CEPT ECC
Electronic Communications Committee

ECC Report 356

Technical analysis to support the ECC Recommendation on receiver resilience to transmission on adjacent frequency ranges

approved 10 May 2024

0 EXECUTIVE SUMMARY

This Report describes the method for receiver resilience “MRR” to calculate the receiver resilience levels, namely the adjacent channel protection ratio (PR) and blocking levels, to be included in the ECC Recommendation on receiver resilience to transmission on adjacent frequency ranges [1].

Section 2 presents the definition of the most important terms used in this Report. Section 3 provides the basic theory of receiver resilience and presents the derivation of the basic equation of MRR. Section 4 presents in detail the use of MRR to calculate/specify the receiver resilience levels to be included in the ECC Recommendation on receivers. The last section of the report, Section 5, presents the conclusions of the Report. Extensive information on the work carried out within the development of MRR is provided in ANNEX 1: to ANNEX 6:. MRR is a technology neutral generic method. It uses four parameters N, M, ILR and FOS to calculate the receiver resilience levels by a single equation derived from fundamental equations describing the mechanism of receiver resilience to a frequency offset interfering signal.

The consistency of MRR with existing relevant standards was assessed by calculating receiver resilience levels defined in eight different ETSI Harmonised Standards for numerous receiver configurations. The results obtained showed good agreement with the adjacent channel PR and blocking levels defined in the relevant standards for six of the eight assessed but can also result in more stringent blocking levels e.g. in the case of DAB and DTTB.

Based on the analysis carried out in this Report, the receiver resilience levels defined in the ECC Recommendation on receivers are determined as follows:

- In the Out-Of-Band (OOB) domain, the resilience levels have been determined from the intra-system interfering signals defined in relevant standards; In the reciprocal spurious blocking domain (RSBD), the resilience levels have been determined based on:
  - i) a reference interfering signal (OFDM signal) having SE (spurious emissions) levels in accordance with the most common SE levels defined in ERC Recommendation 74-01 [2]; and
  - ii) the equivalent blocking levels for a CW interfering signal to allow the conformance tests to be done with a CW signal as an alternative;

It should be noted that MRR is intended to calculate the resilience levels of receivers in the presence of a single frequency offset interfering signal as used in blocking tests in relevant standards. It is not intended to evaluate the impact of undesired signal variations, inherent in receivers, on receiver performance (e.g., to predict intermodulation response rejection or to model non-linear receivers.), or to predict the behaviour of a specific receiver or for system design/development.

There are some types of receivers, e.g. receivers used by passive radio services, and those below 9 kHz where the applicability of the MRR would require further investigation.

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LIST OF ABBREVIATIONS

Abbreviation	Explanation
CEPT	European Conference of Postal and Telecommunications Administrations
ECC	Electronic Communications Committee
ACLR	Adjacent Channel Leakage Ratio
ACS	Adjacent Channel Selectivity
A/D	Analog/Digital
ADC	Analog Digital Converter
BB	Base Band
BS	Base Station
BER	Bit Error Rate
BW	Bandwidth
C/I	Carrier-to-Interference ratio
CW	Continuous Wave
D/A	Digital/Analog
DAB	Digital Audio Broadcasting
DFRS	Digital Fixed Radio Systems
DRM	Digital Radio Mondiale
DTT	Digital Terrestrial Television
DVB-T	Digital Video Broadcasting – Terrestrial
ECSD	Enhanced Circuit-Switched Data
EIRP	Effective Isotropic Radiated Power
ETSI	European Telecommunications Standards Institute
E-UTRA	Evolved Universal Terrestrial Radio Access
FOS	Frequency Offset Selectivity
FPGA	Field Programmable Gate Array
FRB	Fast Radio Burst
FS	Fixed Service
GSM	Global System for Mobile Communications
HS	Harmonised Standards
HTS	High Temperature Superconductor
IF	Intermediate Frequency

Abbreviation	Explanation
ILR	Interference Leakage Ratio
IMT	International Mobile Telecommunications
IoT	Internet of Things
LAN	Local Area Network
LNA	Low Noise Amplifier
LTE	Long Term Evolution
M	Receiver desensitisation
NB	Narrow-Band
N	Noise Floor
NF	Noise Figure
NFD	Net Filter Discrimination
OMT	Orthogonal Mode Transducer
OFDM	Orthogonal Frequency Division Multiplexing
OOB	Out-Of-Band
OOBD	Out-Of-Band Domain
OOBE	Out-Of-Band Emissions
PAF	Phased Array Feed
PAPR	Peak-to-Average Power Ratio
PtP	Point-to-Point
PR	Protection Ratio
QAM	Quadrature Amplitude Modulation
QoS	Quality of Service
RAS	Radio Astronomy Service
RF	Radio Frequency
RI	Reference Interferer
RIR	Receiver Interference Ratio
RLAN	Radio Local Area Network
RMS	Root-Mean-Square
RSBD	Reciprocal Spurious Blocking Domain
RQZ	Radio Quiet Zone
MRR	Method for Receiver Resilience
Rx	Receiver
SD	Spurious Domain
SE	Spurious Emissions
SEAMCAT	Spectrum Engineering Advanced Monte Carlo Analysis Tool
SINR	Signal to Interference and Noise Ratio
S/N	Signal/Noise
SNR	Signal to Noise Ratio

Abbreviation	Explanation
SRD	Short Range Devices
Tx	Transmitter
UWB	Ultra-Wide Band
VGOS	VLBI Global Observing System
WBDT	Wide-Band Data Transmission
VLBI	Very Long Baseline Interferometry
WBSEL	Wide-Band Selectivity
WiMAX	Worldwide Interoperability for Microwave Access

1 INTRODUCTION

This technical report aims to develop a methodology, which will be part of the ECC Recommendation on receiver resilience to transmission on adjacent frequency ranges, to be used for specifying recommended levels of receiver resilience. Two different levels of receiver resilience are specified in the Recommendation; these are receiver adjacent channel PR ($PR_{adj\text{-}ch}$) and blocking level (I_{blk}), both with a related frequency offset selectivity (FOS).

Section 3 provides the basic theory of receiver resilience and section 4 presents the method for receiver resilience (MRR) developed to calculate/specify the receiver resilience levels to be included in the ECC Recommendation on receivers [1].

2 DEFINITIONS

The definitions below, with the exception of “Receiver desensitisation” " Reciprocal spurious blocking domain" and Receiver desensitiation (M)”, are taken from ECC Report 310 [3] with some small modifications and are used in this Report for receivers working in their linear range.

Term	Definition
Blocking	A measure of the receiver capability to receive a wanted signal without exceeding a given degradation due to the presence of an unwanted signal at any frequency other than those of the spurious responses or of the adjacent channels and it is defined as the maximum interfering signal level expressed in dBm reducing the specified receiver sensitivity by a certain number of dBs (desensitisation).
Frequency Offset Selectivity (FOS)	A measure of the receiver ability to receive a wanted signal at its assigned channel frequency in the presence of an unwanted adjacent signal at a given frequency offset from the centre frequency of the assigned channel. In this context, it is defined as the ratio of the receiver filter attenuation on the offset frequency to the receiver filter attenuation on the assigned channel frequency (normally a positive number in dB). FOS is of general use for any mixed wanted and unwanted signal situation.
Interference Leakage Ratio (ILR)	The ratio of the filtered mean power centred on the assigned channel frequency to the similarly filtered mean power centred on a given frequency offset.
Receiver Interference Ratio (RIR)	The ratio of the in-channel interference power on a given frequency offset to the interference power received by the victim receiver.
Receiver desensitisation (M)	Reduction in the signal to noise ratio of the receiver or a reduction in the effective sensitivity in the presence of an interfering signal, given in dB. t corresponds to the ‘noise rise’ due to the interfering signal.
Receiver noise floor (N)	The total noise power at the receiver including the effect of thermal noise and the receiver noise figure.
Reciprocal spurious blocking domain (RSBD)	Implies that the victim receiver channel is in the spurious domain of the interfering transmitter and reciprocally the interfering transmitter channel is in the spurious domain of the victim receiver.

3 BASIC THEORY ON RECEIVER RESILIENCE

It is important to understand the mechanism of receiver resilience to a frequency offset interfering signal before attempting to develop a methodology to be used for specifying recommended levels of receiver resilience to transmission on adjacent frequency ranges, which are blocking (I_{blk}) or adjacent channel protection ratio ($PR_{adj - ch}$).

The resilience of a receiver to a frequency offset interfering signal depends on the receiver's overall performance based on filtering, linearity impairment effects, demodulation and decoding techniques used. Nevertheless, it can be defined by the receiver blocking (I_{blk}), frequency offset selectivity (FOS) or receiver interference ratio (RIR) levels.

Note that in the theoretical development presented in the following sections the receiver intermodulation response rejection is not considered.

Receiver intermodulation response rejection is a measure of the capability of the receiver to receive a wanted signal, without exceeding a given degradation due to the presence of at least two interfering signals at frequencies f1 and f2, with a specific frequency relationship to the wanted signal frequency, which give rise to 2nd and 3rd order intermodulation products that may fall into the victim receiver channel.

However, the receiver intermodulation response rejection test is not included in all ETSI Harmonised Standards, because intermodulation in the RF tuner will result in a degradation in the adjacent channel selectivity, which is extensively tested in many standards. Moreover, real-life experience indicates that interference due to intermodulation is not a widespread problem between different system/services in Europe.

For example, relevant standards dealing with DTT, DAB and fixed service do not include a specific intermodulation response rejection test. The impact of the intermodulation products on the receiver performances are covered by adjacent channel selectivity tests. The method developed in this Report aims to derive blocking or selectivity levels of a receiver in the presence of a single interfering signal as it is done in all relevant standards.

There are some types of receivers, e.g. receivers used by passive radio services, and those below 9 kHz, where the applicability of the MRR for these receivers would require further investigation. Relevant information about the receiver resilience of RAS can be found in ANNEX 8:.

3.1 FREQUENCY OFFSET INTERFERING SIGNAL POWER RECEIVED BY THE VICTIM RECEIVER

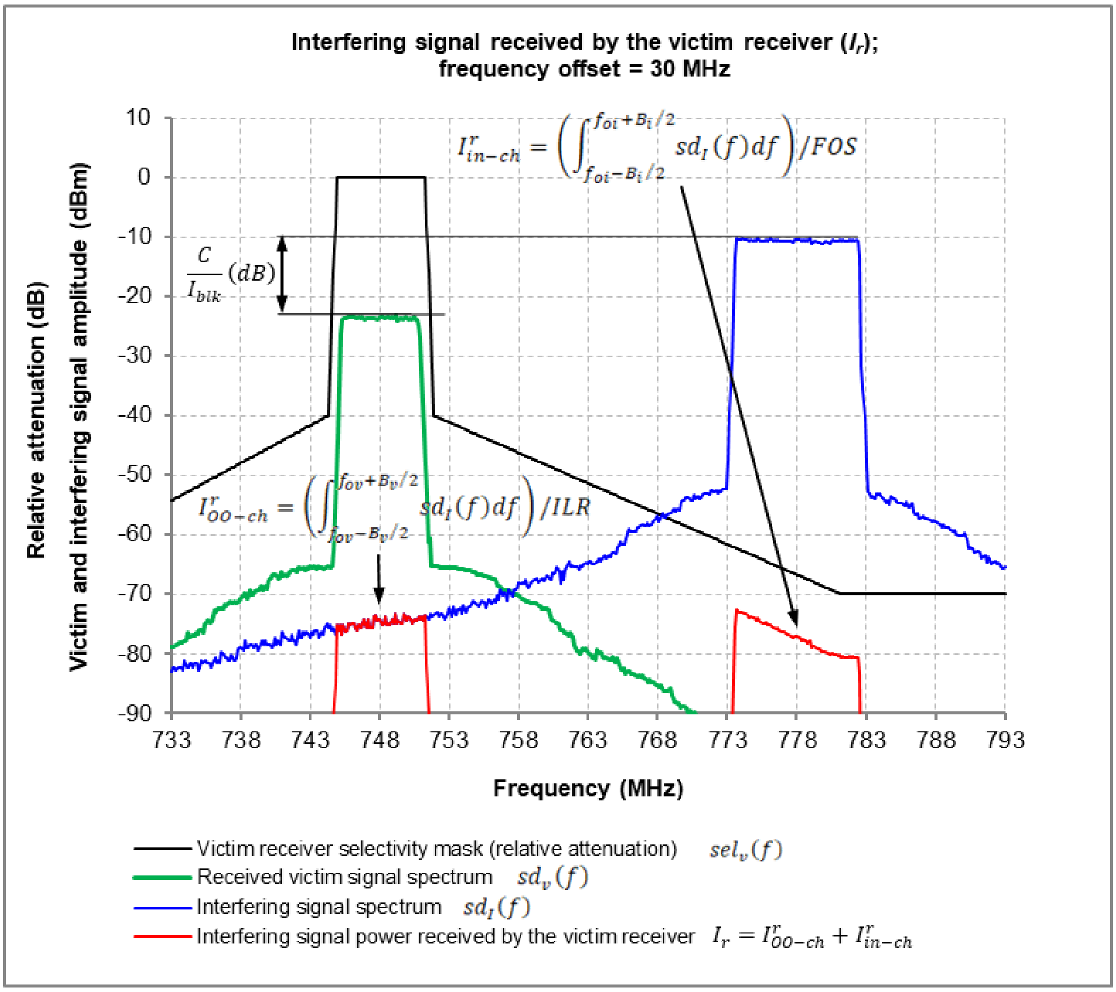

Figure 1: Frequency offset interfering signal power received by the victim receiver

The victim receiver receives both the interfering signal in-channel emissions (I_{in-ch}) attenuated by FOS and the interfering signal out-of-channel emissions (I_{oo-ch}) falling into its channel, which are defined by ILR. The interfering signal power received by the victim receiver is shown in Figure 1 and can be expressed in the linear domain as:

$$\begin{aligned} I_r &= I^r_{oo-ch} + I^r_{in-ch} \\ &= \frac{I_{in-ch}}{ILR} + \frac{I_{in-ch}}{FOS} \end{aligned} \tag{1}$$

In the logarithmic domain:

$$I_r(dBm) = I_{in-ch}(dBm) + 10\log_{10}(10^{-ILR(dB)/10} + 10^{-FOS(dB)/10}) \tag{2}$$

Where:

- I_r: frequency offset interfering signal power received by the victim receiver after RF/IF/BB filtering;
- I_{in-ch}: frequency offset interfering signal in-channel power, measured in its channel bandwidth at the receiver input before RF/IF/BB filtering;
- I^r_{oo-ch}: frequency offset interfering signal out-of-channel power, received by the victim receiver after RF/IF/BB filtering;
- I^r_{in-ch}: frequency offset interfering signal in-channel power received by the victim receiver after RF/IF/BB filtering.

Note that when defining the resilience levels of a receiver, $I_{in\text{-}ch}$ is often referred to as $I_{adj\text{-}ch}$ or I_{blk} depending on the frequency offset between the useful and interfering signals.

Note also that, as shown in Figure 1, the ILR and FOS are, formally, calculated as the ratios of the total in-channel interfering power (I_{in-ch}) to the result of the following integrations:

ILR in the linear domain:

$$ILR = \frac{I_{in-ch}}{I^{r}_{oo-ch}} = \frac{\int_{f_{oi}-B_i/2}^{f_{oi}+B_i/2} sd_I(f)df}{\int_{f_{ov}-B_v/2}^{f_{ov}+B_v/2} sd_I(f)df} \quad (3)$$

ILR in the logarithmic domain:

$$ILR(dB) = 10\log_{10}\left(\frac{\int_{f_{oi}-B_i/2}^{f_{oi}+B_i/2} sd_I(f)df}{\int_{f_{ov}-B_v/2}^{f_{ov}+B_v/2} sd_I(f)df}\right) \quad (4)$$

Where:

- f_{0i}: centre frequency of the interfering transmitter;
- f_{0v}: centre frequency of the victim receiver;
- B_i and B_v: interferer and victim bandwidth respectively.

FOS, which is the frequency offset selectivity of the receiver (equivalently the overall receiver filter attenuation at the related frequency offset), can be expressed in the linear domain as:

$$FOS = \frac{I_{in-ch}}{I^{r}_{in-ch}} = \frac{I_{in-ch}}{\frac{I_{in-ch}}{\frac{1}{B_i}\int_{f_{oi}-B_i/2}^{f_{oi}+B_i/2} sel_v(f)df}} \quad (5)$$

Therefore:

$$FOS = \frac{1}{B_i}\int_{f_{oi}-B_i/2}^{f_{oi}+B_i/2} sel_v(f)df \quad (6)$$

FOS in the logarithmic domain:

$$FOS(dB) = 10\log_{10}\left(\frac{1}{B_i}\int_{f_{oi}-B_i/2}^{f_{oi}+B_i/2} sel_v(f)df\right) \quad (7)$$

Where:

- B_i and B_v: interferer and victim bandwidth respectively;
- $sd_I sd_I(f)$: interferer spectral density in the linear domain;
- sel_v: victim selectivity in the linear domain, in number, relative to the 0 value at its centre frequency f_{ov}.

For a sufficiently large offset frequency, where the interferer relative spectral density and the victim receiver selectivity can be considered as flat and equal to a constant value, the value of FOS is independent of the interference signal bandwidth (B_i) as shown in Equation (7) while the value of ILR varies as a function of the ratio B_i/B_v as shown in Equation (4).

The receiver desensitisation “M” is the reduction in the signal to noise ratio of the receiver or a reduction in the effective sensitivity in the presence of an interfering signal. The receiver desensitisation corresponds to the ‘equivalent noise rise’ due to the interfering signal, which can be expressed in the linear domain as:

$$M = \frac{N + I_r}{N} \tag{8}$$

Where:

- N: receiver noise floor measured in the receiver bandwidth;
- I_r: interfering signal power received by the victim receiver after RF/IF/BB filtering.

Note from Equation (8) that M is always greater than 1. Then, the interfering signal power received by the victim receiver is:

$$I_r = NM - N \tag{9}$$

In the logarithmic domain:

$$\begin{aligned} I_r(dBm) &= 10log(NM - N) \\ &= 10log\left(10^{\frac{N(dBm)}{10}} 10^{\frac{M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) \\ &= 10log\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) \end{aligned} \tag{10}$$

Equation (9) shows the relation between I_r, N and M. For more detailed information on N, M, I_r see ANNEX 3:.

3.2 RELATION BETWEEN FOS, RIR AND ILR

From its definition, the RIR can be expressed in the linear domain as:

$$RIR = \frac{I_{in-ch}}{I_r} \tag{11}$$

By replacing I_r with the expression given in Equation (1):

$$\begin{aligned} RIR &= \frac{I_{in-ch}}{\frac{I_{in-ch}}{ILR} + \frac{I_{in-ch}}{FOS}} \\ &= \frac{1}{\frac{1}{ILR} + \frac{1}{FOS}} \end{aligned} \tag{12}$$

Or, equivalently,

$$\frac{1}{RIR} = \frac{1}{ILR} + \frac{1}{FOS} \tag{13}$$

By rearranging the above equation:

$$FOS = \frac{1}{\frac{1}{RIR} - \frac{1}{ILR}} \tag{14}$$

In the logarithmic domain:

$$\begin{aligned} FOS(dB) &= -10log\left(\frac{1}{10^{\frac{RIR(dB)}{10}}} - \frac{1}{10^{\frac{ILR(dB)}{10}}}\right) \\ &= -10log\left(10^{\frac{-RIR(dB)}{10}} - 10^{\frac{-ILR(dB)}{10}}\right) \end{aligned} \tag{15}$$

Equation (14) shows the relation between FOS, RIR and ILR.

3.3 DERIVATION OF LEVELS OF RECEIVER RESILIENCE TO TRANSMISSION ON ADJACENT FREQUENCY RANGES

The levels of receiver resilience to transmission on adjacent frequency ranges for a given receiver desensitisation "M" can be defined either by I_r, FOS or RIR. These three parameters are linked to each other as demonstrated in the previous sections, I_r and FOS are more commonly used in relevant standards to specify receiver resilience.

Equations (1), (9) and (14) developed in sections 3.1 and 3.2 are the fundamental equations describing the mechanism of receivers resilience to a frequency offset interfering signal. They can be used to develop a method to derive levels of receiver resilience to transmission on adjacent frequency ranges.

Equating equations (1) and (9) (linear terms) leads to the basic equation of MRR, which is used to derive levels of receiver resilience to transmission on adjacent frequency ranges as described in ANNEX 1:, section A1.1:

$$I_{in-ch} = \frac{NM - N}{\left(\frac{1}{ILR} + \frac{1}{FOS}\right)} \tag{16}$$

Or in the logarithmic domain:

$$I_{in-ch}(dBm) = 10log\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) - 10log\left(10^{\frac{-ILR\ (dB)}{10}} + 10^{\frac{-FOS\ (dB)}{10}}\right) \tag{17}$$

4 METHOD FOR RECEIVER RESILIENCE (MRR) USED TO DEREIVE BLOCKING AND SELECTIVITY LEVELS

4.1 USE OF THE METHOD

MRR aims to calculate/specify the receiver resilience levels to be included in the ECC Recommendation on receivers. The basic equation of the method Equation (16) predicts the frequency offset interfering signal level ($I_{in\text{-}ch}$) at the receiver input and consequently the $PR_{adj\text{-}ch}$ and blocking level (I_{blk}) of the receiver based on four input parameters N, M, ILR and FOS.

The basic equation of MRR is not intended to evaluate the impact of undesired signal variations, inherent in receivers, on their performance (or to model a badly-behaved non-linear receiver), nor to predict the behaviour of a specific receiver.

4.2 CONSIDERATION OF DIFFERENT BANDWIDTHS OF VICTIM AND INTERFERER

The parameters ILR and FOS of the basic equation of MRR implicitly take into account the interfering and victim signal bandwidths as explained in ANNEX 3:, section A3.2.3.

4.3 CONSIDERATION OF PEAK-TO-AVERAGE POWER RATIO (PAPR)

In modern communication systems, the PAPR of the majority of transmitted signals is optimised (minimised) to improve the efficiency of transmitters. This enables to prevent high undesirable power dissipation at the transmitter (cost reduction) and to improve the reliability of the transmitted signal (high QoS). The minimisation of the PAPR reduces the impact of interfering signal on the receiver resilience levels.

The blocking level of receivers (I_{blk}) is defined as mean power of the interfering signal. I_{blk} is measured in RMS (root mean square) measurement mode in a bandwidth equal to the interfering signal channel bandwidth at the input of the victim receiver in the presence of the interfering signal under normal operating conditions. In the case of a time varying interfering signal, the I_{blk} is the RMS power of the active portions of the interfering signal measured over active time of the interferer. The I_{blk} predicted by the basic MRR equation is the interfering signal in-block mean power.

4.4 BASICS AND APPLICATION OF MRR

4.4.1 Basics of the method

MRR derives the frequency offset interfering signal level ($I_{in\text{-}ch}$) at the receiver input and consequently the blocking level (I_{blk}) and $PR_{adj\text{-}ch}$ of the receiver, for given values of M, N, ILR and FOS, from the equation below (see ANNEX 1:). $I_{in\text{-}ch}$ is referred as I_{blk} when calculating the blocking level and as $_{Iadj\text{-}ch}$ when calculating the adjacent channel protection ratio (PR), where:

- $I_{in\text{-}ch}$: frequency offset interfering signal power measured in the interferer channel bandwidth at the receiver input before RF/IF/BB filtering;
- I_{blk}: blocking interfering signal power measured in the interferer channel bandwidth at the receiver input before RF/IF/BB filtering;
- $I_{adj\text{-}ch}$: adjacent channel interfering signal power measured in the interferer channel bandwidth at the receiver input before RF/IF/BB filtering.

The bandwidth of the wanted received signal and the bandwidth of the interfering signal need to be determined in advance (see ANNEX 5: for more information on defining receiver and transmitter bandwidth). The frequency offset is the offset between the centre frequency of the wanted received signal bandwidth and the centre frequency of the interfering signal bandwidth.

For the frequency offsets beyond the second adjacent channel, the blocking level ($I_{in-ch} = I_{blk}$):

$$I_{in-ch}(dBm) = I_{blk}(dBm) = 10log\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) - 10log\left(10^{\frac{-ILR\,(dB)}{10}} + 10^{\frac{-FOS\,(dB)}{10}}\right) \quad (18)$$

Or, for first and second adjacent channels frequency offsets, the carrier to interference ratio ($C/I_{in-ch} = C/I_{adj-ch}$) as follows:

$$\begin{aligned} C/I_{in-ch}\,(dB) &= C/I_{adj-ch}(dB) = C(dBm) - I_{adj-ch}(dBm) \\ &= C_{sens}(dBm) + M(dB) - I_{adj-ch}(dBm) \end{aligned} \quad (19)$$

Where:
- N: Noise floor of the victim receiver;
- M: Victim receiver maximum acceptable desensitisation at a given frequency offset;
- FOS: Frequency offset selectivity of the victim receiver;
- ILR: Leakage power ratio of the interfering signal at offset frequency;
- C: useful signal level received by the victim receiver;
- I_{blk}: blocking interfering signal power measured in its channel bandwidth at the receiver input before RF/IF/BB filtering;
- I_{adj-ch}: adjacent channel interfering signal power measured in its channel bandwidth at the receiver input before RF/IF/BB filtering;
- C_{sens}: receiver sensitivity level, specified for the victim receiver.

4.4.2 Application of the method

The steps below should be followed to derive the receiver blocking level (I_{blk}) or the adjacent channel protection ratio (PR_{adj-ch}=C/I_{adj-ch}):

Step 1 Calculate the noise floor (N) of the receiver:

Every radio receiver is subject to a noise floor that can be described using the following equation expressed in dBm:

$$N(dBm) = 10log_{10}(kTB) + NF(dB) + 30 \quad (20)$$

Where:
- N: receiver noise floor (dBm);
- k: Boltzmann constant in Joules per Kelvin (1.381×10^{-23});
- T: temperature in degrees Kelvin (for common terrestrial radio receivers, 290 K can be used);
- B: receiver bandwidth in Hertz;
- kTB: receiver thermal noise in Watts;
- NF: receiver Noise Figure (dB).

In cases where the noise figure is not known, the noise floor may also be derived from the receiver sensitivity provided that the interference signal bandwidth is equal to that of the receiver signal bandwidth:

$$N(dBm) = Receiver\ sensitivity(dBm) - C/I_{co-ch} \quad (21)$$

Step 2 Calculate the receiver desensitisation (M) (go to Step 3 if M is known):

M is often found in relevant standards, either explicitly mentioned, or calculated from the required “Receiver sensitivity” (in absence of interference) and the “Minimum wanted received signal level” required in the blocking test through the equation:

$$M(dB) = Minimum\ wanted\ received\ signal\ level(dBm) - Receiver\ sensitivity(dBm) \qquad (22)$$

Care should be taken such that the value of M is chosen based on the expected linear range of the receiver to prevent overloading.

If it is not possible to find/calculate M from the relevant standards as above, the following typical values: M = 3 dB or M = 15 dB could be used; the M = 3 dB value may be used for systems deployed in a “noise limited” operational environment, while M = 15 dB could be used for systems deployed in an “interference limited” operational environment (see ANNEX 3:, section A3.2.2).

Step 3 Determine the ILR value to be used in Step 4 (see ANNEX 3:, section A3.2.4 for more detailed information):

a) *Identify the interfering signal and interference scenario:*

The relevant interfering signal and interference scenario can be identified based on:

- the existing or planned deployment of victim and interfering systems and the compatibility studies presented in various CEPT/ECC, ETSI and ITU-R technical reports (see annex A3.3)
- or, if the information in the previous paragraph is not available, information may be sourced from either of the following:
  - from the intra/inter-system interfering signals and interference scenarios defined in relevant standards; or
  - from the reference interfering signal and interference scenario defined in ANNEX 5:.

b) *Determine the ILR value:*

Once the relevant interfering signal and interference scenario are identified the following approach can be used to determine the appropriate ILR value[1]:

- If the victim receiver channel is in the **out-of-band** (OOB**)** domain of the interfering signal (see A3.2.4) use either of the following:
  - the ILR (ACLR) value defined in existing EC/ECC Decisions, ECC/ITU-R Recommendations, CEPT/ECC Reports and relevant standards /technical specifications/reports;
  - the ILR value determined by measurements;

- If the victim receiver channel is in the **reciprocal** spurious blocking domain **(RSBD)** (see n A3.2.4), use either of the following:
  - the ILR value derived from the reference interfering signal defined in ANNEX 5:.
  - the ILR value derived from the in-channel power of the interfering signal and the spurious emission levels defined in external documents including: the existing EC/ECC Decisions, and relevant standards/technical specifications/reports, ECC/ITU-R Recommendations (e.g. ERC Recommendation 74-01 [2] and Recommendation ITU-R SM.329-12 [7]);

Step 4 Define the frequency offset selectivity (FOS) of the receiver based on the following options:

Depending on the scenario, FOS can either be derived directly from the relevant standard, or according to:

$$FOS\ (dB) \geq ILR\ (dB) + X\ dB \qquad (23)$$

where the value of X depends on the maximum acceptable increase of the interfering signal power (I_r) at the receiver input due to I^r_{in-ch}, e.g.:

- X = 0 dB would correspond to $I_r = I^r_{oo-ch} + 3\ dB$;
- X = 10 dB would correspond to $I_r = I^r_{oo-ch} + 0.4\ dB$

Where:

[1] The way in which the ILR was derived is described in each annex

- $I_r = I^r_{oo-ch} + I^r_{in-ch}$;
- I^r_{oo-ch}: frequency offset interfering signal out-of-channel power received by the victim receiver;
- I^r_{in-ch}: frequency offset interfering signal in-channel power received by the victim receiver.

FOS is defined based on the scenario as follows:

- If the victim receiver channel is in the OOB domain of the interfering signal and

  the interfering signal and interference scenario are identified based on a relevant standard (i.e. from Step 3):
  - if the interfering signal is an unmodulated continuous wave (CW) signal, the value of FOS can be determined based on the adjacent channel interfering signal limit defined in ETSI Harmonised Standards as follows, noting that this is a rearrangement of equation (18) and the value may have already been available:

$$FOS(dB) = -10log\left(\frac{10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}}{10^{\frac{I_{adj-ch}(dBm)}{10}}} - 10^{\frac{-ILR(dB)}{10}}\right) \quad (24)$$

  - if the interfering signal is similar to the wanted signal, the value of FOS is determined according to equation (23) with the value of X defined according to the relevant standard;

    In this interference configuration the impact of the interfering signal OOBE falling into the receiver channel, added to the interfering signal in-channel emission received by the receiver after filtering, is important. As the emissions falling into the receiver channel cannot be reduced by filtering, it is sensible to reduce as much as possible the interfering signal in-channel emission at the receiver input by filtering to reduce the impact of the interfering signal on the receiver performance. Therefore, setting X = 10 such that FOS=ILR + 10 dB would be an appropriate choice when deriving the adjacent-channel protection ratio of a receiver.

- For all other cases (i.e. the reciprocal spurious blocking domain or interfering signal/scenario not based on an ETSI harmonised standard), the value of FOS is determined by equation (23) with the value of X set according to the maximum acceptable increase of the interfering signal power.

  In most cases with the spurious emissions (SE) level of the interfering signal being low, X=0 dB such that FOS=ILR can be used in the calculation in order to avoid unnecessary constraint on the receiver. X=10 such that FOS=ILR+10 dB would be an appropriate choice if higher selectivity is required.

  If a CW interfering signal is used, instead of the RI signal, to calculate the blocking level of a victim receiver, the receiver FOS value should be equal to the FOS value specified for the blocking level calculation with the RI signal, since the ILR value of the CW signal is assumed to be very high or infinite (≥130 dB).

Step 5 Derive the receiver blocking level (I_{blk}) or adjacent channel protection ratio (PR_{adj-ch}):

In the RSBD:

$$I_{blk}(dBm) = 10log\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) - 10log\left(10^{\frac{-ILR\,(dB)}{10}} + 10^{\frac{-FOS\,(dB)}{10}}\right) \quad (25)$$

Or if the *ILR* of the interfering signal can be considered infinite (very high):

$$I_{blk}(dBm) = 10log\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) + FOS(dB) \quad (26)$$

In the OOB domain:

Calculate PR_{adj-ch} (dB) as follows, with I_{adj_ch} evaluated from equation (25) by replacing I_{blk} with I_{adj_ch}:

$PR_{adj-ch} = C/I_{adj-ch}$ by replacing I_{blk} with I_{adj-ch}:

$$\begin{aligned} PR_{adj-ch}(dB) = C/I_{adj-ch}(dB) &= C(dBm) - I_{adj-ch}(dBm) - I_{adj-ch}(dBm) \\ &= C_{sens}(dBm) + M(dB) - I_{adj-ch}(dBm) \end{aligned} \quad (27)$$

The method is illustrated in the following flowchart.

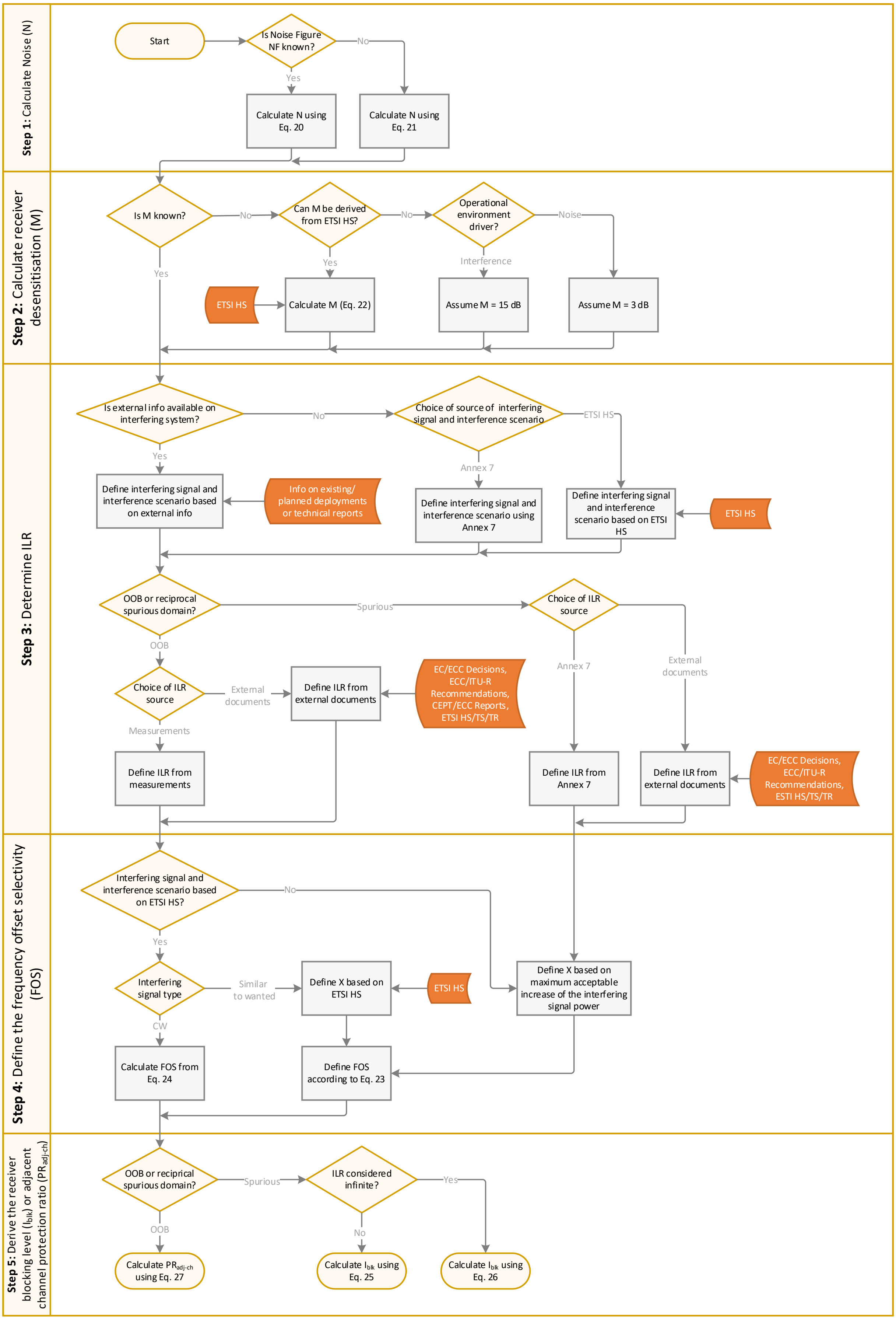

Figure 2: Flowchart describing the method for receiver resilience for a given Rx bandwidth

5 CONCLUSIONS

The method for receiver resilience (MRR) is a technology neutral generic method. It uses four parameters N, M, ILR and FOS to calculate the receiver resilience levels by a single equation derived from fundamental equations describing the mechanism of receiver resilience to a frequency offset interfering signal.

The consistency of MRR with existing relevant standards was assessed by calculating receiver resilience levels defined in eight different ETSI Harmonised Standards for numerous receiver configurations. The results obtained showed good agreement with the adjacent channel PR and blocking levels defined in the relevant standards for six of the eight assessed but can also result in more stringent blocking levels e.g. in the case of DAB and DTTB.

Five distinct steps should be followed to derive the receiver resilience levels when using MRR.

The most important step is Step 3, which aims to identify the interfering signal and interference scenario, and to determine the ILR value to be used in the calculation. This can be achieved, as proposed in this Report, by two different approaches, summarised below, each with advantages and disadvantages:

1. Based on the existing or planned deployment of victim and interfering systems and the compatibility studies presented in various CEPT/ECC, ETSI and ITU-R technical reports or ETSI Harmonised Standards.
   - Advantages:
     - Possibility to choose the interfering signal and interference scenario for each system/service if they are specified in the existing technical reports or relevant standards;
     - Possibility to compare the calculated resilience levels with those assessed in the existing technical reports or defined in the existing relevant standards.
   - Disadvantages:
     - Possible difficulty to choose the most relevant interfering signal and interference scenario for each system/service if they are not clearly specified in the existing technical reports or relevant standards. Note that this difficulty may not apply to all systems/services and can be overcome by dealing with the issue within the competent CEPT or ETSI technical group;
     - Difficulty to choose the most relevant interfering signal and interference scenario for a given system/service if there are no technical reports or relevant standards available;
     - Difficulty to choose the ILR value if it is not clearly defined in the existing technical reports or relevant standards, or if there are no technical reports or relevant standards available.

2. From a reference interfering signal and interference scenario (one example is specified in ANNEX 5::
   - Advantages:
     - No need to choose an interfering signal and interference scenario for each system/service, since a single well specified reference interfering signal and interference scenario are used for all systems/services independent of the radio environment in the field;
     - No need to choose the ILR value, since it is calculated from the well-defined spectrum mask of the reference interfering signal.
   - Disadvantages:
     - The specified reference interfering signal and interference scenario cannot be changed and should be used for all systems/services;
     - The specified reference interfering signal and interference scenario may not be fully relevant to some systems/services, since the real radio environment in the field is not considered in this approach.

Finally, the proposed receiver resilience levels in the ECC Recommendation on receivers are determined as follows:

- In the Out-Of-Band (OOB) domain, the proposed resilience levels have been determined from the intra/inter-system interfering signals and interference scenarios defined in relevant standards.
- In the reciprocal spurious blocking domain (RSBD), the proposed resilience levels have been determined based on:

i) a reference interfering signal (OFDM signal) having SE (spurious emissions) levels in accordance with the most common SE levels defined in ERC Recommendation 74-01 [2]; and
ii) the equivalent blocking levels for a CW interfering signal to allow the conformance tests to be done with a CW signal as an alternative.

This Report presents the main information on how MRR was developed and can be used to calculate/specify the receiver resilience levels in the ECC Recommendation on receivers.

ANNEX 1: DERIVATION OF THE METHOD FOR RECEIVER RESILIENCE (MRR)

A1.1 DERIVATION OF THE BASIC EQUATION OF MRR

The basic equation of MRR shown below allows calculation of the receiver blocking level as a function of N, M, ILR and FOS:

$$I_{in-ch}(dBm) = 10\log_{10}\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) - 10\log_{10}\left(10^{\frac{-ILR\,(dB)}{10}} + 10^{\frac{-FOS\,(dB)}{10}}\right) \qquad (28)$$

This equation has been derived through the following steps. Note that the derivation is carried out in the linear domain and only the final result is transformed from the linear domain to the logarithmic domain:

The victim receiver receives both the interfering signal in-channel emissions ($I_{in\text{-}ch}$) attenuated by FOS and the interfering signal out-of-channel emissions ($I_{oo\text{-}ch}$) falling into its channel, which are defined by ILR.

$$\begin{aligned} I_r &= I^r_{oo-ch} + I^r_{in-ch} \\ &= \frac{I_{in-ch}}{ILR} + \frac{I_{in-ch}}{FOS} \end{aligned} \qquad (29)$$

Where:

- I_r: frequency offset interfering signal power received by the victim receiver;
- I_{in-ch}: frequency offset interfering signal power measured in its channel bandwidth at the receiver input before RF/IF/BB filtering;
- I^r_{oo-ch}: frequency offset interfering signal out-of-channel power received by the victim receiver;
- I^r_{in-ch}: frequency offset interfering signal in-channel power received by the victim receiver. Note that when defining the resilience levels of a receiver, $I_{in\text{-}ch}$ is often referred to as $I_{adj\text{-}ch}$ or I_{blk} depending on the frequency offset between the useful and interfering signals.

The received interfering signal power causing a receiver desensitisation of M can be calculated as follows, where N is the noise floor of the receiver:

$$I_r = NM - N \qquad (30)$$

By equating Equations (29) and (30) :

$$\frac{I_{in-ch}}{ILR} + \frac{I_{in-ch}}{FOS} = NM - N \qquad (31)$$

The above equation can be rewritten as:

$$I_{in-ch}\left(\frac{1}{ILR} + \frac{1}{FOS}\right) = NM - N \qquad (32)$$

and then:

$$I_{in-ch} = \frac{NM - N}{\left(\frac{1}{ILR} + \frac{1}{FOS}\right)} \qquad (33)$$

Transforming from the linear domain to the logarithmic domain results in:

$$I_{in-ch}(dBm) = 10\log_{10}(NM - N) - 10\log_{10}\left(\frac{1}{ILR} + \frac{1}{FOS}\right) \qquad (34)$$

As the values of N, M, ILR and FOS are always defined in the logarithmic domain the above equation can be rewritten as:

$$I_{in-ch}(dBm) = 10\log_{10}\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) - 10\log_{10}\left(10^{\frac{-ILR\,(dB)}{10}} + 10^{\frac{-FOS\,(dB)}{10}}\right) \quad (35)$$

which is the basic equation of MRR.

Note that the receiver frequency offset selectivity can also be expressed as a function of N, M, $I_{in\text{-}ch}$ and ILR from Equation (32):

$$\frac{1}{ILR} + \frac{1}{FOS} = \frac{NM - N}{I_{in-ch}} \quad (36)$$

and:

$$\frac{1}{FOS} = \frac{NM - N}{I_{in-ch}} - \frac{1}{ILR} \quad (37)$$

then:

$$FOS = \frac{1}{\frac{NM - N}{I_{in-ch}} - \frac{1}{ILR}} \quad (38)$$

Transforming from the linear domain to the logarithmic domain results in:

$$FOS(dB) = -10log\left(\frac{NM - N}{I_{in-ch}} - \frac{1}{ILR}\right) \quad (39)$$

and finally:

$$FOS(dB) = -10\log_{10}\left(\frac{10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}}{10^{\frac{I_{in-ch}(dBm)}{10}}} - 10^{\frac{-ILR(dB)}{10}}\right) \quad (40)$$

Note that when defining the resilience levels of a receiver, $I_{in\text{-}ch}$ is often referred to as $I_{adj\text{-}ch}$ or I_{blk} depending on the frequency offset between the useful and interfering signals.

ANNEX 2: EXAMPLES OF APPLICATION OF MRR

A2.1 APPLICATION OF MRR WHERE RECEIVER BLOCKING LEVEL IS INDEPENDENT OF THE ILR FOR THE REQUIREMENT

In a real-life interference scenario, the desensitisation of the receiver will depend on both ILR and FOS as shown in ANNEX 1:, leading to the MRR for deriving receiver blocking described in section 4.4. In a compatibility scenario the impact of both FOS and ILR is important. However, it may be desirable to define a receiver requirement based only on FOS for the cases where the specific ILR of the interfering signal used in a harmonised standard is difficult to determine. A sufficiently high ILR that would not impact the requirement should then be assumed. Achieving adequate ACLR/ILR on test signals is important to reduce the impact on I/C measurements. The ideal configuration uses separate signal generators for the wanted and interferer signals which also allow a band pass filter to be added to the interference path to further improve the ACLR/ILR.

While some harmonised standards only imply such requirements on the interfering signal through the way receiver requirements are defined, other standards have explicit requirements on interfering signals. One example is ETSI EN 303 340 [4], which stresses the importance of improving ACLR in informative Annex D.3 and sets normative requirements for minimum ACLR of the interfering signal in informative Annex F.

In certain scenarios, the choice of a FOS value at least equal or even considerably higher than ILR, may be reasonable in order that the FOS is not the limiting in the compatibility scenario. If FOS >> ILR is assumed, Step 5 of MRR boils down to:

$$I_{in-ch}(dBm) \approx 10\log_{10}\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) + ILR(\mathrm{dB}),\ \text{assuming}\ FOS \gg ILR \quad (41)$$

Such a choice is clearly not desirable, since the blocking level then depends very little on the desired receiver resilience (FOS), but almost completely on the ILR of the interfering signal. Such a requirement is not very useful.

Note that the same reasoning (but reversed) can be made for transmitters. Defining a transmitter requirement using a receiver with a non-adequate FOS would impair the definition of the requirement and make the result depend mostly on the receiver impairments. Clearly the limits in ERC Recommendation 74-01 [2] are not defined that way, but rather assume a sufficiently good spectrum analyser to detect the unwanted emissions independent of receiver impairments.

The solution would be to separate the assumptions of the compatibility scenario from the requirement scenario. In the compatibility scenario, FOS and ILR are analysed to make an informed choice of FOS for the receiver according to Step 4 of Method MRR, for example:

1 FOS based on “balance” between Tx and Rx requirements: FOS (dB) = ILR (dB);
2 FOS defined with some margin above the impairment from unwanted emissions, in cases where it is desirable to not be limited by the receiver: FOS (dB) > ILR (dB).
3 FOS defined with some margin below the impairment from unwanted emissions, in cases where it is desirable to not be limited by the transmitter, e.g. when the allowed complexity is much higher for the transmitter than for the receiver: FOS (dB) < ILR (dB).

Once FOS is selected, the blocking requirement is defined using a different scenario, ensuring that ILR >> FOS. It is reasonable to assume that ILR is infinite and has no impact, leading to a blocking level in Step 5 of MRR independent of the ILR of the interfering (test) signal:

$$I_{in-ch}(dBm) \approx 10\log_{10}\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) + FOS(\mathrm{dB}),\ \text{assuming}\ ILR \gg FOS \quad (42)$$

Where:

- N: Noise floor of the victim receiver (dBm);
- M: Victim receiver desensitisation at a given frequency offset (dB);
- FOS: Frequency offset selectivity of the victim receiver (dB);
- ILR: Leakage power ratio of the interfering signal at offset frequency (dB).

A2.2 GENERIC APPLICATION OF MRR FOR SPECIFYING RECOMMENDED LEVELS OF RECEIVERS RESILIENCE TO TRANSMISSION ON ADJACENT FREQUENCY RANGES IN ETSI HARMONISED STANDARDS

MRR is proposed to be used for specifying recommended levels of receiver resilience to transmission on adjacent frequency ranges. The proposed method is technology neutral and uses N, M, ILR and FOS to derive the receivers blocking level by a single equation derived from the basic equations describing the receiver resilience mechanism (see section 3).

This Annex presents the results of in-depth analysis carried out:

- to understand how receivers adjacent channel selectivity and blocking requirements are defined in different relevant standards;
- to check if the basic equation of MRR can be used to derive the blocking requirements defined in relevantstandards;
- to check if the basic equation of MRR can be used to verify the consistency between different requirements defined in relevantstandards;
- to find out how to choose Rx desensitization value (M) when calculating I_{blk}.

MRR consists of deriving the receiver blocking level (I_{blk}) or the carrier to interference blocking level ratio (C/I_{blk}), for a given acceptable receiver desensitization "M", from the equation below:

$$I_{blk}(dBm) = 10\log_{10}\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) - 10\log_{10}\left(10^{\frac{-ILR\,(dB)}{10}} + 10^{\frac{-FOS\,(dB)}{10}}\right) \quad (43)$$

Or, in terms of C/I_{blk} ratio:

$$\begin{aligned} C/I_{blk}(dB) &= C(dBm) - I_{blk}(dBm) \\ &= C_{sens}(dBm) + M(dB) - I_{blk}(dBm) \end{aligned} \quad (44)$$

Where:

- N: Noise floor of the victim receiver (dBm);
- M: Victim receiver maximum acceptable desensitisation at a given frequency offset (dB);
- FOS: Frequency offset selectivity of the victim receiver (dB);
- ILR: Leakage power ratio of the interfering signal at offset frequency (dB);
- C_{sens}: receiver sensitivity level, specified for the victim receiver.

Note that even if the basic equation is proposed to calculate I_{blk}, the calculated I_{blk} is simply I_{adj} in the case of adjacent channel interference and $I_{co\text{-}ch}$ in the case of co-channel interference, as far as the victim receiver operates in its linear range. This is demonstrated in the analysis presented in this Report.

A2.2.1 Receiver blocking requirements defined in different relevant standards and applicability of MRR to determine the receiver blocking levels

Four ETSI Harmonised Standards have been analysed to understand how the receiver resilience requirements (FOS and I_{blk}) are defined in these standards. The analysed relevant standards cover Broadcasting, Mobile, Fixed services and GSM respectively:

1 ETSI EN 303 340 V1.2.1: Digital Terrestrial TV Broadcast Receivers; Harmonised Standard for access to radio spectrum [4];

2 ETSI EN 301 908-14 V13.1.1: IMT cellular networks; Harmonised Standard for access to radio spectrum; Part 14: Evolved Universal Terrestrial Radio Access (E-UTRA) Base Stations (BS) [8];

3 ETSI EN 302 217-2 V3.3.1: Fixed Radio Systems; Characteristics and requirements for point-to-point equipment and antennas; Part 2: Digital systems operating in frequency bands from 1 GHz to 86 GHz; Harmonised Standard for access to radio spectrum [9];

4 ETSI EN 301 502 V12.5.1: Global System for Mobile communications (GSM); Base Station (BS) equipment; Harmonised Standard covering the essential requirements of article 3.2 of the Directive 2014/53/EU [10].

A2.2.2 Harmonised Standard ETSI EN 303 340 “Digital Terrestrial TV Broadcast Receivers”

Note: the ACLR (or ILR) mentioned in this section refers to the ACLR (or ILR) of the interfering test signal.

A2.2.2.1 General comments

FOS and ILR defined in ECC Report 310 [3]0 have not been taken into account in ETSI EN 303 340 [4] yet. Instead, ACS (adjacent channel selectivity) and ACLR (adjacent channel leakage ratio) are used independently from the frequency offset between the useful and interfering signals in ETSI EN 303 340.

Moreover, the term ACS used in this harmonised standard does not have the same meaning as the term ACS defined in ECC Report 310 and in compatibility studies carried out within CEPT. The term ACS used in ETSI EN 303 340 is equivalent to the measured I/C ratio of the receiver under test.

The receivers ACS and blocking level (I_{blk}) are defined in the presence of an interfering signal with a minimum ACLR value ($ACLR_{min}$) with the assumption ACS = $ACLR_{min}$, which is normative (see ETSI EN 303 340, annex F). This assumption implies that the OOBE of the interfering signal received by the receiver under test is not negligible and consequently will impact the compliance measurement results. However, note that $ACLR_{min}$ means that the value of ACLR used in the measurement can be higher than the value of ACLRmin. This is explicit in ETSI EN 303 340, informative Annex D/D3, where it is proposed to improve the interfering signal ACLR to minimise its impact on measurement results.

The above findings are summarised in Table 1:.

Table 1: Derived ACLR values from ETSI EN 303 340 V1.2.1: Digital Terrestrial TV Broadcast Receivers; Harmonised Standard for access to radio spectrum [4]

Victim receiver	Frequency offset	Requirement	Type of interfering signal	ACLR requirement for the interfering signal (Note 1)	Assumption used to define blocking requirements (Note 1) (Note 2)	ACLR proposed for conformance testing (Note 1) (Note 3)
DVB-T/T (8 MHz)	8 MHz	Receiver ACS (I_{adj}/C)	DVB-T (8 MHz)	43 dB ($ACLR_{min}$)	ACS = $ACLR_{min}$	ACLR ≥ $ACLR_{min}$
DVB-T/T2 (8 MHz)	8 MHz	Receiver ACS (I_{adj}/C)	DVB-T (8 MHz)	47 dB ($ACLR_{min}$)	ACS = $ACLR_{min}$	ACLR ≥ $ACLR_{min}$
DVB-T (8 MHz)	10 MHz	Receiver ACS (I_{adj}/C)	10 MHz LTE 700 MHz BS light load (near idle)	53 dB ($ACLR_{min}$)	ACS = $ACLR_{min}$	ACLR ≥ $ACLR_{min}$
DVB-T (8 MHz)	73 MHz	Receiver ACS (I_{adj}/C)	10 MHz LTE 700 MHz BS light load (near idle)	61 dB ($ACLR_{min}$)	ACS = $ACLR_{min}$	ACLR ≥ $ACLR_{min}$
DVB-T2 (8 MHz)	10 MHz	Receiver ACS (I_{adj}/C)	10 MHz LTE 700 MHz BS light load (near idle)	58dB ($ACLR_{min}$)	ACS = $ACLR_{min}$	ACLR ≥ $ACLR_{min}$
DVB-T2 (8 MHz)	73 MHz	Receiver ACS (I_{adj}/C)	10 MHz LTE 700 MHz BS light load (near idle)	65 dB ($ACLR_{min}$)	ACS = $ACLR_{min}$	ACLR ≥ $ACLR_{min}$
DVB-T (8 MHz)	73 MHz	Receiver blocking level (I_{blk})	10 MHz LTE 700 MHz BS fully loaded	64 dB ($ACLR_{min}$)	ACS = $ACLR_{min}$	ACLR ≥ $ACLR_{min}$
DVB-T2 (8 MHz)	73 MHz	Receiver blocking level (I_{blk})	10 MHz LTE 700 MHz BS fully loaded	66 dB ($ACLR_{min}$)	ACS = $ACLR_{min}$	ACLR ≥ $ACLR_{min}$

Note 1: The ACLR mentioned in this table refers to the ACLR of the interfering test signal

Note 2: The required minimum ACLR level depends on the magnitude of I/C values to be measured. Minimum ACLR values to just pass the tests in the present document are provided in table F.1. These values assume a 3 dB ACLR degradation contributing to on the measured performance (see ETSI EN 303 340 V1.2.1 Annex F (Normative).

Note 3: In Annex D/D3 (informative) it is proposed to improve the interfering signal ACLR to minimise its impact on measurement results.

Also note the measured DTT receiver sensitivities, and not those defined in ETSI EN 303 340 have been used - in all the calculations presented in the following sections. This is justified by the fact that the DTT receiver sensitivities defined in ETSI EN 303 340 are far above the measured (actual) DTT receiver sensitivities (see ECC Report 310 [3]).

A2.2.2.2 Applicability of MRR to determine the receiver blocking levels defined in ETSI EN 303 340

The applicability of MRR to determine the receiver selectivity (I_{adj-ch}/C) and blocking levels (I_{blk}) defined in ETSI EN 303 340 [4] has been checked. The calculations have been made for two cases:

- FOS = ILR = $ACLR_{min}$;
- FOS = $ACLR_{min}$ and ILR = 100 dB (ILR>>FOS).

When verifying the applicability of MRR to determine the receiver blocking levels defined in ETSI EN 303 340, no calculations have been made when the values of FOS (or ACS) and ILR (or ACLR) are not defined/provided and they cannot be calculated independently from the I_{blk} defined in the harmonised standard.

A2.2.2.3 Results of the calculations

The results of the calculations , showed that for DVB-T/T2 receivers the C/ $I_{adj\text{-}ch}$ and I_{blk} derived by using MRR are quite close to those defined in ETSI EN 303 340 [4], the difference being less than 2 dB, except 4 values (over 16) where the value of FOS used in the calculations was different than ILR. This can be explained by the fact that in ETSI EN 303 340 C/$I_{adj\text{-}ch}$ and I_{blk} are defined with the assumption of ACS (FOS)=$ACLR_{min}$ (ILR).

The value of M to be used in the conformity tests is not directly defined in ETSI EN 303 340. Nevertheless, it can easily be calculated as follows:

$$M(dB) = Rx\ usefu l\ signal\ level(dBm) - Rx\ sensitivity(dBm) \quad (45)$$

$$= C(dBm) - Rx_{sens}(dBm)$$

The calculated values of M are about 13 and 14 dB for receivers blocking tests and vary from 25 to 34 dB for receivers selectivity (C/ $I_{adj\text{-}ch}$) tests.

A2.2.3 Harmonised Standard ETSI EN 301 908-14 [8] “IMT cellular networks”

Note: the ACLR (or ILR) mentioned in this section refers to the ACLR (or ILR) of the interfering test signal.

A2.2.3.1 General comments

Note that while the receiver selectivity requirement in the harmonised standard is called ACS, it is expressed with an interferer level and not with an ACS value as used in compatibility studies within CEPT (see ECC Report 310 [3]). The term ACS used in ETSI EN 301 908-14 [8] is equivalent to the measured interfering signal level (I_{adj}).

The receivers ACS and blocking level (I_{blk}) are defined in the presence of a 5 MHz E-UTRA interfering signal for frequency offsets lower than 20 MHz and a CW interfering signal for frequency offsets higher than 20 MHz from the uplink operating band. There is no explicit requirement for interfering signal ACLR in ETSI EN 301 908-14. Nevertheless, the $ACLR_{min}$ of the interfering signal to be used for conformance testing should be equal to ACS+10.2 dB (see ETSI TS 136 141 [12]). This condition implies that the OOBE of the interfering signal received by the receiver under test are very low and consequently will barely impact the compliance measurement results (0.4 dB additional rise in interference,).

In the case of ACS and narrow-band blocking, the ACLR values to be used in compliance measurements are defined in ETSI TS 136 141. ACLR is not applicable to CW interfering signal.

The above findings are summarised in Table 2:.

Table 2: ACLR values derived from ETSI EN 301 908-14 V13.1.1: IMT cellular networks; Harmonised Standard for access to radio spectrum; Part 14: Evolved Universal Terrestrial Radio Access (E-UTRA) Base Stations (BS)

Victim receiver (10 MHz)	Frequency offset	Require-ment	Type of interfering signal	ACLR requirement for the interfering signal (Note 1)	Assumption on ACS/ACLR used to define blocking requirements (Note 1) (Note 2) (Note 3)	ACLR requirement for conformance testing (Note 1) (Note 2) (Note 3)
Wide Area IMT BS	±2.5075 MHz	Receiver ACS (Iadj)	5 MHz E-UTRA signal	No requirement	$ACLR_{min}$ =56 dB ACS= $ACLR_{min}$ - 10.2 dB=45.8 dB	$ACLR_{min}$=56 dB $ACLR_{min}$=ACS+10.2 dB
Wide Area IMT BS	±(347.5 + m*180), m = 0, 1, 2, 3, 4, 9, 14, 19, 24	Receiver narrowband blocking level (I_{blk})	5 MHz E-UTRA signal, 1 RB (Note 4))	No requirement	$ACLR_{min}$ =59 dB ACS= $ACLR_{min}$-10.2 dB=48.8 dB	$ACLR_{min}$=59 dB $ACLR_{min}$=ACS+10.2 dB
Wide Area IMT BS	(F_{UL_low}-20) to (F_{UL_high}+ 20) (Note 5)	Receiver blocking level (I_{blk})	5 MHz E-UTRA signal	No requirement	$ACLR_{min}$ =ACS+10.2 dB	$ACLR_{min}$ =ACS+10.2 dB
Wide Area IMT BS	1 to (F_{UL_low}-20) (F_{UL_low}+20) to (F_{UL_high}+ 12750) (Note 5)	Receiver blocking level (I_{blk})	CW signal	No requirement	ACLR not applicable	ACLR not applicable

Note 1: The ACLR mentioned in this table refers to the ACLR of the interfering test signal.

Note 2: The contribution from the Test equipment ACLR is calculated to give a 0.4dB additional rise in interference. This corresponds to a Test equipment ACLR which is 10.2 dB better than the BS ACS (see ETSI TS 136 141 Table 4.1.2-2, 7.5 Adjacent Channel Selectivity (ACS) and narrow-band blocking; Note 2: d and e)

Note 3: The contribution from the Test equipment ACLR is calculated to give a 0.4dB additional rise in interference (see ETSI TS 136 141 Table 4.1.2-2, 7.6.5.1 Blocking (General requirements))

Note 4: Interfering signal consisting of one resource block is positioned at the stated offset, the channel bandwidth of the interfering signal is located adjacently to the lower/upper Base Station RF Bandwidth edge. Frequency offsets are such that the interfering signal is outside the channel

Note 5: F_{UL_low} and F_{UL_high} are the lowest and highest frequencies of the uplink operating band

A2.2.3.2 Applicability of MRR to determine the receiver blocking levels defined in ETSI EN 301 908-14

The applicability of MRR to determine the receiver blocking levels (I_{blk}) defined in ETSI EN 301 908-14 [8] has been checked. The calculations were carried out for two cases:

- FOS = ILR = $ACLR_{min}$;
- FOS = $ACLR_{min}$ and ILR = 100 dB (ILR>>FOS).

The ACS of the victim receiver being defined by the interfering signal level I_{adj}, this level has also been calculated by using the basic equation of MRR.

When verifying the applicability of MRR to determine the receiver blocking levels defined in ETSI EN 301 908-14, no calculations have been made in when the values of FOS (or ACS) and ILR (or ACLR) are not defined/provided and they cannot be calculated independently from the Iadj or Iblk defined in the harmonised standard. This is the case for the receiver blocking level tests.

In this section and the following section calculations have only been performed for 1.4 MHz, 5 MHz and 10 MHz Wide Area IMT BS receivers.

A2.2.3.3 Results of the calculations

The results of the calculations showed that for 1.4, 5 and 10 MHz Wide Area IMT BS receiver the I_{adj-ch} and narrowband I_{blk} derived by using MRR are very close to those defined in ETSI EN 301 908. Their difference is less than 1 dB. I_{blk} cannot be calculated since there is no information on the ACS and ACLR values used to define this parameter in ETSI EN 301 908 [8] and TS 136 141 [12].

The value of M to be used in the conformity tests is defined in ETSI EN 301 908 [8]. It varies from 6 to 27 dB depending on the channel BW and type of the IMT BS under test.

A2.2.4 Harmonised Standard ETSI EN 302 217-2 [9] "Fixed Radio Systems"

Note: the ACLR (or ILR) mentioned in this section refers to the ACLR (or ILR) of the interfering test signal.

A2.2.4.1 General comments

Neither ACS or FOS nor ACLR or ILR are used in ETSI EN 302 217-2 [9].

In this harmonised standard, the receiver selectivity is specified in terms of receiver sensitivity degradation (i.e. equal to M) in presence of a wanted signal like interfering signal of predefined protection ratio (PR=C/Iref) in the adjacent channels and in the presence of a generic unmodulated (CW interference) signal anywhere in a large portion of the spurious domain (blocking and spurious response requirement). Then, a kind of wide-band selectivity (WBSEL) is derived (i.e. based on the NFD method for FOS assessing referred in sSection A4.1 of this document) by comparing the adjacent channel C/I_{adj-ch} ratios, or the C/I_{blk} at given BER threshold degradation (M) to the co-channel C/I_{co-ch} ratio producing the same BER degradation.

In the presence of a wanted signal like-modulated interfering signal up to 2nd adjacent channel:

$$WBSEL_{adj-ch}(dB) = \frac{C}{I_{co-ch}}(dB) - \frac{C}{I_{adj-ch}}(dB) \tag{46}$$

and, in the presence of a CW interfering signals beyond 2nd adjacent channel.

$$\begin{aligned} WBSEL_{CW}(dB) &= \frac{C}{I_{co-ch}}(dB) - \frac{C}{I_{CW}}(dB) \\ &= \frac{C}{I_{co-ch}}(dB) - \frac{C}{I_{blk}}(dB) \end{aligned} \tag{47}$$

Note that when defining the resilience levels of a receiver, I_{in-ch} is often referred as I_{adj-ch} or I_{blk} depending on the frequency offset between the useful and interfering signals.

It is considered that this kind of wide-band selectivity (WBSEL) response is comprehensive of all effects (linear and not linear) that define the overall response of the receiver to interference; therefore, it is intended as the real selectivity of the digital receiver.

It is worth noting that WBSEL is equivalent to RIR (receiver interference ratio) or ACIR (adjacent channel interference ratio) within the adjacent channel frequency ranges (see ECC Report 310 0).[3]0). Note also that for ILR>>FOS, WBSEL is equal to FOS. Consequently, it can be safely assumed that $WBSEL_{CW}(dB) = FOS\ (dB)$, in the presence of CW interfering signals beyond 2nd adjacent channel.

As WBSEL is directly linked to the PR of digital fixed radio systems (DFRS) receivers, for the sake of simplicity, MRR has been used to derive C/I_{co-ch}, C/I_{adj-ch} and C/I_{blk} defined in ETSI EN 302 217-2.

The above findings are summarised in Table 3:.

Table 3: ILR values derived from ETSI EN 302 217-2 V3.3.1: Fixed Radio Systems; Characteristics and requirements for point-to-point equipment and antennas; Part 2: Digital systems operating in frequency bands from 1 GHz to 86 GHz; Harmonised Standard for access to radio spectrum [9]

Victim receiver	Frequency offset	Requirement	Type of interfering signal	Assumption on ILR requirements (Note 1)	Assumption on FOS/ILR used to define blocking requirements (Note 1)	ILR requirements for conformance testing (Note 1)
Digital Fixed Radio Systems (DFRS)	0 to 2nd adjacent channel	Selectivity ($C/I_{co\text{-}ch}$-$C/I_{adj\text{-}ch}$)	Wanted signal like-modulated interfering signal	Wanted signal like-ILR (Note 2)	FOS=ILR (1st adjacent) FOS ≥ ILR+ 10 dB (2nd adjacent)	No requirements
Digital Fixed Radio Systems (DFRS)	beyond 2nd adjacent channel	Selectivity ($C/I_{co\text{-}ch}$-C/I_{CW}) (Note 3)	CW carrier	No requirements	ILR >> FOS	No requirements
Note 1: The ILR mentioned in this section refers to the ILR of the interfering test signal Note 2: Derived from the spectrum mask requirement, tested up to the 2nd adjacent channel included, in ETSI 302 217-2 V3.3.1 (see also ANNEX 3: of this document) Note 3: WBSELCW range evaluated as C/IC - C/ICW; requirement in clause 4.3.3.3; value is valid from ±3rd CS (as centre frequency of the channel fully within the CW requirement range) and up to the frequency range where the CW test is defined by clause 7 of ETSI EN 301 390 [13]. It should also be understood that WBSELCW value is applicable on real interference environment only if the interfering signal emission exhibits a corresponding reduction of its OOB and spurious emissions within the victim DFRS RX bandwidth (see ETSI EN 302 217-2 V3.3.1, Figure P.1, NOTE 4).						

A2.2.4.2 Applicability of the basic equation of MRR for determining the receiver protection ratios defined in ETSI EN 302 217-2

The applicability of MRR for determining the PR ($C/I_{co\text{-}ch}$, $C/I_{adj\text{-}ch}$ and C/I_{blk}) of receiver defined in ETSI EN 302 217-2 [9] has been checked. The calculations were carried out for two cases:

- in the presence of a co- channel wanted signal like-modulated interfering signal. In this case Equation (1) was simplified as FOS and ILR have no impact on the calculation of $I_{co\text{-}ch}$:

$$I_{co-ch}(dBm) = 10log\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) \quad (48)$$

  in the presence of a 1st adjacent channel wanted signal like-modulated interfering signal. In this case there is no information on receivers FOS in ETSI EN 302 217-2. Nevertheless, with the interfering signal as a wanted signal like-modulated signal, it is sensibly assumed that FOS=ILR (or equivalently ACS=ACLR).

When verifying the applicability of MRR to determine the receiver blocking levels defined in ETSI EN 302 217-2, no calculations have been made when the values of FOS (or ACS) and ILR (or ACLR) are not defined/provided and they cannot be calculated independently from the I_{blk} defined in the harmonised standard. This is the case for the receiver blocking level tests with a wanted signal like-modulated interfering signal at the 2nd adjacent channel and in the presence of a CW interfering signal beyond the 2nd adjacent channel.

In this section, calculations have only been performed for 7 MHz and 14 MHz channels DFRS Class 2 receivers operating respectively in the bands 3.5 GHz and 26 GHz have been considered. On the other hand, the spectral emission requirements (spectrum mask) depend on the channel size and are equal for frequency bands from 3.5 to 57 GHz; in addition, DFRS modulation-demodulation techniques, optimised for best S/N, apply the Nyquist principle of equal split between Tx and Rx of the overall square-cosine pulse shaping filtering (rendering parametric as well the RX filtering). Therefore, the results can be intended valid for any channels size and frequency bands.

A2.2.4.3 Results of the calculations

The results of the calculations show that in the 3.5 and 26 GHz bands the 7 and 14 MHz DFRS receiver $C/I_{co\text{-}ch}$, and $C/I_{adj\text{-}ch}$ derived by using MRR are close to those defined in ETSI EN 301 390 [13]. Their difference is less than 3 dB.

Two single M values, 1 and 3 dB, are used in the conformity tests defined in ETSI EN 302 217-2 [9].

A2.2.5 Harmonised Standard ETSI EN 301 502 [10] “Global System for Mobile communications (GSM)”

Note: the ACLR (or ILR) mentioned in this section refers to the ACLR (or ILR) of the interfering test signal.

A2.2.5.1 General comments

Neither ACS or FOS nor ACLR or ILR are used in ETSI EN 301 502 [10].

In this harmonised standard, the receiver selectivity is specified in terms of receiver sensitivity degradation in presence of a wanted signal like interfering signal of predefined protection ratio ($PR=C/I_{ref}$) in the adjacent channels (1st and 2nd). The blocking level (I_{blk}) is defined beyond the 2nd adjacent channel and in a large portion of the spurious domain (blocking and spurious response requirement) in the presence of a generic unmodulated (CW interference) signal. Consequently, MRR has been used to derive $C/I_{co\text{-}ch}$, $C/I_{adj\text{-}ch}$ and I_{blk} defined in ETSI EN 301 502.

The above findings are summarised in the following table.

Table 4: ILR values derived from ETSI EN 301 502 V12.5.1: Global System for Mobile communications (GSM); Base Station (BS) equipment; Harmonised Standard covering the essential requirements of article 3.2 of the Directive 2014/53/EU [10]

Victim receiver	Frequency offset	Requirement	Type of interfering signal	Assumption on ILR requirements (Note 1)	Assumption on FOS/ILR used to define blocking requirements (Note 1)	ILR requirements for conformance testing (Note 1)
GSM 400/GSM 900/DCS 1800 BTS	0 to 2nd adjacent channel	Selectivity (Note 2) ($C/I_{adj\text{-}ch}$); M=20 dB	Wanted signal like-modulated interfering signal	Wanted signal like-ILR	FOS=ILR for 1st adjacent channel	No requirements
GSM 400/GSM 900/DCS 1800 BTS	Beyond 2nd adjacent channel	Blocking (I_{blk}); M=3 dB	CW signal	No information provided	No information provided	No requirements
Note 1: the ILR mentioned in this table refers to the ILR of the interfering test signal						
Note 2: Adjacent channel interference rejections for circuit switched channels except Enhanced Circuit-Switched Data (ECSD)						

It is also important to note here that no information has been found on GSM BS noise figure (NF) as well as SNR in ETSI HS or TS. A GSM BS NF of 8 dB is used in ECC Report 146 [14]. This value has also been used in this Report to calculate the noise floor of GSM receivers.

A2.2.5.2 Applicability of the basic equation of MRR for determining the receiver protection ratios defined in ETSI EN 301 502

The applicability of MRR for determining the PR ($C/I_{co\text{-}ch}$, $C/I_{adj\text{-}ch}$ and C/I_{blk}) of receiver defined in ETSI EN 301 502 [10] has been checked. The calculations were carried out for two cases:

- in the presence of a co- channel wanted signal like-modulated interfering signal. In this case Equation 1 was simplified as FOS and ILR have no impact on the calculation of $I_{co\text{-}ch}$:

$$I_{co-ch}(dBm) = 10log\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) \tag{49}$$

3

- in the presence of a 1st adjacent channel wanted signal like-modulated interfering signal as well as in the presence of a 2nd adjacent channel wanted signal like-modulated interfering signal. In these cases, there is no information on receivers FOS in ETSI EN 301 502 . Nevertheless, GSM BS 1st and 2nd adjacent channels FOS values provided in ECC Report 146 [14] has been used in the calculation. The interfering signal being wanted signal like-modulated signal, it is sensibly assumed that FOS=ILR in the case of 1st adjacent channel interference.

When verifying the applicability of MRR to determine the receiver blocking levels defined in ETSI EN 301 502, no calculations have been made when the values of FOS (or ACS) and ILR (or ACLR) are not defined/provided and they cannot be calculated independently from the I_{blk} defined in the harmonised standard. This was the case for the receiver blocking level tests in the presence of a CW interfering signal beyond the 2nd adjacent channel.

In this section, calculations have only been performed for GSM 400/GSM 900/ER-GSM 900/DCS 1800 Macro-BTS (GMSK modulated carrier) type receivers have been considered.

A2.2.5.3 Results of the calculations

The results of the calculations show that for 200 kHz GSM BST receiver the $C/I_{co\text{-}ch}$ and $C/I_{adj\text{-}ch}$ derived by using MRR are very close to those defined in ETSI EN 301 502 [10]. Their difference is less than 1 dB, except the two $C/I_{adj\text{-}ch}$ calculated with assumption of ACS=ACLR. In these latter cases, the difference is 3 dB. This difference is probably due to the assumption made on the GSM ACS in ECC Report 146. It seems that in ECC Report 146 [14] the GSM ACS values were derived as follows:

$$ACS(dB) = \frac{C}{I_{co-ch}}(dB) - \frac{C}{I_{adj-ch}}(dB) \tag{50}$$

which is RIR (dB) and not ACS (dB) unless ILR>>FOS.

RIR is equal to FOS if and only if ILR>>FOS. Consequently, it can be safely assumed that $RIR_{blk}(dB) = FOS\ (dB)$ in the presence of CW interfering signals beyond the 2nd adjacent channel.

The value of M to be used in the conformity tests is defined in ETSI EN 301 502 [10]. They are 3 and 20 dB respectively for blocking and selectivity tests.

A2.2.6 Conclusions

- The results of the in-depth analysis carried out on four ETSI Harmonised Standards show that, when accurate information on FOS and ILR is provided in relevant standards, the receiver blocking level (I_{blk}) or the carrier to interference blocking level ratio C/I_{blk} derived by MRR are very close to those defined in HS. This confirms that MRR is an accurate and robust method for deriving blocking levels of receivers for the future recommendation on receivers. Moreover, the method can also be used to check the consistency between different requirements defined in HS;
- The main obstacle to using MRR to check the relevance of the receiver blocking requirements defined in HS is the lack of information on the ACS and ACLR values used to define those requirements. It would be sensible that this important information is provided in HS.

ANNEX 3: GENERAL GUIDANCE ON PARAMETERS M, N, ILR AND FOS

A3.1 INTRODUCTION

Administrations have a duty to ensure efficient use of spectrum. There needs to be sufficient and complete information to assess the ability of services to coexist both on the same and adjacent frequencies and bands. This is an important consideration when services, systems and applications are seeking spectrum to use and applies to both new and existing services, systems and applications.

Administrations now face the challenge of stretching spectrum utilisation with greater numbers of users. Administrations require accurate information on the relevant receiver parameters in addition to the transmitter parameters.

All the following parameters need to be considered when designing of radio receivers: sensitivity, selectivity, desensitisation, blocking and noise performance. To achieve efficient use of the spectrum, services should only use sufficient signal power levels necessary to provide a satisfactory quality of service. This required quality of service will vary: some services absolutely must work immediately during an emergency while others may require a less reliable service where a message is delivered even after a long delay.

A common way to overcome interference is to increase the transmitter power levels for the wanted service to increase the C/I ratio. However, this will increase the likelihood of interference to other users driving higher signal level requirements into other systems. This is clearly not energy-efficient neither cost-effective in the long term.

Some services operate with a very low field strength. This implies that services in adjacent frequencies should operate at very low levels of unwanted emissions to avoid interference. Furthermore, very sensitive receivers need to be able to avoid overload if its receiver filtering does not remove all the adjacent signals. Table 5 is an extract from ECC Recommendation (02)01 [15] and gives an overview of the negative effects on spectrum use and compatibility when the performance of receivers is poor in one or several of its characteristics:

Table 5: Impact of Reference receiver performance parameters on spectrum utilisation and efficiency of radio equipment

Reference receiver performance parameters	Impact on spectrum utilisation and efficiency of radio equipment with poor receiver performance parameters
Sensitivity	Increase of number of transmitters (base stations) Increase of transmitter power Increased spectrum demand if number of transmitters and transmitter powers cannot be changed Increased difficulty to elaborate channel plans which leads to more interference to other services and system capacity loss and therefore an inefficient spectrum use
Blocking, desensitisation, spurious response, protection ratio, co-channel rejection, receiving mask, selectivity and adjacent band rejection	Decrease of number of transmitters of the interfering service and Decrease of transmitter power of the interfering service which leads to system capacity loss for the interfering service and consequently more spectrum for the other service and increase of the interfering probability to the wanted radio service.

Radio services very often have the capacity to dynamically adapt the transmission modulation and operating characteristics based on the radio channel environment. Where a service has an adaptive capability, its technical specification needs to provide sufficient detail about the equipment state during each individual measurement of the dynamic adaptation (including modulation, throughput, etc.). Lack of clarity in the technical details for each measurement will result in difficulties for carrying out compatibility studies. The characterising documentation should provide this additional information alongside the values measured in each case. Similarly, the lowest signal level at which the radio service can or is designed to work at usually conveys a low capability (for example, low data rate streams or low audio quality telephony codecs). Higher capability, e.g. throughput, is usually only achieved at higher quality received signal levels. Clarity around received signal levels and system performance is needed in the documented values in order to facilitate administrations sharing and compatibility studies requirements. They should not be masked inside a large margin or, for example by omitting the margin at a particular measurement to describe a failure point. These values need to have clear explicit additional notes indicating the performance capability or operating state for each measurement and if they incorporate a safety margin or are to the point of full failure.

Administrations also need to consider the expectation of equipment users. Measurements lacking clarity on which modulation mode is in use and that mode's capability, can also make sharing and compatibility study calculations very difficult without additional contextual information. This, for example, may then require additional bench measurement campaigns to supplement the available documentation. Clarity in how a service/application equipment works as a system is required at least for the RF air interface capability where dynamic mode changes may occur in both transmitters and receivers.

In summary, both the environment and the whole set of equipment performance parameters need to be considered in the assessment, with existing examples of services and services as well as new ones. The receiver noise figure, sensitivity, dynamic range and its signal filtering all play a part, and not just the filtering alone. Similarly, the filtered final emission characteristics of the transmitter also directly interact in the assessment.

A3.2 OVERVIEW OF RELEVANT RECEIVER PARAMETERS TO CONSIDER IN THE METHODOLOGY

This section includes details about how the four different receiver parameters under discussion (M, N, ILR and FOS as defined in Section 2) are interlinked in the context of receiver resilience against harmful interference.

As defined in Section 3.1, the interfering signal power received by the victim receiver is, in the logarithmic domain:

$$I_r(dBm) = 10\log_{10}\left(10^{\frac{N(dBm)}{10}}10^{\frac{M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) \quad (51)$$
$$= 10\log_{10}\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right)$$

When the interfering signal is adjacent to the victim, the signal power received by the victim is also defined as:

$$I_r(dBm) = I_{in-ch}(dBm) + 10\log_{10}(10^{-ILR(dB)/10} + 10^{-FOS(dB)/10}) \quad (52)$$

Equating (51) and (52), the four parameters are related to the interference signal as:

$$10log\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) \quad (53)$$
$$= I_{in-ch}(dBm) + 10\log_{10}(10^{-ILR(dB)/10} + 10^{-FOS(dB)/10})$$

A3.2.1 Receiver noise floor (N)

Every radio receiver is subject to a noise floor that can be described using the following equation and expressed in dB[2]:

$$N(dBm) = 10\log_{10} kTB + NF + 30 \quad (54)$$

Where:

- N is the receiver noise floor;

[2] Also defined in ECC Report 252 (SEAMCAT handbook) as the level of noise introduced by the receiver system below which the signal that is being captured cannot be isolated from the noise (see ECC Report 252, section 1.2.3) [16])

- k is the Boltzmann constant in Joules per Kelvin (1.381×10^{-23});
- T is the temperature in degrees Kelvin (for common terrestrial radio receivers, 290 K can be used);
- B is the receiver bandwidth in Hertz;
- kTB is the receiver thermal noise in dBW;
- NF is the receiver Noise Figure in dB.

Compatibility studies require a figure for, or be able to derive, the composite value of “N”. Without this essential figure, the overall baseline of the systems operating characteristics has no reference. “N” is an aggregate noise value that includes:

- The receiver thermal noise, which is always present from thermal considerations by the classic k, T and the receiver bandwidth (B) and;
- the receiver’s own internal system noise contribution or noise figure (NF).

The NF is mainly defined by the first active device (e.g. an LNA or a simple diode for direct signal conversion/demodulation) of the front end, added to the attenuation of any passive circuitry (e.g. a filter) between the antenna port and that active device itself. The design target of the NF depends on the expected performance of the radio system in terms of the operational range (e.g. maximum distance or cell size diameter between Tx and Rx in the field); therefore, NF target should be balanced by the corresponding Tx EIRP.

Usually the NF is minimised in order to reduce the costlier Tx power; however, in some cases, e.g. in SRDs, the operational range might be significantly low, justifying a high (and cheaper) NF target. Clearly, the higher is the NF and the less sensitive is the Rx to high level of “in-channel” interference (i.e. the same level of interference produces a smaller desensitisation than on a receiver with better NF). When off-channel” interference is considered, the NF plays a lesser role, provided that low NF receivers are designed with better selectivity.

If not known through the receiver characteristics given by the manufacturer, the value of NF may have to be derived from not one, but indirectly from several measurements when testing a particular product. Without the knowledge or a good estimation of NF, there is no baseline reference for any studies.

A3.2.2 Receiver desensitisation (M)

A3.2.2.1 General considerations

Elaborating from ANNEX 1: (Equation (30)) the desensitisation M (dB) can be defined as:

$$M = 10\log_{10}\left[\frac{I_r + N}{N}\right] \tag{55}$$

Where:

- I_r is the total interfering signal power (in W) (see Figure 1);
- N is the receiver noise floor (in W) measured in the receiver bandwidth.

To ensure proper operation, the operational range of the radio system is designed to include a margin equal or lower than M which allows it to tolerate a certain level of interference (I) in the listened channel. This can be caused by co-channel and/or non-co-channel interference sources. When running a radio network or a radio link, the objective is to maintain the signal to interference and noise ratio SINR as close as possible to the signal to noise ratio (SNR), in any case within the factor M difference representing the operational margin to interference.

Like N, M is an aggregate value that includes the common protection ratio to ensure a stability of the service reception. This protection ratio in addition to the receiver dynamic performance need to be clearly defined.

M corresponds to the ‘equivalent noise rise’ due to the interfering signal (in-band and off-band) at the input stages of the receiver. Less predictable Receiver desensitisation also occurs for non-linear effects when a strong off-channel signal overloads a receiver front end and thus reduces the sensitivity to weaker on-channel wanted signals. This effect is caused for example by reciprocal mixing, due to phase noise, and A/D or D/A performance ranges.

Use of large values of additional margin may overprotect a service unnecessarily; low values of additional margin may under-protect a service and make it prone to interference. The proper choice of balanced “N” and “M” values depends on the service requirements. This must include other external effects (for example propagation fading and building signal path loss variations etc) that may require consideration as well as the expectation of the user of the equipment.

In many cases of radio services where stringent QoS (Quality of Service) is required (e.g. in fixed service links), M should not be considered as a “variable”, but as the “maximum acceptable degradation” of the victim receiver due to interference. Its value depends on (or it is related to) the QoS of the specific victim service receiver and, in some cases it is fixed by the ITU-R or other international regulation (e.g. for FS links by ITU-R Recommendation F.1094 [5] and its practical application guidelines in Recommendation F.758 [6]).

A3.2.2.2 Practical approach

The desensitisation values (M) used for receivers blocking conformity tests in the following ETSI Harmonised Standards are presented in Table 6:

- ETSI EN 303 340 V1.2.1: Digital Terrestrial TV Broadcast Receivers; Harmonised Standard for access to radio spectrum [4];
- ETSI EN 301 908-14 V13.1.1: IMT cellular networks; Harmonised Standard for access to radio spectrum; Part 14: Evolved Universal Terrestrial Radio Access (E-UTRA) Base Stations (BS) [8];
- ETSI EN 302 217-2 V3.3.1: Fixed Radio Systems; Characteristics and requirements for point-to-point equipment and antennas; Part 2: Digital systems operating in frequency bands from 1 GHz to 86 GHz; Harmonised Standard for access to radio spectrum [9];
- ETSI EN 301 502 V12.5.1: Global System for Mobile communications (GSM); Base Station (BS) equipment; Harmonised Standard covering the essential requirements of article 3.2 of the Directive 2014/53/EU [10].

Table 6: Desensitisation values (M) used for blocking conformity tests in the analysed ETSI Harmonised Standards

Desensitisation values (M) used for blocking conformity tests in the analysed ETSI Harmonised Standards	
Harmonised Standard	M (dB)
DTT	13 and 14
IMT cellular networks: ▪ Blocking and NB blocking ▪ NB blocking for NB-IoT wanted signal	6 and 14 (Note 1) 6 (Note 2), 8, 10, 11 and 12
Fixed Radio Systems	1
GSM	3
Note 1. Used only for Home BS which has a sensitivity 8 dB higher than that of other BS categories Note 2. Used in the majority of the NB-IoT wanted signal blocking tests	

As shown in Table 6, the value of M to be used in blocking requirement conformity tests may vary from one HS to another. It may also vary depending on the interference configurations considered, as is the case in the HS dealing with IMT cellular networks. Consequently, the choice of Rx desensitisation value (M) when calculating receiver blocking levels by MRR may seem to be difficult at first, and it is worthwhile to explore this matter in more detail. This is done in the following section.

A3.2.2.3 Variation of blocking level as a function of M

In this section Equation (30) has been used to evaluate the variation of receivers blocking level (I_{blk}) as a function of the value of M.

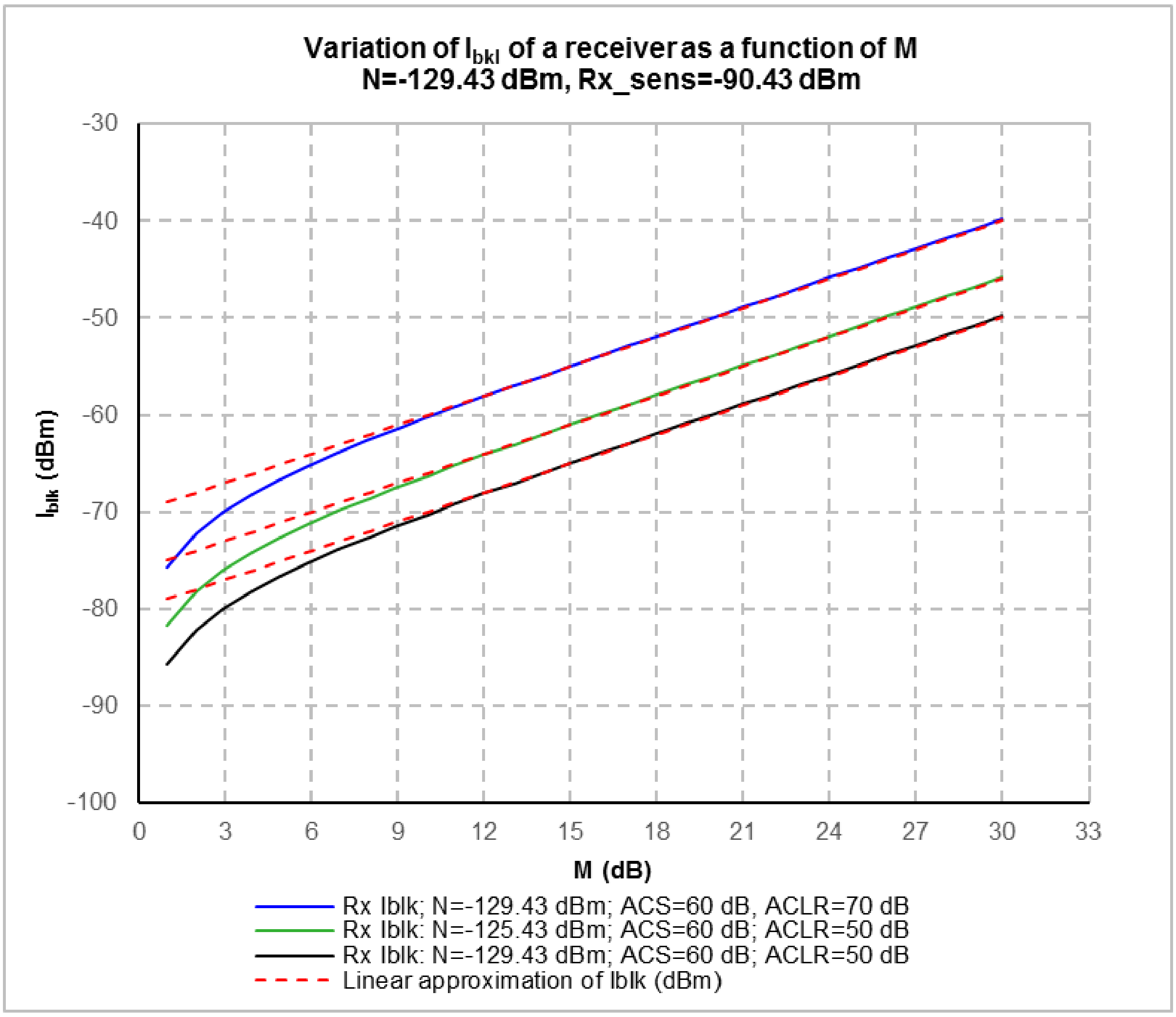

Figure 3: Variation of blocking level as a function of M

Figure 3 shows three different blocking curves, calculated from Equation (30), representing the variation of I_{blk} as a function of M for three different sets of N, ACS and ACLR values.

These blocking curves show that:

- while the value of I_{blk} depends on N, M, ACS and ACLR, the gradient (or slope) of a blocking curve is identical for all possible sets of N, ACS and ACLR values;
- for the values of M greater than 9 dB the gradient of the curve is equal to 1. Consequently, above this value of M the blocking curves can be approximated to unity slope lines;
- for the values of M smaller than 9 dB the blocking curve deviates from the unity slope line due to the impact of the receiver noise floor (noise power). Actually, a low M value means a low useful signal level thus an increased impact of noise on the receiver performances, while a high M value means a high useful signal level that reduces the impact of noise on the receiver performances ($M(dB) = C(dBm) - Rx_{sens}(dBm)$).

In conclusion, independently from the technology used, the blocking curve of receivers varies, as a function of M, linearly with unity slope for M > 9 dB and non-linearly with a slope depending on M for M < 9 dB. For a given M, the gradient (or slope) of the blocking curve is identical for all possible sets of N, ACS and ACLR values.

A3.2.2.4 Choice of the Rx desensitisation value M

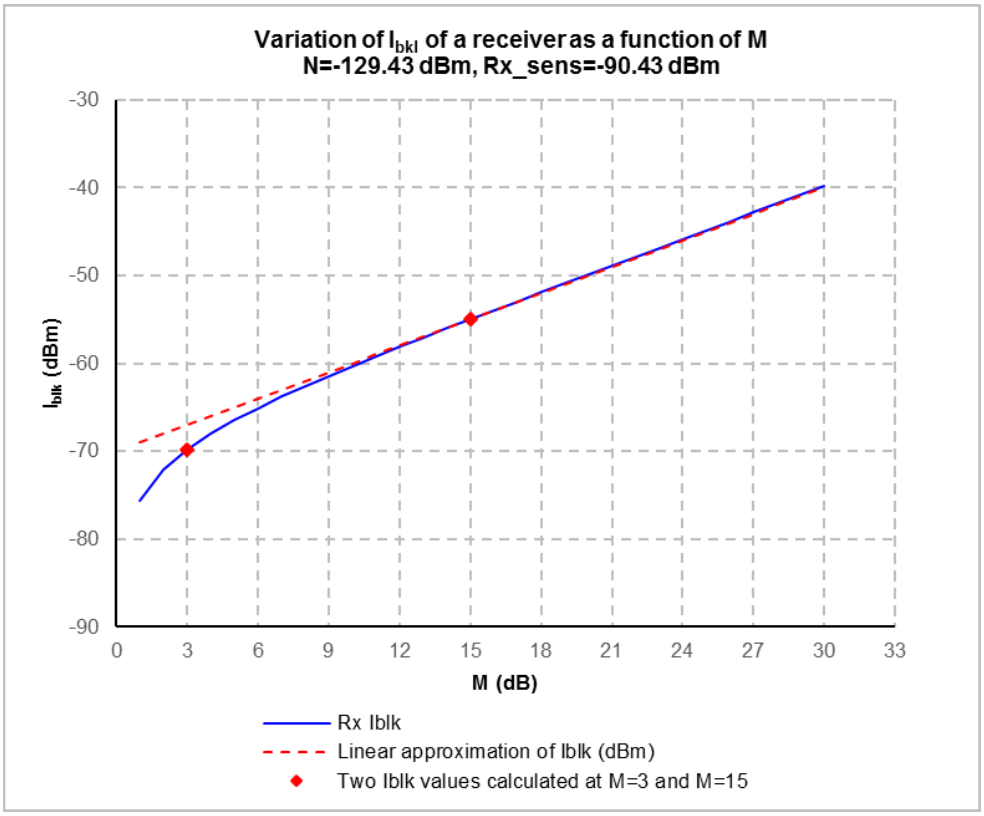

Figure 4 Choice of the Rx desensitisation value M

As shown in Figure 4 and explained in the previous section, for the values of M greater than 9 dB receivers blocking level varies linearly as a function of M, while for the values of M smaller than 9 dB its variation is not linear due to the impact of the receiver noise floor (noise power). Considering this behaviour, which is identical for all receivers, it would be sensible to define the blocking requirements of a receiver for two different values of M, one below and one above 9 dB.

For example, M=3 dB (historical desensitisation value), which takes into account the impact of the receiver noise floor on I_{blk} and M=15 dB, which takes into account high useful signal levels where the receiver noise floor does not have any impact on I_{blk}, might be a reasonable choice. This choice is shown in Figure 4 (red points on the blocking curve). When the blocking level of a receiver is defined with these two values of M, its blocking curve can easily be predicted thanks to the unity slope line.

Consequently, if M is not given or cannot be calculated from $M(dB) = C(dBm) - Rx_{sens}(dBm)$, it is proposed to define the blocking requirements of receivers for M = 3 and/or 15 dB in the ECC Recommendation on receivers.

A3.2.3 Frequency offset selectivity and interference leakage ratio

A3.2.3.1 Consideration of different bandwidths of victim and interferer

The parameters FOS and ILR of the basic equation of MRR implicitly take into account the interfering and victim signal bandwidths as described in this section.

However, when the FOS and ILR are derived from ETSI standards, it might be possible that their definition in the standard is relative to the same bandwidth of the system subject of the standard, either the victim or the interferer. Therefore, care should be taken when such ETSI data are used for compatibility study.

FOS is of general use for any mixed wanted and interfering signal situation and can be calculated as follows:

$$FOS(dB) = -10\log_{10}\left(10^{\frac{-RIR(dB)}{10}} - 10^{\frac{-ILR(dB)}{10}}\right) \quad (56)$$

Where:

- *FOS*: Frequency Offset Selectivity;
- RIR: Receiver Interference Ratio at offset frequency;
- ILR: Leakage Power Ratio of the interfering signal at offset frequency.

RIR and ILR can be expressed as follows:

$$RIR = \frac{I}{I_r} \quad (57)$$
$$ILR = \frac{I}{I_{OOB}}$$

Where:

- I: Interfering signal power measured in its bandwidth;
- Ir: Total interfering signal power received by the victim receiver;
- IOOB: Interfering signal out-of-block or out-of-band power falling into the victim receiver bandwidth.

The above mathematical equations clearly show that ILR and therefore FOS can be calculated for all type of useful and interfering signal bandwidths. However, if ILR is measured in a bandwidth that is different than the victim receiver bandwidth under consideration, and there is no possibility to carry out new measurements, it is necessary to use an appropriate correction factor to define the ILR in the victim receiver bandwidth.

If ILR is measured in the bandwidth of victim receiver A and will be used to calculate I_{blk} of victim receiver B at the same frequency offset from the interferer and there is no abrupt variation of the interferer emission mask around this frequency offset, then the ILR in the victim receiver B bandwidth can be expressed as follows:

$$ILR_B \cong ILR_A - 10 \times \log_{10}(\frac{BW_{VB}}{BW_{VA}}) \quad (58)$$

Note also that a FOS calculated for a given interferer bandwidth may not be applicable to an interferer that have a different bandwidth, which is also true for the FOS (or ACS) values defined in ETSI Harmonised Standards.

It is also important to note that FOS implies attenuation ratios and not power ratios as in the case of ILR. Consequently, it is not possible to apply any bandwidth ratio correction factor to FOS. For example, if the frequency offset between victim and interferer is very large, then the receiver FOS may be flat. Under such conditions the interfering signal bandwidth will have no effect on the value of FOS. This can also be the case when two interferers with different bandwidth have the same offset frequency from the victim and there is no abrupt variation of the receiver FOS.

In conclusion, the bandwidths of victim and interfering systems in EC/ECC Decisions, ECC Recommendations and ETSI Harmonised Standards should be carefully used and possibly corrected on case by case.

A3.2.4 Out-of-band and spurious domains and derivation of ILR value

Out-of-band (OOB) domain (of an emission) is the frequency range, immediately outside the necessary bandwidth but excluding the spurious domain, in which OOB emissions (OOBE) generally predominate. The spurious domain (of an emission) is the frequency range beyond the OOB domain in which spurious emissions generally predominate.

The boundary between the OOB and spurious domains is the separation between the centre frequency of an emission and the beginning of its spurious domain. According to Recommendations ITU-R SM.329-12 [7] and Appendix 3 of the Radio Regulations, the boundary between the OOB and spurious domains is generally the centre frequency of the emission by 250% of the necessary bandwidth of the emission as shown in Figure 5 extracted from ECC Recommendation (02)05 [17].

Note that for single carrier systems, including systems using OFDM/OFDMA, channel bandwidth or channel separation can be used as a substitute for necessary bandwidth provided that they are found in ITU-R Recommendations or in relevant regional and national regulations. Multicarrier systems are not relevant to the determination of receiver resilience levels.

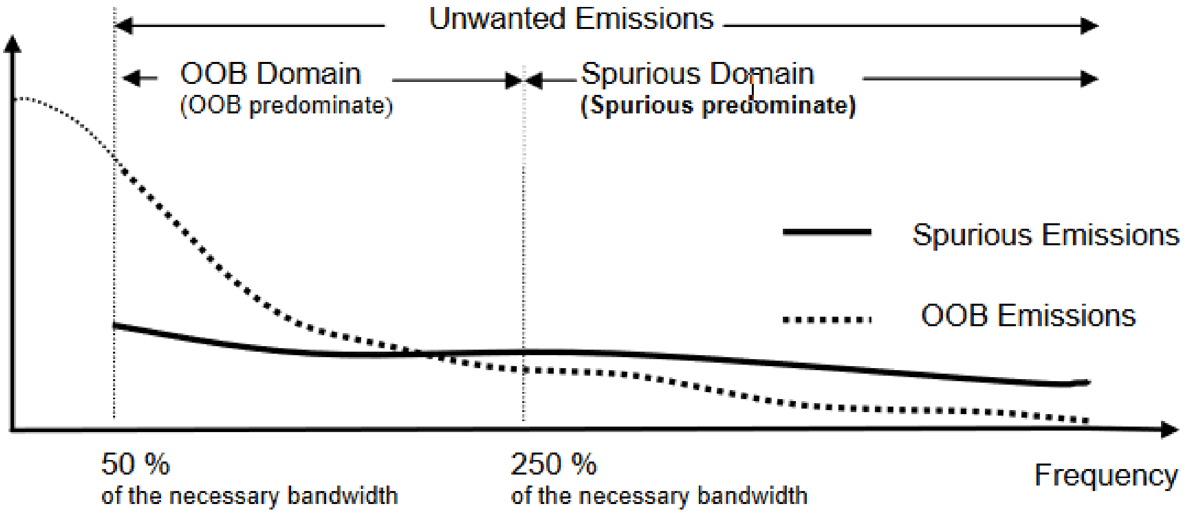

Figure 5: Illustration of the OOB and Spurious Domains [17]

Note: The crossover point between OOB and spurious emissions will vary and Figure 5 shows only an example.

However, the OOB/spurious domain boundary needs to be modified for narrow-band and wideband (including multicarrier) systems, and certain other situations. This modification is very straightforward as described in Table 2 of Recommendation ITU-R SM. 1539-1 [11].

As in OOB domain OOB emissions generally predominate, while in spurious domain spurious emissions generally predominate, it is proposed to choose the ILR value to be used in MRR according to the OOB/spurious domain boundary of the interfering signal. Since:

- If the victim receiver channel is in the interfering signal OOB domain, the interfering signal ILR would be similar to its ACLR values defined on the 1st and 2nd adjacent channels. Consequently, the interfering signal ILR can be found in the existing EC/ECC Decisions, ECC/ITU-R Recommendations, CEPT/ECC Reports and ETSI Harmonised Standards/technical specifications/reports or can be easily determined by measurements;
- If the victim receiver channel is in the interfering signal spurious domain, the interfering signal ILR can be derived from the spurious emission levels defined in the existing EC/ECC Decisions, ECC/ITU-R Recommendations (e.g. ERC Recommendation 74-01 [2] and Recommendation ITU-R SM.329-12 [7]) and ETSI Harmonised Standards.

A3.3 GUIDANCE TO DETERMINE RELEVANT INTERFERING SIGNAL AND INTERFERENCE SCENARIO

The use of reference interfering signal and a single interference scenario as proposed in ANNEX 5: will overcome the difficulty of identifying the relevant interfering signal and interference scenario for each service/system.

Otherwise, the most relevant interfering signal and interference scenario can be defined between the competent CEPT and ETSI technical groups.

Moreover, if needed, a guidance on how to determine relevant interfering signal and interference scenario for a given system can be included in the Recommendation on receivers.

ANNEX 4: EXAMPLE METHODS FOR TESTING

A4.1 NFD ORIENTED TEST OF FOS

RIR is the generic extension to any frequency offset of the ACIR defined for adjacent channel offset when a channel arrangement is identified; ECC Report 310 (Equation (7)) [3] also considers the following equivalence:

$$ACIR(dB) = \left(\frac{C}{I}\right)_{co-ch}(dB) - \left(\frac{C}{I}\right)_{adj-ch}(dB) \tag{59}$$

Where the C/I ratios are considered at the same desensitisation M; in addition, the report recognises that: "ACIR is practically equivalent to the Net Filter Discrimination (NFD) which may be extended to any wanted and interfering frequency separation;" (e.g. to the frequency offset (Δ_f) where the FOS is defined)

Therefore, for the purpose of this test method, when FOS evaluation can be made through C/I requirements, the following formula can also be considered:

$$RIR(dB) = 10\log_{10}\left(\frac{1}{\frac{1}{ILR} + \frac{1}{FOS}}\right) = \left(\frac{C}{I}\right)_{co-ch}(dB) - \left(\frac{C}{I}\right)_{\Delta_f}(dB) = NFD_{\Delta_f}(dB) \tag{60}$$

and, when ILR << FOS:

$$FOS(dB) \cong RIR(dB) = \left(\frac{C}{I}\right)_{co-ch}(dB) - \left(\frac{C}{I}\right)_{\Delta_f}(dB) = NFD_{\Delta_f}(dB) \tag{61}$$

Therefore, whenever a radio equipment standardises also a $\left(\frac{C}{I}\right)_{co-ch}$ behaviour and assuming that the required FOS is derived as in Step 3 of Section 4.4.2, the tests can be made on the basis of NFD defined as in the above equation (64).

This implies a $\left(\frac{C}{I}\right)_{\Delta_f}$ (usually with a CW interfering signal) at the required Δ_f with the same M desensitisation used for the $\left(\frac{C}{I}\right)_{co-ch}$ test.

It is important to note that NFD can be considered as a sub-set of MRR. The major shortcoming of NFD, compared to MRR, is its lack of ability to directly take into account N and M, and therefore the results of NFD are accurate and identical to those obtained by MRR only for high M values (M ≥ 9 dB).

MRR and NFD can be compared simply by comparing the results of Equations (62) and (63) when calculating the adjacent channel PR (PR_{adj-ch}) of a receiver based on the receiver parameters presented in Table 7:

MRR:

$$PR_{adj-ch}(dB) = C(dBm) - 10\log_{10}\left(10^{\frac{N(dBm)+M(dB)}{10}} - 10^{\frac{N(dBm)}{10}}\right) + 10\log_{10}\left(10^{\frac{-ILR\,(dB)}{10}} + 10^{\frac{-FOS\,(dB)}{10}}\right) \tag{62}$$

NFD:

$$PR_{adj-ch}(dB) = PR_{co-ch}(dB) + 10\log_{10}\left(10^{\frac{-ILR\,(dB)}{10}} + 10^{\frac{-FOS\,(dB)}{10}}\right) \tag{63}$$

Table 7: Receiver parameters used to calculate the adjacent channel protection ratio PR_{adj-ch} of a receiver, for comparison between MRR and NFD

Rx N (dBm)	Rx sensitivity (dBm)	C/N (dB)	PR_{co-ch} (dB)	ILR (dB)	FOS (dB)	M (dB)
99.43	90.43	9	9	50	60	1 to 30

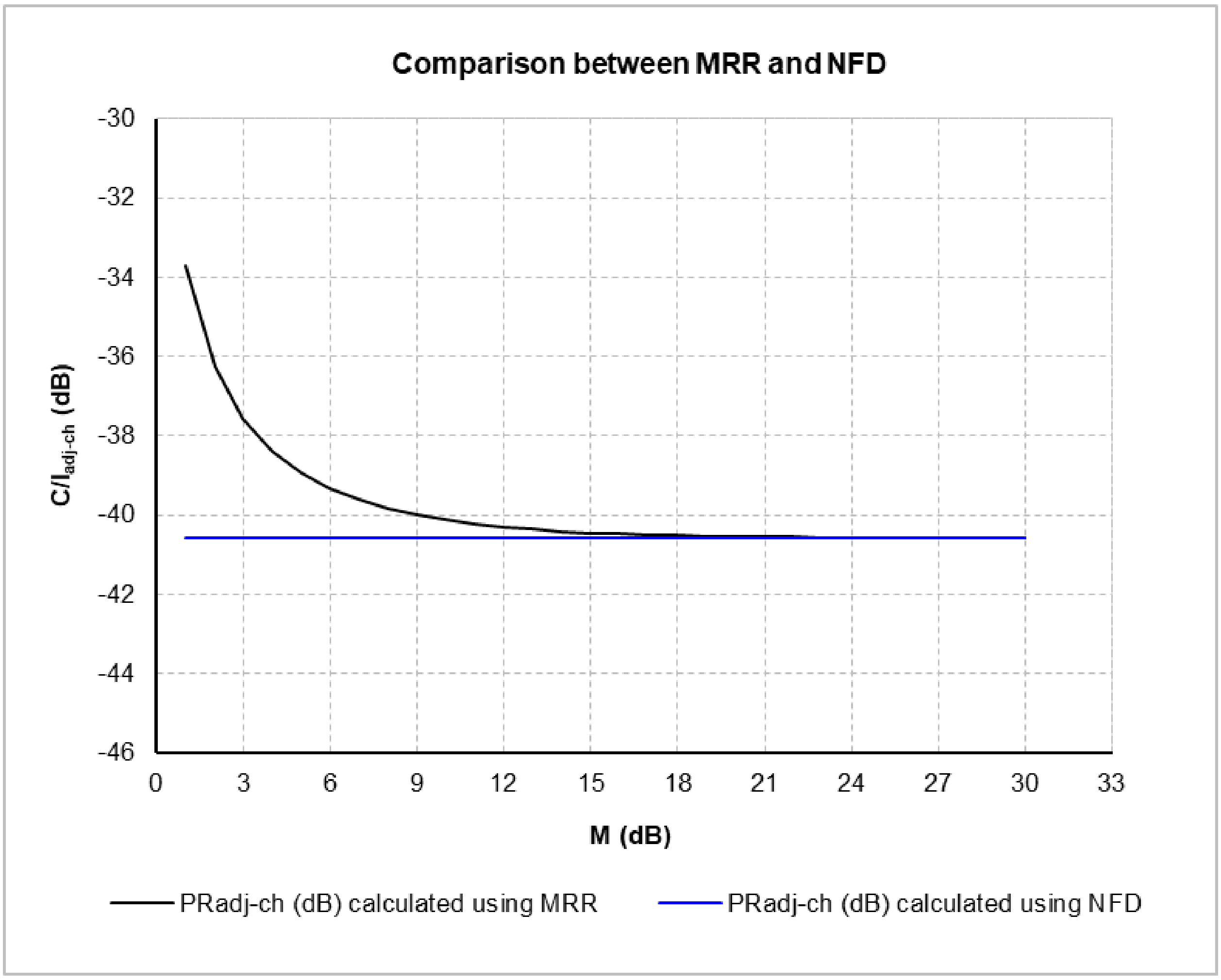

Figure 6: Comparison of MRR and NFD

The results of the calculations are shown in Figure 6, which make it obvious that the results of NFD are accurate only for high M values (M ≥ 9 dB), since the basic equation of this method does not contain any information about N and M. Consequently, for M < 9 dB, the adjacent channel PR, including blocking level, and their variation as a function of N and M cannot be evaluated using NFD unless for a given N value the co-channel PR of the receiver is calculated/measured for different M values and the calculated/measured values are used step by step in Equation (61) to calculate the adjacent channel PR as a function of M. Consequently, NFD can be considered as a simplified form (sub-set) of MRR.

ANNEX 5: REFERENCE INTERFERING SIGNAL AND A SINGLE INTERFERENCE SCENARIO TO BE USED IN THE CALCULATION OF RECEIVERS RESILIENCE LEVELS USING MRR

A5.1 INTRODUCTION

When using MRR to define the adjacent channel protection ratio (PR) and the blocking level of receivers, the relevant interfering signal and interference scenario are determined from existing or planned deployment of victim and interfering systems and the compatibility studies presented in various CEPT/ECC, ETSI and ITU-R technical reports, or, if the information in the previous reports is not available, from the intra/inter-system interfering signals and interference scenarios defined in ETSI Harmonised Standards or from the reference interfering signal and interference scenario defined in this annex.

The blocking levels are calculated using MRR and based on the reference interfering signal (RI) described in this section. Consequently, the RI signal can be used to check the conformity of receivers with the related blocking levels. Additionally, the equivalent blocking levels for a CW interfering signal are derived to allow the conformance tests to be done also with a CW signal.

A5.2 APPROACH USED TO DEFINE A REFERENCE INTERFERING SIGNAL BASED ON AN OFDM SIGNAL GENERATED USING AN RF SIGNAL GENERATOR

In nearly all relevant standards, a wanted signal like modulated adjacent channel interfering signal is used to define receiver resilience levels for intra-system compatibility.

On the other hand, for large frequency offsets beyond the operating band of a given system/service, in most cases an unmodulated continuous wave (CW) interfering signal is used to define receiver resilience levels for inter-system compatibility based on the assumption that beyond a large frequency offset the victim receiver FOS is constant, thus independent of the interfering signal bandwidth. This approach avoids the difficulty of identifying the most relevant interfering signal on a case-by-case basis, which is not an easy task due to the increasingly densely used radio spectrum.

As most standards use a CW test signal, a representative interfering signal is necessary to define receiver resilience levels for inter-system/service compatibility and to calculate receiver resilience levels using MRR.

A5.2.1 Proposed reference interfering signal

The proposed reference interfering signal has been derived based on an OFDM signal generated using an RF signal generator and has the following characteristics:

- Signal type: OFDM 3GPP continuous, fully-loaded, FDD-LTE DL. OFDM is the most common modulation type (e.g. IMT (LTE), DVB-T, DAB, DRM, Wireless LAN, WiMAX, etc.);
  A wide range of OFDM signal generators are available on the market, and such a signal can be generated using the practical approaches explained in ANNEX 6: for test purposes, as part of a larger test configuration;
- Channel Bandwidth: 5 MHz. This is a good compromise between potential narrowband and wideband OFDM interfering signals.

  Note: other bandwidths maybe be needed for frequencies outside of the range considered in ECC Recommendation (24)01 (30 MHz and 5725 MHz)
- Transmission bandwidth configuration: highest transmission bandwidth allowed for uplink or downlink in a given channel bandwidth, expressed in MHz. For downlink: BW_Config = 15 kHz + NRB × 180 kHz in the downlink (See also Figure 3.1-1 in ETSI EN 301 908-14 V15.1.1 (2021-09)).
- OOBD/SD boundary: at ±12.5 MHz offset from the centre frequency of the interfering signal (± 250% of the interfering signal bandwidth).

  OOBE level joins spurious emission (SE) level at the OOBD/SD boundary;
- ILR/out-Of-band (OOB) domain:
  - 1st adjacent channel ILR (ACLR) = 48 dB/(5 MHz)

    Value approximated for IMT BS (44 dB) in ETSI EN 301 908-14 [8] + 3 dB margin (see Figure 8);

  - 2nd adjacent channel ILR = 67 dB/(5 MHz) and 68 dB/(5 MHz), for frequency ranges of 9 kHz ≤ 1 GHz and 1 GHz < f < ≤ 5.725 GHz respectively.

    Value obtained by the method described in section A5.2.2;
  - 1st adjacent channel ILR to OOBD/SD boundary = 70 dB/(5 MHz) and 74 dB/(5 MHz), for frequency ranges of 9 kHz ≤ 1 GHz and 1 GHz < f ≤ 5.725 GHz respectively.

    Value obtained by the method described in section A5.2.2 based on the spurious emissions (SE) limits defined in ERC Recommendation 74-01, (Annex 2, Table 6, line 2.1.1) [2] – 3 dB margin.
- <u>Spurious domain:</u>
  - -39 dBm/(100 kHz) for 9 kHz ≤ 1 GHz and
  - -33 dBm/MHz for 1 GHz < f ≤ 5.725 GHz

    as defined in Rec. ERC Recommendation 74-01 – 3 dB margin, (ERC Recommendation 74-01, Annex 2, Table 6, 2.1.1).

These limits are the most common limits to radio transmitters including IMT base stations. Note that the maximum SE integration bandwidth is limited to 2 times of the bandwidth of the reference interfering signal ($2*BW_{RI}$) to prevent from calculating non-realistic spurious emission levels for a 5 MHz interference signal over a wide frequency range.

A5.2.2 Step-by-step derivation of the spectrum mask of the reference interfering signal

The spectrum mask of the reference interfering signal described in A5.2.1 has been derived according to the following steps:

Step 1: A 5 MHz OFDM signal having a 1st adjacent channel ILR (ACLR) of 47 dB/5 MHz was generated using an OFDM signal generator. The value of 47 dB/5 MHz is the IMT BS 1st adjacent ILR (ACLR) defined in ETSI EN 301 908-14 [8] increased by 3 dB. The 2nd adjacent channel ILR of this signal is 64 dB/(5 MHz) as shown in Figure 7.

The ILR value measured for a given receiver channel bandwidth should be equal to the ILR value calculated from the reference interfering signal spectrum mask ±1 dB, defined in Table 11.

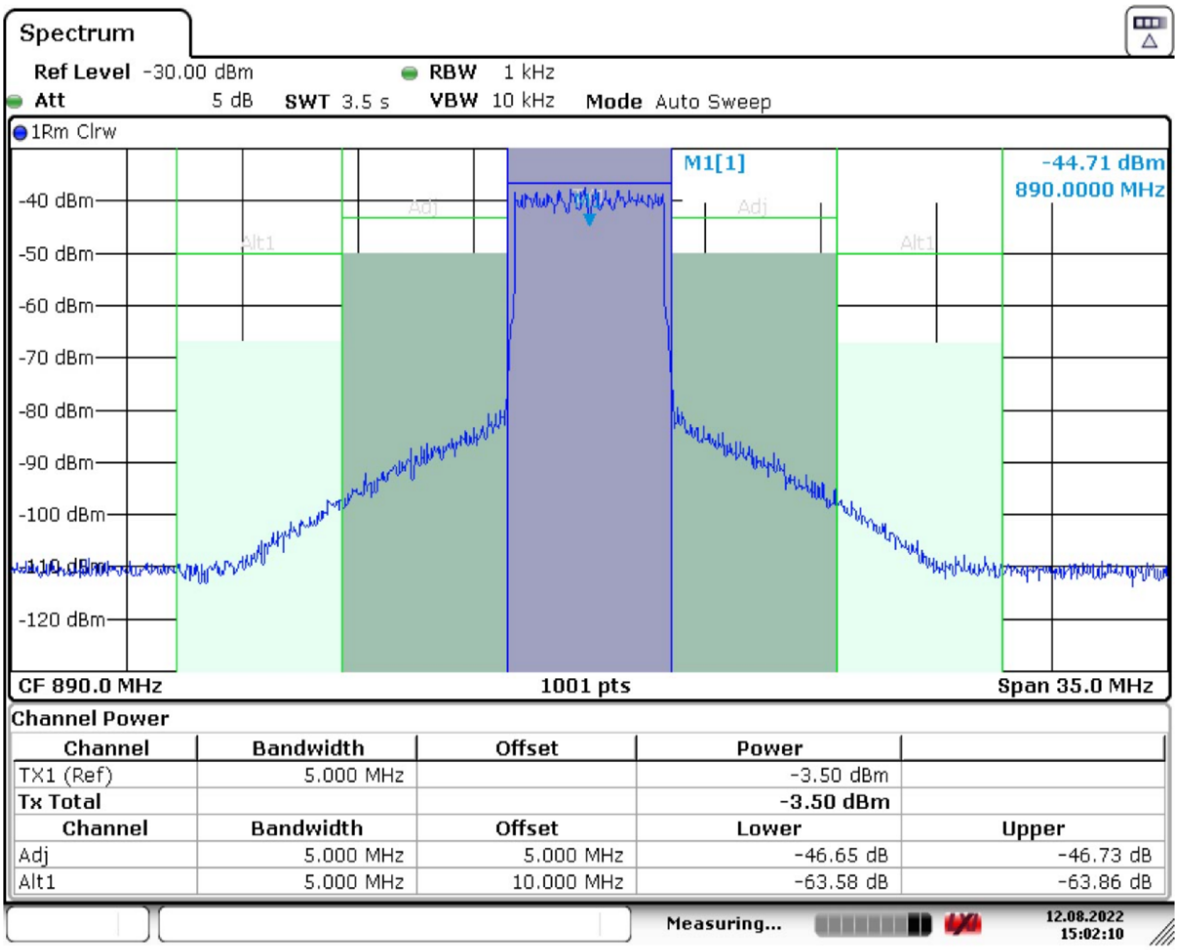

Figure 7: 5 MHz OFDM signal
(QPSK/4-QAM modulation, PAPR = 11 dB, Spectrum analyser noise floor = -117.5 dBm/kHz, generator output power = 18 dBm RMS, external attenuation 21.5 dB)

A numerical integration method based on the trapezoidal rule has been used to calculate the ILR of the spectrum mask of the reference interfering signal as follows:

$$ILR(dB) = 10\log_{10}\left(\frac{I_{in-ch}}{I^{r}_{oo-ch}}\right) = 10\log_{10}\left(\frac{I_{in-ch}}{\int_{f_{ov}-B_v/2}^{f_{ov}+B_v/2} sd_I(f)df}\right) \tag{64}$$

Where:

- I_{in-ch}: frequency offset interfering signal in-channel power at the receiver input;
- B_v: victim receiver bandwidth;
- f_{ov}: centre frequency of the victim receiver channel;
- $sd_I(f)$: frequency offset interfering signal spectral density in the linear domain.

$$\int_{f_{ov}-B_v/2}^{f_{ov}+B_v/2} sd_I(f)df = \Delta f\left(\sum_{k=1}^{N-1} sd_I(f_k) + \frac{sd_I(f_N) + sd_I(f_0)}{2}\right) \tag{65}$$

Where:

$$\Delta f = \frac{B_v/2 - (-B_v/2)}{N} = \frac{B_v}{N} \tag{66}$$

The validity of the integration method was checked as presented in Table 8.

Table 8: Comparison of measurements and numerical integration results (Measurement BW=Integration BW=5 MHz)

Offset type	Integration BW (MHz)	ILR obtained by measurement (dB)	ILR obtained by numerical integration (dB)	Difference (dB)
Co-channel	5 MHz	0	0	0
1st adjacent channel	5 MHz	46.73	47.27	0.54
2nd adjacent channel	5 MHz	63.86	63.80	-0.06
>2nd adjacent channel	5 MHz	70	70.28	0.28

Step 2 (for 30 MHz < f < 1 GHz): A reference interfering (RI) signal mask has been derived from the measured 5 MHz OFDM signal spectrum of which 4.5 MHz is the transmission bandwidth configuration (60 W/(4.5 MHz), 47.78 dBm/(4.5 MHz), Normalisation factor (F_{Norm}) = Power in transmission bandwidth (dBm) – 10*log (transmission bandwidth (Hz)/1000) = 11.25 dBm/kHz, normalising the whole mask to the in-band power spectral density of 0 dBm/kHz. For the spurious emissions, -39 dBm/(100 kHz) = -59 dBm/kHz and then normalised by applying the 11.25 dB to give -70.25 dBm/kHz) using linear interpolation to join the OOBE level to SE level of -36 dBm/(100 kHz) decreased by 3 dB at the OOBD/SD boundary. The derived spectrum mask has a set of break points as shown in Figure 8. The ILR of the RI signal has been calculated by numerical integration of its power (W) in a bandwidth of 5 MHz, as described above, at three different frequency offsets as shown in Table 9.

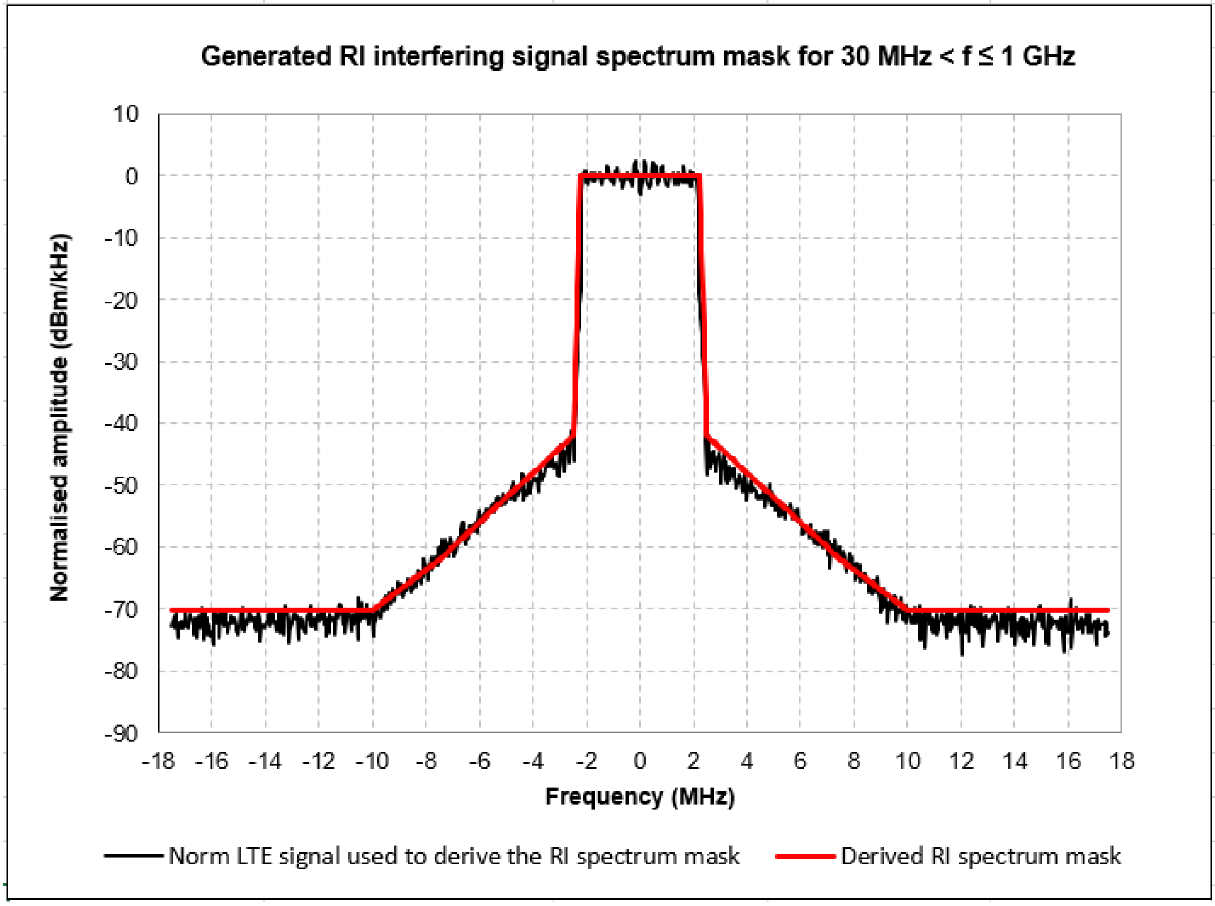

Figure 8: Approximation of the 5 MHz OFDM signal to a reference interfering signal for 30 MHz < f ≤ 1 GHz

Table 9: ILR of the reference interfering signal for 30 MHz < f ≤ 1 GHz (60 W/5 MHz conducted Tx power normalised to 0 dBm/kHz; Fnorm=11.25 dB)

Channel BW (MHz)	Power per channel BW (W)	Power per channel BW (dBm)	Offset type	ILR of the RI signal (dB)	ILR of the measured signal (dB)
5	4.54	36.57	Co-channel	0	0
5	6.61E-05	-11.80	1st adjacent channel	48	47
5	9.25E-07	-30.34	2nd adjacent channel	67	66
5	4.69E-07	-33.29 (Note 1)	1st adjacent channel to OOBD/SD boundary in the SD	70	71
Note 1: -33.29 dBm/5MHz+11.25 dB =-39dBm/100kHz, which is the spurious limit of -36dBm/(100 kHz) minus a 3 dB margin					

Step 3 (for 1 GHz < f ≤ 5.725 GHz): A reference interfering signal mask has been derived from the measured 5 MHz OFDM signal spectrum used in Step 2 (see that step for how the normalisation was calculated) using linear interpolation to join the OOBE level to SE level of -30 dBm/MHz decreased by 3 dB at the OOBD/SD boundary. The ILR of the RI signal has been calculated by numerical integration of its power (W) in its bandwidth of 5 MHz at three different frequency offsets as shown in Table 10.

Table 10: ILR of the reference interfering signal for 1 GHz < f < ≤ 5.725 GHz (60 W/(5 MHz) conducted Tx power normalised to 0 dBm/kHz; Fnorm=11.25 dB):

Channel BW (MHz)	Power per channel BW (W)	Power per channel BW (dBm)	Offset type	ILR of the RI signal (dB)	ILR of the measured signal (dB)
5	4.54	36.57	Co-channel	0	0
5	6.61E-05	-11.80	1st adjacent channel	48	48
5	7.22E-07	-31.41	2nd adjacent channel	68	66
5	1.87E-07	-37.29 (Note 1)	1st adjacent channel to OOBD/SD boundary in the SD	74	71
Note 1: -37.29 dBm/(5 MHz)+11.25 dB =-33dBm/MHz, which is the spurious limit of -30dBm/MHz minus a 3 dB margin					

A5.3 APPROACH USED TO DEFINE A SINGLE INTERFERENCE SCENARIO

A5.3.1 Rationale

The rationale behind the idea to have a single interference scenario is:

- To treat all the services, systems and applications with a common methodology;
- To define the recommended receiver resilience levels only for the most relevant frequency offsets, thus to keep the size of the ECC Recommendation on receivers reasonable;

A5.3.2 Detailed description of the proposed interference scenario

According to the rational presented in Section A5.3.1, it is proposed to define the receiver resilience level for different frequency offsets between the interfering transmitter and the victim receiver:

- In the OOB domain, the first adjacent channel to the interfering signal, for the purpose of receiver adjacent channel PR calculation/measurement;

- In the RSBD of the interfering signal, provided that the interfering signal is beyond the second adjacent channel of the victim receiver, for the purpose of receiver blocking level calculation/measurement.

The frequency offsets between the reference interfering signal (RI) centre frequency and the victim receiver (Rx) centre frequency are calculated as follows:

- In the OOB domain:

  Frequency offset (MHz) = $BW_{RI}/2$ (MHz) +$BW_{Rx}/2$ (MHz)
- In the RSBD:

  If $BW_{RI} > BW_{Rx}$

  Frequency offset (MHz) ≥ $2.5BW_{RI}$ (MHz) + $BW_{Rx}/2$ (MHz)

  Else

  Frequency offset (MHz) ≥ $2.5BW_{Rx}$ (MHz) + $BW_{RI}/2$ (MHz)

For channelised systems, the above calculations define that the victim receiver channel is in the spurious domain of the interfering transmitter and reciprocally the interfering transmitter channel is in the spurious domain of the wanted transmitter of the victim receiver, whatever their bandwidth (BW) is.

For single carrier and multi-carrier receivers the bottom channel at the lower edge and the top channel at the upper edge of the receiver operating band is to be defined and tested. For receivers supporting multiple channel bandwidths, the smallest and the largest channel bandwidths should be defined and tested.

For non-channelised systems the occupied bandwidth is considered as a single channel (see Figure 9).

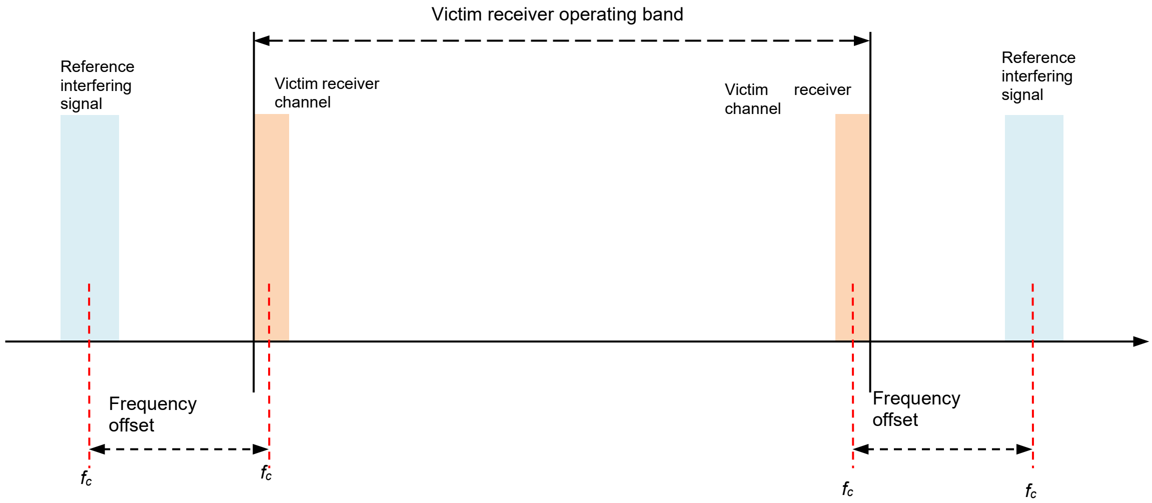

Figure 9: Channel of the receiver to be defined and tested

This interference scenario including reference interferer (RI) and the two offset frequencies is depicted in Figure 10 for the example of a victim receiver having a bandwidth of 2 MHz ($BW_{RI} > BW_{Rx}$).

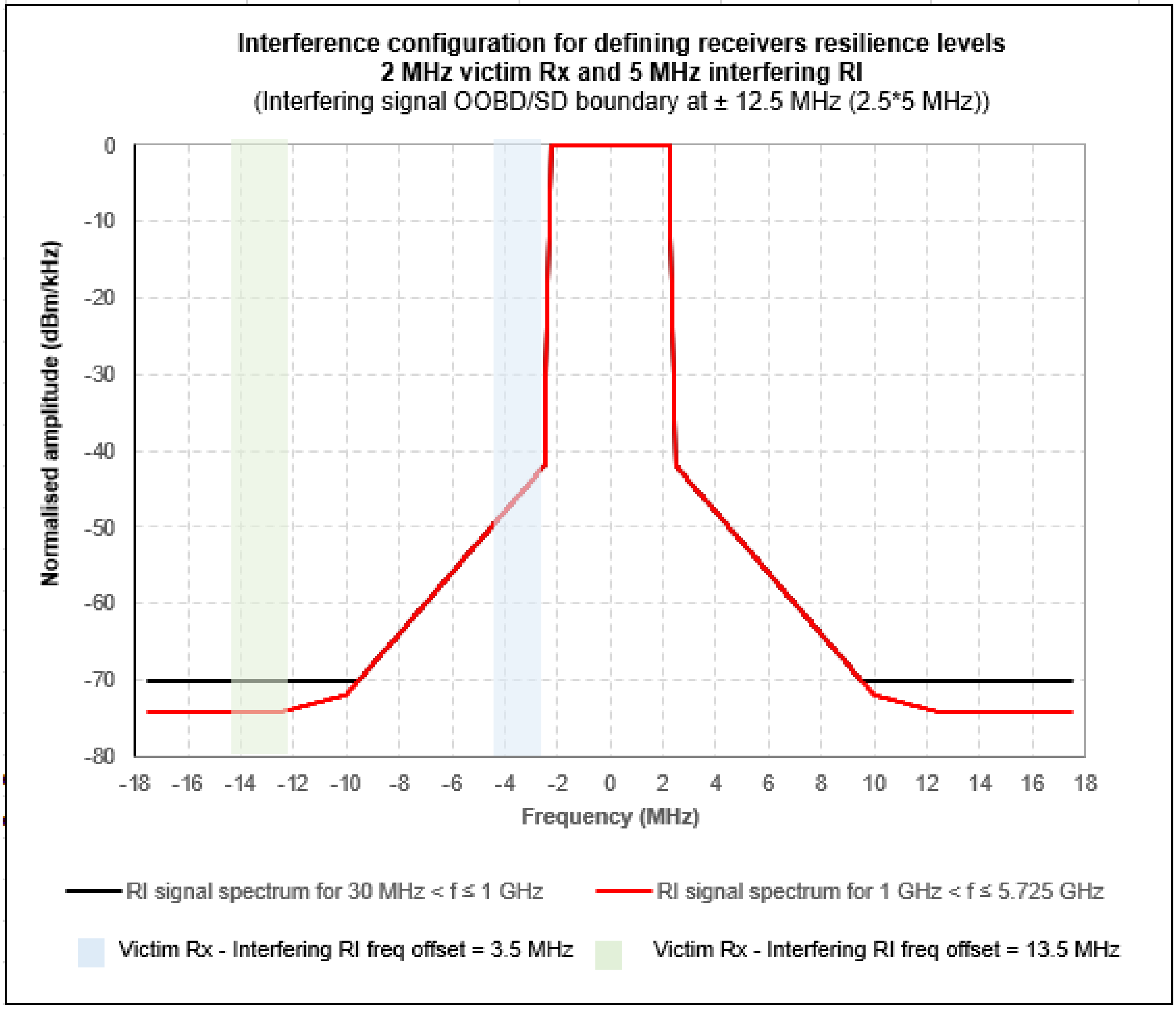

Figure 10: Proposed single interference scenario to be used for all systems/services when calculating receivers resilience levels using MRR

The resulting offset frequencies from the centre of the RI to the centre frequency of the victim receiver are for the 2 MHz receiver:

- 3.5 MHz: 2.5 MHz (BW_{RI}/2)) + 1 MHz (BW_{Rx}/2);
- 13.5 MHz: 12.5 MHz (2.5x BW_{RI})) + 1 MHz (BW_{Rx}/2).

The ILR values for the two offset frequencies are then to be derived by numerical integration from the spectrum masks defined in Table 11: depending on the victim receiver bandwidth according to Equation (4). Note that the maximum SE integration bandwidth (BW) of the RI signal is assumed to be 10 MHz ($2BW_{RI}$). This can be implemented by using the equation: $10\log_{10}(BW_{Rx}$ (MHz)/10 (MHz))..

Table 11: Reference interfering signal spectrum mask

Offset (MHz)	RI signal spectrum mask for frequencies of 9 kHz ≤ f ≤ 1 GHz (dBm/kHz)	RI signal spectrum mask for frequencies of 1 GHz < f ≤ 5.725 GHz (dBm/kHz)
-17.5	-70.25	-74.25
-12.5	-70.25	-74.25
-10	-70.25	-72
-7.5	-62	-62
-2.5	-42	-42
-2.25	0	0
+2.25	0	0
+2.5	-42	-42
+7.5	-62	-62
+10	-70.25	-72
+12.5	-70.25	-74.25
+17.5	-70.25	-74.25

For measurement purpose, the reference interfering signal can be generated using an OFDM 3GPP FDD LTE RF signal generator which is capable to produce the necessary distortion to the pure OFDM signal so that the desired spectrum mask is achieved. The recommended modulation schemes are QPSK/4-QAM (data rate ≅ 8 Mbps, sub-carrier spacing ≅ 15 kHz), with preferably a PAPR =11 ± 3 dB at the RF generator output. Higher constellations and data rates can also be used provided that the generated signal meets the ILR values calculated from the interfering signal spectrum mask defined in this table as well as the PAPR value. In all the cases, the ILR value measured for a given receiver channel bandwidth should be equal (± 1 dB) to the ILR value calculated from the reference interfering signal spectrum mask defined in the table.

Alternatively, the RI signal can also be generated using a synthesised signal provided in a file that can be read by most of the modern RF signal generators. The practical generation of the RI signal is described in detail in ANNEX 6:.

Some calculation examples of RI signal ILR values are provided in the following section.

A5.3.3 Derivation of the reference interfering signal ILR to be used in the calculation of receivers resilience levels using MRR

The ILR values of the reference interfering signal to be used to calculate the receiver resilience levels for some systems/applications have been derived by using numerical integration. These systems are DAB, DTT, IMT, Fixed service, GSM, 2.4 GHz WBDT systems, 5 GHz RLAN and 30 MHz-1GHz SRD. The results obtained are presented for information in Section A5.3.5.

A5.3.4 Choice of the FOS value

Based on the definition of the reference interfering signal, it is proposed to choose the value of FOS to be used in the calculation, when using MRR to calculate the receiver resilience levels, as follows:

- FOS = ILR + 10 dB, when the victim receiver channel is adjacent to the interfering transmitter channel (see Figure 10);

Reason: This configuration is critical since the interfering transmitter channel is very close to the victim receiver channel resulting in a high OOB emission level in the receiver input, which cannot be reduced by filtering. Consequently, a good adjacent channel filtering would be necessary to minimise the interfering signal in-block power at the receiver input;

- FOS = ILR dB, when the victim receiver channel is adjacent to the OOBD/SD boundary of the interfering signal in its SD (see Figure 10). If a CW interfering signal is used, instead of the RI signal, to calculate the blocking level of a victim receiver, the receiver FOS value should be equal to the FOS value specified for the blocking level calculation with the RI signal, since the ILR value of the CW signal is assumed to be very high or infinite (≥130 dB).

  Reason: This configuration is less critical compared to the first configuration, since the interfering transmitter channel is far away from to the victim receiver channel resulting in a lower OOBE emission level in the receiver input. Consequently, it is sensible to put less constraint on the receiver FOS.

A5.3.5 Derived reference interfering signal ILR to be used in the calculation of receiver resilience levels

Table 12: ILR values for systems/applications as a function of the victim channel bandwidth and frequency offset: 30 MHz < f ≤ 1 GHz

System	Victim channel BW (MHz)	Power (W)	Power (dBm)	Frequency offset	ILR (dB)
RI	4.97	4.54	36.57	Co-channel	0
DAB (1.54)	1.54	5.05E-05	-12.97	1st adjacent channel	50
DAB (1.54)	1.54	1.45E-07	-38.37	1st adjacent channel to OOBD/SD boundary in the SD	75
DTT (8 MHz)	7.60	5.70E-05	-12.44	1st adjacent channel	49
DTT (8 MHz)	7.60	7.17E-07	-31.44	1st adjacent channel to OOBD/SD boundary in the SD	68
IMT (1.4 MHz)	1.09	4.21E-05	-13.76	1st adjacent channel	50
IMT (1.4 MHz)	1.09	1.02E-07	-39.90	1st adjacent channel to OOBD/SD boundary in the SD	76
IMT (3 MHz)	2.70	6.12E-05	-12.13	1st adjacent channel	49
IMT (3 MHz)	2.70	2.54E-07	-35.94	1st adjacent channel to OOBD/SD boundary in the SD	73
IMT (5 MHz)	4.52	2.91E-05	-15.36	1st adjacent channel	52
IMT (5 MHz)	4.52	4.26E-07	-33.70	1st adjacent channel to OOBD/SD boundary in the SD	70
IMT (10 MHz)	9.00	4.43E-05	-13.54	1st adjacent channel	50
IMT (10 MHz)	9.00	8.49E-07	-30.71	1st adjacent channel to OOBD/SD boundary in the SD	67
IMT (15 MHz)	13.51	3.26E-05	-14.86	1st adjacent channel	51
IMT (15 MHz)	10.01	9.45E-07	-30.25	1st adjacent channel to OOBD/SD boundary in the SD	67
IMT (20 MHz)	18.03	2.92E-05	-15.34	1st adjacent channel	52

System	Victim channel BW (MHz)	Power (W)	Power (dBm)	Frequency offset	ILR (dB)
IMT (20 MHz)	10.01	9.45E-07	-30.25	1st adjacent channel to OOBD/SD boundary in the SD	67
GSM (0.2 MHz)	0.18	9.88E-06	-20.05	1st adjacent channel	57
GSM (0.2 MHz)	0.18	1.65E-08	-47.82	1st adjacent channel to OOBD/SD boundary in the SD	84
SRD (0.1 MHz)	0.11	6.12E-06	-22.13	1st adjacent channel	59
SRD (0.1 MHz)	0.11	9.91E-09	-50.04	1st adjacent channel to OOBD/SD boundary in the SD	87
SRD (0.5 MHz)	0.49	2.41E-05	-16.17	1st adjacent channel	53
SRD (0.5 MHz)	0.49	4.63E-08	-43.35	1st adjacent channel to OOBD/SD boundary in the SD	80
SRD (1.7 MHz)	1.68	5.25E-05	-12.80	1st adjacent channel	49
SRD (1.7 MHz)	1.68	1.59E-07	-38.00	1st adjacent channel to OOBD/SD boundary in the SD	75
SRD (0.05 MHz)	0.04	2.63E-06	-26.77	1st adjacent channel	62
SRD (0.05 MHz)	0.04	4.13E-09	-54.81	1st adjacent channel to OOBD/SD boundary in the SD	90
SRD (0.025 MHz)	0.04	1.32E-06	-28.81	1st adjacent channel	65
SRD (0.025 MHz)	0.04	2.06E-09	-56.85	1st adjacent channel to OOBD/SD boundary in the SD	93

Table 13: ILR values for systems/applications as a function of the victim channel bandwidth and frequency offset: 1 GHz < f ≤ 5.725

System	Victim channel BW (MHz)	Power (W)	Power (dBm)	Frequency offset	ILR (dB)
RI	4.97	4.54	36.57	Co-channel	0
DAB (1.54 MHz)	1.54	5.05E-05	-12.97	1st adjacent channel	50
DAB (1.54 MHz)	1.54	5.79E-08	-42.37	1stadjacent channel to OOBD/SD boundary in the SD	79
DTT (8 MHz)	7.60	5.69E-05	-12.,45	1st adjacent channel	49
DTT (8 MHz)	7.60	2.85E-07	-35.44	1st adjacent channel to OOBD/SD boundary in the SD	72
IMT (1.4 MHz)	1.09	4.21E-05	-13.76	1st adjacent channel	50
IMT (1.4 MHz)	1.09	4.08E-08	-43.9	1stadjacent channel to OOBD/SD boundary in the SD	80
IMT (3 MHz)	2.70	6.12E-05	-12.13	1stadjacent channel	49

System	Victim channel BW (MHz)	Power (W)	Power (dBm)	Frequency offset	ILR (dB)
IMT (3 MHz)	2.70	1.01E-07	-39.94	1st adjacent channel to OOBD/SD boundary in the SD	77
IMT (5 MHz)	4.52	2.91E-05	-15.36	1st adjacent channel	52
IMT (5 MHz)	4.52	1.70E-07	-37.7	1st adjacent channel to OOBD/SD boundary in the SD	74
IMT (10 MHz)	9.00	4.41E-05	-13.56	1st adjacent channel	50
IMT (10 MHz)	9.00	3.38E-07	-34.71	1st adjacent channel to OOBD/SD boundary in the SD	71
IMT (15 MHz)	13.51	3.22E-05	-14.92	1st adjacent channel	51
IMT (15 MHz)	10.01	3.76E-07	-34.25	1st adjacent channel to OOBD/SD boundary in the SD	71
IMT (20 MHz)	18.03	2.85E-05	-15.45	1st adjacent channel	52
IMT (20 MHz)	10.01	3.76E-07	-34.25	1st adjacent channel to OOBD/SD boundary in the SD	71
FS (1.4 MHz)	1.75	5.34E-05	-12.73	1st adjacent channel	49
FS (1.4 MHz)	1.75	6.58E-08	-41.82	1st adjacent channel to OOBD/SD boundary in the SD	78
FS (1.4 MHz)	7.00	6.67E-05	-11.76	1st adjacent channel	48
FS (1.4 MHz)	7.00	2.63E-07	-35.8	1st adjacent channel to OOBD/SD boundary in the SD	72
FS (1.4 MHz)	14.00	6.70E-05	-11.74	1st adjacent channel	48
FS (1.4 MHz)	10.01	3.76E-07	-34.25	1st adjacent channel to OOBD/SD boundary in the SD	71
SRD (0.1 MHz)	0.11	6.12E-06	-22.13	1st adjacent channel	59
SRD (0.1 MHz)	0.11	3.95E-09	-54.04	1st adjacent channel to OOBD/SD boundary in the SD	91
WLAN (20 MHz)	19.99	4.45E-05	-13.52	1st adjacent channel	50
WLAN (20 MHz)	10.01	3.78E-07	-34.23	1st adjacent channel to OOBD/SD boundary in the SD	71
DCS (0.2 MHz)	0.18	9,88E-06	-20.05	1st adjacent channel	57
DCS (0.2 MHz)	0.18	6,58E-09	-51.82	1st adjacent channel to OOBD/SD boundary in the SD	88

ANNEX 6: PRACTICAL GENERATION OF THE REFERENCE INTERFERING TEST SIGNAL

The reference interfering signal described in section A5.2 can be generated using an RF signal generator:

- equipped with an option to generate LTE signals;
- be able to read signal files with extension ".WV" and generate the synthetic reference interfering signal "Golden waveform" developed by the Joint Research Centre (JRC) and be found here.

A6.1 REFERENCE INTERFERENCE SIGNAL GENERATED USING A DIGITAL SIGNAL PROCESSING TOOL TO OBTAIN TWO "GOLDEN WAVEFORM" SIGNAL FILES

In this example the RI signal has been generated using an RF signal generator, which was able to read signal files with extension "WV" and the "Golden waveforms" developed by the Joint Research Centre (JRC) and provided in the files below. This method is expected to be easier to implement than that in A6.2.

The two "Golden waveforms (GW)" have been generated using a digital signal processing tool, one for frequencies below 1 GHz (GW1) and one for frequencies above 1 GHz (GW2). In both waveforms, a central 5 MHz part was allocated within a 20 MHz LTE DL structure. Each subcarrier, that was defined in this central part, was QPSK modulated with random symbols. A transition window of 10 µs was used to minimize the discontinuities between two consecutive symbols. Then, a power amplifier distortion was added to the generated signal to obtain the desired adjacent channel ILR values. In the case of GW1, an additive white Gaussian noise was added to increase the ILR value in the spurious domain. The comparison between the generated RI signal spectrum and the RI mask defined in this Recommendation is presented in Figure 11 for GW1 (30 MHz < f ≤ 1 GHz) and Figure 12 for GW2 (1 GHz < f ≤ 5725 MHz):

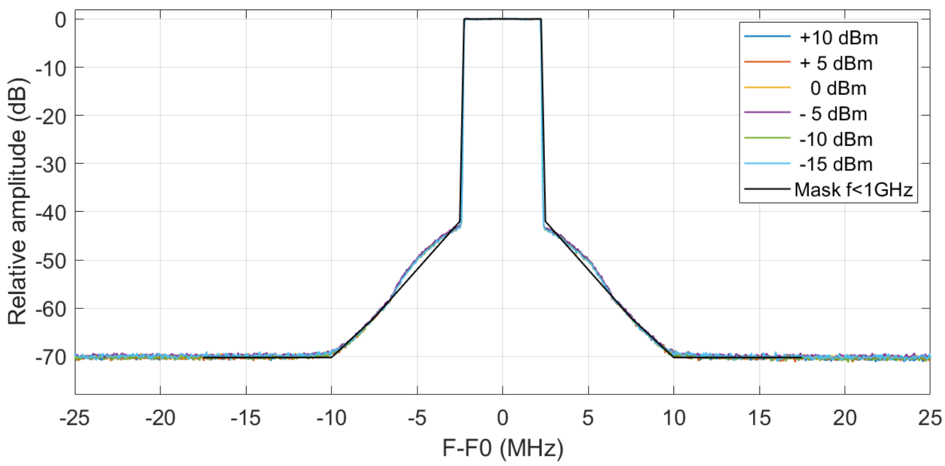

Figure 11: Comparison between the generated RI signal spectrum (GW1) and the RI signal spectrum mask defined in this recommendation (30 MHz < f ≤ 1 GHz)

Table 14: Measured ILR values of the generated RI signal (30 MHz < f ≤ 1 GHz)

Measurement BW (MHz)	Offset type	Target Values (note 1) (dB)	Measurement results (dB)	Difference (dB)
5	Co-ch	0	0	0
5	1st adj-ch	48	48	0
5	2nd adj-ch	67	67	0
5	3rd adj-ch <	70	70	0
Note 1: Values calculated from the RI signal spectrum mask defined in Table 9				

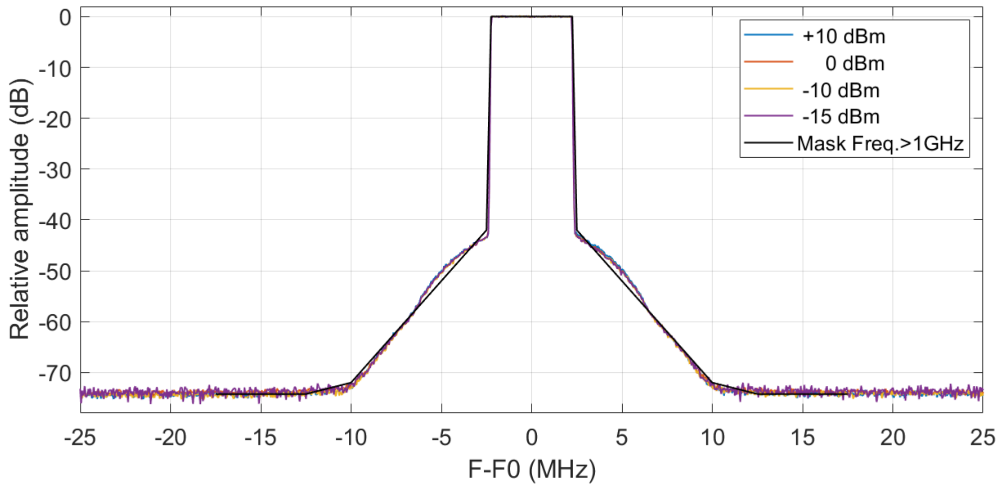

Figure 12: Comparison between the generated RI signal spectrum (GW2) and the RI signal spectrum mask defined in this recommendation (1 GHz < f ≤ 5725 MHz)

Table 15: Measured ILR values of the generated RI signal (1 GHz < f ≤ 5725 MHz)

Measurement BW (MHz)	Offset type	Target Values (note 1) (dB)	Measurement results (dB)	Difference (dB)
5	Co-ch	0	0	0
5	1st adj-ch	48	48	0
5	2nd adj-ch	68	68	0
5	3rd adj-ch <	74	74	0
Note 1: Values calculated from the RI signal spectrum mask defined in Table 10				

A6.2 REFERENCE INTERFERENCE SIGNAL GENERATED USING AN RF SIGNAL GENERATOR EQUIPPED WITH AN OPTION TO GENERATE LTE SIGNALS

In this example the RI signal has been generated using an RF signal generator equipped with an option to generate LTE signals. The LTE parameters used to generate the RI signal are the followings:

- System: 5 MHz Eutra LTE;
- Modulation QPSK, 16-QAM and 64-QAM;
- Link: FDD/DL;
- Number of used frames/allocations: 10/4;
- Filtering / Clipping: Balanced EVM and ACP / off;
- Signal generator output power: 17 dBm (PAPR=11 dB).

The comparison between the generated RI signal spectrum and the RI mask defined in this recommendation is presented in Figure 13:

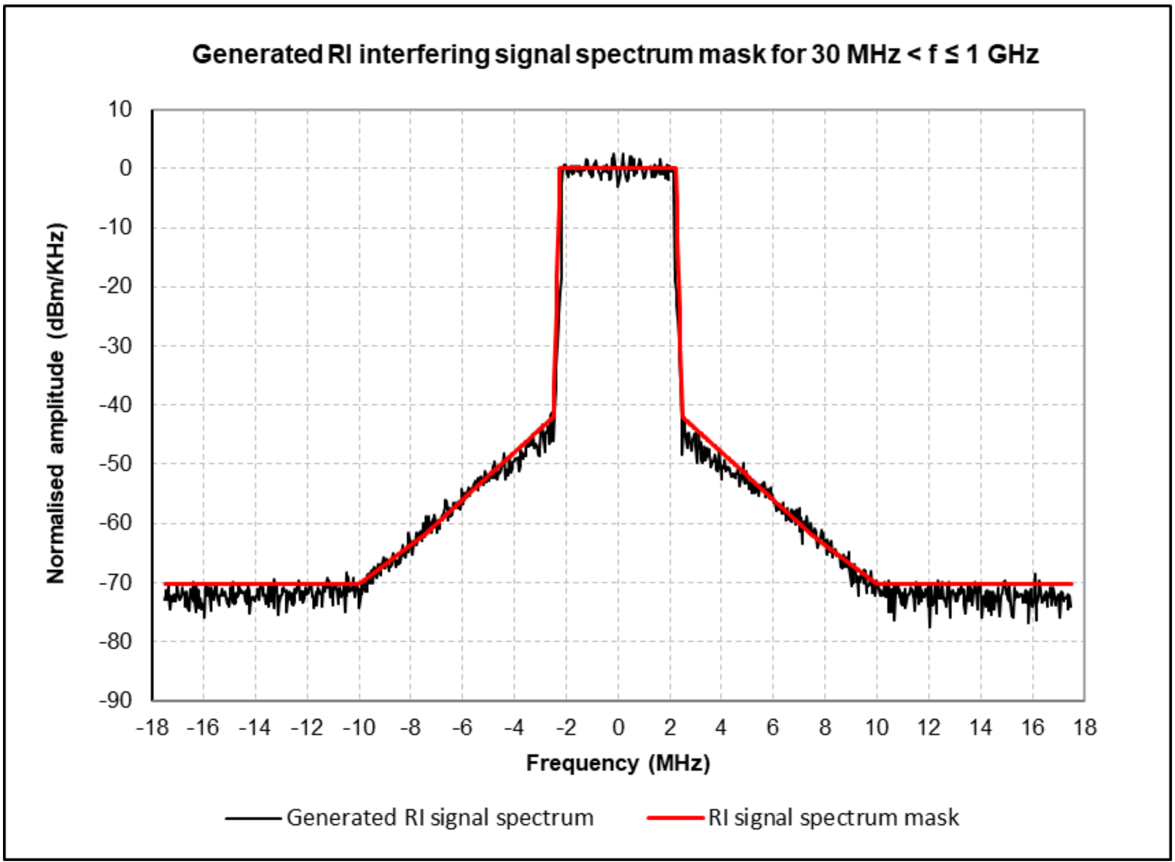

Figure 13: Comparison between the generated RI signal spectrum and the RI signal spectrum mask defined in this Recommendation (30 MHz < f ≤ 1 GHz)

Table 16: Measured ILR values of the generated RI signal (30 MHz < f ≤ 1 GHz)

Generated 5 MHz RI signal ILR values (30 MHz < f ≤ 1 GHz)				
Measurement BW (MHz)	**Offset type**	**Target Values (note 1) (dBm)**	**Measurement results (dB)**	**Difference (dB)**
5	Co-ch	0	0	0
5	1st adj-ch	48	48	0
5	2nd adj-ch	67	66	1
5	3rd adj-ch <	70	71	1
Note 1: Values calculated from the RI signal spectrum mask defined in Table 9				

ANNEX 7: POINT-TO-POINT FS EMISSION MASKS FOR ILR/ACLR EVALUATION

A7.1 INTRODUCTION

The proposed RI mask for bands < 1 GHz and for bands from 1 to 5.725 GHz was derived as OFDM 5 MHz bandwidth (widely used in broadcasting and mobile radios, which are the most used in those bands).

In higher bands, one of the most used "licensed/authorised" service is the fixed service, of which the most popular worldwide application is point-to-point (PtP) used in many fixed core and mobile backhauling applications.

The allocated bands with major usage (or usable span), with a few exceptions, range from 5.925 GHz to 43.5 GHz, 48.5 GHz to 57 GHz and 71 GHz to 86 GHz. Higher bands up to 174.7 GHz are already considered usable in ECC Recommendation (18)01 [18] and ECC Recommendation (18)02 [19].

The major differences from the OFDM case proposed for ≤ 5.725 GHz are:

1 The modulation used is X-QAM (with X from 4 to 2048 and higher) with relatively tight roll-off; this implies that the spectrum emission is not rectangular.
However, while the regulations for broadcast and mobile allocated bands do not provide guard-bands (hence the need for quasi rectangular spectrum), the channel arrangements for FS PtP allocated bands always provide internal guard-bands.

2 The PtP fixed service operates using high directivity antennas, in line of sight between fixed transmitter and receiver station locations, therefore along the path no interfered radio system (infrastructure or persons) may be present;

3 Consequently, while the broadcasting and mobile BS EIRP is generated by low gain antennas (then the actual emitted power at antenna port is similarly high); the PtP EIRP is the result of high directivity antennas and the actual transmit power, at antenna port, is far less (at least 40 dB) than the EIRP.

In addition, the RF power devices capability, in terms of maximum power, rapidly decreases while the operating band increases.

The approximately needed (at antenna port) 60 Watts considered in the RI for broadcasting and mobile BSs, are reflected in needed 1 Watt (decreasing with frequency to about 0.25 W) generated by FS transmitters at antenna port.

Therefore, the comparison between broadcasting/mobile BS transmit power and FS PtP power (at the antenna port) is here given only for discussing the suitability of using other relative RI emission masks.

4 The available system bandwidth increases with the operating band; which fits with the constantly increasing demand for backhauling traffic; while there might be relatively narrow bandwidth legacy links (e.g. 7 or 14 MHz channels), the vast majority of new links have channel bandwidth of:
   i) 28, 40 or 56 MHz (in bands from 5.925 GHz to 15.35 GHz);
   ii) 56, 112 and possibly 224 MHz (in bands from 17 GHz to 43.5 GHz);
   iii) 500, 1000 and possibly up to about 2 000 MHz (in bands from 71 GHz to 115 GHz).

5 The transmitted relative spectral attenuation decreases accordingly with the actual transmit power; in addition, the filtering capability is relative to the pass-band; therefore, at a fixed distance from the channel edge, the capability decreases for the wider bandwidth systems.

ANNEX 8: BACKGROUND INFORMATION ON RAS RECEIVERS

Receiving signals above the linear regime of a radio astronomy telescope frontend can result in a variety of effects, such as saturation of the LNAs or other devices (such as ADCs and optical fibre links), generation of intermodulation products, or even worse, reaching the amplifiers' break point. While the use of filters may look as a potential solution to these issues, it comes associated with many challenges, which are reviewed in this Annex.

This Annex aims to provide more detailed technical information on RAS receivers. While these receivers are not within the scope of the ECC Recommendation (24)01 on receiver resilience levels, it is important to capture their unique characteristics and requirements to highlight the need for a different approach for non-market scientific equipment, including RAS receivers. It should be also noted that there are some fundamental differences between modern digital-based communication systems and nature-observing ones (e.g. lack of BER characteristic).

Modern radio astronomy is heavily dependent on both microwave front-end receiver and digital signal processing technology. To reach the required levels of sensitivity, a combination of huge collecting areas (radio telescopes), extremely low-noise amplifiers, cryogenically cooled passive microwave components, and digital acquisition systems is required. Besides state-of-the-art equipment, long integration times (often hours) and wide frequency bands are used by radio astronomers to reduce the thermal noise and to detect the faintest celestial signals. As the scientific instrumentation requires major investments, usually from governments, it requires appropriate protection from active radio transmitters. Astronomers do their part towards this protection by building new telescopes in very remote sites, often taking advantage of natural terrain shielding (see Recommendation ITU-R RA.611 [14]). Several administrations have also established radio-quiet zones (RQZ) or coordination zones around the RAS stations in their countries (see Report ITU-R RA.2259 [21]).

In order to justify the high cost of operation, resulting from the heavy investments, astronomy stations need to stay internationally competitive and they need to maximise the scientific output of their instruments. Therefore, observing time is granted by independent refereeing committees, which will assign time slots only to those scientists who submitted the highest-ranking observing proposals. Furthermore, most radio observatories are heavily overbooked, i.e., the amount of requested time is much higher than the available time, making the loss of observing time a critical factor for scientific development through radio astronomy observations.

One way to improve the overall observing efficiency and sensitivity, is the installation of broad-band receivers, which can measure astronomical signals in several allocated RAS bands at the same time. The merit of observing large bandwidths is not restricted to gaining extra sensitivity. Many important astrophysical sources can only be properly investigated if a significant fraction of their spectra is acquired. The relative strength of the electromagnetic signal as a function of wavelength or frequency allows us to determine the radiation mechanisms within the celestial sources. Likewise, spectral line ratios can be used to measure physical properties of the interstellar gas, such as temperature or pressure. Other motivations to perform multi-broadband observations are to measure the dispersive effect of the interstellar medium or calibrating high frequency observations by using low frequency targets. By measuring the delay of a pulsed (brief) signal at different frequencies, astronomers can estimate the density and distribution of the interstellar medium. Time-domain astronomy, which looks out for individual rare events such as fast-radio bursts (FRB), cannot even observe the different bands at different times, because such events are singular.

An example of a modern receiver is the Ultra-Broadband Receiver at the 100 m telescope at Effelsberg (Germany) offering an observing band between 1 and 5 GHz, simultaneously, with analogue-digital converters (ADC) installed at the front-end. A second example of applied wide-band technology is the VGOS system, which uses four or more frequency bands in the range from 2.5 GHz to 14 GHz at the same time for highest precision and to calibrate out atmospheric delays.

Unfortunately, the extreme sensitivity of RAS receivers is also a burden. Even if more effort is made to increase the dynamic range, it is technically impossible to build a receiver capable of observing the faintest signals from the universe (reaching the sub-μJy to mJy regime, with 1 Jy = 10-26 W/m2/Hz = –260 dB(W/m^2/Hz)) and at the same time be robust against strong anthropogenic signal transmissions, which can easily cause levels 100 dB above the noise floor at the RAS receiver input, if active services and the RAS are not well-coordinated. Usually, the RAS and active services do not share allocations in the same bands, especially in the case of airborne and spaceborne transmitters. In addition to the few shared bands, out-of-band and spurious emissions from active services in adjacent and nearby bands may generate harmful interference in protected radio astronomy bands.

However, even if the active transmitter frequencies are well separated in frequency from the RAS observing band and out-of-band and spurious emissions are well coordinated, serious issues can still emerge. Receiving signals above the linear regime of a radio astronomy telescope front-end can result in a variety of effects, such as saturation of the low noise amplifiers (LNAs) or other active devices (such as second stage amplifiers, ADCs and optical fibre links), generation of intermodulation products, or even worse, reaching the amplifiers' break point, leading to (see Report ITU-R RA.2188 [22]). It is also worth noting that today a fair fraction of LNAs used in radio astronomy receivers, especially below about 100 GHz, are often off-the-shelf components. These devices usually come with a relatively large input bandwidth already.

If this worst-case situation occurs, the replacement of the damaged device is required to repair the receiver. This incurs a cost, which is reflected in the purchase of new components, their installation and the added cost of having the receiver out of operation for the duration of this process. But even in less drastic scenarios, mitigation techniques may need to be implemented to suppress the strongest signals, which is still costly, and as the spectrum environment is very dynamic and is subject to changes, this solution implies a continuous effort for the observatories.

While the use of filters may appear as a potential solution to improve compatibility between strong transmitters and a RAS station, it is not always technically feasible and it can have significant impact on sensitivity. First, analogue filters have finite suppression factors, so their effectiveness depends on the technology used and the frequency separation between active services and RAS bands. Second, frontend (notch-) filters installed in front of the first LNA add significant noise to the system, as their signal gets amplified by tens of dB in the signal chain. If at least the analogue part of the receiver is kept in the linear domain then, in principle, digital filters can be used further downstream to still use the unaffected frequencies.

In conclusion, it is important to preserve unique operational conditions of scientific equipment, and especially RAS operations, to enable future scientific breakthroughs. RAS emission sources are natural sources of cosmic origin, which are of analogue nature and are not only out of control, but also out of humankind's reach. Therefore, radio astronomy receiving systems require a certain level of general radio quietness in the environment of an observatory in order to protect them.

A8.1 RAS RECEIVER SENSITIVITY

For telecommunication applications using a receiving system linked to an emitter, the artificial signal of interest is usually above the receiver system noise and so the signal-to-noise ratio (SNR) is greater than 0 dB. This is not the case for high sensitivity detection applications such as radio astronomy for which the natural signal of interest (for example a radio source as a star or a galaxy) can be 30 to 60 dB below the instantaneous thermal noise. One way to detect such a tiny stationary signal within uncorrelated noise is to decrease the noise fluctuation level using long integration times.

A8.1.1 Sensitivity and the radiometer equation

The goal of a radiometer is to convert the energy of the incident radio wave into an electrical signal, feasible to be analysed and processed in a computational system. In general, a radiometer is a high-sensitivity microwave receiver that detects and measures noise-like signals. The setup of a classical radiometer is composed of an antenna and its front-end amplifier (a mixer or a bolometer in millimetric frequency), a radio frequency conditioning system (filtering and a data acquisition system. A radio telescope can be viewed as a very high performance radiometer. In astronomy, the sources of radiated emissions can be celestial bodies, such as planets, stars (including the Sun) or meteors, but also the interstellar or intergalactic medium. One can distinguish the celestial sources by spectral properties, e.g. continuum vs. spectral-line emission, but also by temporal properties (constant vs. intermittent or even pulse-like). Often the received signal has similar properties to the noise generated by the receiver itself or to terrestrial natural background signals coupled to the receiver, such as ground or atmospheric radiation. In rare cases receivers also detect diffuse anthropogenic signals that can be hard to distinguish from celestial signals of interest.

Assuming an antenna pointing to a celestial source, the received power at the output port of the radiometer is P (W) expressed as:

$$P = kBG(T_A + T_R) = kBGT_{sys} \tag{67}$$

Where:

- k is the Boltzmann constant (1.38 × 10-23 J/K);
- B is the receiver bandwidth in Hertz;
- G is the receiver gain ;
- T_R is the receiver noise temperature, which represents the equivalent internal noise generated by the receiver itself brought to the input.

Moreover, the antenna temperature ($T_{\square}TT_{\square}$) is the convolution between the sky brightness temperature and the antenna pattern (spatial distribution of the antenna response). In simple terms, the brightness temperature, $T_{\square}TT_{\square}$of any source is the physical temperature of a black body which would produce the same power flux density at the RAS receiver as the source. The sky brightness includes all emitting factors seen by the radio telescope, such as the background signals from the universe (cosmic background noise or galactic centre or other strong celestial emitters), natural ground contribution, atmospheric emission etc. The sum of the two noise contributions, $T_{\square}TT_{\square}$ and $T_{\square}TT_{\square}$, is usually denoted as the system temperature, $T_{\square}TT_{\square}$.

The sensitivity of a radio receiver refers to its ability to detect weak signals and is usually defined as the minimum detectable temperature difference (ΔT) in the antenna temperature that can be measured by the receiver in a given bandwidth B and integration time τ. The sensitivity, is given by the radiometer equation:

$$\Delta T = \frac{T_{sys}}{\sqrt{B\tau}} \tag{68}$$

Where:

- B is the bandwidth of the observation and τ is the integration time.

The square root in the denominator of this equation is important. If the noise level of a measurement shall be reduced by a factor of two, the integration time needs to be quadrupled. Thus, even small degradations in the sensitivity have drastic consequences on the necessary integration times to reach a certain noise level, increasing the operation cost of a telescope accordingly.

It is common for radio astronomers to observe signals that have mean power levels several times less than background noise. Therefore, long integration exposure and wide integration band (for continuum observations) are commonly applied to increase the sensitivity.

In a radio astronomy receiver, uncorrelated noise, which is mainly caused by thermal noise, is the dominant type of noise added to the signal of the observed source. This noise is generated by the random motion of electrons in a conductor and is present in all electronic devices, contributing to receiver noise temperature. Additionally, thermal noise can also come from natural sources such as the atmosphere, ground emissions picked-up by the telescope antenna, and background signals from the universe, which may not be of primary interest to the observer. Because the random nature of the noise, integration reduces the level of fluctuations by averaging the signal over a longer period of time. The integration time used in radio astronomy can vary from seconds to several hours or even more. Longer integration times enable the detection of weaker and more distant sources. However, for some scientific applications, such as to detect fast transient phenomena, the integration time must be kept short to reach the desired time resolution. In this case only an increased bandwidth can be used to reduce the noise as described in the radiometer equation.

The calculation method and typical technical parameters that define the sensitivity of radio astronomy observations over a large range of frequencies can be found in Recommendation ITU-R RA.769 [23]. To quote some example values, power flux density levels that must not be exceeded to protect RAS observations range from about -260 dB(W/m^2/Hz) at lower frequencies (e.g. at 150 MHz) to approximately -220 dB(W/m^2/Hz) above 200 GHz.

A8.1.2 Noise figure

The noise figure is a standard way to express the noise temperature of a receiver. It is a metric for the amount of additional noise added by an electronic device to the input signal. It is typically expressed in decibels (dB) and is calculated as the difference between the input SNR and the output SNR of the receiver:

$$F = SNR_{in} - SNR_{out} \tag{69}$$

In a multi-stage system with multiple filtering stages, the overall noise figure can be calculated using Friis' formula:

$$NF_{Total} = NF_1 + \frac{NF_2 - 1}{G_1} + \frac{NF_3 - 1}{G_1 G_2} + \cdots \tag{70}$$

Where:
- NF_1 is the noise figure of the first stage;
- NF_2 is the noise figure of the second stage, etc.
- G_1, G_2, etc. are the gains of the first, second, and subsequent stages.

Therefore, the noise figure of a multi-stage system cannot be lower than the noise figure of the first stage and also that subsequent stage's noise figures are decreased by cumulative gains of previous stages. Therefore, it is essential to minimise the noise figure of the first stage and use subsequent stages only when necessary, in order to achieve the desired gain.

For high sensitivity receiving systems, as those used in radio astronomy, the receiving chain is usually composed by the so-called feed-system consisting of passive components to couple the incoming radio wave reflected by the mirrors of the radio telescopes into a guided transmission line. Such components typically consist of the following elements: feed-horn, polariser, ortho-mode transducer and noise marker coupler. Since all these components come before the first amplifier, their contribution to the overall receiver noise is significant. Therefore, they are cooled to cryogenic temperatures to minimise their ohmic attenuation together with the first low noise amplifier (LNA). Positioning a microwave filter to increase immunity to interference would add further losses in this sensitive section of the chain and therefore it should be considered as a last resource mitigation measure. Filters are typically placed after the first stage of amplification, lowering their noise contribution and also because the last stages of the receiver are more prone to get saturated due to the additional gain provided by the previous stages.

As a rule of thumb, modern LNAs achieve noise temperatures of 0.25–0.5 K times f0 (GHz), e.g. a 4-8 GHz LNA would have between 2 to 4 K when cooled to cryogenic temperatures of 15 K, and thus provide a very sensitive receiver. This implies that its dynamical margin cannot cope with very strong active-service transmissions (e.g. radars or telecommunication towers) in the immediate vicinity. The typical compression level at the input of an LNA is around -50 to -40 dBm noting that this value depends on the material of the transistors of the LNA, the polarisation of the different stages, the frequency, etc. Figure 14 shows the gain and output power vs. input power for a typical LNA used in radio astronomy.

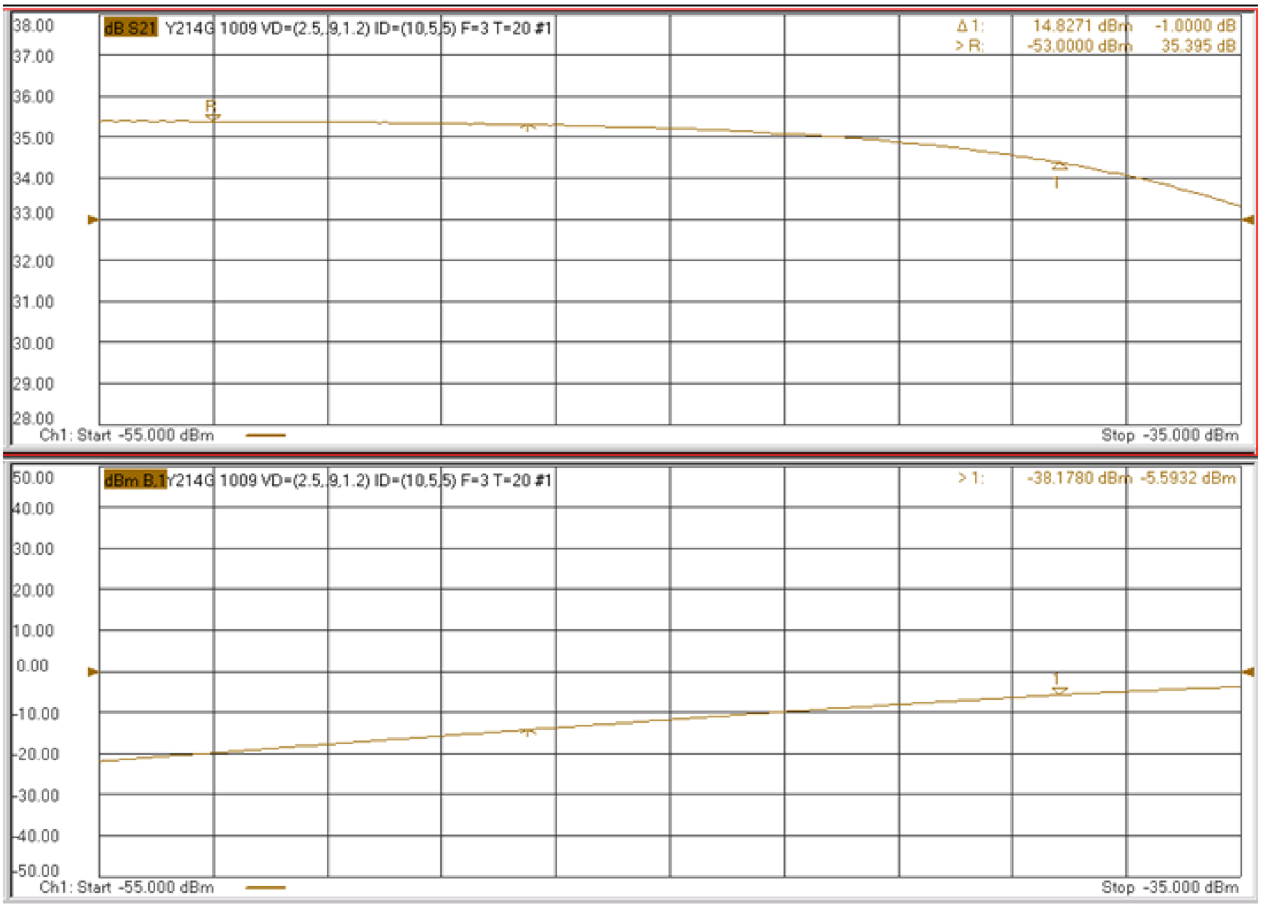

Figure 14: Characterisation of a broadband LNA (2–14 GHz) for a VGOS receiver. Top: gain vs. input power. Bottom: output power versus input power, the marker indicates the P1dB level (–38 dBm at the input)

A8.2 NON-LINEAR EFFECTS

An important figure of merit in radio astronomical receivers is the linearity. If the active stages do not operate in linear regime some distortions will occur affecting the astronomical observation. Signals at new frequencies would appear at the output of the receiver and might be interpreted as actual radiation entering in the antenna. Nowadays, this is of primary concern in radio astronomy due to the presence of strong artificial emitters.

The amplifier's burning point and compression points are crucial levels of input power for the LNA. The burning point is the maximum output power level that an amplifier can sustain without damaging itself due to excessive heat generation. This is also called the amplifier's maximum power handling capability.

Report ITU-R RA.2188 [22] provides guidelines for the maximum power flux-density and equivalent isotropically radiated power (e.i.r.p.) levels that could potentially cause damage to radio astronomy receivers in the frequency range of 1 to 275 GHz. Tables 1 and 2 in RA.2188 list the maximum allowable power flux-density and e.i.r.p. levels in units of dB(W/m²) and dB(W/Hz) for two different technologies in the active stages.

A8.2.1 Compression point and intermodulation points

The saturation points are the points at which the amplifier enters the non-linear regime. Two types of saturation points should be considered at minimum, the compression point and the intermodulation point, considering amplitude and frequency, respectively, of the non-linear effects.

A8.2.1.1 1 dBc compression point

At the compression point, the output power level stops to increase, even if the input signal level is increased further. An amplifier is said to be “saturated” at this point and can no longer provide any additional gain. Further increases in the input signal level will result in distortion of the output signal. This results in a distortion of the received signal and a loss of information because the receiver's output signal becomes clipped.

The 1 dBc compression point represents the power level at which the output power of the LNA is reduced by 1 dB with respect to a linear scale due to compression. One refers to the input and output power levels associated with the 1 dB compression points as IP1 and OP1, respectively. Higher 1 dB compression points indicate better linearity and less distortion of the signal.

A8.2.1.2 Intermodulation points

When operating in the non-linear regime, the amplifier can also produce harmonics or intermodulation products that emerge above the noise floor. The latter effect is typically the most problematic, due to the proximity of the fundamental tones to be amplified. From the different tones, the third order intermodulation product is generally the most problematic, by falling inside the frequency band of reception. It should be noted that, considering the low level of noise in RAS receivers, intermodulation products can be detected well before the power level reaches the 1 dB compression point.

A common figure of merit for intermodulation products is the IP3, which describes third-order effects. Its value is derived as the intercept point of the fundamental (in log scale has 1 dB slope) and the third order product curve (3 dB slope). In essence, it shows the maximum amplitude of a signal that can be applied to the device before intermodulation distortion occurs.

A8.3 MITIGATION MEASURES

A8.3.1 Filtering and insertion loss

One possibility to improve the resilience of a receiver is to use filters. First active stages in radio astronomical receivers are designed primarily for optimum noise figure and secondarily for high gain, because the radio astronomical power level entering the receiver is very low. As a consequence, initial active stages lack for high robustness performance. On the other hand, subsequent amplifiers in the receiving chain usually have relatively high 1 dB compression points to handle the amplification of the previous stages. Depending on the specific architecture of the receiver, the weakest stage to get compressed earlier in the chain can be either at the initial or at end of the chain.

Filters are commonly used to increase a receiver's resilience to interference by attenuating the entire band that contains interference. However, this solution is not always applicable in radio astronomy. In particular, it is not feasible to place narrow passband filters just around RAS bands. This is especially true if the RAS band is on the first or second adjacent channel of a strong transmitter.

In electronic filter design, the insertion loss is a figure of merit that is frequently employed. This metric is defined as the ratio between the signal level without the filter (V_1) and the signal level with the filter installed (V_2), and it is calculated in decibels. The insertion loss can be calculated as follows:

$$Insertion\ Loss\ [dB] = 10 \log_{10} \frac{|V_1|^2}{|V_2|^2} \tag{71}$$

Insertion loss can have a significant impact on the noise figure of a system, which can influence the overall sensitivity and performance of the system. When a component with insertion loss is added to the system, it can reduce the signal power, leading to a decrease in the SNR at the input of the component.

A8.3.2 Ultra-low insertion loss filters

As a tentative solution, high-temperature superconductor (HTS) filters have been used at some radio telescopes. The idea is that such a filter can even be placed before the cryo-cooled LNA without affecting the sensitivity too much – HTS filters have very low insertion loss. The HTS technology is maturing, but is still costly and not flexible enough compared to the quick changes that occur in some parts of the radio spectrum. HTS filters are static components, whose response cannot be adjusted/tuned based to the given scenario. Furthermore, component tests cannot be performed at room temperature, which makes designing the filters more challenging than classical analogue filters. During the last few years a number of HTS microwave filters have been designed and fabricated for three receivers installed on the 64 m Sardinia Radio Telescope (SRT). All these filters are based on an Yttrium Barium Copper Oxide (YBCO) film laid on a Magnesium Oxide (MgO) substrate with a thickness of 0.5 mm and a dielectric constant of 9.65. The front-end receivers for which they have been developed are: a P-band receiver (305-410 MHz), a “High” C-band receiver (5.7-7.7 GHz), and a “Low” C-band receiver (4.2-5.6 GHz). Although there have been strikingly successful and hopeful deployments of HTS microwave filters at a few observatories, they cannot be seen at the moment as the universal solution for the receiver problems associated with very powerful out-of-band signals. High gain wide-band LNAs often have frequency intervals where they are not unconditionally stable, and together with low loss filters being LC circuits attenuating via impedance mismatch, this can lead to spurious oscillations, often outside the pass band of the amplifier and difficult to detect, but which can severely degrade performance. Many specific technical factors determine the feasibility of using appropriate input filters. Retrofitting such filters to existing radio astronomical receivers is often impossible and the question of their use for new receivers has to be evaluated on a case-to-case basis for each band and for each observatory.

A8.3.3 Microwave filters

All optics systems (including the primary reflector and feed horn) necessarily act as filters to an extent, so the feasibility of filtering out harmful transmissions in the optical domain rather than the coaxial domain is also considered. The principle of optical filtering relies on waveguide components attenuating radiation below the cut-off frequency, which is inversely proportional to the dimensions of the waveguide. That is, a length of waveguide may be used as a high-pass filter, providing strong attenuation of any signals below the cut-off frequency. The caveat is that access to all frequencies below the cut-off frequency is lost, and not just the frequencies occupied by an exemplary troublesome transmission that needs to be filtered out. Waveguides cannot be employed in the application of bandstop filtering, because there is no upper cut-off frequency.

A key challenge in traditional single-pixel receiver design is the need to improve the impedance match between free space and the transition to the coaxial domain. This requires careful design of the optics system, and while the benefit of waveguide filtering is the minimisation of loss for propagated modes, the degradation of the impedance match with free space will introduce other aspects of undesirable performance reduction, such as bandpass ripple.

In conclusion, a RAS receiver is best optimised when the minimum of required waveguide components is used, such as the feed horn and OMT, and extra waveguide components introduced in an attempt to counter powerful terrestrial transmissions should be avoided.

A8.4 NEW DEVELOPMENTS IN RECEIVER DESIGN

Until a couple of decades back, most existing receivers at European radio telescopes were single- or dual-pixel receivers, owing to the physical tension imposed by the cooling process. These receivers typically consisted of a single chain for each of the two polarisations needed to sample the ensemble of randomly polarised photons that are focused onto the phase centre of the receiver's feed horn. However, in order to increase the operational capacity of radio telescopes, radio receiver development has tended towards placing an array of receivers on the telescope’s focal plane. There are two basic configurations of such an array, multi-beam receivers and phased array feeds, which use a planar aperture of dozens of antenna elements (similar to active antenna systems in telecommunications, but here there is a large reflector attached – the radio telescope).

A multi-beam receiver takes the form of an array of independent traditional single-pixel receivers. Each pixel of the array is positioned on the focal plane of the telescope in such a way as to point not just along the boresight axis, but also at positions away from the boresight axis. Despite the loss of forward gain associated with placing a feed horn's phase centre off-axis, the advantages in increasing the instantaneous field-of-view far outweigh the disadvantages in required engineering effort, mostly by enhancing the survey speed especially of large single dish radio telescopes. Examples of very successful multi-beam receivers include the Jodrell Bank Observatory (JBO, UK) Multi-beam that operates in the protected L-band, and the Effelsberg (Germany) 7-Beam, which is used extensively as a survey instrument, for both spectral lines and pulsar research.

Phased array feed (PAF) receivers (sometimes also called focal-plane arrays, or colloquially radio cameras) are a development of the multi-beam concept and take the form of a planar array of detectors that populate the focal plane. The principle of the PAF is to maximally sample the complex voltage field on the focal plane, such that the axis of peak forward gain may be determined in post-processing by appropriately weighting the complex voltages received by every antenna element of the PAF, e.g., in FPGAs. This allows for an arbitrary number of overlapping beams independent of any optical limitations that a feed horn may impose reducing off-axis effects significantly.

Multi-beam receivers and phased array feed receivers are, by definition, optimised to maximise the number of beams in order to maximise science capability, and are typically limited by size and weight considerations. As such, their cryostats are densely populated with RF components, which further drives the development of each antenna's chain of RF components towards simplicity; a typical PAF will consist of an antenna, an LNA, a bandpass filter, and either a digitiser or further electronics to interface with an RF-over-Fibre system. As such, PAFs are particularly susceptible to being driven nonlinear.

In the context of the potential future requirement to retrofit one or more band-stop filters in order to reject specific powerful transmissions, such a retrofit may be impossible without a substantial redesign. While an alternative approach is to limit the fraction of sky in which such a receiver may be operated, this places heavy limits on the science capability of the instrument. The potential future omnipresence of such transmissions may simply make operation of the receiver unfeasible.

In summary, placing a requirement on the RAS to make their receivers robust to present and future strong active transmitters will eventually direct the RAS away from developing towards maximising the science capability of their instruments – their *raison d'etre* – instead towards optimisation for maximum interoperability with commercial users of the spectrum, with necessarily heavily-limited science capability.

A8.5 PRACTICAL EXPERIENCE

Jodrell Bank has been badly affected by powerful ground-based transmissions. The L-band receivers on the 76 m Lovell, and 25 m Mark II Telescopes each have an unfiltered observing bandwidth ~1 GHz, however long-standing transmissions necessitate increasingly strong bandpass filtering to their current designs with ~500 MHz bandwidth centred on ~1500 MHz.

IMT allocations in the 1452-1472 MHz band for LTE downlinks became active in 2018. The transmissions are sufficient to force the receiver system into the nonlinear regime for all directions except those for which the primary reflector itself provides shielding from the transmitter. In other words, more than half of the sky was lost.

It was later discovered that backend components were the first to be driven nonlinear, and as such it was possible to fit bandstop filters behind the LNA rather than ahead of them. These filters act in addition to the existing band-defining filters and their use necessarily degrades system performance at the ~10% level, which reduces the observatory's effectiveness in its science mission.

The extent of the problem the 1452–1472 MHz transmissions caused can be summarised as follows:

1 The bandstop filters had to be custom-designed and could not be obtained off-the-shelf.
2 Consequently, the unit price was high, and the lead time was long (~months).
3 Further, the cryostats for each receiver needed to be modified to accommodate the bandstop filters, which required extensive engineering time.
4 The total expenditure required to adapt the L-band systems to co-operate with the new transmissions in hardware and effort is estimated at ~5% of the observatory's total operating budget, which is unsustainable for any potential future requirement to repeat this operation for other transmissions.

At the 100 m radio telescope at Effelsberg (Germany), similar experience was made with a very broad L-band receiver (called UBB). Originally, it was designed to offer frequencies between 600 MHz and 3 GHz, but strong interferers below 1 GHz drove the system into non-linearity and measurements were impossible. Later the receiver was significantly modified and now covers a range between about 1.2 and 5 GHz. Several well-known IMT bands, e.g. at 1800 and 2100 MHz are filtered out with stop-band filters. Fortunately, the LNA of the UBB is robust enough to allow placing these filters after LNA, such that the overall system performance is not heavily affected. However, since late 2022 a nearby IMT base station was upgraded to provide LTE/4G supplemental downlinks at 1452-1472 MHz, which causes intermodulation products throughout the observing bands. Even worse, for the Effelsberg “powerhorse”, the 21 cm 7-feed receiver, the system is completely saturated for more than half of the accessible sky. This receiver is used for more than 25% of the time, as it provides superb survey capabilities. Equipping this system with new filters would not only be very costly (as there are 14 independent channels), but it would also require enormous efforts to ensure that the calibration can be done consistently for old and newer data. Surveys can take several years, and it is usually not possible to change the system properties in the middle of the observations. It is currently not even clear whether stop-band filters after the LNAs would suffice. If filters needed to be put before the LNAs, this would have a huge impact on the sensitivity and require hundreds if not thousands of additional measurement hours to finish the surveys.

Some radio telescopes in Europe have suffered from the high-power signal received from radars very close to the radio telescopes. In Ny-Ålesund (Norway) and in Onsala (Sweden) observatories, the signal received from ship-borne radars close to the radio telescopes were so powerful that it drove the LNAs into their breaking point and, hence, were destroyed. In the RAEGE Santa María station (Azores, Portugal) a nearby radar makes it impossible to use the broadband receiver (VGOS radio telescope, 2-14 GHz) without any preventive measure to avoid driving the LNAs into the breaking point. Measurements were only possible after the installation of a filter in the cryostat at the input of the LNAs. This solution implies a degradation of 17% of its sensitivity. VGOS radio telescopes are planned to observe 4 different sub-bands of 1 GHz width in the 2-14 GHz range. With the installation of the filter, the lowest sub-band has been sacrificed to be able to observe in the three other ones.

ANNEX 9: LIST OF REFERENCES

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[4] ETSI EN 303 340 V1.2.1: "Digital Terrestrial TV Broadcast Receivers; Harmonised Standard covering the essential requirements of article 3.2 of Directive 2014/53/EU ";

[5] Recommendation ITU-R F.1094-2: "Maximum allowable error performance and availability degradations to digital fixed wireless systems arising from radio interference from emissions and radiations from other sources"

[6] Recommendation ITU-R F.758-7: "System parameters and considerations in the development of criteria for sharing or compatibility between digital fixed wireless systems in the fixed service and systems in other services and other sources of interference"

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[9] ETSI EN 302 217-2 "V3.3.1: "Fixed Radio Systems; Characteristics and requirements for point-to-point equipment and antennas; Part 2: Digital systems operating in frequency bands from 1,3 GHz to 86 GHz; Harmonised Standard covering the essential requirements of article 3.2 of Directive 2014/53/EU ;

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[15] ECC Recommendation (02)01: "Specification of reference receiver performance parameters", approved February 2002

[16] ECC Report 252: "SEAMCAT Handbook" approved April 2016 Note! The most recent version of the SEAMCAT handbook can be found online at SEAMCAT Handbook

[17] ECC Recommendation (02)05: "Unwanted emissions", approved February 2002, latest amended 30 March 2012

[18] ECC Recommendation (18)01: "Radio frequency channel/block arrangements for Fixed Service systems operating in the bands 130-134 GHz, 141-148.5 GHz, 151.5-164 GHz and 167-174.8 GHz", approved April 2018

[19] ECC Recommendation (18)02: "Radio frequency channel/block arrangements for Fixed Service systems operating in the bands 92-94 GHz, 94.1-100 GHz, 102-109.5 GHz and 111.8-114.25 GHz", approved September 2018

[20] Recommendation ITU-R RA.611: "Protection of the radio astronomy service from spurious emissions"

[21] Report ITU-R RA.2259: "Characteristics of radio quiet zones"

[22] Report ITU-R RA.2188: "Power flux-density and e.i.r.p. levels potentially damaging to radio astronomy receivers"

[23] Recommendation ITU-R RA.769: "Protection criteria used for radio astronomical measurements"