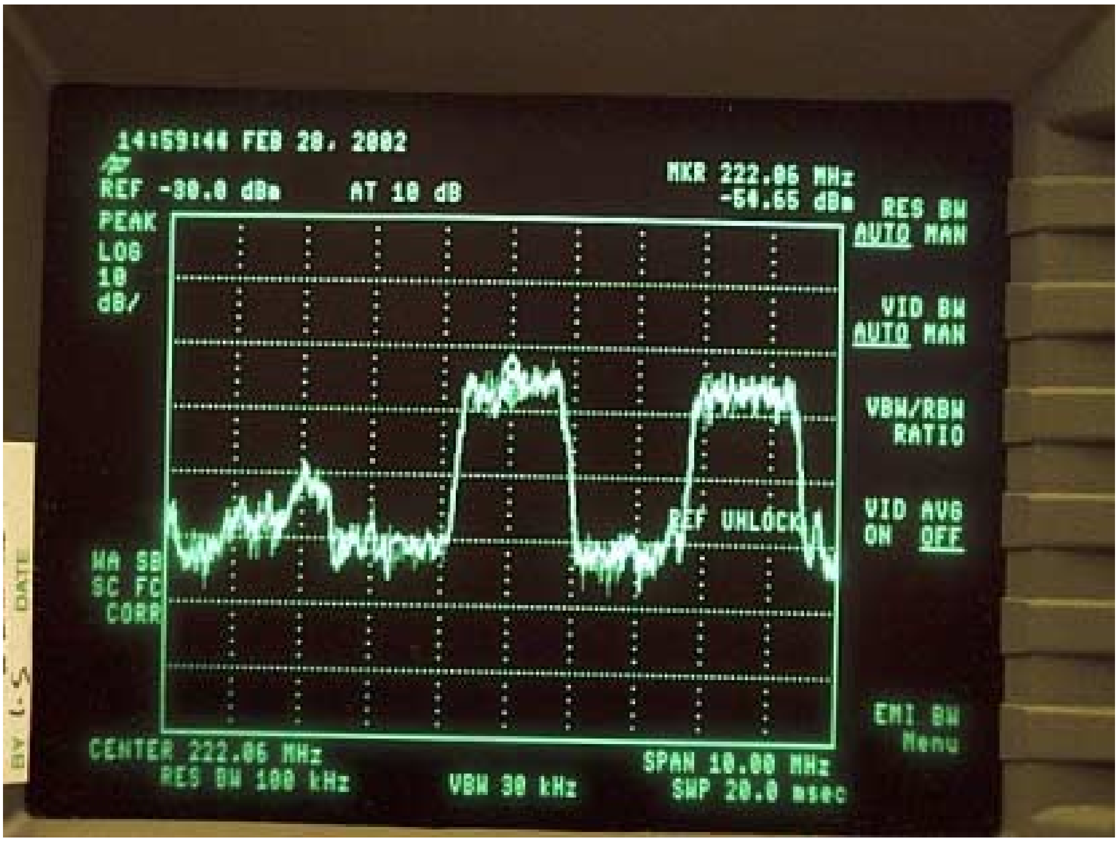


Figure 4.16: The DAB multiplex with no interference signal

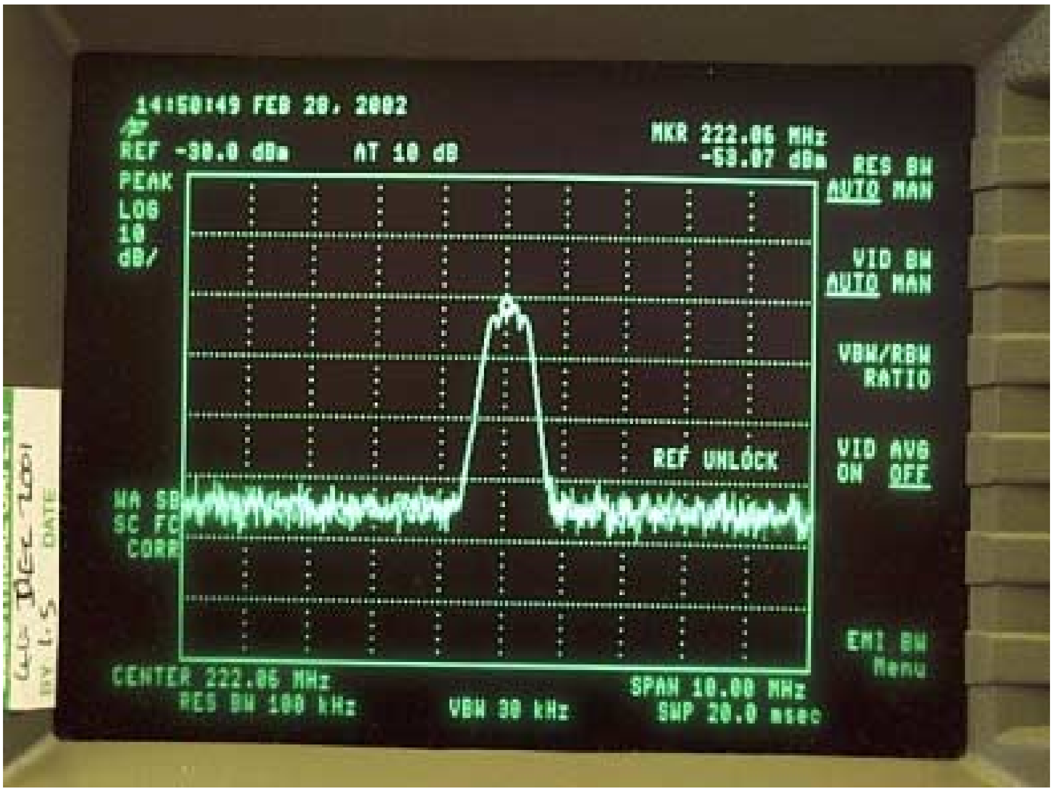


Figure 4.17: 0.5% Lexmark interfering signal sufficient to cause the DAB receiver to mute its audio output

Figure 4.18: DAB multiplex with the Lexmark interfering signal sufficient to cause the DAB receiver to mute its audio output

Figure 4.16 to Figure 4.18 show pictures of the spectrum analyser: centred on the DAB multiplex with no interference signal; a 0.5% Lexmark interfering signal sufficient to cause the DAB receiver to mute its audio output; the DAB multiplex with the Lexmark interfering signal sufficient to cause the DAB receiver to mute its audio output.

Modulation	Deviation (% of carrier frequency)	Rate (kHz)	RF level from signal generator (dBm)	Errors at DAB receiver Viterbi	User perception of error
Lexmark	1	30	-42	101 to135	Audible disturbance
Lexmark	1	30	-39	189 to 211	Cuts out
Lexmark	0.5	30	-43	66 to 74	Audible disturbance
Lexmark	0.5	30	-38	121 to 155	Cuts out
None	None	None	-35	31 to 61	Audible disturbance
None	None	None	-34	Up to 500 plus 3 CRC	Cuts out

Table 4.4: Test results for DAB immunity test

Table 4.4 summarises the test results. Two levels are recorded for each type of interfering signal: the level at which interference could be heard and the level at which the receiver muted its audio output. The errors indicated at the receiver are also recorded. A significant variation of the error rate over time was observed, typical variation is indicated in the table. In general there were no CRC errors except for the case where the multiplex was subject to CW interference.

4.4 Analysis of Measurement Results

4.4.1 DVB-T

The main conclusions from the DVB-T measurement programme are:

1. For a fixed total interference power in a clock harmonic overlapping a DVB-T multiplex dithering the clock by 0.5-1 % significantly reduces the immunity of DVB-T receiver. The reduction in immunity is highly EUT dependent because COFDM is extremely robust to CW interference and the immunity of the receiver is therefore determined by its implementation and not the modulation scheme;

2. The total co-channel interference power required to cause DVB-T to fail is of the order 6-10 dB higher than that needed for PAL-I when DCO is enabled, and somewhat greater than 10 dB for CW interference. DVB-T therefore has higher immunity to both CW and DCO interference than PAL-I;
3. However, since the emission limits are specified in a 120 kHz measurement bandwidth a DCO enabled EUT could radiate more than 10 dB greater total interference power into a UHF channel and still comply with the EMC limit. This potentially offsets the increased immunity of DVB-T over PAL-I. The measurements demonstrate that the increased total interference power which dithering the clock allows would have an equally significant impact of PAL-I receivers which appear to be sensitive to the total interference power in the UHF channel;

It is instructive to calculate the signal to interference ratio (SIR) at which both CW and the DCO interference causes the digital and analogue receivers to fail. For comparative purposes it is useful to measure the interference as both the RMS level in the entire bandwidth (yielding SIR_{RMS}) and as the quasi-peak level in 120 kHz (denoted SIR_{QP}). The DCO signal level in different measurement bandwidths and detectors can be estimated by applying correction factors given in Table 4.. The results from a selection of the measurements undertaken in this study are presented in Table 4.5 and Table 4.6 compared to estimates made from published measurements by Philips [21] and the University of Hertfordshire [10]. Note that the UHF channel frequencies used and the dithering deviations and waveforms do not exactly match up between the different sets of measurement, however, broadly there is reasonable agreement for “narrow-band” (~0.5 %) and “wide-band” (~1.0 %) dithering.

Interference	**YES Ch. 52**		**Philips Ch. 39**	**Hertfordshire Ch. 33**
	SIR_{RMS}	**SIR_{QP}**	**SIR_{QP}**	**SIR_{QP}**
CW	3.3	3.3	8.3	-5
0.3 % DCO	-	-	23	17
0.5% DCO	13.8	23.3	-	-
1.0% DCO	14.8	28.8	-	-
1.2% DCO	-	-	26	27

Table 4.5: Comparison of YES, Philips and Hertfordshire measurement results for DVB-T. The YES measurements are the average values for two STBs

Interference	**YES Ch. 51**		**Philips Ch. 69**	**Hertfordshire Ch. 26**
	SIR_{RMS}	**SIR_{QP}**	**SIR_{QP}**	**SIR_{QP}**
CW	49	49	52	43
0.3 % DCO	-	-	61	42
~0.5% DCO	50	60	-	-
1.0% DCO	50	64	-	-
1.2% DCO	-	-	62	41

Table 4.6: Comparison of YES and Philips measurement results for PAL-I

For CW interference there is a wide variation in the SIR for DVB-T showing that the effect of CW interference is EUT dependent and determined by the implementation of the DVB-T receiver. For example the effect of the interference on automatic gain control systems will be highly dependent on the receiver architecture. For analogue reception the results are much less variable and consistent with a required co-channel protection ration of around 50 dB for CW interference. Hertfordshire appear to have used a more relaxed criterion for PAL-I interference than YES and Philips resulting in a lower SIR.

For DCO interference to analogue reception there is reasonable agreement between the YES and Philips measurements. If the interference is measured in a wide bandwidth then a similar protection ratio of about 50 dB appears to be appropriate for both CW and DCO interference. However, if the interference is measured in a narrow bandwidth such as 100 kHz then a higher protection ratio of about 60 dB may be required for DCOs with >~ 1% deviation. For DCO interference to DVB-T the three sets of measurements are also in good agreement. The results suggest that DVB-T is less robust to DCO interference than it is to CW for the same total interference power in the UHF channel. When the interference is measured in a 120 kHz QP detector the DCO also allows more interference power to be radiated into the DVB-T channel. This results in a much higher protection ratio for DVB-T of about 25-27 dB if the interference is assessed using a 120 kHz QP detector.

The predictions of the theoretical analysis and simulations for SIR_{RMS} appear to be somewhat lower (by about 7 dB) than the measured immunity levels. This could be caused by a number of factors including:

1. The presence of a large analogue signal in the adjacent channel during the measurements, which may reduce the immunity to other types of interference;
2. The modelling does not account for any effects that may occur in the RF tuner of the receiver, such as disruption of the automatic gain control systems.

Table 4.7 summarises the required protection ratios estimated from the measurements for DVB-T and PAL-I. These protection ratios assume that the DCO interference is bounded by the parameters considered in this study (approximately triangular modulation waveform, modulation rate of 3-50 kHz and frequency deviation < 2%).

System	**SIR**	**DCO**	**CW**
PAL-I	SIR_{RMS}	50	50
	SIR_{QP}	65	50
DVB-T	SIR_{RMS}	15	5
	SIR_{QP}	30	5

Table 4.7: Summary of required signal to interference ratios for Band 5 DVB-T and PAL-I subjected to both CW and DCO interference

4.4.2 T-DAB

T-DAB is less susceptible to interference than DVB-T due to the use of more robust modulation and coding schemes. This increased immunity is required to provide reliable reception in the more demanding portable and mobile markets which are subject to strong fading and highly variable penetration losses into buildings. The YES measurements described above can be translated into signal-to–interference ratios using the same techniques as applied to DVB-T. Table 4.8 summarises the SIR at which failure was found to occur where the signal level is the RMS level in the entire DAB multiplex and the interference is either the RMS level in the multiplex (SIR_{RMS}) or the quasi-peak level in 120 kHz (SIR_{QP}). The SIR_{QP} results measured by Hertfordshire are also shown for comparison [10].

Interference	**YES 222.063 MHz**		**Hertfordshire 225.648 MHz**
	SIR_{RMS}	**SIR_{QP}**	**SIR_{QP}**
CW	-13	-13	-5*
0.3 % DCO	-	-	0
0.5% DCO	-5	1	-
1.0% DCO	-6	2.2	-
1.2% DCO	-	-	10

Table 4.8: Comparison of YES and Hertfordshire measurement results for T-DAB (median over multiplex)

The two sets of measurements are generally consistent, showing high (EUT dependent) immunity to CW interference and increasing susceptibility to DCO interference with increasing frequency deviation. With CW interference, failure of the receiver appeared to be caused by disruption of the automatic gain control system, which could be monitored on the front panel. Again T-DAB appears to be more susceptible to DCO interference than CW for the same total interference power in the multiplex as indicated by the increased SIR_{RMS} when the dithering is enabled. The dithering also allows more interference power to be radiated, which is manifested in the even greater increases in SIR_{QP} with the dithering present.

5 INTERFERENCE SCENARIOS AND EMISSION LIMITS

5.1 Radio System Characteristics

The interference generated by DCO enabled equipment may be of particular concern for wideband digital radio systems. Such systems typically use spread spectrum or multi-carrier modulation schemes in order to distribute the signal energy over a wide band of frequencies (frequency diversity) and thus achieve relatively high immunity to narrow band interference sources. There are many wideband digital systems currently in use, or planned for the near future, including digital broadcast systems, wireless LANs and third generation mobile telecommunication systems. This study will concentrate on the potential interference to two digital broadcast systems (DAB-T and DVB-T) and the new third generation mobile telecommunication system (UTRA). The systems considered, with their key protection requirements, are summarised in Table 5.1.

Band (MHz)	System	Minimum Field Strength (dBµV/m)	Noise Bandwidth (kHz)	Protection Ratio (dB)
217-230	Commercial DAB Broadcasting (UK) Band 3	58 [28]	1.5	14
470-590	Digital Terrestrial Television (DTTV) Band 4	44 [22]	7.6	20.0
598-854	Digital Terrestrial Television (DTTV) Band 5	48 [22]	7.6	20.0
1452-1492	Commercial DAB radio (Europe) L Band	>50 [23]	1.5	14
1920-2170	UTRA	43	3.840	5

Table 5.1: System parameters of key broadband digital radio services. For DVB and DAB the minimum field strength is for 10 m agl

5.2 Terrestrial Digital Video Broadcast (DVB-T)

5.2.1 UK DVB-T Network Parameters

Terrestrial Digital Video Broadcast (DVB-T) uses a COFDM modulation scheme. The parameters used in the UK for Band 4 and Band 5 DVB-T broadcasts are summarised in Table 5.2. The maximum BER before interference occurs is about 2×10^{-4}.

Coverage is planned on the basis of rooftop reception at a height of 10 m above ground level, with antenna directivity and cross-polar discrimination as defined in ITU-R Rec 419-3 [24] (10 dB at 500 MHz and 12 dB at 800 MHz). The minimum field strengths are 44 dBµV/m for

Band 4 and 48 dBμV/m for Band 5. An implementation margin of 3 dB is assumed to allow for multipath conditions. Coverage is much more statistical in nature than for analogue television services; the planning criteria is for service to be protected from interference 99 % of the time [25,26].

Although the UK DVB-T network is not planned on the basis of indoor reception, a useful level of indoor reception is expected to be achieved in many areas. Typical field strengths at a height of 10 m required to give 90 % coverage within 50 % of ground floor rooms are of the order 79 dBμV/m falling to 74 dBμV/m for first floor rooms. For 90 % coverage in all ground floor rooms the required field strength rises to about 90 dBμV/m [25,26]. This should be compared to the minimum field strength required for outdoor reception of 45 dBμV/m. Based on a number of studies [22] recommends assuming a building penetration loss of 7 dB with 6 dB standard deviation at UHF frequencies. Clearly indoor reception is only possible in strong signal areas close to the transmitter, particularly on a ground floor. For portable reception it is typically assumed that the antenna is 1.5 m above the floor of the room.

Noise Bandwidth	7.6 MHz
Modulation	64-QAM
Error Coding Rate	2/3
Guard Interval	7 μs
Carriers	1705 (2k-FFT)
Sub-carrier Spacing	4.464 kHz (2k-FFT)
Data Rate	24.13 Mbit/s
System C/N	20 dB

Table 5.2: UK DVB-T COFDM system parameters

5.2.2 Domestic Band 5 Interference Scenario

The first scenario considers the potential interference to the reception of both DVB-T and analogue PAL services in Band V (800 MHz) using an external roof mounted antenna due to a DCO enabled PC in a neighbouring house – see Figure 5.1. Both the digital and analogue receivers are assumed to be in a ground floor room of a two-story house with the RF feed supplied by a cable with 5 dB of loss from a Yagi antenna mounted at a height of 10 m on the roof. The DCO enabled PC is situated at a distance of about 10 m from the antenna in a loft conversion of the neighbouring house. The PC is assumed to just meet the 37 dBμV/m Class B EN55022 limit. The separation of the PC and antenna is taken to be 10 m and the interference is assumed to suffer a penetration loss of of 5 dB through the roof of the house.

The gain of the Yagi antenna in the direction of the PC is taken to be 10 dB lower than the maximum gain in the direction of the TV transmitter.

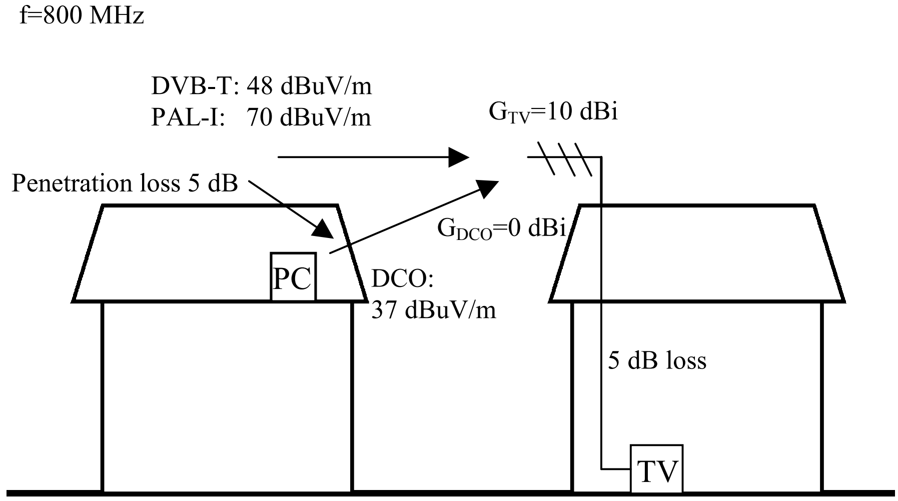


Figure 5.1: Worst-case domestic DVB-T interference scenario

	DVB-T	PAL-I	DCO
Field strength at antenna (dBμV/m)	48	70	32
Antenna gain in direction of source (dBi)	10	10	0
Antenna aperture (dBm^2)	-9.51	-9.51	-19.5
Power flux density (dBW/m^2)	-97.76	-75.76	-113.76
Received power at antenna (dBm)	-77.27	-55.27	-103.26
Cable loss (dB)	5	5	5
Received power at TV (dBm)	-82.27	-60.27	-108.26
Received voltage at TV (dBμV)	26.5	48.5	0.54
SIR_{QP} (dB)	26	48	-

Table 5.3: Domestic DVB-T interference analysis

The derived SIR for DVB-T is for the intentional RMS level signal in the whole multiplex to the quasi-peak interference level in 120 kHz. For PAL-I the signal is the RMS level of the vision carrier in 300 kHz.

For CW interference the estimated SIRs should be adequate to prevent interference to both DVB-T and PAL-I. For DCO interference the situation is more marginal based on the protection ratios estimated from the measurement given in Table 4.7, which are somewhat higher than for CW interference. Some perceptible effect could be apparent on PAL-I, though it is unlikely to be serious. However, for DVB-T and a DCO harmonic that occupies a significant proportion of the UHF channel (e.g 8^{th} harmonic of a 100 MHz or 6^{th} harmonic of a 133 MHz PC front side bus using 0.5-1 % dither) the SIR is close to the estimated limits for at which total loss of service is possible.

This scenario is an extreme worst-case for roof top DVB reception and any additional attenuation of the interference signal due to shielding of the building or polarisation mismatch will reduce the potential for interference in practice. It should also be noted that the received signal voltage level for the DVB-T signal in this example is somewhat lower than the minimum value of 45 dBμV recommended for reliable reception by the DTG [27] (this suggests that meeting the DTG requirement with the minimum field strength of 48 dBμV/m probably requires the use of a mast-head amplifier). It should also be noted that the effects of noise have not been considered. The measurements presented in [29] show that as the threshold of visibility with respect to noise is approached the required SIR_{QP} for DCO interference increases.

Clearly for indoor reception the additional interference potential for DCO equipment is much more significant; marginal indoor reception is likely to be a common occurrence with current DVB-T transmission powers and therefore any increased interference from DCOs is likely to have an impact.

5.2.3 Industrial Band 5 Interference Scenario

According to EN55022 class A equipment can radiate 10 dB more interference than class B domestic equipment (the limit above 230 MHz is 47 dBμV/m at 10 m). This fact, together with the potentially dense usage of DCO enabled equipment in a commercial environment such as an office block, makes it interesting to consider if a dense deployment of DCO enabled IT equipment poses a threat to broadcast TV reception.

First consider a single DCO enabled PC that radiates 47 dBμV/m into a DVB-T channel. If placed at a range of 30 m from the roof-top aerial described in the domestic scenario above and subjected to 5 dB penetration loss and free space propagation to the antenna then the level of interference in the DVB-T receiver will be approximately the same. At closer separations the increased emission power allowed for Class A equipment would potentially pose a greater threat than the worst-case domestic situation: Class A DCO equipment used in exactly the same configuration as the worst-case domestic scenario leading to an SIR_{QP} of 16 would present a very high risk to reliable DVB-T reception.

For multiple DCO sources the probability of the DCO harmonics overlapping both each other and the UHF channel at risk must also be factored into the assessment. The variation in the distance of the sources and the statistics of the penetration loss out of the commercial premises and the use of more appropriate propagation models for the commercial environment also make the determination of a reasonable worst-case situation much more difficult. For harmonics that do overlap it is reasonable to assume that there is no coherence between the clocks in each item of equipment and therefore the overall effect of the combined dithered clock interference in the overlapping band can be estimated by power addition. The ensemble signal also will be more ‘noise like’ in nature so it may be more appropriate to apply the protection ratio for Gaussian white noise for cumulative DCO emissions. For UK DVB-T parameters the protection ratio for white Gaussian noise is SNR=20 dB where both the signal and noise are RMS levels in the entire bandwidth of the multiplex.

If we assume that only the closest DCO equipment can contribute to the emissions (since the sources at greater distances are rapidly attenuated due to the high path loss in a typical urban environment) and that of the closest sources only a handful have commensurate penetration losses out of their buildings due the large statistical variation of the penetration loss at UHF (mean=9 dB, standard deviation=8 dB) then it is likely that only a very few DCO harmonics with significant power could arrive at the antenna. If we then factor in the probability that the harmonics actually overlap both each other and an active UHF channel then it would seem extremely unlikely for adverse effects due to cumulative DCO emissions to have a significant risk. Further work is necessary to determine the statistical distribution of the key parameters in order to quantify the risk in this scenario.

5.3 Terrestrial Digital Radio Broadcast (DAB-T)

DAB-T also uses COFDM modulation to provide robust digital radio broadcast from a network of terrestrial radio transmitters. In the UK DAB-T transmission currently only uses Band 3 (Mode I) with 1536 carriers at a spacing of 1 kHz. In the future L Band will also be available using Mode II with 384 carriers at a spacing of 4 kHz. Unlike DVB-T the more rugged differential QPSK modulation scheme is used on the individual sub-carriers.

The target CNR is typically defined to give a BER of about 5×10^{-5} with failure occurring at a BER of about 10^{-2}; a reasonable edge of service target is a BER of 10^{-4}. The relation between the BER and CNR depends on propagation conditions resulting in a large range of required CNRs depending on channel. For a static Gaussian channel a CNR of 8 dB may give good reception (with 6 dB corresponding to failure). However in a multipath (Rayleigh) channel a CNR of up to 14 dB may be needed in an urban environment with the receiver moving at moderate speed.

The UK target field strength for fixed DAB-T reception in Band 3 is 58 dBμV/m at 10 m agl; the receivers are expected to work with field strengths as low as 37 dBμV/m [28]. For DAB-T both portable (e.g. indoor portable stereo with integrated antenna) and mobile (e.g. vehicle system with antenna on roof) reception are important markets. Interference from DCOs is most likely to be an issue for portable reception. For Band 3 portable receivers it is likely that whip antennas with a gain of –6 to –2 dBd (-8 to –4 dBi) could be employed, whilst for L Band portable radios an antenna gain of –1 dBd (1.1 dBi) may be more realistic [23]. For indoor reception, penetration loss into the building may be as high as 9-20 dB. Studies of indoor reception have found an average penetration loss of 8 dB with a standard deviation of 4 dB.

5.3.1 Installed External Aerial Scenario (Band 3)

This scenario, illustrated in Figure 5. 2, is very similar to the Band 5 DVB-T scenario above. The DCO is assumed to just meet the EN55022 emission limit of 30 dBμV/m 120 kHz QP at 225 MHz. The analysis is presented in Table 5.4.

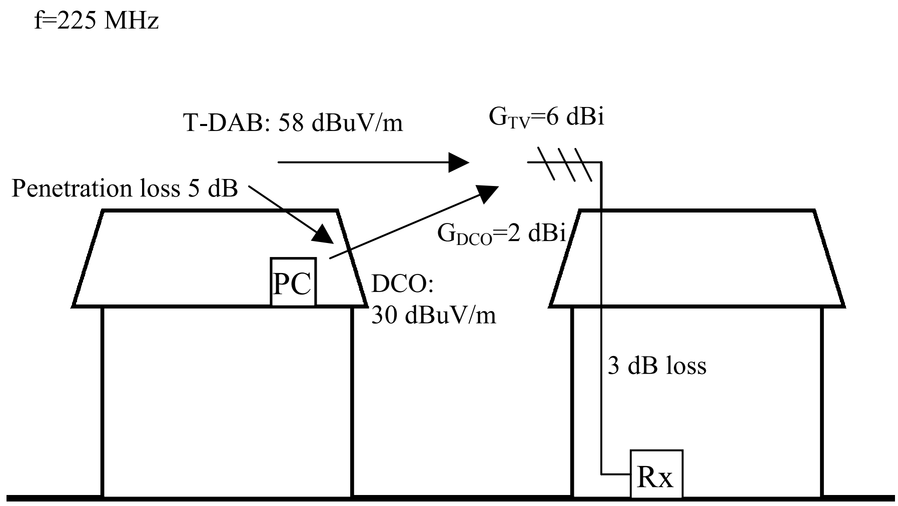


Figure 5. 2: Worst-case domestic Band 3 T-DAB interference scenario with external antenna

	T-DAB	**DCO**
Field strength at antenna (dBμV/m)	58	25
Antenna gain in direction of source (dBi)	6	2
Antenna aperture (dBm^2)	-2.49	-6.49
Power flux density (dBW/m^2)	-87.76	-120.76
Received power at antenna (dBm)	-60.25	-97.25
Cable loss (dB)	3	3
Received power at TV (dBm)	-63.25	-100.25
Received voltage at TV (dBμV)	43.75	6.75
SIR (dB)	37	-

Table 5.4: Domestic Band 3 T-DAB interference analysis for external antenna

In Section 4.4.2 the worst-case SIR required for T-DAB was estimated to be around 10 dB. This suggests that there is a large margin of reliable reception of T-DAB using external aerials even in a worst-case DCO scenario.

5.3.2 Indoor Band 3 Reception Scenario

Indoor reception is a much more critical market for T-DAB than it is for DVB-T. In this scenario the potential for interference between an indoor Band 3 T-DAB receiver and a DCO enabled PC in a neighbouring house is considered (see Figure 5.3). The intentional field strength at the T-DAB antenna is assumed to be 37 dBμV/m, which is the minimum field strength for which the receiver is expected to work [23]. This could be achieved with a 10 m agl field strength of 58 dBμV/m and a building penetration loss of 18 dB.

f=225 MHz

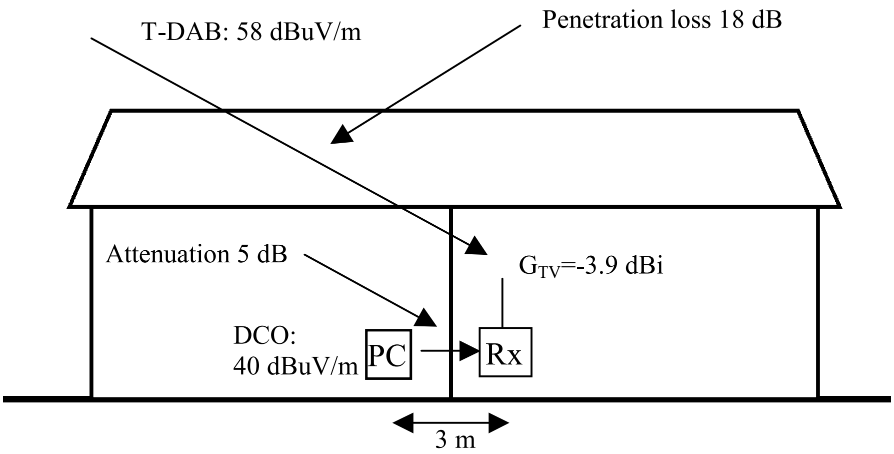


Figure 5.3: Worst-case domestic indoor Band 3 T-DAB interference scenario

The PC is assumed to just meet the EN55022 emission limit of 30 dBμV/m at 10 m which translates to about 40 dBμV/m at 3 m. An attenuation of 5 dB through the wall is also included giving an interference field strength of 35 dBμV/m at the T-DAB antenna. The analysis is given in Table 5.5.

	T-DAB	DCO
Field strength at antenna (dBμV/m)	37	35
Antenna gain in direction of source (dBi)	-3.9	-3.9
Antenna aperture (dBm^2)	-12.39	-12.39
Power flux density (dBW/m^2)	-108.76	-110.76
Received power at antenna (dBm)	-91.15	-93.15
Antenna loss (dB)	1	1
Received power at TV (dBm)	-92.15	-94.15
SIR_{QP} (dB)	2	-

Table 5.5: Domestic indoor Band 3 T-DAB interference analysis

The predicted SIR_{QP} of only 2 dB is adequate to protect the service against CW interference in the T-DAB channel. For worst DCO interference (say, 0.6% dither on the 2nd harmonic of a 112.5 MHz clock or 8th harmonic of a 28.125 MHz clock) an SIR of about 10 dB is required [10] suggesting a serious interference potential for indoor reception of T-DAB.

5.3.3 Indoor L Band Reception Scenario

f=1470 MHz

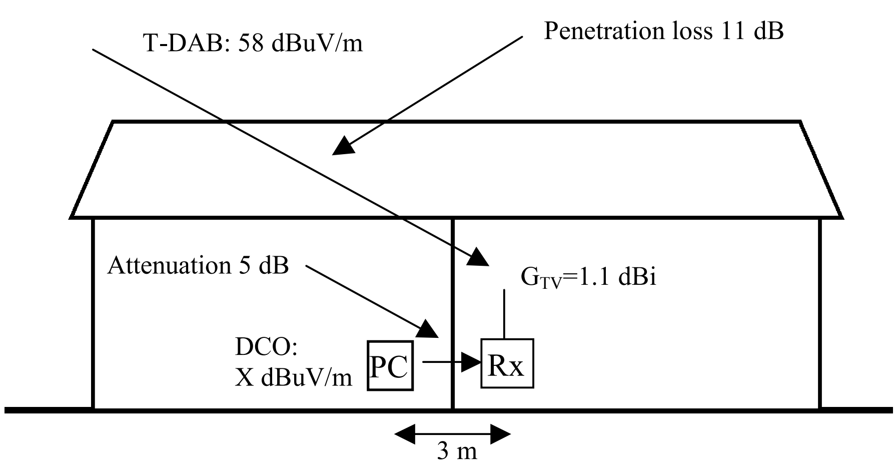


Figure 5.4: Worst-case domestic indoor L-Band T-DAB interference scenario

Although L band DAB is not currently used in the UK it is interesting to consider the implications of DCO emissions in this band. The same geometry is used as in the scenario above with the DAB-T parameters and fields strengths (Figure 5.4) taken from the optimised L Band analysis in [23].

Currently there are no enforced standards for radiated emissions above 1 GHz in the UK. There are however two relevant standards of interest:

1. FCC Part 15 1999 defines a radiated emissions limit above 960MHz for unintentional radiators of 54dBμV/m at 3m measured in a bandwidth of 1 MHz using a peak detector.
2. CISPR11 (1998 + A1 1999) - ISM equipment defines an emission limit for Group 2 class B for equipment operating at frequencies above 400MHz of 70dBμV/m (peak) in the band 1-18 GHz (60dBμV/m with video bandwidth of 10Hz) measured at 3m in a resolution bandwidth of 1 MHz.

The measurement bandwidth of 1 MHz used in these standards encompasses most of a T-DAB multiplex in L band. If we take the CISPR 60 dBμV/m limit then the analysis is given in Table 5.5 can be performed.

	T-DAB	DCO
Field strength at antenna (dBμV/m)	47	60
Antenna gain in direction of source (dBi)	1.1	1.1
Antenna aperture (dBm^2)	-23.70	-23.70
Power flux density (dBW/m^2)	-98.76	-71.76
Received power at antenna (dBm)	-92.45	-79.45
Antenna loss (dB)	2.5	2.5
Received power at TV (dBm)	-95.0	-82.15
SIR (dB)	-12.9	-

Table 5.6: Domestic indoor L Band T-DAB interference analysis

In order to relate this to the critical SIRs for T-DAB measured earlier the peak level of the DCO interference in 1 MHz has to be related to the QP level in 120 kHz. Estimating:

$$I(QP,120\text{ kHz}) = I(QP, 1\text{ MHz}) - 10\log_{10}(1000/120) = I(QP, 1\text{ MHz}) - 9.2\text{ dB}$$

$$I(QP,120\text{ kHz}) = I(Pk, 120\text{ kHz}) - 3\text{ dB}$$

We find that

$$SIR_{QP} = SIR + 12\text{ dB} \sim -1\text{ dB}.$$

This suggests that the CISPR limit would be sufficient to protect L – Band DAB from CW interference but may be 3-10 dB too low for DCO emissions.

5.4 Measurement Environments and Emission Limits

5.4.1 Discussion of DCO Interference Issues

The interference from DCO enabled equipment consists of a harmonic spectrum of components $f_n=n \cdot f_{clk}$ (n=1, 2,3, …) at multiples of the clock frequency f_{clk}. Each harmonic is itself composed of sub-harmonics separated by the DCO modulating frequency f_{rep}. The fractional bandwidth, $\beta=\Delta f_n/f_n$, of each harmonic is the same and equal to the fractional dithering deviation. The number of sub-harmonics, with significant energy in each harmonic can be estimated to be $N_n=\Delta f_n/f_{rep}= \beta f_n/f_{rep}$. The exact envelope of the DCO sub-harmonic spectrum depends on the modulating waveform and the receiver characteristics. In measurement bandwidths greater than the dithering rate the discrete nature of the DCO spectrum is not readily apparent since multiple sub-harmonics are detected simultaneously. The occupied bandwidth of DCO emissions increases with frequency according to $\Delta f_n=\beta f_n=\beta n f_1$; higher order harmonics therefore attain greater apparent level suppression in a CISPR QP detector than the lower ones. A detailed discussion of these effects can be found in [10]. The discrete nature of the spectrum of a DCO signal may have an important impact on the interaction between DCO interference and wideband digital radio receivers. In particular, the protection ratios for radio receivers to Gaussian white noise and CW interference are not directly applicable to DCO interference.

DCO interference raises two important questions with respect to protecting radio services from interference:

1. Does dithering (with a constant total interference power in a radio channel) affect the immunity of broadband radio services?
2. What are the implications of the increased total radiated interference power that dithering a clock makes possible?

Based on the measurements and simulations undertaken in this study the answer to the first question for DVB-T and T-DAB is *yes*. Although the symbol period of each COFDM sub-channel is not long enough (compared to the dithering period) for a frequency domain picture to be sufficient to describe the interference mechanism, the sub-channels close to sub-harmonic frequencies of the DCO do receive significantly more interference power from the DCO signal than the rest. If the dithering rate is a multiple of the sub-channel bandwidth then all the interference power is concentrated into a small number of sub-channels, though this is a special case. The COFDM demodulator is in principle able to cope with very high interference levels in a few tens of sub-channels due to the redundancy in the error coding schemes employed. In a static propagation channel, DCO interference at relatively high levels in a number of sub-channels can therefore be tolerated. However, in a fading channel, where a number of sub-carriers in the multiplex are received with insufficient power to be reliably demodulated, the redundancy in the coding is required to mitigate against the fading. Any further loss of sub-channels due to DCO interference is therefore likely to be more significant in a fading channel. For worst-case DCO parameters the simulations suggest that a signal to interference ratio of about 8 dB is the critical value below which DVB-T reception fails. Here, both the signal and interference are measured in their entire bandwidth using a RMS detector. If the interference is measured with a CISPR quasi peak detector the appropriate value is SIR_{QP}=22 dB.

The measurements verify that DVB-T is extremely robust to CW interference. In fact the modulation scheme itself is infinitely robust to CW interference since it can tolerate the complete loss of a number of sub-channels. In practice, the receiver implementation will limit the immunity of the system to CW and it is therefore not surprising that there was found to be significant differences between the two set top boxes measured (see section 4). As predicted DVB-T is less robust to DCO interference than it is to CW. The total interference power required to cause failure with DCO enabled was 5-18 dB lower than for CW. The wide range represents the large difference in immunity to CW interference between the two set top boxes tested. In absolute terms the total DCO interference power required to cause failure was very similar for the two STBs tested and showed little variation (+/-1 dB) with the modulation shape, rate or deviation. For worst-case DCO parameters (i.e. frequency deviation equal to the multiplex bandwidth) the STBs failed at a signal to interference level of about SIR_{QP}~30 dB. This SIR_{QP} is consistent with other measurements on DVB-T [10,29], though it is higher than predicted by the theoretical analysis and simulations. Similar trends were found in the T-DAB measurements; however, T-DAB is generally more robust than DVB-T. The SIR_{RMS} for critical BER was measured to be about –5 dB for worst-case DCO parameters, corresponding to SIR_{QP}~2.2 dB. This is somewhat lower than the value of SIR_{QP}=10 dB measured by Hertfordshire [10].

Since the interaction between DCO interference and digital radio systems is quite complex further work is necessary to rigorously determine the effect on other systems such as UMTS.

To answer the second question posed above it is instructive to consider the effect of DCO interference on PAL-I receivers. The measurement results show that PAL-I is relatively insensitive to the dithering, at least when the interference is centred on the vision carrier; the same total interference power in the PAL-I channel causes a similar level of degradation (though the form of the pattering is somewhat different) regardless of whether or not the clock is dithered. The main effect of the dithering is therefore to allow higher interference power to be radiated into the radio channel and therefore increase the potential for interference. This is equally true for DVB-T and T-DAB, which therefore suffer a two-fold adverse effect: increased interference power and heightened sensitivity to the modulated (versus CW) interference.

Initial analysis suggests that UMTS is likely to be susceptible to total interference power in the channel and therefore also at increased risk from DCO interference. Further work is necessary to quantify any effect and determine if it is significant.

While the use of dithering clearly has an adverse effect on the performance of DVB-T and T-DAB receivers when compared to a pure CW signal it must be considered that DVB-T/T-DAB are both extremely robust to CW interference and therefore that other comparisons may be more appropriate when assessing the practical interference potential of DCOs. For example, the measurements on PAL-I reception demonstrate that analogue television has less immunity to both CW and DCO interference than DVB-T. For CW interference over 12 dB less total interference power was required to cause interference to PAL broadcasts than was needed for DVB-T. With dithered interference the difference was reduced to about 8 dB. This change was entirely due to the reduced immunity of DVB since analogue reception was relatively unaffected by the dithering, requiring the same total interference power to cause comparable observable effects with both CW and various DCO signals. Nevertheless DVB-T remained more robust than PAL to the interference, by about 8 dB, even with dithering enabled. This raises a third question:

3. What are the “real world” consequences of clock dithering and are revised EMC measurement methods and/or limits necessary?

This is a very difficult question to answer in general since it crosses into uncertain and political areas such as coverage, borderline reception and the more statistical nature of reception in broadband digital systems such as DVB-T. A number of scenarios for DVB-T and T-DAB reception have been analysed to attempt to answer this question.

For reception of broadcast television services using an external antenna the scenarios considered suggest that although DCOs have an adverse effect on both PAL-I and DVB-T the higher power levels of analogue transmission and high immunity of DVB-T mean that only in a worst-case is roof-top reception likely the be less reliable due dithering. Without dithering there is significant margin for reliable reception for both PAL-I and DVB-T. Although indoor reception is not a planned service it is expected than in many areas close to the transmitter a reasonable level of indoor coverage may be available. The potential for adverse affects from DCOs to indoor reception are much greater and likely to be significant.

Roof top reception of T-DAB will be unaffected by DCO emissions (assuming the system is correctly installed). T-DAB is much more robust than DVB-T to DCO signals, however, the increased immunity is required to offset the much more demanding portable and mobile reception markets in which T-DAB is designed to operate. T-DAB must cope with more severe fading and higher variability in the propagation conditions than DVB-T. The indoor Band 3 T-DAB reception scenario demonstrates that for the lowest intentional signal strengths for which T-DAB is expected to work the use of a DCO enabled PC in a neighbouring house could cause interference to reception. This should be compared to the same scenario in which dithering is not used in the PC where there is a significant margin for reliable reception.

The interference scenarios have assumed that the level of emissions into the receiver channel is the maximum allowed by the current emission standards and that the DCO interference optimally overlaps the radio channel. Worst-case, though realistic coupling scenarios between the DCO equipment and radio receiver have been assumed. The practical impact of DCOs will depend on how likely these circumstances are to occur.

The following sections provide details of emission measurement techniques and limits which could be applied to limit the effects of DCO on radio services.

5.4.2 Wideband Emission Measurements

The most direct approach to devising a measurement procedure that limits the potential increases in radiated emissions from DCO enabled equipment is to specify a larger measurement bandwidth and thereby offset the effect of the dithering. For example, many spectrum analysers have a 3 MHz resolution bandwidth setting. This approach is a move towards a total power measurement and effectively limits the total power in the DCO interference across the band regardless of the amount of dithering employed. For spread spectrum systems such as UMTS, which are likely to be sensitive to the total interference power in their channel, a wideband emissions measurement could be mapped directly onto the relevant protection requirement and hence a limit. Systems using COFDM and similar modulation schemes, which have a complex response to DCO interference, would also be afforded increased protection though the actual limit applied would be dependent on the distribution of energy within the measurement bandwidth and therefore the dithering algorithm. Provision would also have to be made for situations in which a number of narrow band interferes lie in the measurement bandwidth potentially causing the EUT to fail when it would have passed a conventional emissions test. However, for the reasons outlined below a wideband emissions test may be a highly contentious issue.

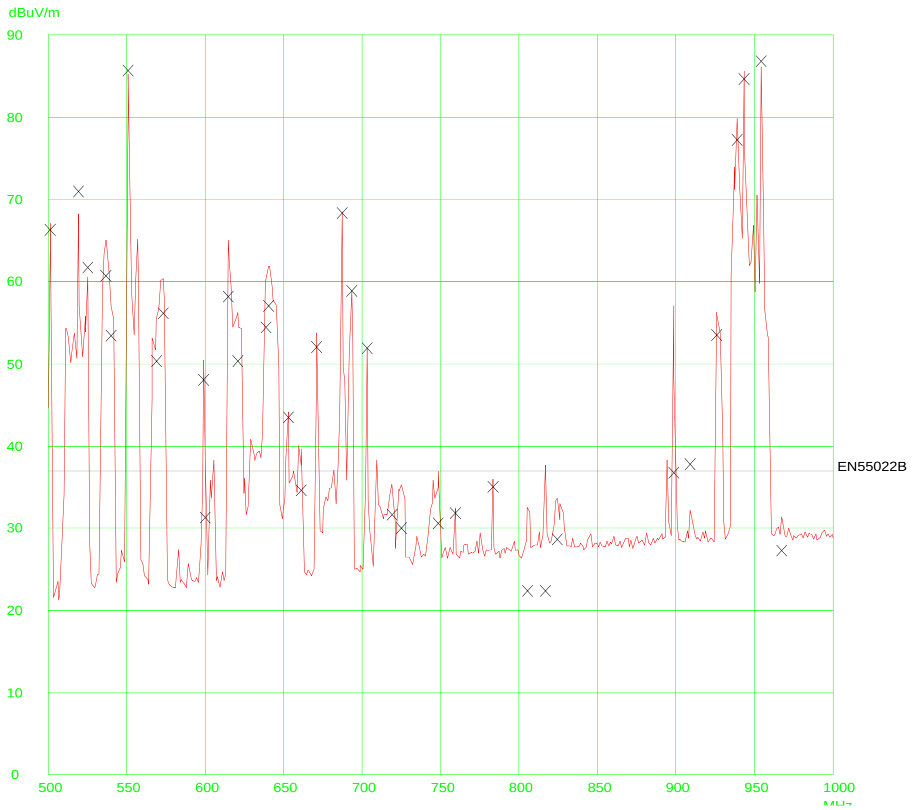


Figure 5.5: Horizontal ambient on typical UKAS accredited UK test site

Figure 5.5 shows the ambient signal level at a typical UK test site for a peak measurement using a 120kHz bandwidth. Also shown on the figure is the EN 55022 class B limit. At 1GHz the noise floor of the measurement system is only 7dB below the limit. The measurement system has an uncertainty of 4.3dB at 1GHz. Consequently there is only 2.7dB between the bottom of the measurement uncertainty and the noise floor of the measurement system. This is the minimum practical margin for a UKAS accredited test laboratory. If the margin were any less it would be almost impossible to say with certainty whether an EUT had passed the test or not, which would be an impossible situation for a test house to be in. The situation gradually improves as the frequency decreases below 1GHz. Increasing the measurement bandwidth from 120kHz to 3MHz would significantly increase the noise floor of the measurement system. On a Hewlett Packard 8594E spectrum analyser, the noise floor of the instrument increases by 8dB, when the bandwidth is changed from 120kHz to 3MHz. Clearly, this increase could not be tolerated for a 10m measurement distance. Decreasing the measurement distance to 3m would increase the limit by 10dB so a standard based on a 3m measurement may be possible. For a 3m measurement distance there is some issue with the possibility of near field effects for large EUTs [30]. EN 50147-3 [31] deals with this precisely by stating maximum EUT dimensions for given test distances. At 3m, for example, the maximum EUT diameter is 1.5m. Limiting the EUT size implies that for a large EUT, only the testing of subsystems would be possible. Increasing the measurement bandwidth would therefore lead into a debate on ensuring EMC for large EUTs which is a highly contentious issue.

5.4.3 Modified 120 kHz Measurement Procedures and Limits

An alternative approach proposed in [10] is to amend the existing EN55022 measurement protocols and limits to account for the level suppression of emissions from DCO equipment in a 120 kHz bandwidth CISPR quasi-peak detector. In essence the method seeks to enforce a lower limit for emissions that are broadband in nature while retaining the existing limit for narrow band emissions. Narrow band and broadband (DCO) emissions are discriminated by identifying emissions which are close enough to the limit (within 10 dB) that they could exceed the limit if dithering were to be disabled. In order to assess the nature of such emissions the bandwidth, Δf, of the emission is estimated by a prescribed procedure. If Δf is <~150 kHz the usual limit (L dB) specified in the standard is applied, otherwise the limit is reduced by 10 $\log_{10}$(Δf/120 kHz) or 10 dB, whichever is the lower reduction.

This method has been demonstrated to be just about practical on an OATS with standard EMC test equipment where measurements at levels much lower than 10 dB below the current limits are not feasible*. The technique is effective at limiting the total power in a DCO emission with a bandwidth less than 1.2 MHz to the about the same level that it would have if dithering was not employed. This should protect radio systems like PAL-I that are affected by the total interference power in their channel bandwidth but are not particularly sensitive to the dithering itself.

For COFDM systems whose immunity appears to be sensitive to the dithering it could be argued that even more protection may be appropriate. While the reduced limit outlined above may in fact be adequate for protection of DVB-T reception using a roof-top aerial in most cases it would not precluded the possibility of interference. Roof-top reception of T-DAB does not appear to be at risk. For indoor reception of both DAB-T and DVB-T a further reduction in the limit of about 10 dB for worst-case DCO parameters may be necessary to protect the service. This would realistically require an anechoic chamber in order to reduce the noise floor to a level at which the measurement procedure could be carried out.

In general we can therefore say that the application of a reduced limit of

$$L_{\mathrm{new}} = L - 10\,\log_{10}\left[\frac{\Delta f}{120\ \mathrm{kHz}}\right]$$

where Δf is the bandwidth occupied by a DCO emission (in kHz) and L is the existing EN55022 limit (in dB) appears to be a reasonable approach to protecting most wideband radio services from the increased interference power of DCOs. In [10] the reduction in the

* Presumably the figure of 10dB ignores the issue of measurement uncertainty discussed in section 5.4.2.

emission limit is a maximum of 10 dB if the bandwidth of the DCO emission exceeds 1.2 MHz. However, for COFDM systems which are also much more sensitive to the particular characteristics of DCO interference an even stricter new limit of

$$L_{\text{new}} = L - 20\log_{10}\left[\frac{\Delta f}{120\ \text{kHz}}\right]$$

may be necessary to provide sufficient margin for the reliable indoor reception of DVB-T and T-DAB in worst-case situations. This limit reduction could also be limited to a maximum of 20 dB if the DCO emission bandwidth exceeds 1.2 MHz.

These limits are likely to be very difficult for equipment manufacturers to achieve, after all DCOs were developed because of the increasing difficulty of meeting EN55022 with modern high-speed electronic systems. It is therefore important that some attempt is made to estimate the actual probability of interference occurring in practice to determine if the increased burden of the stricter limits can be justified. For COFDM systems this is not a straightforward exercise as many factors would need to be considered including the following: detailed statistical information on coverage levels, propagation conditions, atmospheric variation, penetration losses, fade margins, population distributions relative to transmitter locations.

5.4.4 Power Spectral Density Measurements

The theoretical analysis of COFDM argues in favour of protecting each sub-channel (~4 kHz bandwidth for DVB-T and 1 kHz for DAB-T) from interference across the entire multiplex to ensure that reception is not adversely affected by DCO interference. This approach would be a move towards a power spectral density type emissions limit (i.e. a measurement in a relatively small bandwidth). In order not to penalise undithered EUTs (or undithered subsystems within a single EUT) provision would have to be made to allow for higher levels of emissions from narrow band sources. This would require a procedure to similar to the above for discriminating between narrow band and broadband emissions. The theoretical analysis suggests that DVB-T reception is reliable providing the signal to interference level in a sub-carrier bandwidth, SIR_{sub}, is greater than about 5 dB. Developing an emission measurement standard along these lines would represent a major departure from the current practice for the reasons outlined below.

Reducing the measurement bandwidth has a highly beneficial effect on the measurement system noise floor. Reducing the bandwidth from 120kHz to 1kHz on a Hewlett Packard 8594E spectrum analyser gives a 28dB reduction in measurement system noise floor. On this basis it would be quite possible to significantly reduce the EN 55022 class B limit and still be able to make meaningful measurements on an UKAS accredited open field test site at 10m distance. However, there are two major problems with doing this. Firstly the measurement time would be vastly increased. Some investigation of current receiver technology is required to determine whether:

1. receivers with the required bandwith in the 30MHz to 1GHz range are available (standard 30MHz –1GHz EMC test receivers do not possess a 1kHz rbw)
2. the measurement time associated with a single receiver (if it exists) would be a practical proposition

Secondly if the class B limit is reduced then the apparent level of ambient signals with respect to the limit is increased. This would imply that no sensible measurements of emission could be made on an open field test site. Any measurements would need to be made in an anechoic chamber probably in accordance with EN 50147-3.

6 CONCLUSIONS

1. The immunity of DVB-T receivers to DCO clock oscillator interference or CW interference is not as great as the modelling suggests it should be. This is thought to be due to design specific effects in the RF tuners not included in the model. It seems that potentially the immunity of the receivers to DCO interference could be significantly improved. Similar problems are thought to exist for the DAB receiver investigated. Note that the simulation assumes no white noise, while in the trials noise is present, albeit at a much lower level than the DCO interference. However the structure of the interference means that the receiver might be susceptible to low levels of white noise mixed with the interference.

2. The use of DCOs allows manufacturers to emit more power with respect to current EMC test limits. For existing analogue PAL receivers, the consequence of this is that they are subject to greater interference power than from non-DCO sources. For COFDM receivers, such as DVB-T, not only are the receivers subject to greater power than from non-DCO sources they are also more susceptible to DCO noise than CW noise. The use of DCOs therefore increases the chances of interference to both DVB-T and PAL when compared with using a non-dithered clock.

3. The probability of serious interference to TV reception is greater with the pairing of DVB-T and DCOs than with the pairing of PAL and non-DCO computers, therefore some action should be taken. Four courses of action exist:

    a. Improve the immunity of the DVB-T receivers (see point 1). The immunity of the receivers seems to be influenced by the tuner design. This suggests that an alternative approach to reducing the emissions limit would be to formulate an in band immunity test for receivers. In this way, the design of the receiver would be expected to approach the modelled limit. However, further work is required to ascertain the susceptibility of the ideal receiver to a mixture of white noise and DCO.

    b. Increase the power of DVB-T transmission. This is already happening to some extent for other reasons.

    c. Impose more stringent EMC requirements for DCO enabled products. Imposing significantly more stringent EMC requirements for DCO enabled products (10dB or more) would require radical departures from the current test methods and would potentially lead to manufacturers reverting to non dithered clocks. Existing test methods could stand a reduction of 3dB in the limit over the DVB-T frequency range (470MHz to 854MHz) and on the analyses presented should significantly reduce the likelihood of reported interferences.

If it is possible to build IT equipment that meets these more stringent requirements then this is one way of improving the EMC of DVB-T receivers in the short to medium term.

d. Tolerate an increase in the number of reported cases of interference.

4. Both indoor DAB reception and L-Band DAB reception are likely to be seriously impaired by the presence of DCO enabled computers. The DAB receivers are generally more immune but the interference sources are potentially closer. The situation appears less marginal than the DVB-T case so any reduction in the emissions limit would have to be larger than that suggested for DVB-T. In the case of DAB a significant reduction in the current limit is possible based on the noise floor of a typical measurement system relative to the current limit. A reduction of 8dB in the EN 55022B limit would be possible, measurements could still be performed using conventional measurement equipment and give much improved protection for DAB. It is proposed that the reduction would be over the DAB frequency range of (217MHz to 230MHz). Once again, if it is possible to build IT equipment that meets these more stringent requirements, then this is one way of ensuring the EMC of DAB receivers in the short to medium term. Although the reduction could be introduced as an amendment to EN 55022 a reasonable quality measurement could only be made in a shielded anechoic room. This is very much the case with current measurement practice, so the 8dB reduction should be entirely practical.

7 FURTHER WORK

It has been concluded that DCO enabled equipment has a smaller margin of Electromagnetic Compatibility between it and broadcast services than none DCO enabled equipment. Furthermore it was found that the digital receivers investigated were less immune than might be expected and their immunity to both DCO and CW interference was thought to be dependent on the tuner design. Two possible approaches to improving the EMC margin were suggested and further work would therefore investigate:

1. Practical tests with the proposed new emissions limits should be undertaken at a UKAS accredited laboratory. Tests on currently available IT equipment should be conducted to assess the likely impact of such limits on the IT industry. The tests would need to be conducted in partnership with a major IT industry so that it can be established if it is practical to build equipment that can meet the requirements of the new proposed limits.
2. Tuner design needs to be investigated to determine if the modelled values of DCO immunity can be obtained in practical designs. This would involve further modelling work to determine if white noise superimposed on DCO noise could cause the reduced immunity observed in practical receivers. If improved levels of immunity can be obtained in practical designs, then an in-band DCO immunity test should be devised, with the intention that it is implemented as a European standard so that digital receiver immunity to DCO noise is maximised. This activity would have to be carried out with the close involvement of a major manufacturer of set top boxes and/or DAB receivers.

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